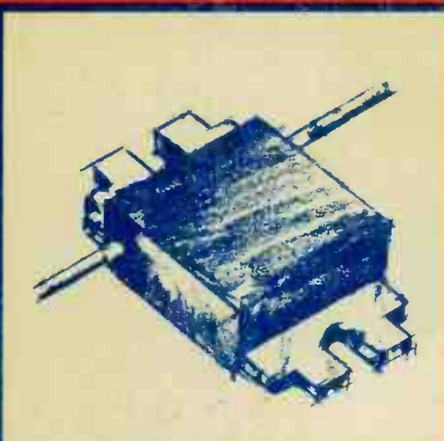
TAB BOOKS / No. 568 $4.95

IC Projects

for Amateur & Experimenter



Edited by
Wayne
Green

IC PROJECTS

FOR THE AMATEUR AND EXPERIMENTER

An anthology of some of the more interesting IC construction projects from 73 Magazine, selected to be of value both to the radio amateur and the experimenter.

Editor: Wayne Green W2NSD/1

TAB BOOKS
Blue Ridge Summit, Pa. 17214

FIRST EDITION

FIRST PRINTING—AUGUST 1971
SECOND PRINTING—SEPTEMBER 1973

Printed in the United States
of America

Hardbound Edition: International Standard Book No. 0-8306-1568-7

Paperbound Edition: International Standard Book No. 0-8306-0568-1

Library of Congress Card Number: 70-170666

Preface

In the 1950's you needed a shop to turn out homemade ham equipment. It took special tube socket punches, large steel chassis, relay-rack panels, a good drill press and substantial power supplies even for the simpler devices such as VHF converters, audio filters, etc.

The transistor replaced tubes for most applications during the 1960's and the workshop shrank from a room full of machines to a desk or table with some hand tools and perhaps a small electric drill. The big 150 watt soldering irons had long since worn out and had been replaced by instant soldering guns.

As more and more companies tooled up to turn out integrated circuits, they began to replace transistors in the more common applications and the desk-top workshop of the 60's gave way to a drawer of the 70's. The soldering guns have been replaced by little soldering pencils.

Can this foolishness go any further? Will the ham of the 80's keep his workshop in a matchbox and solder with a soldering pin? Will those parts bins of the 50's which gave way to parts cabinets in the 60's and miniature parts drawers of the 70's be replaced by thimble-sized parts holders?

Grab your soldering pencil and a small sheet of styrofoam and get busy on some of these projects, wiping a tear of nostalgia away for those of us left with shelves full of power transformers and 4000 volt 4 μF capacitors and other worthless memorabilia of days gone by.

Wayne Green W2NSD/1

Contents

Getting Hep to IC's

Integrated circuits are here to stay and the next few years will show a massive transition in industry and consumer products to complete integration. Based on this assumption, this article has been prepared for the amateur as well as professional user of integrated circuits (ICs). It includes tips on wiring, soldering, cross referencing, and simple projects using the ICs contained in the Motorola HEP kits.

A lengthy discussion of IC construction will not be covered in detail, as this information can be found in many IC textbooks. However, in order to be better able to know the advantages and limitations of those microcircuits, the reader should know what is contained in the basic IC and how these devices differ from other solid-state components.

As the name implies, an integrated circuit is a collection of many different components. The quantity and types of components vary from one IC configuration to another. A particular IC could contain active components (transistors, diodes) and passive components (resistors, capacitors). If all the components of the circuit are contained on the same "chip" or substrate, the unit is said to be "monolithic" (single crystal). The monolithic type is the most common and the least expensive to build. Other construction types are: thin film, thick film, hybrid, multichip. A discussion of these types can be found in almost any book that deals with the subject of ICs.

As an illustration of the extreme size reduction possible with integrated circuits, consider the Motorola 4-bit memory core, which contains 524 different components on a chip 50 x 70 mils. The average IC is much smaller, usually 40 mils square (1 mil = .001 in.). As the above example indicates, the race is on to see how much circuitry can be crowded into the smallest space. This effort is known in the trade as LSI (large-scale integration). Manufacturers are already starting to produce ICs that contain FET tunnel diodes, and even power transistors!

It is unfortunate, but many people are resisting the changeover from discrete (individual) components to ICs. This resistance could largely be due to the fact that people tend to shy away from circuits they are not familiar with.

The advantages of ICs over discrete components greatly outweigh the disadvantages. Size and weight reduction are obvious advantages but cost savings should also be considered. Consider the HEP 583, which contains 21 transistors and 27 resistors. If you had to buy all these parts individually and build this unit using a breadboard or printed circuit board you would indeed feel the pinch on both your pocketbook and your time. Other disadvantages that are not so obvious are as follows:

Repetition

If you need a circuit containing 20 J-K flip-flops, it would be a difficult task to build 20 of these, each containing 21 transistors and 27 resistors. This adds up to 420 transistors and 540 resistors! With ICs, only 20 TO-5 packages are necessary. Here is where cost, size, and time advantages come through again.

Repeatability

Because of the way ICs are constructed with components located in close proximity to each other, tolerances are much finer and parts are better matched, thus making up a device that functions as a complete unit. Power drain is lowered, there is less spurious noise pickup, and there is less noise generated within the unit.

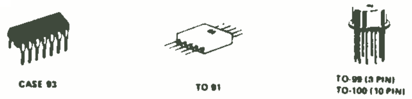


Fig. 1. Typical IC case styles.

Fig. 2. The split supply contain four series batteries grounded at the connect point. Where possible to use, the single supply offers the advantage of simplicity.

Reliability

Many manufacturers are turning to ICs because of their high reliability. Devices built under almost clinical conditions are bound to be better than a circuit built on a workbench. As an example, consider building the electricaequivalent to the HEP 583, using the 21 transistors and 27 resistors: It would be necessary to make 80 to 90 solder connections, a real source for potential trouble.

In addition to the advantages listed, replacement is simple. Schematics are easy to read, especially for the beginner. Areas yet to be conquered in the construction of ICs are: How to built inductors, large-value capacitors, and high-value resistance on an IC chip. It is presently necessary to connect these components externally.

ICs can be mounted on perforated board or printed-circuit board by either soldering to terminals or using sockets. Sockets are definitely recommended, especially for the hobbyist who will, generally, use the IC over and over in different applications. Constant soldering and unsoldering of the leads weakens them and could cause the wires to be broken, or internal damage could result due to excessive heat from the solder iron.

The HEP 580 thru 583 (devices included a Motorola IC kit) are mW RTLs. This logic family is considered the easiest for the hobbyist, experimenter, and IC novice to "cut their teeth" on. The HEP 584, 570, 571, 572 are MRTLs – also a good family for the beginner. The HEP 553, 554, 556, 558 are ECLs – not the easier to work with, but the best logic family for high frequency and noise rejection.

IC Packaging

Integrated circuits can be found in a variety of packages. At the present time, there are more than 120 case types made by some 70 companies around the world. Of these many case styles, three types are dominant. (In terms of quantity of devices on the market, in a given case type, about 90% of their quantity can be found in some variation of one of these three case types). As yet, no definite standardization has been set up among the manufacturers regarding packaging, pin numbers, and locations, so carefully check the basing before you plug that IC into the socket. Three popular case styles are pictures in Fig. 1.

Use a low-wattage soldering iron! 25 to 40 watts is a good range. Excess heat could "kill" the IC.

Keep component leads short! Excess lead length could cause spurious or parasitic oscillations or no operation at all.

If you are using a power supply (other than a battery), it is a good idea to bypass the power leads. Connect a 0.05 or 0.1 capacitor from the power input to ground at or near the input terminal of the IC.

Fig. 3. Four basic logic element configurations used in IC diagrams.

Power Supplies

For projects using 1 or 2 ICs, batteries are usually the best supply. On larger projects, an ac supply is better. The power supply requirements for the various logic functions have been standardized as follows:

RTL	3AV ± 10% (2.6 to 3.3V) and 3.6 ± 10% (3.24 to 3.96V)
MwRTL	3V ± 10% (2.6 to 3.3V) and 3.6 ± 10% (3.24 to 3.96V)
DTL	4V ± 10% (3.6 to 4.4V)
MDTL	5V ± 10% (4.5 - 5.5V)

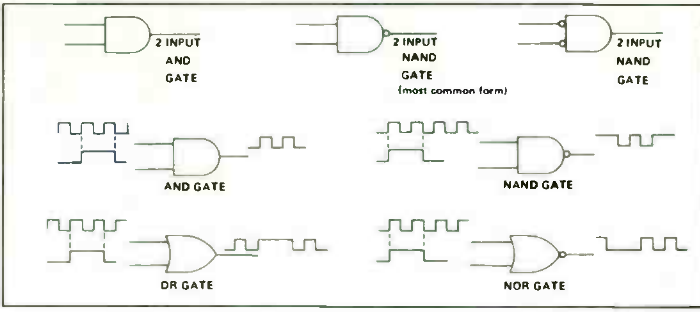


Fig. 4. Basic gates with their input/output waveforms.

VTL	±4V to ±10V (8 - 20V)
ECL	5.2V ± 10% (4.5 - 5.5V)
TTL	5V ± 10% (4.5 - 5.5V)
HTL	18V

Obviously, batteries in some of these odd voltage ranges are not available: however, experimentation has categories as follows: many of the devices were found to work well from 1.5 to 12 V! *Very few* did not – but after all, they are only rated from 2.6 to 4V (approximately). This makes it possible to use many of these ICs over a wide voltage range. Usually an IC rated at 3.2V minimum works well on 3V and one rated at 5.5V maximum works at 6V.

ICs can be connected in one of two ways, using one or two supplies. The dual or split supply is most common in linear circuits. The two supplies are shown in Fig. 2.

There are some applications where the split supply is advantageous but generally it involves more complicated circuitry. The novice in ICs is likely to be a novice in the area of computer logic also. The logic symbols are to digital ICs what schematic symbols are to resistors, capacitors, etc. Some of the more common types are shown in Feb. 3. These symbols have recently been standardized by the government. Before that time, each manufacturer had his own set of symbols.

Basic Logic Types

Most computers work on the binary principle. Binary stands for "two",– two states or conditions, which are either *on* or *off*, *high* or *low* or *1* and *0*.) Consider the condition where we have zero or near zero volts at the input to a gate, flip-flop, amplifier, etc; with positive logic it is an *off* condition. If this voltage goes positive, let's say to 1 or 2V, it is now in an *on* condition.

The common functions in digital ICs are:

- Gates - control the passage of signals.
- Buffer - amplifies power of signals to be able to drive more units.
- Inverter – reverses the logic from + to - or - to +.
- Expander – affords additional inputs to a gate.

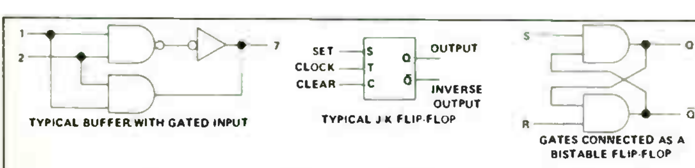


Fig. 5. Basic amplifier and flip-flop configurations.

- Adder – provides the *summ* and *carry* operations on two input signals.
- Shift Registers – provides bistable storage.
- Flip-flop - provides division or *count*. One flip-flop divides by 2, provides one output change in state for every two input charges.

Gates. This function comes in a wide variety of configurations. There are 2, 3, 4, or more inputs and 4 categories as follows:

- *And:* When all inputs go to 1, output will go to 1.
- *Nand:* Output will be 1 except when all inputs go to 1.
- *Or:* When any input goes to 1, output will be 1.
- *Nor:* Output will be 1 except when any input goes to 1.

Nand and *nor* differ from *and* and *or* in that inversion has taken place. Refer to Fig. 4. Note the small o at the input or output of some of the examples. This o indicates that inversion has taken place.

Gates can be connected to operate in a wide variety of applications other than those for which they were designed. Some applications are free-running multivibrators, bistable, one-shot, amplifiers, and audio mixers.

Occasionally the time arises when the hobbyist needs something in the way of gates other than what he has or what is available. For example, you need a 3-input gate and you have a 4-input gate; simply ground one input. Ground two inputs to obtain a 2-input gate. If you have a dual 2-input gate, such as the HEP 580, and you need a 4-input gate, tie pins 6 and 7 together and this becomes the output; inputs are then on pins 1, 2, 3, and 5.

Amplifiers. In digital work it is referred to as a buffer. Its original use is to increase "fan-in" or "fan-out" capability; that is, the number of other units that can be connected in parallel to the input (fan-in) or output (fan-out). By adding proper external biasing it is possible to connect this unit to linear (audio-rf) usage.

Flip-Flop. There are a number of types of flip-flops available. As mentioned previously, a flip-flop (multivibrator) can be "made up" by cross connecting two gates. The R-S flip-flop is one example. The J-K flip-flop is similar but has the added function known as "clock input" shown as "T" on the logic block of Fig. 5.

Fundamentally, flip-flops divide by two. By proper connection, division by 3, 4, 5, etc. can be obtained using a few ICs as shown in Fig. 6.

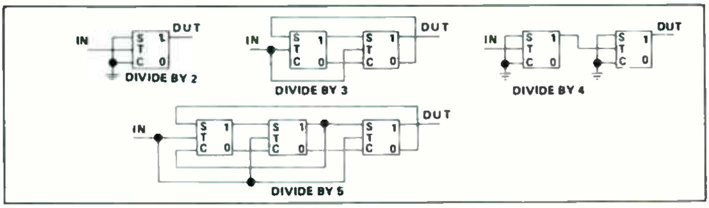


Fig. 6. Flip-flops connected as dividers.

IC Receiver Accessory

John J. Schultz, W2EEY

Audio frequency integrated circuits provide the opportunity to develop very useful circuit functions—in many instances at a far lower cost than would be possible using discrete components. On the other hand, the inherent nature of the IC housing and the conditions under which it must operate also make the practical realization of a circuit using IC's, in some instances, just as complicated as a circuit using discrete components. This seeming contradiction was experienced by the author in developing the receiver accessory unit described in this article. So, even if one has no immediate need to build the unit described, it may prove interesting to still read about and appreciate some of the considerations involved in the use of audio integrated circuits.

General

The accessory unit described was designed as a compact, solid-state unit that could be plugged into a medium to high impedance headphone jack and function as a self-contained unit to provide tunable af selectivity, agc and about 1 watt of power for direct loudspeaker operation. By the addition of a few diodes a simple noise limiter can also be added. For compactness and simplicity, only a resistor-capacitor network was used without any inductive elements to obtain audio selectivity. The resultant selectivity is not as sharp as that provided by a bulky inductor but is quite usable on CW, especially with a transceiver that already contains a steep-skirted crystal or mechanical SSB filter. The fact that the af selectivity is tunable also adds to its usefulness. The agc feature is not absolutely necessary but was added since many transceivers, although their avc cannot be disabled as such, do not provide full agc on CW when the rf gain control is at some intermediate setting, as would be normal if one were using the rf gain on CW as the "volume" control while tuning. The audio output amplifier was included to eliminate the need for going back into the receiver in order to use its audio output stage. Thus, no modification whatever is required to the receiver and the accessory unit can simply be unplugged from the headphone jack when it is not desired for operation.

Power amplifier IC mounted in heat sink is shown on left. FET and associated components are at upper right side of heat sink. Other components are conveniently arranged around dual stage FET. Potentiometers shown on right are miniature type mounted directly on perforated board stock.

Circuit

A block diagram of the accessory unit is shown in Fig. 1. The basic stage functions are relatively simple. The incoming audio is split into two parts, each going ½ of the HEP592 (a hobbyist version of the MC1535 dual operational amplifier sold as a stereo preamplifier). One section of the HEP592 serves as a level detector with an adjustable threshold. When the positive going portion of the input signal exceeds the threshold level, an output voltage is produced which when rectified is coupled to the gate of a

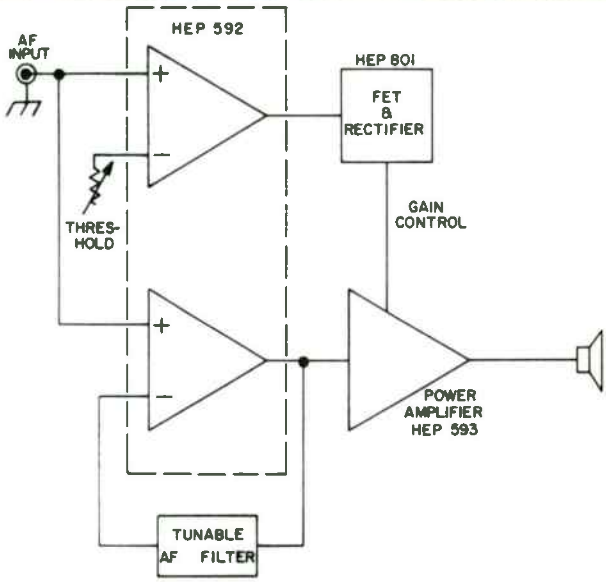


Fig. 1 Block diagram of accessory unit stage functions. Other similar IC operational amplifiers can be used to perform the same functions.

HEP801 FET. The drain-source resistance of the FET is normally quite low but increases to several thousand ohms with increasing negative gate-source potential. This wide resistance variation is used to regulate the gain of the power amplifier IC. This "forward" method of obtaining agc is somewhat different from the usual "backward" method where part of the final output af voltage is rectified and then used to control the gain of a preceeding stage. However, it works just as well and with some experimentation of the circuit time constants, it may even appeal to some as providing more responsive agc action than the "backward" system.

The other section of the HEP592 serves as a tunable af filter. Advantage is taken in constructing the filter circuit of the fact that most "operational amplifier-turned-hobbyist" IC's provide a differential input. That is, so-called inverting (-) and non-inverting (+) inputs. A portion of the input signal is fed to the non-inverting input. The other input is coupled to the output via a tunable audio filter which passes all audio frequencies *except* the one to which it is set. Thus, all other audio frequencies are fed back to the input to oppose wall input frequencies except one frequency. The overall result is an audio peaking stage at the one frequency to which the audio filter is set.

The power amplifier stage is necessary to raise the output level to drive a small loudspeaker and produces about 3/4 watt output.

The wiring diagram of the unit is shown in Fig. 2. A 10K potentiometer is used with the agc portion of the HEP 592 IC to set the threshold value at which agc action starts. It can either be brought out as a panel control or left as a trimmer adjustment. The tunable af filter is of the bridged-T type. As shown, a three-unit potentiometer is needed to cover the complete audio range up to a few thousand cycles. One could possibly make only two of the resistor legs variable, but the frequency range will be restricted to a few hundred cycles over which the network is effective. Fig. 3A shows an alternative filter network which can be connected between terminals 7 and 10 of the HEP 592 and which requires only one potentiometer. It is rated to be effective from 70 to 10,000 cycles, but only the components specified should be used.

Fig. 3B shows a simple noise limiter which can be connected either before or after the HEP 592. Advantage is taken of the fact that the IC units are operated with a dual-polarity power supply to allow biasing of the diodes for symmetrical clipping.

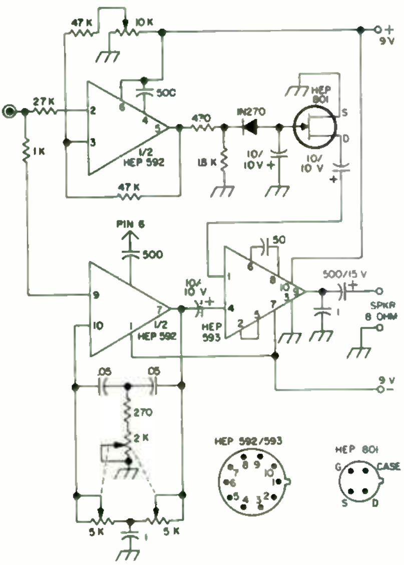


Fig. 2. Wiring diagram of accessory unit. Resistors are ½ watt. Triple section potentiometer used in audio filter is IRC type 45D502MD502MD202.

Construction

The photograph shows one construction layout possible for the unit. Basically, the circuit components are grouped around each IC and directly soldered together. The potentiometers for the agc and af filter are mounted directly on the perforated board stock. The potentiometers shown are actually of a miniature type but normal, less expensive ¼ inch shaft types are quite adequate. Also, although not shown, an IC socket should be used for the HEP592 unit, instead of directly wiring it, for protective purposes.

The use of the HEP 593 power amplifier appears simple from Fig. 2 and although its connections are not involved, its placement in a unit does present some problems. A heat sink *must* be used and Motorola specifically advises against the simple slip-on fin-type head sinks available for TO-5 transistor cases. They suggest a 2 inch x 2 inch x 1/8 inch piece of aluminum with a center hole drilled to snugly fit over the transistor case. Not having any material available, the author used a heat sink found on a surplus IBM computer board. The index tab was removed from the HEP 593 and a clean contact area between the flange of the IC and the heat sink established. The IC was then glued into place. Overall, the relatively low voltage gain (18-35) of the IC power amplifier, its cost and the necessity of using a heat sink does not compare too favorably with using two larger case audio transistors which do not require heat sinks at the 1 watt level.

Two 9 volt transistor batteries were used to power the unit shown. The battery potential slightly exceeds the 8 volt operating voltage recommended for the units but seems to cause no difficulty.

Adjustment & Operation

There are no real adjustments that should be necessary if proper layout procedures are followed. Some experimentation of the feedback resistors in the agc amplifier circuit may be necessary to obtain the best action and "bread-boarding" the circuit initially will save time later. The IC's did show some tendencies toward instability if too much coupling were allowed to exist between the various input and output circuits. This is understandable when one realizes that although the IC's are called "audio" type, their actual response extends up to several hundred khz. So, an rf feedback loop can exist which will overload the units but yet not produce an audible indication.

In operation, the receiver af output level is adjusted to produce adequate audio output, but not to the point of overloading the unit. The agc threshold control is set as desired to produce the best agc action when going from a weak to a strong signal.

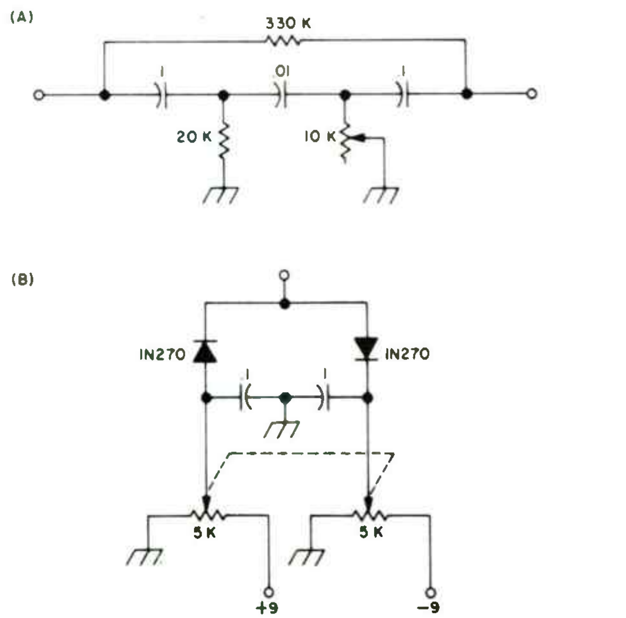


Fig. 3. Simpler audio notch filter (A) requiring only one potentiometer. Capacitors are Aerovox type P123ZNG. Symmetrical noise clipper (B) which can be used at input to unit or at input to power amplifier (pin 4).

Summary

Various other operational or audio type IC's can be used for this type of accessory unit so long as the device ratings are not exceeded. The main criteria is that the agc/af filter IC have differential inputs. One could, of course, also use separate IC's for these functions, each with a differential input.

Such a unit adds a significant degree of improvement to the operating possibilities of a SSB transceiver used on CW which does not have real provisions for CW reception. It also provides some very interesting exercise and experience in the many versatile uses to which IC's can be put.

Build an 8 Transistor Code Oscillator With Just One IC

Ken W. Sessions Jr., K6MVH

A code practice oscillator is practically always a simple affair. But with a good IC, such as Motorola's HEP 570, the project can be made even simpler while the circuitry itself – within the integrated circuit – stays complex enough to assure plenty of audio gain, good stability, and excellent quality of tone.

The oscillator described here was designed by experts at Motorola, who allowed for four "discrete" stages, each with a two-transistor capability, within the framework of the lone flatpack IC package. One of the nicest features of the HEP 570 as the oscillator element is its economy of external parts. Apart from the power switch, the telegraph key, the speaker, and the battery, the only additional components required are two resistors, a pot, and three capacitors. What could be simpler?

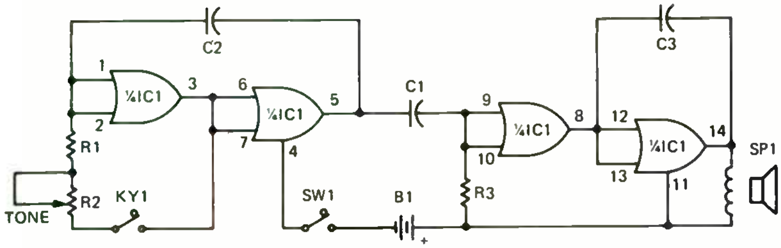
Construction is as simple as the schematic (Fig. 1) makes it seem. One item not shown on the schematic, however, could make the job easier and save the insecure builder a great deal of grief: a suitable socket for the IC. An advantage of using IC sockets is that if all does not go well right off, the IC remains isolated from the rest of the circuit while changes are incorporated. Also, most hams use their ICs again and again, for any number of appropriate projects. Soldering of an IC directly into a circuit will seriously curtail its universal utility. The package can't stand too many solder/unsolder operations before the leads give out. But with a socket, the IC stays like new and may be used in as many projects as the builder has sockets for. And changing the IC from one project to another is no more difficult than making a tube change in an old-fashioned rig.

Your junkbox will undoubtedly yield the resistors and capacitors necessary to complete the oscillator project, but it is unlikely that you'll find the right speaker kicking around the shack. Since the output of the amplifier/oscillator is in the vicinity of 50Ω, a standard "intercom-type" speaker is required. If the expense of such a speaker proves a bit much, there are other routes that might prove entirely satisfactory, such as scrounging up matching transformers that can be used to drive either a speaker or an external amplifier arrangement.

One method that has proved adequate is to use an ordinary low-voltage power transformer to couple from the IC to the speaker. A transformer with a primary winding of 120V and a secondary of 12V provides a reasonable approximation of the proper turns ratio, and will deliver a fairly healthy audio signal to a 3.2Ω voice coil. Since power-handling capability is no consideration, you can use the smallest physical size of transformer you can get your hands on. The only disadvantage with this approach is that it plays hob with any attempt toward miniaturization.

A small chunk of perforated board makes an ideal mounting bed. If the small intercom speaker is in your list of "availables," you can easily mount the whole affair in an enclosure no larger than a tiny portable radio. If you have to use the power transformer, you'll have to just poke around for a chassis with enough bulk to accommodate everything.

An obvious "extra" that will enhance the usefulness and attractiveness of your oscillator are matching jack and plug for the sending key. This will also simplify the use of another code oscillator for two-way operation.



PARTS LIST:

IC1	HEP 570	R3	10 K	SW1	SPST	SP1	45 Ω
R1	3.9 K	C1, 2	.05 µF	KY1	MORSE KEY		INTERCOM
R2	10 K POT	C3	.001 µF	B1	3 VDC		4" SPEAKER

Fig. 1. Schematic of the "8-transistor" code practice oscillator made with a Motorola HEP 570 IC.

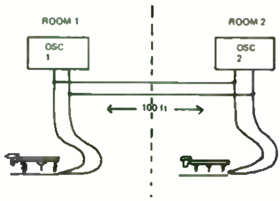


Fig. 2. If the distance between operating stations is held down to a reasonable value a CW intercom setup can be made with two or more code oscillators. If you've got kids in the family who can't find time to practice their code for the Novice exam, this arrangement is sure to turn the trick.

To couple a pair of code oscillators for two-way use, it is better to parallel the keys rather than the speakers themselves. This approach keeps each unit from the labor of driving more than its fair share of the load. Figure 2 illustrates an excellent room-to-room interconnect method that has already been used to bring two new Novices to the bands. (Learning the code is fun if there is some incentive to study; and a room-to-room CW intercom will work wonders with harmonics who might otherwise be reluctant to practice.)

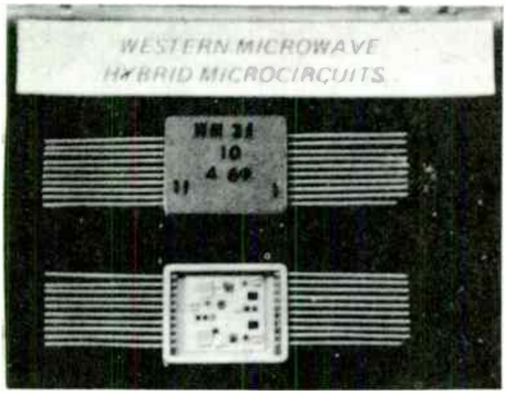
Integrated Circuit Audio Filter

John J. Schultz, W2EEY/DL

One item that has been used in innumerable pieces of amateur equipment and accessories over the years is an audio filter. Such filters, particularly if they were used for audio selectivity purposes, could get to be very elaborate and large with multiple section designs. It was probably only inevitable that the current stream of progress toward the micro-miniaturization of electronic components would also reach audio filters. However, the miniaturization of audio filters that has been achieved is not just simply a miniaturization of inductors and transformers. None of these components are used in the audio filters to be described, and these integrated circuit filters offer adjustment versitility that could never be achieved with inductors. Such filters open up the possibility for the construction of numerous compact pieces of accessory equipment that can be used to improve the operation of receivers and transceivers.

Background

Many attempts have been made to do away with the use of inductors in filters both because of size and cost factors. Some of these attempts date back quite a few years and all revolve about the use of rc networks in place of inductors. For instance Fig. 1(A) shows an rc high-pass filter. As the frequency of the input signal increases, the reactance of the capacitors decrease and more voltage appears across the output. Fig. 1(B) shows a low-pass filter that works in a similar manner. If you combined the filters, using an amplifier for each, you could form a bandpass filter. You could also combine various forms of rc networks to form notching or peaking filters, as shown in Fig. 1(C). Such filters by themselves, of course, are crude and provide poor selectivity. Usually such filters are used together with amplifier stages to compensate for the filter attenuation and



A view of an encased WM3A filter and one with part of the casing removed to show the internal hybrid integrated circuitry. The encased filter measures only 8/10" x 7/10" and is about 3/16" thick. Photo courtesy Western Microwave, Los Gatos CA 95030.

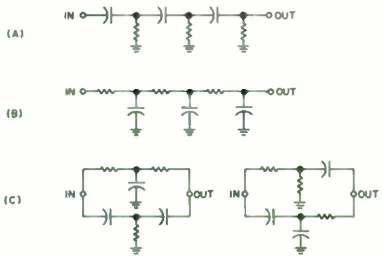


Fig. 1. Basic rc networks allow filter circuits to be built without the use of inductors. High-pass circuit (A), low-pass circuit (B), and notching and peaking circuits (C).

also in feedback arrangements so the filter circuits are not loaded down. Unfortunately, by the time you combine sufficient discrete component rc networks and transistor amplifier stages to have the rc filter duplicate the performance of an inductor network, the rc filter can be as large and as costly as the latter. The advent of integrated circuits has changed all that, however. High gain amplifier circuits and multiple rc networks can be incorporated in one physically compact unit.

The block diagram of the integrated circuit filter is shown in **Fig. 2.** Three multi-transistor operational amplifiers and the necessary rc networks in a feedback arrangement are combined in the hybrid filter unit. Three external resistors are used and can be manipulated to change the operational characteristics of the filter. One can see some of the filter components in

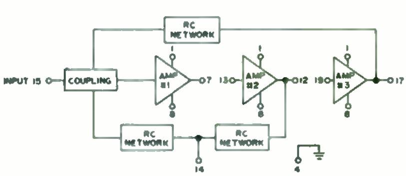


Fig. 2. Block diagram of the Western Microwave WM3 filter. ALL of the blocks shown are contained in the single filter unit shown in the photograph. Each amplifier is, in fact, a separate integrated circuit amplifier. The numbers refer to the terminal connections.

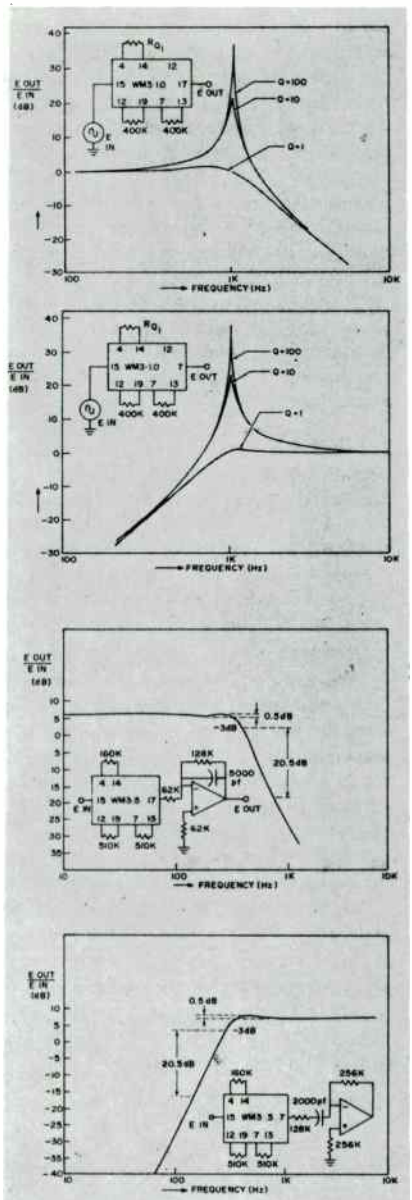


Fig. 3. The WM3 filter can be used alone [(A) and (B)] to function as a low-pass or hugh-pass filter or in conjunction with a postamplifier [(C) and (D)] to produce a steeper slope at the cutoff frequency.

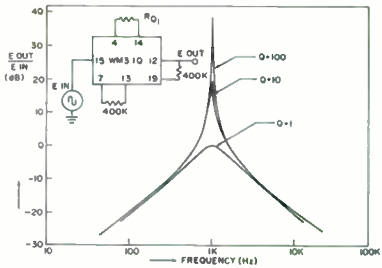
the unencased view of the filter in the photograph. The encased filter measures 0.8 inches x 0.65 inches x .15 inches thick. The surface area is about that of a 25 cent piece, and it is hardly any thicker.

Performance

The filter really begins to shine when one investigates its performance possibilities. It can be used as a high-pass filter, low-pass filter, peaking filter, notching filter, etc. The center frequency can be adjusted as desired by an external potentiometer as well as the Q if desired. The unit can also be set up so you can switch select a variety of different filter effect outputs.

Graphs portray the performance of this type of filter best. **Fig.** 3 illustrates the output versus frequency characteristic of the filter in several low-pass and high-pass circuits. The filter can be used alone for these functions, or its output used to drive another ic operational amplifier power stage to further increase the slope of the frequency response at the cut-off frequencies.

Fig. 4 shows the filter used as a peaking or notching filter. Note the extreme sharpness of the response at the nominal center frequency of 1 khz. Actually, both the center frequency as well as the sharpness of the filter response can be tuned by making various of the external resistors variable as shown in **Fig.** 5. Thus, the center frequency of a nominal 1khz filter can be tuned from about 500 to 1500 cycles. The Q can be varied from about 1 to a maximum of 100.



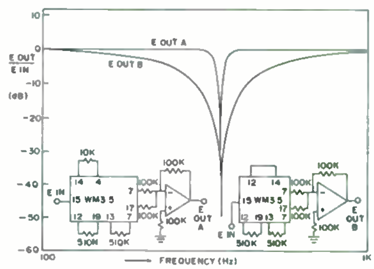


Fig. 4. The extremely sharp peaking and notching characteristics of the filter make it ideal for use as an cw selectivity device. The peaking and notching frequencies can be tuned from 500 to 1500 cycles.

As might be imagined from looking at the arrangement of the external resistors and the output terminals used for each specific application of the filter, you can devise various switching arrangements to select different outputs, different specific center frequencies, etc. The possibilities in this direction are pretty well only limited by your imagination. **Fig.** 6 shows one

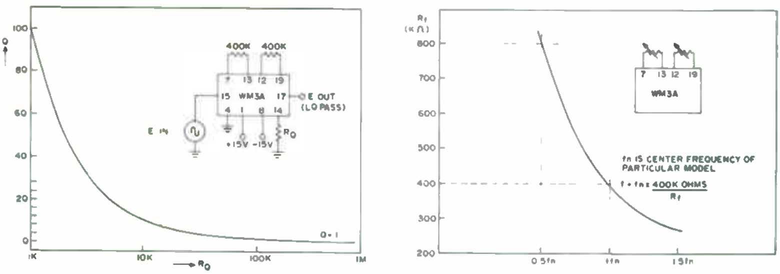


Fig. 5. The above graphs indicate how by simple external potentiometer control, the frequency as well as the Q of the filter can be varied.

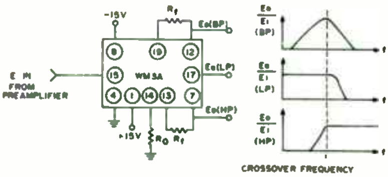


Fig. 6. The above diagram illustrates how the filter can be used to simultaneously produce different output characteristics. Rf and Rq are chosen according to Fig. 5.

simple circuit which provides simultaneous or switch selected different output possibilities.

Summary

The type of filter described is available now from firms such as Western Microwave. The price of such a filter—about $30 depending upon the type of casing used—will restrict its use in amateur equipment until greater sales will invariably bring the price down to that of regular integrated circuits. However, even though one may not be using such a filter tomorrow, such filters will be the type of component that will become common in amateur equipment as the micro-miniaturization of components for use if communications circuits continue.

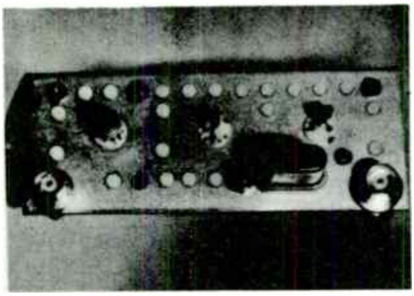
. . . W2EEY

Integrated Circuit 6 Meter Converter

Edward Levy, WB4KMB

At the present time, amateur 6-meter converters are built around nuvistor tubes or field-effect transistors. The former requires a b-supply in the order of 100V and filament power; both types usually require neutralization, which is tedious and requires additional components.

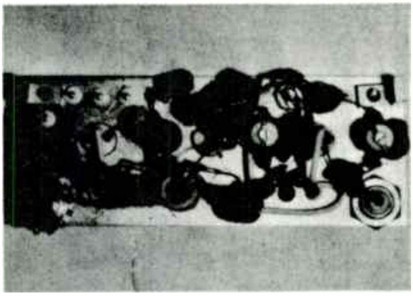
The 6-meter converter described in this article uses an integrated circuit in the front-end, eliminating the need for neutralization and yielding far better gain characteristics than the nuvistor or all-FET counterparts.



Circuit Description

This converter uses a CA3028A integrated circuit in the rf amplifier, which is connected in cascode configuration. The cascode circuit behaves as a pentode tube, and thus prevents tendencies toward oscillation. These tendencies are further reduced by using toroidal coils, which prevents generation of stray magnetic fields and eliminates the need for shielding the input from the output tank circuits.

The amplified 6 meter signal is mixed at the gate of a 2N3819 FET mixer with the local oscillator signal, generated by a crystal controlled 2N3819 FET overtone oscillator operating at 49.5 MHz This produces a difference frequency of 1.0 MHz which is in the middle of the broadcast band, making the converter ideal for mobile applications. If a 7 MHz difference frequency is desired, a 43.5 MHz crystal is used and the 3900 pf capacitor is the mixer drain tank circuit is changed to 100 pF



The converter has a gain of 36 dB up to the mixer drain tank circuit. However, since the tank circuit is high impedance, a link coupling must be used to match to the receiver input impedance, which is usually in the order of 50 ohms. The link consists of three turns of no. 22 wire around the mixer drain tank coil. Matching losses are unavoidable, but a gain of 24 db can be achieved without much difficulty. Typical bandwidth is 100 kHz.

Alignment

With a VTVM connected at the oscillator gate, tune the oscillator tank circuit for maximum negative voltage. Remove the VTVM, and connect an oscilloscope at the mixer drain terminal. With a 1mV 50.5 MHz signal connected at the input, tune the output tank circuit of the rf amplifier for maximum signal. The signal at the mixer drain terminal should be sinusoidal, at a frequency close to 1 MHz if it is not sinusoidal,

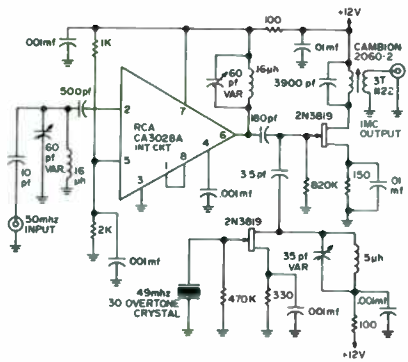


Fig. 1. Schematic

detune the oscillator plate tank circuit slightly. Now tune the mixer drain tank circuit for maximum signal, and tune the rf amplifier input and output tank circuits for maximum signal.

Construction Details

The converter was built on a piece of aluminum, with teflon sockets and terminals used for mounting components, although if built on a punched phenolic board with standard terminals the circuit will work as well. The only precautions to observe are making the connections as short as possible and to keep the coils well separated from each other.

The toroidal coils are not visible in the photographs because they are mounted under the trimmer capacitors.

The converter costs less than $20.00 in parts and outperforms by far many of the commercial units that are presently available for almost twice the cost.

Digital Counters

Robert Suding, W0LMD

"Your frequency is 3637.490 kHz, OM. Would you please move 10 Hz higher to get on the teletype auto-start frequency?"

You'll never hear this on the ham bands? Wait 'till you contact me!

Ever dream of building a receiver/transmitter, or maybe an rf signal generator that would give a digital readout of frequency to the nearest hertz? Well, just you read this article about the digital indicators, and put a little tin on your dream rig.

In building such a project as this, there are 4 main considerations that must be kept in mind:

1) Usages, actual and possible.
2) Reliability (bulb burnout, etc.).
3) Component availability (readout Indicators, transistors, integrated circuits, resistors).
4) How complex; number of parts, amount of wiring (transistors vs integrated circuits).

Having come this far, I now wish to show you the different ways you can get a 0-9 number, then discuss various driver circuits to light up your chosen indicator.

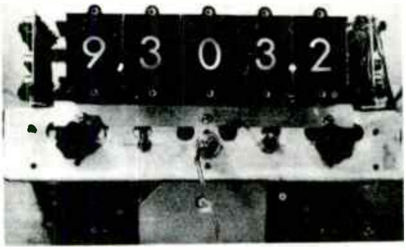
Numbered Light Bulbs

The simplest of all digital displays would be those of either **Fig. 1a** or 1b. **Fig. 1a** is merely a piece of clear plastic painted black on the front except where the number is. Lamps are placed behind each number, and shielded from showing through to the adjacent number.

Fig. 1b shows a similar method, only the light bulbs were encapsulated in plastic and little plastic letters were glued over the top of the appropriate bulb. Other arrangements, such as bulbs in a circle, etc., are also a possibility in this general type.

Advantages: Low cost—only the ten bulbs and a little plastic need be bought. Simplicity.

Disadvantages: Relatively short bulb life; small image, necessitating close reading; Out of line reading of long numbers, making the reading slower.



Front view of my digital frequency counter with a range of 50 Hz to 100 kHz and beyond, with a readout of 1/10 of a cycle. Projected image type of readouts are shown, driven by 5 digital decoder #2's.

Light Bar Matrix

Fig. 1c shows a system used by the Simpson III digital voltmeter and others, which presents several advantages while adding a few disadvantages too.

Seven lights are placed so as to form 10 distinct numbers by lighting up various combinations of light bars. To get a "1", bar C & F are lit; "2" has bar A, C, D, E, & G lit to give 2; "3" has bar A, C, D, E, F, & G lit to give 3; "4" has bar B, C, D, F, & G lit to give 4; "5" has bar A, B, D, F, & G lit to give 5; "6" has bar A, B, D, E, F, & G lit to give 6; "7' has bar A, C, & F lit to give 7; "8" has all bars lit; "9" has all bars lit except E to give 9; "0" has all bars lit except D to give 0.

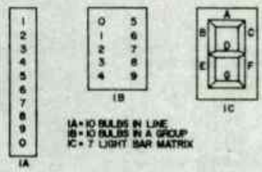


Fig. 1. 1a) 10 bulbs in line; 1b) 10 bulbs in a group; 1c) 7 light bar matrix.

These lights may take a number of forms. They may be a piece of plastic backlighted to give a bar on the front. Another possibility is to use long lamps similar to those used in car dome lamps, only with low amperage. Even NE-2 neon bulbs may be used, along with high voltage switching transistors.

Advantages: Larger Number
Brighter number
Inline reading of long numbers
Cheaply made in comparison with commercially available readouts.

Disadvantages: Proper lamps difficult to find.
More difficult assembly as compared to simple numbered lights.
Brightness varies according to number of bulbs lit.
More current is drawn due to a number of lamps being lit.
Electronic driver circuit is more complex, though not much more expensive.

Edge Lighted

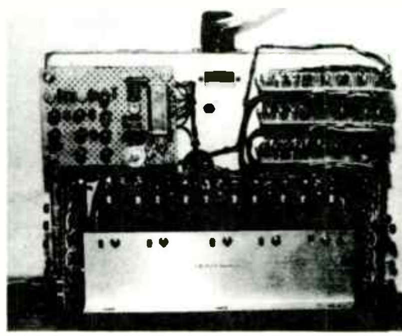
Some manufacturers have produced digital readouts which consist of 10 concentrically placed pieces of plastic, each with a number placed on it which is edge lighted to show the desired number. An example of this type is the readout sold by Radio Shack Corp. for $9.95.

Advantages: Very compact
Well formed number
Inline reading of long numbers.

Disadvantages: Dimmest of all displays.
Small angle of viewability.

Projected Image

This system consists of 12 light bulbs, 0-9 and + and -, which are projected on the back of a ground glass screen. Various bulbs may be used to achieve different levels of illumination. A #44 bulb will give the most brilliant display, but it requires a rather high current from the switching transistor, 250 mA. On the other hand, an 1819 bulb @ 40 mA is rather easy on the transistor, but is quite dim, having a relative character brightness of 15 as opposed to 145 for the #44 bulb.



Top view, showing how I mount the various decoders. Section at back left is the crystal oscillator and gating circuitry. Not shown, but beneath this section are the 24 flip-flops in the ' divide by one million" section.

By reducing the voltage 10%, bulb life of 3000 and 1000 hours, respectively, will be increased 5 times, for 15,000 hours for the #44. Relative brightness will be cut in about half. The popular #47 at 150 mA will perform at about ½ the brilliance of the #44.

Cost of these units varies considerably. New ones run in the neighborhood of $30 each. However, various surplus stores do have them at prices from $6 to $10, the former being about the most that I would pay. At that price, a bank of 5, as I myself use, makes a very attractive display unit for $30.

Advantages: Nice looking image
Ease of mounting

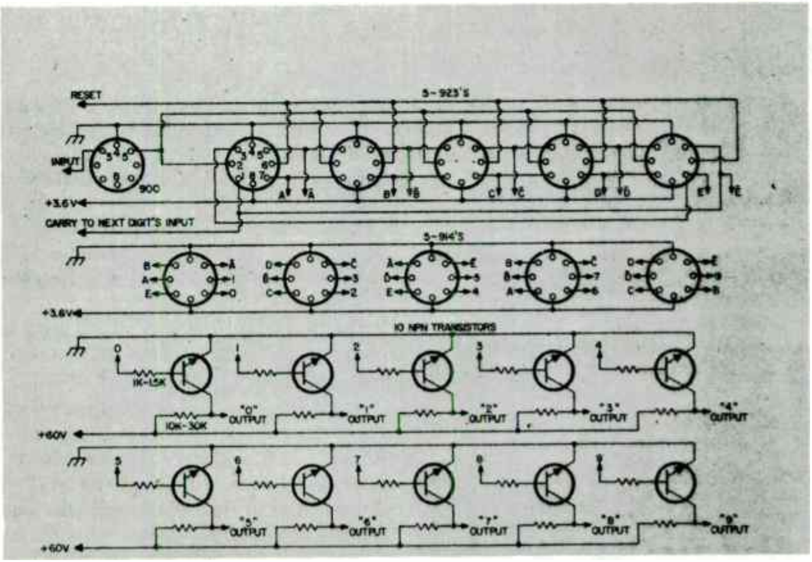


Fig. 2. Digital decoder #1. Reset to 0 by momentarily placing +3.6 v on reset line. All resistors of a similar function are the same value.

RESET TO ZERO BY MOMENTARILY PLACING 3.6V ON RESET LINE.
ALL RESISTORS OF A SIMILAR FUNCTION ARE THE SAME VALUE.

Simple and cheaper driver circuits
Low cost (if surplus)
Inline reading of long numbers

Disadvantages: Bulb burnout
Bulky

Nixie Tubes

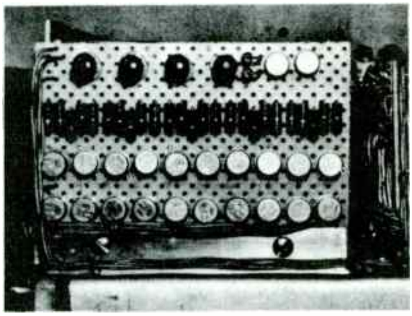
These readouts are a gas filled, cold cathode type tube. They are somewhat similar in basic idea to a vr tube or a neon tube, only they display a given number depending on which of a number of cathodes is hooked up to negative voltage. When this happens, the gas around this element ionizes and glows. Prices on these devices vary according to size and construction. Units of interest to amateurs run from $8 to $30 in price, the $8 one looking like a miniature tube, giving numbers .6" high. These units require special sockets running about $1 apiece. I have seen Nixie tubes on the surplus market as low as $3 apiece. For information, write to: Burroughs Corp., Electronics Components Division, Plainfield NJ 07061.

Advantages: Very long life (200,000 hours)
Bright and easy to read
Compact
Wide angle of viewability
Simple driver circuit
Different sizes available

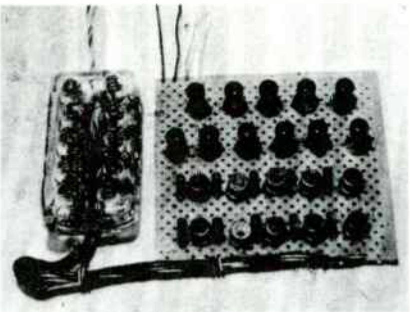
Disadvantages:
Require power supply of about 200V
Special circuit needed for dimming
(I don't like their red colored number.)

Pixie Tubes

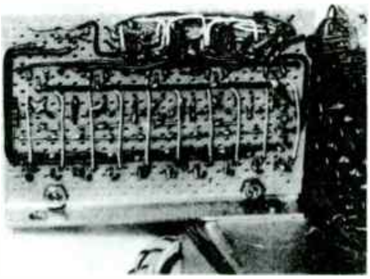
These are units similar to the Nixie, but instead of seeing a relatively large number lit up, a small number is visible through a perforated plate above the lit-up cathode. The advantage of these units over the more common Nixie lies in the fact that they are much cheaper, costing only $5 new, and much less surplus. The main disadvantage is that the number images are so small, about 3/16", that it is very difficult to read them at a distance greater than 6 feet. Besides this,



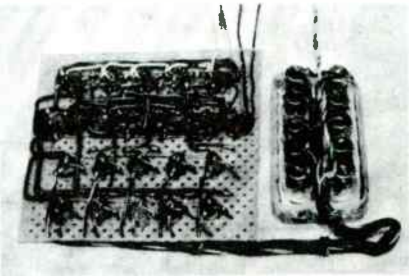
Top view of digital decoder #2. At the top are the IC and 2 transistor gates. In the middle are the other gates. At the bottom are the drivers.



Top view of digital decoder #1. At the top are the 5 flip-flops. The 2nd row has the 900 driver at the left, and the other 5 IC's are the 914's. The other transistors are the drivers. At the left is a digital display which consists of 10—#49 light bulbs encapsulated in plastic.



Bottom view of digital decoder #2.



Bottom view of #5.

a long number would be read slower due to the individual digits out of line.

High Vacuum Readouts

Though similar to a Nixie, these are tubes which project an image quite similar to the way a CRT works. For more information, write to: Industrial Electronic Engineers, Inc., 7720 Lemona Ave., Van Nuys CA 91405. This company can also give you more information on the projected image type readouts, which they make.

Advantages: Long life if brilliance reduced
Other advantages of the Nixie
Can be dimmed by a simple circuit

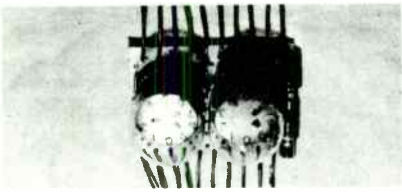
Disadvantages: Hv supply of 1-3 kV required, plus 1.1V filament supply
Smaller image and slightly higher cost than the Nixie.

Digital Drivers

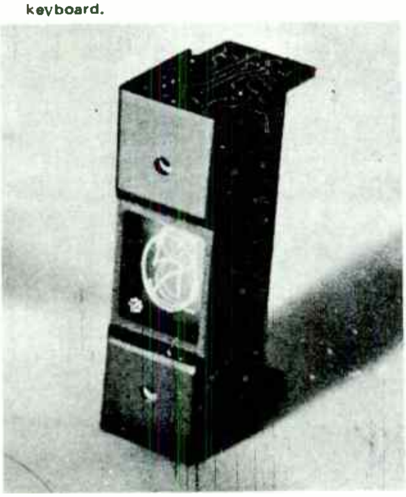
So that you can arrive at the desired digital signal to your selected type of readout, you have 3 items to consider now.

First, you must have a set of flip-flops which will have 10 different states, and then start over. There are a number of different ways to hook up 4-6 flip-flops and additional gating transistors to achieve this. In this article, I have selected 2 ways as being the simplest to work with.

Secondly, you must select the outputs of the flip-flops and steer the proper voltage to the driver stage, which comprises the third part of the whole digital decoder.



Two Pixier indicators, mounted on some keyboard.



An edgelit indicator showing internal construction.

Digital Decoder #1

This decoder needs 31-41 parts: 5–JK flip-flops (Fairchild 923 IC); 1–Driver/Buffer (Fairchild 900 IC); 5–Dual 2 input gates (Fairchild 914 IC); 10–1K-1.5K ½ watt resistors; 10–10K-30K ½ watt resistors (only if Nixie/Pixies used); 10–Driver xstrs (Surplus NPN. or, if Nixie/Pixie used, hv xstr as Fairchild 2N3568)

Cost: $20 if you use Nixie/Pixies and buy the resistors; $12 is you use lamps and have the resistors.

Wiring time: about 6 hours from start to finish.

To build this circuit, a great deal of care is necessary to avoid errors. The way that I wire them is to put all of the ICs in the order shown in the diagram, then wiring the common pins, 4, 8, and 6 of the 923's. Next, wire the common pins of the driver and gating transistors and ICs. After this, put in the resistors and complete the wiring except the wiring of the 914 decoding gates which is the last to be wired and the most prone to error.

As you can see in the parts list, there are a few different ways to build this decoder. If you plan on using the Nixie/Pixie type of indicator, then you will need to use such transistors as the Fairchild 2N3568 or others with a 60V or more collector-to-emitter voltage rating. The 60V line and 10 resistors to the driver transistors is only necessary when Nixie/Pixies are used also. Fig. 3a shows how to hook up lamps to the decoder, and 3b shows how to hook up Nixie/Pixies.

Digital Decoder #2

This is the unit which I use in my present digital counter. It has only 4 ICs, so cuts down on the cost of the unit, but uses 22 NPN computer transistors, which, while cutting down on the cost, adds to the complexity.

Parts used are as follows: 4–JK flip-flops (Fairchild 923 IC); 22–NPN switching transistors (Surplus computer); 2–2.7K ½ watt resistors; 5–470-680Ω ½ watt resistors; 10–4.7K-6.8K ½ watt resistors; 10–1K-2K ½ watt resistors.

Cost: Well under $10.

Wiring time: About 6 hours.

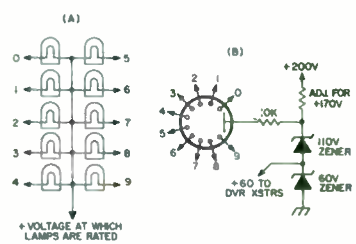


Fig. 3. 3a) How to hook up lamps to the decoder; 3b) How to hook up Nixie/Pixies.

As you can see, there is a definite cost advantage to decoder #2, if you can come by the transistors and resistors cheaply. I find the wiring of decoder #2 easier, due to less wires in the decoder, which is the most confusing part. Should you desire to use decoder #2 with Nixie/Pixie readouts, then change the driver transistors to hv types, and add a 10–30 kΩ resistor to each driver's

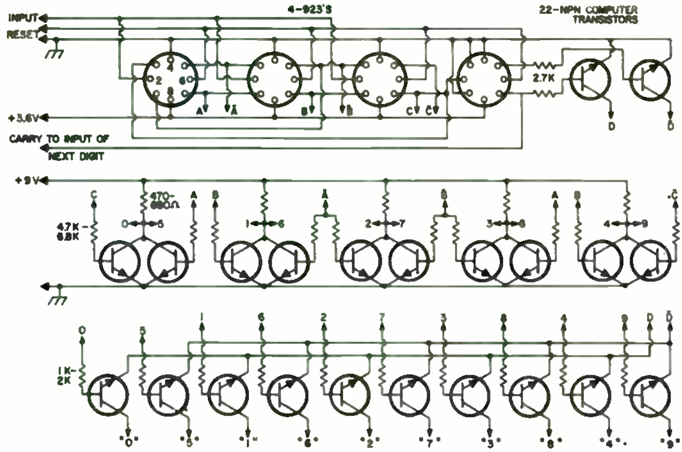


Fig. 4. Digital decoder #2. All resistors of a similar function are the same value. To reset to 0, place +3.6 v on reset line momentarily.

collector from a +60V source, as in decoder #1. Please note on the decoder #2 diagram, **Fig.** 4, that the outputs are not in numerical order. This is due to the fact that the ICs are in a biquinary count configuration, which means that they count to 5 twice to reach 10, then start over.

Digital Decoder #3

This particular unit is made up of either of the two preceding decoders plus a diode

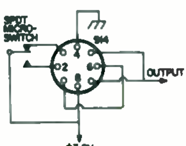


Fig. 5. A simple count checker.

matrix to drive the 7 light bars. As it is rather involved, I am not including it here, but those interested may obtain the schematic by sending me a SASE. The cost will approach $15 to $25, depending on parts availability, and the construction time will be about 10 or more hours.

Count Checker

A simple count checker which can be used to see that everything is counting correctly is shown in **Fig.** 5. Before you hook up the counter to read a multidigit number, make sure you have each digit counting correctly!

Improved Input Sensitivity

Fig. 6 shows the addition of an emitter follower to my original input circuit; this considerably improves the low frequency performance.
mance.

Digital Frequency Divider

As many have mentioned to me regarding the binary counter, it would be a shame to depend on the 60 Hz line frequency for a timing standard for such an accurate instrument as a digital counter. Therefore, I am now using a 100 kHz crystal as the standard. **Fig.** 7 shows that I am using one of Jim Fisk's little crystal circuits into a 914 monostable pulse shaper. This in turn drives 24

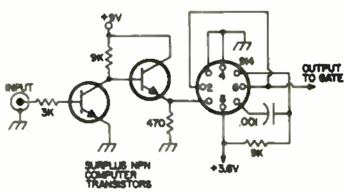


Fig. 6. Addition of an emitter follower to original input circuit of the binary counter.

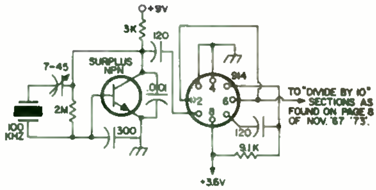
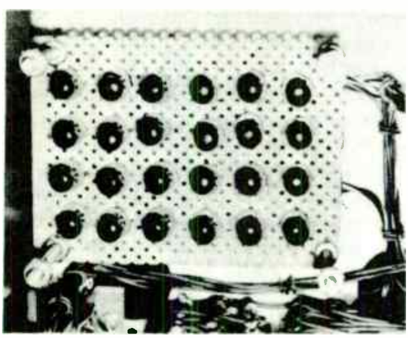


Fig. 7.

923 flip-flops hooked up as 6 "divide by 10" frequency dividers. This results in a "divide by one million," giving me a final output frequency of one hertz in ten seconds. Using this to trigger the count controller and gate, I take a count of all pulses in 10 seconds, thereby giving me an accuracy to one tenth of a hertz. The decimal point is inserted as hertz. The decimal point is inserted as shown in the picture. Various other frequencies are tapped off the divide by one million set of flip-flops to achieve other values of counts. This is described in my previous article.



Top view of the 24 923's in the ' divide by one million" section.

Flip-flops

A number of people have inquired about other types of ICs which might be used. Suppose that you wanted to count up higher in frequency. To do this, you would have to change the time constant of the input shaper by putting in a smaller condenser, and then using higher frequency ICs Say you wanted to read the output of a 7 MHz vfo. Simple. Use Fairchild 926s (freq. to 8 MHz as opposed to 2 MHz of the 923). Make a frequency divide section which will give you the desired accuracy, say 100 Hz. This will require a divide-by-1000 section, or 12 flip-flops. Now you can read out the frequency of that vfo to, say, 7034.8 Hz. How about 20, 15, and 10? Simple too! for 20 meters, add one more flip-flop to the 1000 divider to get 50 Hz. This way, you get double the count, or 14069.6 Hz. For 15, add a divide-by-3 to the divide-by-1000 to get 21,104.4 Hz. For 10 meters, add two divide-by-2 for a divide-by-4. . Exciting? Certainly is! Accurate? Just as accurate as when you zero that 100 kHz crystal with WWV. Of course you don't have to read out the whole number, but can read out the last 3 or 4 only, since you should know your frequency to within a megahertz. Where do you do from here? Well, I've given you some ideas on the fundamentals. The rest I leave to you.

. . .WØLMD

Bibliography

For those people who are interested in the theory behind digital readouts and decoders, I would like to recommend the following free literature.

Nixie/Pixie tubes: Bulletin 1104A
Bulletin 1095
#616E (General Catalog & Applications.

Write: Burroughs Corp., Electronic Components Div., Plainfield NJ 07061.

High Vacuum/Projected Image readouts: Request information from:

Industrial Electronic Engineers, Inc., 7720 Lemona Ave., Van Nuys CA 91405.

Fairchild IC's: Applications brief 36 on 9960 series
Circuit Notes RTL 1 through 5.
App-120/2, App-118/2 & SL-218

Write: Fairchild Semiconductor, 313 Fairchild Dr., Mountain View CA.

Fairchild parts and information: Request "Designing with Integrated Circuit Components"

Write: Hyer Electronics Co., Denver Technological Center, PO Box 22227, Denver CO 80222.

General Digital information:

Write: Interstate Electronics Corp. 707 E Vermont Ave., Anaheim CA–Request S-139A and Updates. Digital Equipt Corp., Technical Publications Dept., 146 Main St., Maynard MA 01754–Request "Logic Handbook."

Unique Digital TTY Accessories

J. A. Murphy, K5CBA

This article describes three accessories for the RTTY station. The first is a regenerative repeater; you put highly distorted, biased signals in one end and get nice, clean, properly timed signals out the other. The second forms the basis of an electronic stunt box; it performs the "cleaning up" function of the first, plus converting the serial TTY signal to a 5-line parallel signal which can be used to perform various functions on receipt of a specified group of characters. The third performs all the functions of the first two plus speed conversions; with this little goodie you can put 100 wpm gears in your machine for copying the commercials and use the speed conversion function to operate at 60 wpm on the ham bands.

In order to understand the operation of these three devices, let's take a moment to review the manner in which a TTY printer decodes a signal. The start element of the code drops the selector magnet, initiating a mechanical timing cycle. Five times during this cycle the machine mechanically "samples" the condition of the selector magnet and positions the code bars accordingly. In a 60 wpm machine these samples are 22 ms apart and about 4.4 ms long. The length of time between the beginning of the start element and the first sample can be varied with the range adjustment. With the range control set at 60 the samples occur 33 ms, 55 ms, 77 ms, 99 ms, and 121 ms after the beginning of the start element.
The Regenerative Repeater

The Regenerative Repeater

The repeater electronically samples the TTY signal in much the same manner as the printer. A start element initiates a series of seven sampling pulses. The input signal is sampled in the middle of the start element, in the middle of the five signal elements, and 11 ms into the stop element. The condition of the signal at the time of each sample is loaded into a flip-flop memory where it is stored until the next sample is taken. Thus, the output of the flip-flop is a perfectly timed TTY signal, delayed 11 ms from the input signal, with signal elements corresponding to the condition of the input at the time of sampling (see Fig. 1). After the stop element is sampled, the circuit resets and waits for the next start element.

The logic required to perform these functions may be implemented in many different ways. The diagrams included here show 803 series DTL integrated circuits. It should be noted, however, that the same functions could just as well be accomplished with RTL, TTL, ECL, or HTL ICs, discrete transistors, or even tubes or relays, to name a few! with this series of DTL a high logic level is approximately +5V and a low logic level is approximately ground.

Referring to the logic diagram of Fig. 2 and the timing diagram Fig. 3, a start element (space) puts *nor* gate U4 input pin 1 at a low level, forcing output pin 3 and inverter U5 input pin 1 high. Output U5 pin 2 goes low, starting the 91 Hz synchronous clock.

"Synchronous clock" is simply a high-powered term for an oscillator that can be turned on in an orderly fashion; that is, the first cycle after the start command has the same period as all the following cycles. A simple oscillator of this type is shown in Fig. 4. It produces a series of narrow positive-going pulses at 11 ms intervals, the first occurring 11 ms after the input goes low.

The clock output drives divider flip-flop U3, the output of which is a square wave with a period of 22 ms. *Nand* gate U4 picks out every other clock pulse and drives inverter U5 input pin 3. The signal at U4 pin 4 is the string of positive going sampling pulses. The first of these pulses causes JK flip-flop U3 output pin 6, which has been high until now, to go low. As long as the K input, pin 3, of this flip-flop remains low the following sample pulses will have no effect on its output. The low output at pin 6 goes to *nor* gate input U4 pin 2, causing the timing cycle to continue regardless of the signal at the input. Note, however, that this happens only after 11 ms of continuous spacing signal. This means that a spacing condition must exist at the input for at least 11 ms to initiate a timing cycle. This provides protection against noise on the signal line.

The first sample pulse also loads the start element into output flip-flop U1 pin 5 and causes the counter outputs, U1 pin 9, U2 pin 5, and U2 pin 9 to step from all (lll), the reset condition, to (Hll). The second sample pulse loads the first signal element into the output flip-flop and steps the counter to LHL. This process continues for pulses three through six. At this time the last signal element has been loaded into the output flip-flop and the counter outputs are LHH. This makes both U4 pin 9 and U4 pin 10 high, so U4 pin 8 and U5 pin 9 go low. Inverter output U5 pin 8 JK flip-flop U3's K input, pin 3, go high. The next sample pulse, which loads the stop element into the output flip-flop, also causes JK flip-flop output U3 pin 6 to go high. This, together with the condition on

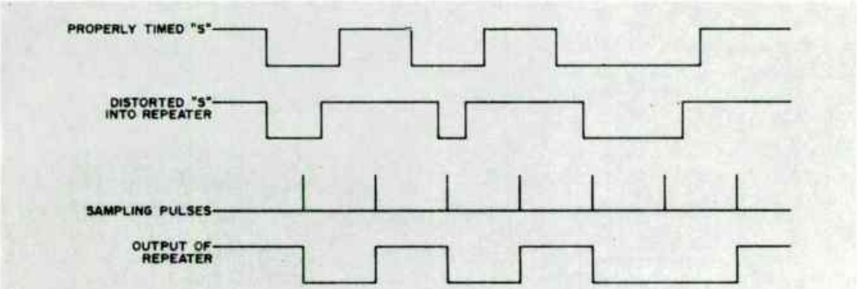


Fig. 1. Simplified timing diagram for regenerative repeater.

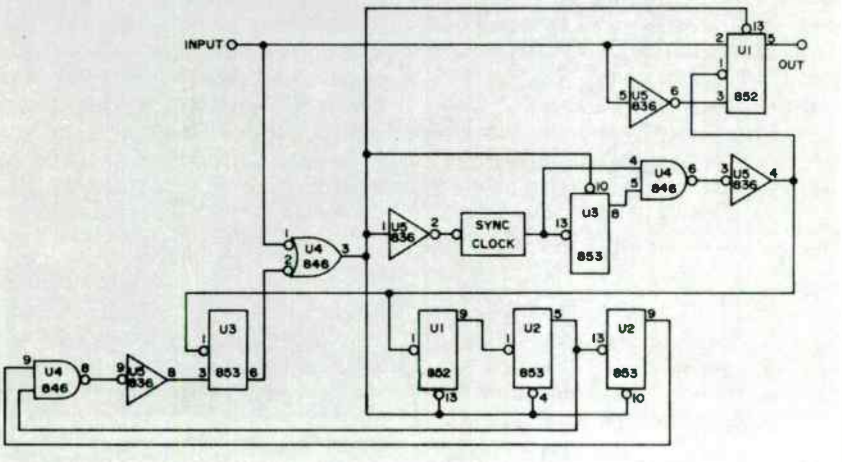


Fig. 2. Logic diagram for regenerative repeater.

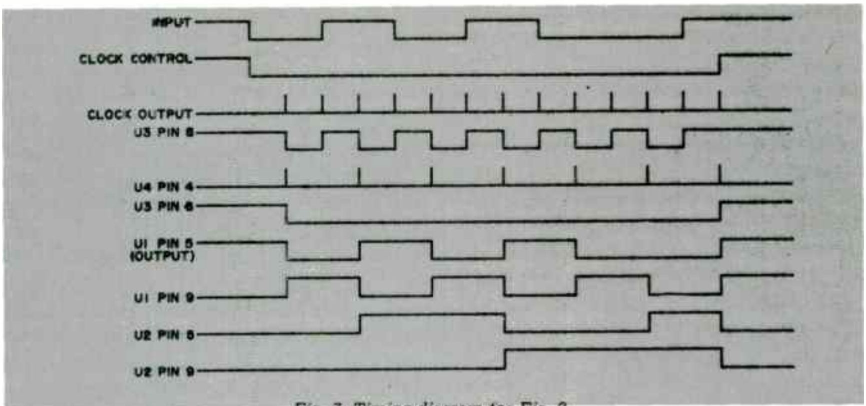


Fig. 3. Timing diagram for Fig. 2.

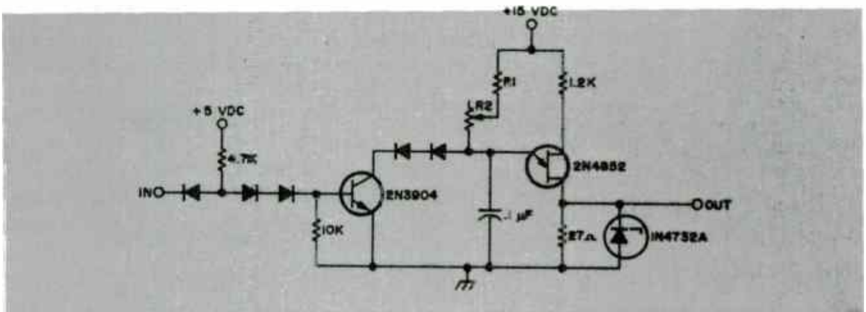


Fig. 4. Synchronous clock.

the signal line, causes *nor* output U4 pin 3 to go low, stopping the clock, resetting the counter to LLL, and forcing the output flip-flop to remain in the mark condition until the next start element initiates a new timing cycle. Just add some simple level converters to make your TU, printer, keyboard, and keyer DTL compatible and you're all set to clean up distorted received signals and to transmit perfectly clean signals from your aging keyboard!

The Stunt Box

The stunt box is a relatively simple expansion of the repeater, requiring only a few more parts and the rerouting of a few wires. Comparing the stunt box (Fig. 5) and the repeater, (Fig. 2), we find three differences: First, the output flip-flop is now the first stage of a six-stage shift register; when the first signal element is loaded into U1 pin 5 the start element moves to U1 pin 9, and so on until after the sixth sample, when the start element is present at U6 pin 9 and the five signal elements are stored in the first five stages of the register.

Second, instead of counting out seven sample pulses with a counter, we now wait for the start element to move into the last stage of the register, causing U6 pin 8 and U3 pin 8 to go high. Now, as before, the seventh pulse samples the stop element and resets the whole works.

Third, 11 ms after the sixth pulse we find that U5 pin 9, U5 pin 11, and U5 pin 13 are all low, allowing U5 pin 8, 10, and 12, and U4 pin 12 to go high and forcing U4 pin 11 low. Thus, we get both a positive and a negative strobe pulse which occur once each character, and at a time when the entire character code is stored in the register. This is all the information we need to decode a character.

A simple stunt box decoder is shown in Fig. 6. This decoder is set up to respond to the two character sequence ZB. When the positive strobe pulse occurs the two flip-flops sample the *nand* gate outputs U7 pin 6 and U7 pin 8. If the character in the register is anything other than a Z, U7 pin 6 will be high when strobed and U8 pin 6 will remain low. A Z forces U7 pin 6 low and U8 pin 6 goes high after the strobe pulse. If the next character is anything other than a B, U7 pin 8 and U7 pin 6 will be high at the time of the strobe pulse and U8 pin 8 will remain low. If, however, the character after the Z is a B, U7 pin 8 will go low and U8 pin 8 will go high and remain high until the next character is received. Using this decoder as a starting point, much more complex decoders can be built to respond to any number of character sequences of any length and used to set and reset latches to turn your printer, tape unit, or coffee pot on and off on command.

The Speed Converter

Now that we have taken a 60 wpm TTY signal and shifted it into a register where we have it temporarily stored, all that remains to be done to make it print on a 100 wpm machine is to load it into another resister, in parallel form, and shift it out to the printer with shift pulses that occur every 13.5 ms. The hardware required to accomplish this is shown in Fig. 7.

Before a character is received U6 pin 8 and the input to the 74.2 Hz synchronous clock are low, the clock is running, and the register is shifting. But since the input to the first stage of the resister, U10 pin 3, is held low and the output to the printer, U14 pin 6, remains high, or marking. When the start element of a character shifts into the last stage of the first register U6 pin 8 goes high and the 74.2 Hz clock stops. When the strobe pulse occurs, 11 ms later, U5 pin 10 goes high, loading the five signal elements into the first five stages of the second register. At the same time U4 pin 11 goes low, forcing a space, or start element, into the last stage of the second register. The output to the printer remains in the marking condition, however, since U6 pin 9 is holding U14 pin 4 and the input to the output *nand* gate at a low level. When the stop element has been sampled and the first register resets, U6 pin 9 goes high, allowing the start element in the last stage of the second register to appear at the output to the printer. At the same time U6 pin 8 goes low, restarting the 74.2 Hz clock and allowing the character to shift out of the register to the printer. As the character shifts out the register fills

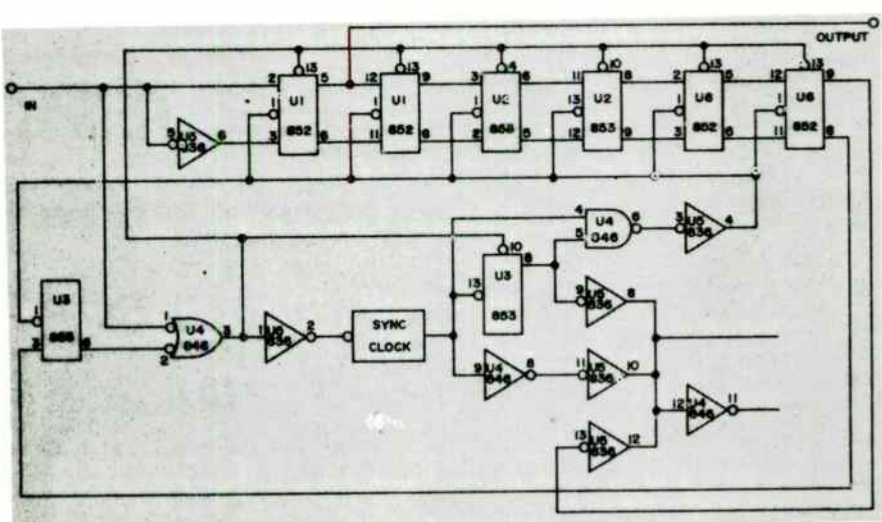


Fig. 5. Basic stunt box.

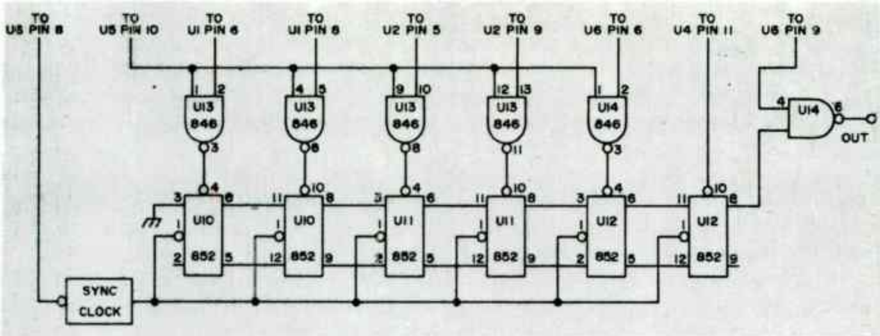


Fig. 6. Simple stunt box decoder.

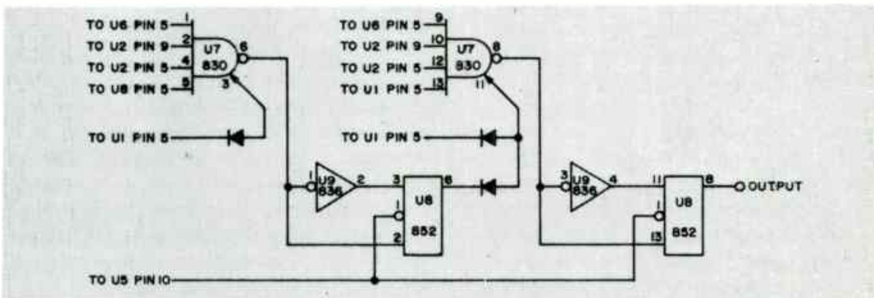


Fig. 7. Speed converter adapter for stunt box.

up with marks so that 13.5 ms after the last signal element is sent to the printer the output goes to a marking condition and stays there until the next character arrives.

To convert the 100 wpm output of the keyboard to a 60 wpm input for your keyer it is only necessary to reverse the input and output leads of the converter and switch the clock frequencies from 91 Hz and 74.2 Hz to 148.4 Hz and 45.5 Hz. Of course, the keyboard may only be operated at typing speeds up to 60 wpm, but that shouldn't cramp the typing style of too many of us!

Understanding and Using Integrated Circuits

Rbt. A. Hirschfield, W6DNS

The transistor was born amid predictions that vacuum tubes would soon disappear; then, some ten years ago, the first integrated circuits led to speculation that both tubes and transistors would be replaced. While neither of these predictions has yet been completely fulfilled, the last few years have seen substantial shifts towards exclusive use of ICs. It is becoming increasingly profitable for equipment manufacturers to substitute inexpensive, complex "subsystems" for the array of components formerly used, eliminating at the same time much of the assembly work.

For amateur radio, particularly the "homebrew" man, the wide availability of cheap ICs opens up new possibilities. Commercially built ham equipment should simultaneously improve in performance and decrease in cost. As spectrum space becomes more precious, and more sophisticated modulation methods become necessary, ICs will allow uncomplicated design and construction of transmitters and receivers which would have once caused casual weekend experimenters to give up in despair.

What Is an Integrated Circuit?

There's no reason to be mystified. An IC is simply a garden variety transistor circuit, in which a number of transistors,

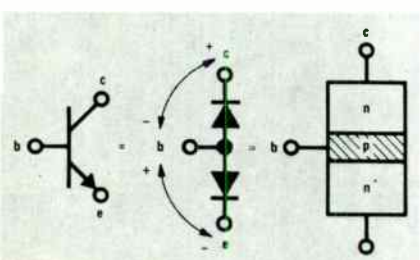


Fig. 1. An NPN transistor as two back-to-back diodes.

diodes and resistors have been made at the same time, by photographically and chemically treating a piece of the base transistor material – silicon. The IC, or microcircuit, is a natural outgrowth of the process used to make single transistors, so let's use the transistor as an IC primer. In its simplest definition, the NPN transistor of Fig. 1 consists of a forward-biased diode (emitter–base junction), and a reverse-biased diode (collector–base junction), which can be made from a slab of silicon by "diffusing" P- or N-type chemicals, in a high-temperature furnace, into selected regions of the slab.

Assuming that we start with a slab already containing N-type chemicals, Fig. 2A, and chemically "grow" a thin layer of glass (silicon dioxide) on top, the glass layer is now a shield against penetration of the slab by other chemicals. Using the same kind of photo-etching techniques used in the making of newspaper type and photo plates, we make a "window" in the protective glass, Fig. 2B, and then place the slab in a furnace containing P-type chemical gases. The result is Fig. 2C, a "diffused" region under the window in which the majority of atoms are now P-type, plus a regrown layer of glass over the window. A second window, smaller than the first, is opened, and the process is repeated in an N-type furnace, giving the structure of Fig. 2D. Windows are then opened, and contacted by a thin layer of aluminum, for each of the three regions in the slab, Fig. 2E. Attached to wires, and brought outside a can enclosing the slab, these three regions become the collector, base, and emitter. We now have an NPN transistor!

Why make just one transistor? Why not use a photo-etching process that makes many transistors, side by side, on the same slab? This is in fact how most single

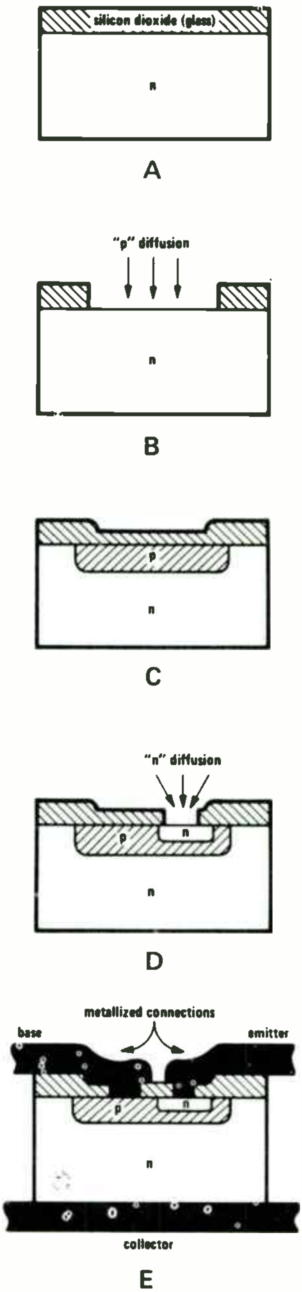


Fig. 2. Construction of a silicon NPN transistor (cross-section view).

transistors are made today. Starting with one N-type collector region, many bases and emitters are diffused, with the resultant thousands of transistors, being cut apart afterwards and put in individual packages.

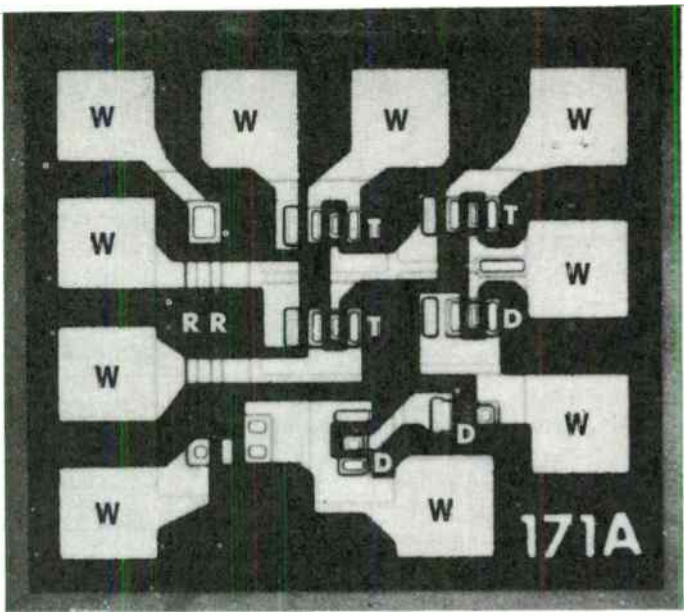
Suppose we want to build other components – for example, a resistor – using the same processing steps employed in making transistors. Why not use the built-in resistivity of the chemically treated silicon as a resistor? And why not cut apart the transistors in interconnected groups, rather than into individual pieces? One problem exists: isolating one element from another. Remember, in a slab, or "wafer," of NPN transistors, all collectors are in one region, and are thus shorted together. The solution is a fourth, P-type region, separating each transistor or resistor region (Fig. 3). If we bias this "isolating" region so that it always forms a reverse-biased P-N junction with everything it surrounds, each "component" in the IC will be effectively isolated by a nonconducting, open diode. We can now form any combination of transistors, resistors, and diodes we like, simply by choosing which "windows" we open at each step in the fabrication process.

What Are the Advantages of ICs?

Economy – Since we must go through the photochemical process to make transistors anyway, it costs very little more to make several of them; putting many transistors in one package is more economical than separating them. In the long run, an IC gives a lower "cost per transistor" than a discrete transistor circuit.

Ease of Use – Most of the signal processing and dc biasing is done within the microcircuit, so less wiring is needed to bring signals in and out of individual transistor stages, or to interconnect bias voltages.

Performance – As we've seen, there's nothing magic about the individual transistors used in an IC; if anything, they may not be quite as good at high frequencies as their discrete counterparts, because of the capacitance to the "isolation" region. Their real advantage in performance is that we are no longer limited to designing circuits



This photomicrograph of an IC rf amplifier shows NPN transistor elements (T), diodes (D), resistors (R) and the wire bonding pads (W).

to use a minimum number of expensive transistors; we can design a better circuit that uses four or five times as many transistors, at less total cost than the discrete circuit. Moreover, interstage coupling and capacitance is much less of a problem than with "discretes," because of the microscopic lead lengths between components. Finally, because all parts of the IC are made simultaneously, by accurate photographic techniques, excellent component matching is guaranteed. For this reason, integrated circuits are available which simply are impractical using single, hand-matched transistors and resistors.

What Are the IC's Limitations?

Large Component Tolerances – manufacturers, despite their most careful controls, cannot prevent a wider variation in resistance, transistor beta, etc., than would be allowed in conventional circuits. A 1 kΩ resistor might well be 750 kΩ on

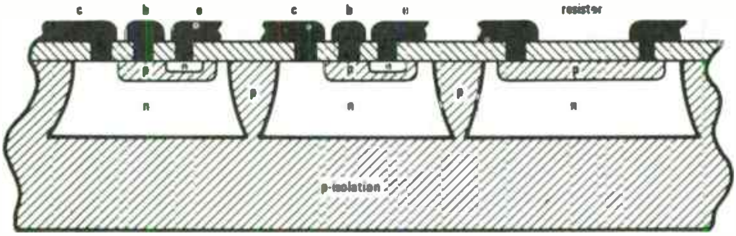


Fig. 3. Cross-section view of an IC, showing two NPN transistors and a resistor.

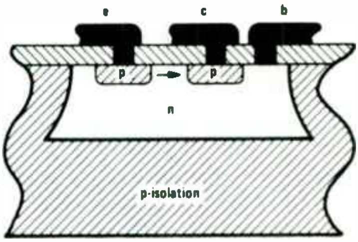


Fig. 4. Lateral PNP (cross section).

units from a given production run, and as high as 1.25 kΩ on the same type of device from a different run. This was considered a problem in the early days of ICs, but designers soon discovered circuit techniques which overcame uncontrolled tolerances. While *value* of monolithic elements varies from run to run, the *ratio* of adjacent components is very accurately controlled. For this reason, most ICs are designed to rely mainly on ratios, rather than component values.

Limited Types of Components – Since the basis for the IC process is the same as that used for NPN silicon transistors, we might expect NPNs to be the best components resulting from that process. Adding resistors, diodes (which are really transistor base–emitter junctions), and zener diodes (which are reverse-biased base–emitter junctions), there is a much more limited selection of practical components available to the IC designer than with conventional circuits. Capacitors of any reasonable value take up too much chip space, which increases the IC's cost. A 0.01 μF capacitor, made either by using the capacitance of a large, reverse-biased PN junction, or by using the "glass" layer as a dielectric, with a metalized top plate would be several times larger than the largest ICs commercially made today. Thus, when a capacitor is needed for an IC, it must be placed external to the IC itself. In an attempt to minimize the number of external capacitors, many new ICs depart from the usual RC-coupled interstages, and are instead completely direct-coupled.

Similarly, inductance on a chip is impractical, so that ICs designed for tuned amplification generally require external tuned circuits.

While PNP transistors are used, they are not capable of nearly as good performance as NPNs, since the chemical processing is not optimized for them. Two types of PNP are available: the *lateral* PNP (Fig. 4) and the *substrate* PNP (Fig. 5). The lateral PNP

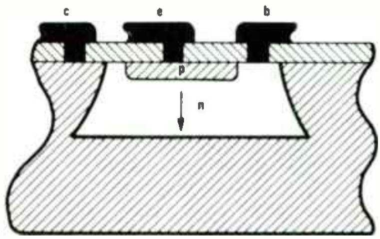


Fig. 5. Substrate PNP (cross section).

uses the NPN collector region as base, with P-type emitter and collector formed by two adjacent diffused-base regions. Current flows sideways, or laterally, from emitter to collector. Since the width of the base

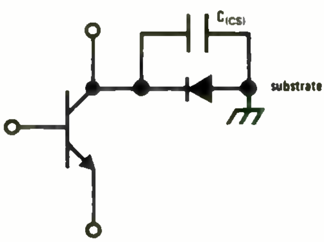


Fig. 6. The "Parasitic" collector-substrate diode.

region is much wider than used for an NPN, the lateral PNP has poor frequency response and a low beta. Thus, lateral PNPs are not used as gain stages, but only as dc level shifters, where necessary. The substrate PNP uses the same type of emitter and base as the lateral; however, its collector is the P-type isolation region, or substrate, which limits the usefulness of this transistor, since its collector is always grounded, and its characteristics as poor as those of the lateral PNP.

Medium-power NPN transistors are now practical; they take up a great deal of chip area because of the need to spread their dissipation over a larger region. Their collector saturation resistance is not as good as very inexpensive "discrete" power transistors, because IC collector contacts are necessarily made from above, while ordinary power transistors, which don't have isolation regions, have bottom collector contacts, with a shorter current path, and hence lower resistance. For most practical designs, an IC can handle power outputs to a watt or so, but must drive an external power transistor to obtain power levels. This also keeps IC power dissipation within the limitations of its small package.

Parasitics – Here is where the actual characteristics of an IC depart in unexpected ways from those of the same circuit in discrete form. The culprit is the P-type isolation region, which separates the various IC elements; it is a reverse-biased PN junction (Fig. 6), which is assumed to be an open circuit. But such a junction produces a parasitic shunt capacitance from NPN collector to ground, which can deteriorate high-frequency performance of the IC. While IC schematics rarely show this PN junction, it must be remembered by the user who applies supply voltage other than those recommended; since the substrate is commonly connected to the most negative point in the circuit, normal operation does indeed keep it reverse-biased, but an attempt to operate an NPN collector more negative than this voltage will forward-bias the junction, drawing possibly destructive currents. Finally, this seldom-mentioned junction can produce undesired substrate PNP action, or even act as an SCR.

Recalling that one definition of a transistor is that it contains a forward-biased and a reverse-biased junction, it can be seen that if the collector–base junction of the NPN is inadvertently forward-biased, it will form the base–emitter junction of a substrate PNP whether we like it or not. Moreover, this condition produces exactly the same trigger condition in the four-layer *NPNP* IC structure as is used to turn on a four-layer SCR. Under some circumstances, this produces "latchup," which holds until the supply is turned off, or draws enough current to damage the IC. In properly designed ICs, used according to manufacturers' recommendations, parasitic problems will not arise. Experimenters, using nonstandard external hookups, however, may find "parasitics" the explanation for the IC's unexpected behavior.

Conclusion

We've peered inside the marvelous, widely acclaimed IC, and found it to be a logical extension of known circuit designs and transistor fabrication techniques. It has a few peculiarities, but offers convenience, performance, and economy in building all types of ham equipment. A basic understanding of what goes into an IC, its advantages and inherent limitations, will prepare us to look at available ICs, and learn how to put them to work.

. . . W6DNS ■

IC's for Amateur Use

Bill Hoisington K1CLL

You may not like parts of this article, but then a lot of real life isn't all that likable either. I'm just going to put down my viewpoint of ICs as related to amateur use at the present time as I find them. Some can be very useful, and some are not so useful, even though they work well for their original purpose.

I hope this article will help you decide which kind to experiment with, and how to tell the useful ones from the kind that are good for computers only.

Integrated circuits are simply very small transistors, diodes, and resistors, constructed on tiny dice, sometimes called chips, of substrate material.

Capacitors and inductors are not generally included in these devices; therefore, in communications systems the external components are often many times larger than the IC itself. The benefits of ICs can be considerable, however, as in an example such as the HEP 590, which has high gain, low internal feedback, and absence of detuning effect over the entire avc range.

Computer ICs

This is where the whole thing started, and it was a fascinating story, for the science of electronics at least. In a computer there is a tremendous need for hundreds, thousands, and millions of memory cells, switches, gates, adders, shift registers, "scratch" pads, inverters, counter–dividers, delay lines, digital-to-analog converters, analog-to-digital converters, flip-flops, buffers, comparators, parity generators, current drivers, differential and operational amplifiers, binary decoders, and a host of other device types – all of which are geared not for the ham but for the computer specialist.

In spite of the cumbersome logic that early Scientific American articles declared to be about ready to topple from sheer unreality, some computers *can* do things that cannot be done by men, such as computing an astronaut's course correction in a few seconds, solving vast and complex urgent equations also in seconds, etc.

But what concerns us as amateur–experimenters building new amateur communications equipment is the IC itself. Today's computers use, as one example, a 14-pin flatpack about ¼ in. long by about 1/8 in. wide, and *very* thin. The particular one I have in mind contains some 32 transistors and is the great-grandchild of a two-transisor flip-flop multivibrator, which can "stand on one leg or the other but not on both." To make the pesky little thing foolproof, they included a variety of "constant-current transistors, lockup prevention transistors, phase reversers, (which reverse the "truth table"), and heaven knows what else.

I did detail all the functions in there at one time, a few years ago, and believe it or not the specs on this thing do include a truth table. They do work, even though there is not one capacitor in there. Everything is direct-coupled. With 5V at several hundred amps total on a computer power supply they're happy with 1V in the off state and 4.5V in the on. They work fast, too, like being clocked (pulse driven) by 2,10, or 20 MHz oscillators, or even higher as each year goes by. Yes, this does mean exactly what you're wondering about 2,10, or 20 *million* pulses, or "clicks of the clock" per second.

So the computer can work fast – so fast even that they're now worrying about the time it takes for a signal to travel over printed ribbon connectors from one tray to another. Sound familiar? Like short leads at VHF and UHF?

MSI and LSI

Just another word to let you know that the big manufacturers are not content with just 30 transistors in an IC. No siree!

Medium-scale integration, or MSI, which is just a way station, puts a lot more than 30 devices – really invisible to the human eye now – in that little can.

Large-scale integration (LSI) really gets to be high-density. I have one with 156 leads and 774 announced functions on a single chip.

There are manufacturers who make ICs that amateurs can use, as well as some transistors for TV front ends that look great for VHF and UHF.

Adding another transistor to the design of a chip is a matter of around 3¢ or 4¢, maybe even less today. You draw up a set of masks and reduce them down photographically to where each individual transistor-to-be cannot be seen at all except with a good microscope.

Then these masks are used one after the other to diffuse various materials going onto the chip, such as properly doped silicon, aluminum, gold, etc. – and eventually you have a wafer with fifty or a hundred, or some other larger number of ICs on it. One transistor more or less is thus only a matter of dividing the time involved in drawing it once – plus the engineering time of thinking about how to do it, and how to test it afterwards. This is one of the main reasons for the seemingly large numbers of active devices in some ICs. If there's any possible advantage, in they go! Why not?

Communication ICs

Here is a different story right away. Practically every "radio set" for communication work that I've ever seen or heard of has coils in it (or at least resonators if you go to UHF and microwaves). Now in a little flatpack only a few mils high how are you going to put in any tuned circuits? The answer is, of course, you don't. They go outside. So now where are you? You're up against a conflicting set of requirements.

There are some "natural" divisions of rf, af, and i-f circuitry where attention must be paid to the proximity of the components, as illustrated in Fig. 1. Shielding, or great care, or both, should be used at these points. They are;

1. Rf amplifier input–output.
2. I-f input–output, in particular from the diode region back to the front end – particularly touchy on certain i-f harmonics.
3. Af input–output.
4. Overall feedback from the af output to the front end, even speaker to antenna, this being often just a loopstick and close by.

In the following rf–i-f IC example you will see that certain things *can* be done on those tiny chips which are very interesting – not only for i-f work but for rf also.

Motorola's HEP 590

Motorola has a very interesting IC device for high frequency amplification, the HEP 590. Two outstanding advantages can be noted: low internal feedback, even when using the maximum gain of over 30 dB, and large avc action without detuning of the circuits

It is packaged in a 10-pin can some 5/16 in. in diameter. The leads can be soldered or inserted in a socket. The present suggested net price is $3.99, which is quite low considering what it accomplishes.

How the HEP 590 Works

Figure 2 shows the internal schematic. When avc voltage is applied to the base of Q2, Q3 is turned off and the ac gain will be at a minimum. This action takes place *without* noticeable change in the operating point of Q1, whose input impedance remains constant, with very little detuning effect on the input tuned circuit even with maximum avc voltage on Q2.

The configuration of Q1 and Q3 reduces the internal feedback to a low figure, which is generally immeasurable up to several hundred megahertz. With the 30 dB of gain obtainable at 60 MHz, this is a great advantage.

Diode D1 is for dc biasing of Q1, under conditions of varying temperature. Being laid out on the same silicon die as Q1, their currents will be closely similar, even with severe changes in temperature, with consequent dc stability.

The noise figure of a two-transistor pair is lowest when the input device uses the

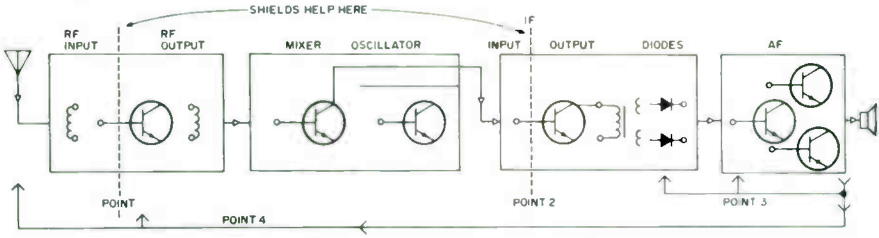


Fig. 1. Block diagram of typical receiver showing "trouble spots."

common-emitter connection. The second device will then have little effect on the noise figure.

The actual measured gain at 60 MHz is over 30 dB in suitable circuits, which consist of no more than tuned and impedance-matched inductors.

By the use of two of these devices in cascade, bandwidths of over 10 MHz may be obtained at 60 MHz, which shows considerable possibility for amateur microwave amplifier service.

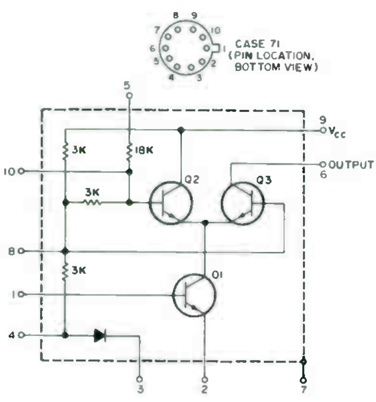


Fig. 2. Internal structure of Motorola's HEP 590 and pin identification data.

The HEP 590 on the Breadboard

The 40 meter amplifier circuit of Fig. 3 was set up on a copper-clad baseboard with soldered connections to terminal strips. The antenna was brought in to a connection on L1 at 16 turns from the cold end and the 590 input lead (pin 1) tapped at 5 turns. The 100 pF capacitors (C1 and C2) are too large for in-band operating units but are all right for experimental use.

Manual control of the gain was accomplished with the 5 kΩ pot, avc tests to be done later in i-f service.

The output of the device (pin 6) was connected to the top of L2, and 6 turns of small wire wound around the cold end.

I started out with loose-coupled antenna, base input, and output link, but soon discovered this is *not* the way to go with the 590. It likes *lots* of coupling, on both input and output. When this was done, the really large gain of the device became evident. Various signals from 80 to below 40 meters sounded like reception with a superhet receiver (except for the selectivity).

When I used the device as an rf amplifier in front of my lab receiver, I had to

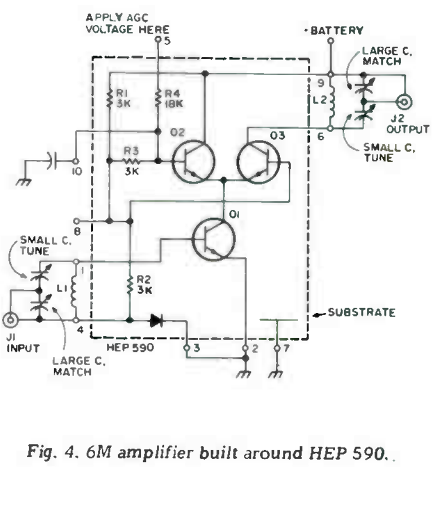


Fig. 4. 6M amplifier built around HEP 590.

reduce the gain of that receiver by a large amount. That HEP 590 really has a lot of sock.

Putting the HEP 590 on 6 Meters

After removing the 40 meter coil and putting in the required values for 6 meters, signals came in right away. Plenty of stations were heard, and in particular, using a 100 ft wire, Q5 signals were heard that could not even be found on the dial without the 590 preamp. It works! The 6m values for the inductors are as follows: L1 – 4 turns at 8 turns per inch, 9/16 in. diameter; L2 – 5 turns at 4 per inch, 9/16 in. diameter L3 – 4 turns, wound on L2. Figure 4 shows the 6 meter version of the HEP 590 amplifier.

So the Motorola HEP 590 IC linear amplifier works as claimed. It is a relatively easy IC to practice on, with only three active devices inside, and can be built up into a circuit and tested in a day's time.

From working with it so far, I can verify that excellent reduced-size, high-frequency i-f amplifiers for amateurs can be made with it.

The Amperex TAA300

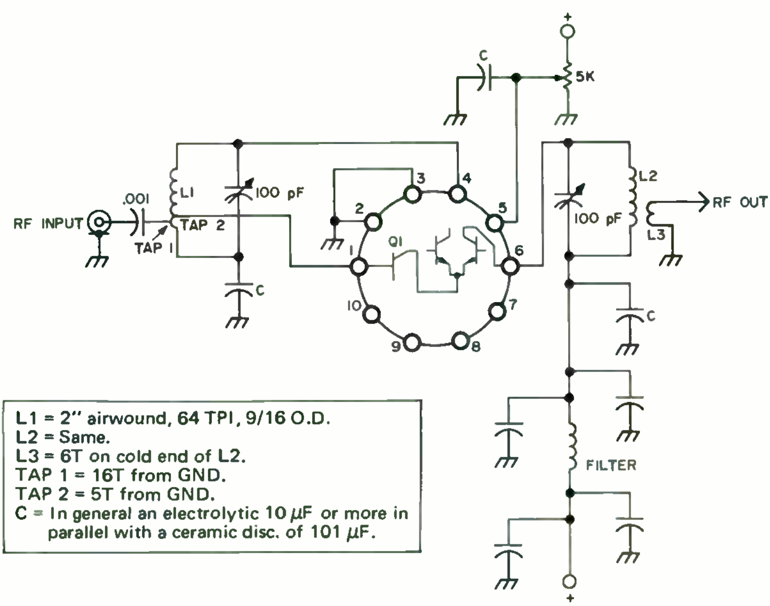
The home of Amperex being not too far away in Rhode Island, I visited down there for half a day and was well rewarded with some new VHF transistors that look swell, and also in meeting some topnotch engineers in the lab.

The TAA300 is a complete af amplifier in one of those little 10-pin cans, and puts out a watt of audio when required – enough to modulate a couple of watts of rf on 6, 2, or 432 – which, along with the exciter section, will take just about all of the dc power of two lantern batteries, rated at 12V and 0.5A.

And it's good for the receiver also. Plug the i-f diode output into it and there's your loudspeaker banging out a watt. Let's take a quick look at one of these.

With ICs, allow for lots of time. Some of this time will be spent in puzzling out three things: the internal circuit, the external circuit, and how to draw one schematic that has both.

I am counting on the TAA-300 to do a lot for use as a 1W af amplifier; but once again, remember that these little gems are



L1 = 2" airwound, 64 TPI, 9/16 O.D.
L2 = Same.
L3 = 6T on cold end of L2.
TAP 1 = 16T from GND.
TAP 2 = 5T from GND.
C = In general an electrolytic 10 μF or more in parallel with a ceramic disc. of 101 μF.

Fig. 3. 40m amplifier built around a HEP 590.

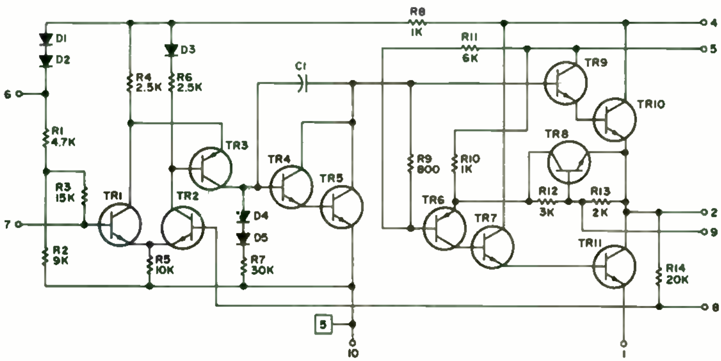


Fig. 5. Amperex TAA300 – a 1W audio amplifier IC.

not primarily made for experimenters – they're made to be wavesoldered into small radios and TVs by the hundred thousands.

An engineer in a large company can afford to sit down at his desk and spend a week or two figuring how to best put this device into a set because that week can be spread cost-wise over (hopefully) a large number of sets if he does a good job. Can you do this just for one item?

Does the circuit of Fig. 5 look like an audio amplifier? Where's the input and the output? What are those five diodes doing in there? Why 11 transistors just for an af amplifier when the "All-American Five" design will give you mixer, oscillator, i-f, af driver, and af power stage?

Don't think I'm running this thing down, because I intend to use it. I just want you to be prepared for a little "new thinking."

With printed circuitry it has always been my feeling that why should anyone build just one? To learn the printing process involved, yes. For building one circuit, no. Now some of these little 10-pin jobs *are* good for experimenters, and you may also be interested in learning about them for business reasons, too – or for size, although to really cut down in size calls for some pretty expensive external components. Look at Fig. 6, the external circuit. There is a .64 μF capacitor on pin 6, 400 μF from pin 2 to pin 5, a 25 μF capacitor on pin 8, and 47 μF from pin 2 to ground. With "ordinary" size electrolytics three of these values are each four times the size of the device itself, but I expect with voice frequencies some of these can be cut down a bit.

Then there are a few application notes to think about, like the stability question,

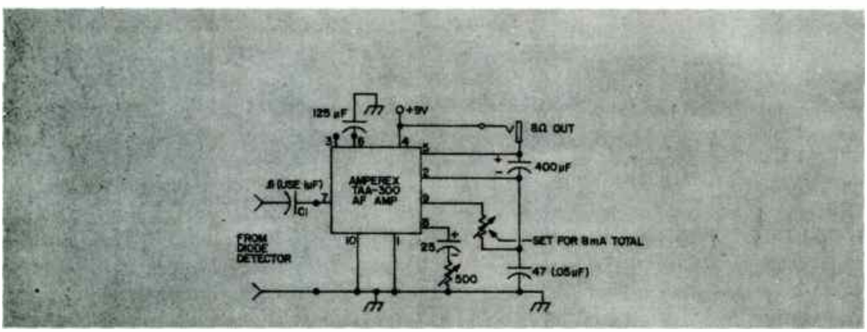


Fig. 6. External connections required to get the TAA300 operating as a complete amplifier.

and a few others, but they are not bad.

All worth-while things take time, and this is one of them. Be advised, and allow yourself time enough to study it out before you pick up that little soldering iron.

Amperex TAD100

Ambitious, knowledgeable, hardworking people have looked at the several-million-per-year market for just plain old radios (new ones, of course) and thought about making ICs for them. And they made some. I have worked with one of them, the Amperex TAD100, and here is the story as I found it.

There are a dozen or so transistors in the little plastic box only 3/4 in. long, as you can see in Fig. 7, the internal circuitry of this one. These transistors are pretty close together in there – much too close for an amateur experimental unit, as you will see.

The manufacturer, ambitious as he was, did produce large numbers of excellent and very tiny receivers with this IC. It has a mixer, oscillator, i-f, detector, and an af amplifier – but no power stages. Almost everything you might need, except you have to add a few things on the outside, of course, as shown in Fig. 8.

There are some good ideas incorporated into this baby – like using two transistors for the mixer. So are three for one i-f stage. I don't particularly go for the four transistors in the af stage, without even the power output unit yet, but it does work.

What doesn't go for the amateur experimenter is the fantastic proximity of the input, the mixer, the oscillator, the i-f and the af – all in there together on that one tiny chip. Just too, too close for me. When everything is running right with all the precautions advised and an exact copy of the original printed board is made up and all the components are in exactly proper places, it does work as a BC set.

One of the precautions listed: "The oscillator must be limited to 100 mV; otherwise it will get into the i-f and the af." It did! I worked 10 days on this particular IC and to make a long story short, the mixer and the i-f ran okay at times (and at times not), but when the internal local oscillator was used, I kept running into trouble.

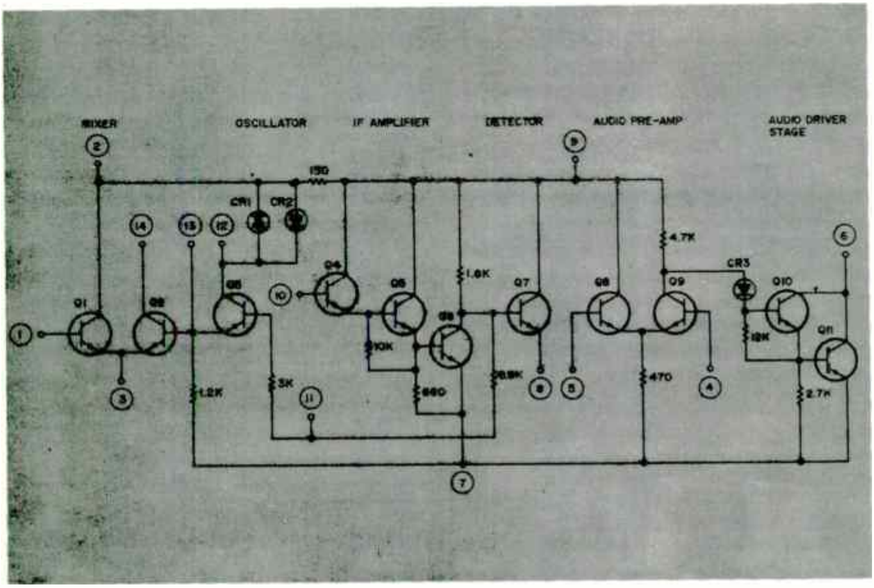


Fig. 7. Amperex TAD100 is a complete broadcast receiver in one tiny package.

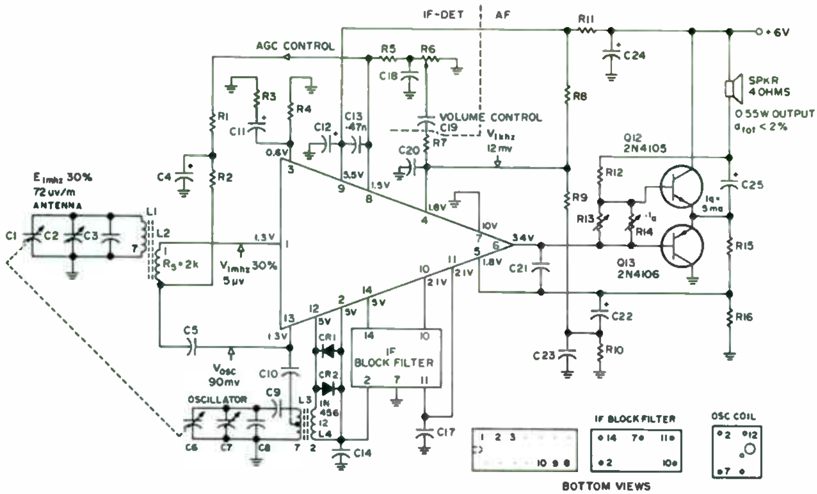


Fig. 8. External circuitry. B.C.–I.C.

Bear in mind, such an IC does work fine for mass production of BC sets. It's just that for an amateur experimenter as a single unit, there is too much feedback involved. Also, the oscillator is strictly limited to the BC band service, which precludes using it as an amateur converter. The people handling this unit are excellent engineers and have some other *very* good devices; and they are still working with newer and better devices for a complete set (including an FM job!), so we will almost certainly see them again soon.

Repeater Audio Mixer

Ray Pichulo W1IRH

Paul Hoffman, W1ELU

The audio mixer described in this article, although designed primarily for repeater use, can be used anywhere it is desired to mix a number of audio inputs with a high degree of isolation between inputs. The mixer is adaptable to almost any configuration which may be required to suit the individual's requirements. The number of inputs can be increased by a factor of two or three to suit the user's needs. The isolation between individual inputs of over 40 dB makes it possible in repeater operation to have tone command information on one channel not be affected by another input. The mixer shown in this article is the one designed for use in the WA1KFY repeater. It has eight inputs – three of them squelched, the other five continuously on.

Operation

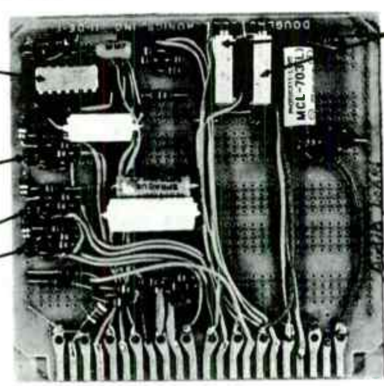
The amplifier uses a single 709D opamp plus one FET for each squelched input. The audio inputs as shown in Fig. 1 are applied to the opamp's inverting input. The output is fed back through 150 kΩ resistor R1. Notice that the signal from each input is applied through 150 kΩ also. The resultant voltage at the input terminal of the opamp is the combination of the input signal plus the out-of-phase feedback voltage. Since both are applied through equal series resistances, the resultant voltage is zero. This condition results in the opamp's having an extremely low (almost zero) input impedance. This resultant low impedance, together with the high series resistance on each input, accounts for the high degree of isolation between inputs. The squelched inputs use a FET across each input as a switch. With zero volts on the gate, the FET exhibits a drain-to-source resistance of about 350Ω, effectively shorting its associated input to ground. When +15V is applied to the gate, the FET switches off, enabling the input channel. The 75 kΩ resistors between each FET and the input bus prevent the input bus from being shorted by the FETs.

709D

2N4360

2N4360

2N4360



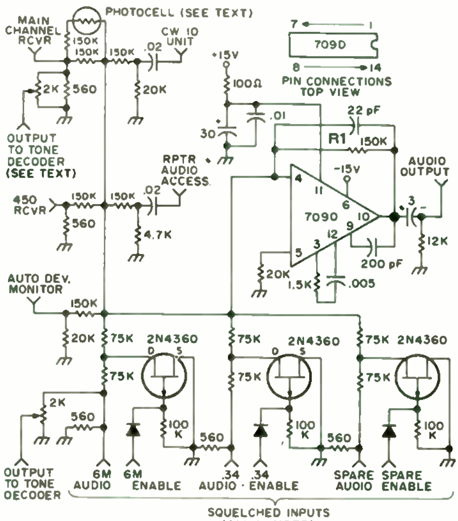
CONTROL OUTPUT LEVEL ADJ.

The gain of the opamp is determined by the ratio of the feedback resistance to the input series resistance. In this case, the gain on all the inputs is unity. If more gain is desired on a particular channel, the input resistance can be lowered to change the ratio (and therefore the gain). For example, it was found necessary to increase the gain from the main channel receiver when the phone patch was connected in order to increase the level into the phone line. (The phone patch does not load down the output; rather, the gain had to be raised in order to properly drive the line.) The gain change is accomplished by a photocell–lamp assembly, with the photocell in series with a second 150 kΩ.

The frequency response of the amplifier is essentially flat from dc up to about 8 kHz. Beyond that point, it rolls off. The 22 pF capacitor across the feedback resistor determines the rolloff frequency characteristc. The audio output voltage swing can go as much as ± 10V. This is more than ample to drive 10 or more high-impedance inputs. The audio amplifier in WA1KFY is presently used to drive three transmitters, a phone patch, and a monitor earpiece. The number of inputs can be expanded to suit individual requirements. The unsquelched inputs require only an additional 150 kΩ series resistor and load resistor for each leg. The squelched inputs each require an additional FET switch in addition to the load resistor and series resistors.

The opamp requires 15V (positive as well as negative) to operate it. The current requirements are approximately 30mA. The components shown in the detail schematic (1.5 kΩ 0.005 mF, and 200 pF) are used to compensate the amplifier against instability. The 100Ω resistor, 30 μF capacitor and 0.1 μF capacitor on the +15V line form a decoupling network.

Two trimpots at the top edge of the board (see photo) are used to provide audio outputs for tone-operated command functions. They are connected across the load resistors of the six meter receiver and the main-channel two-meter receiver as shown in Fig. 1. The arms of both pots connect to pins on the edge connector and



Notes:

1. Resistance values in ohms, capacitance values in microfarads unless otherwise noted.
2. Diodes are IN457'S.

Fig. 1. Audio summing amplifier.

go off the board to their associated tone decoders on another board. These pots are mounted on the audio mixer board as a matter of convenience and would not be used if it is not required to bring audio to another point in the system for tone command or other functions. The photocell–lamp assembly for gain changing on the main channel input is on the right side of the board. Similarly, it could be deleted if this feature is not required.

Summary

This audio mixer provides the repeater owner a high-quality audio mixer system which has minimal space and power requirements. It is especially attractive for use in more sophisticated repeater systems where several receiver/transmitter combinations are used. However, if the same construction techniques as those described in this article are used, it is just as good to use with a simpler repeater system because there is more than adequate room for expansion at a later time.

...W1ELU,W1IRH■

The SST-1, Solid State Receiver for 40 Meters

Calvin Sondgeroth, W9ZTK

This is the age of the transceiver in ham band operation. Although there are many medium- to high-power commercial SSB packages available with self-contained transmitter and receiver, there is not much to choose from for low-power CW work. This article describes a small unit which is solid-state and provides about 2W input with a self-contained battery power supply. The receiver is simple but quite effective for CW work, and its performance matches that of the cheaper superhets. All the components are readily available and no special parts were used in the design. Since the transmitter is crystal-controlled, this little rig would make an ideal beginning station for the Novice.

Receiver

A QST article by D. DeMaw (May 1969) described a solid-state direct conversion receiver using an RCA linear integrated circuit. This design requires a minimum of parts for CW (and sideband) reception since the incoming signal is mixed with a local oscillator at the signal frequency for audio output directly from the product detector. Although the selectivity is not as good as a superhet with i-f conversion, the results are entirely adequate for general operation. The receiver far outperforms any superregenerative set I have built, and I've tried many circuits.

The front end is double-tuned to prevent responses on strong signals outside the 40 meter band; two toroids tuned with a 140 pF double-section variable capacitor accomplish this, as shown in the transceiver schematic, Fig. 1. The incoming signal is lightly coupled to the IC product detector to prevent loading the input circuits, and the local oscillator uses an identical toroid in its tuned circuit. The oscillator circuit is tuned by a 50 pF variable with a small trimmer in series to adjust the bandspread to cover the full 180 degrees on the dial.

Audio from the detector is coupled out through a small transistor interstage transformer with a 10 kΩ potentiometer across the secondary as a gain control. A single stage of audio amplification precedes the audio amplifier, which is the push-pull audio section salvaged from an old transistor broadcast radio. This generates plenty of audio to drive a small loudspeaker (also from the BC set); and if just headphone operation is desired, the 2N3391A will provide enough output by itself.

The original article on this receiver used an audio bandpass filter between the detector and the audio amplifier. This had two large toroids and, in the interest of size reduction, a low-pass filter consisting of a single LC section provides adequate cutoff of high-frequency hiss and noise which is present without any filtering at all.

The inductance in the filter is the secondary of an interstage audio transformer shunted with a 0.1 μF capacitor to ground; this arrangement cuts off around 2000 Hz. The capacitance value can be adjusted to provide proper cutoff with the particular audio transformer used. The receiver is usable without the filter, but the high-frequency components in the detector output become annoying after an extended period of operation.

It will be noted that two 9V batteries are shown on the schematic to power the receiver section. This was done because the *rf–first audio* portion used a negative ground system while the push-pull output of most of the small imported radios uses a positive ground system, and the output transformer secondary has one side tied directly to ground. Various ground arrangements were tried to eliminate one of the batteries, but the two-battery setup was finally decided upon. This does split up the load on the batteries somewhat and increases their life. The rf portion draws about 15 mA and the audio section anywhere from 10 to 50 mA on strong audio peaks.

CW Monitor

A unijunction audio oscillator is included on the audio output module for monitoring while transmitting. The monitor output is fed into the audio output amplifier after the gain control, and the monitor level is set for suitable volume which is independent of the receiver volume control setting. Voltage for the monitor is obtained from the transmitter supply and it is keyed along with the transmitter.

The 100 kΩ resistor and the 0.01 μF capacitor in the emitter lead of the unijunction provide an audio tone of about 600 Hz; this can be raised by lowering the value of the emitter resistor if you prefer a monitor pitch of higher frequency.

Transmitter

The transmitter, crystal-controlled for simplicity, used three 2N697 transistors – two in the rf section and the third as a switch for keying. The keying switch was added to reduce the current through the key contacts, although it could easily be eliminated and the final connected to the negative supply continuously, since it does not draw any current without drive from the oscillator. The keying arrangement shown was a result of using the transmitter which had already been built when the transceiver idea came up.

The oscillator is connected in a Pierce circuit with the crystal between collector and base. The 1 kΩ potentiometer in the emitter controls the drive to the final. The oscillator collector uses a slug-tuned coil with a small link wound over the cold end to the couple into the final amplifier.

The final, operated without any bias, runs class C. It uses a *pi* network in the output circuit.

SWR Bridge and Meter Circuit

For tuneup while operating portable it was considered desirable to include some means of monitoring final collector current as well as some way of indicating when a match to the antenna was obtained, since random-length antennas are convenient. A small pilot light could be used to indicate relative current in the final, but the addition of the 0–1 mA meter and the swr bridge has proved its worth in the field.

The meter is set by a three-position switch to read either final-stage emitter current or forward or reflected power. For emitter-current readings, the voltage across a 10Ω resistor in the emitter lead gives a full-scale reading of approximately 100 mA with the meter specified. A 20 kΩ potentiometer is connected in series with the meter for swr measurements and must be set to zero resistance for final current indications.

The swr bridge is standard in design except for the two transistors added as dc amplifiers. They were found necessary to get indications with the low power output of the transmitter. The gain mismatch between transistors does not really allow the swr to be measured accurately, but a good indication of a proper match to a 50 or 72Ω coax line can be obtained by adjusting for minimum reflected power and maximum forward power. A two-section switching arrangement could be used to eliminate one of the transistors, thus providing more accurate readings. The bridge conductor is a piece of ¼ in. copper tubing; the pickup wires are of 14 AWG solid wire. The assembly is built in a small channel bent up out of aluminum sheet metal as shown in Fig. 2. In order to get readings with the bridge it must be mounted off the main chassis with an insulating spacer and the channel connected to ground via the coax shield on the input and output only.

Construction

The main housing for the transceiver is a Bud SC-3030, which is 6 x 10 x 7 in. This volume allows construction without crowding yet keeps the unit small enough to be easily portable. The transmitter is built on a separate minibox 2¼ x 2¼ x 5 in. and can be put together as a separate unit with a couple of leads provided for a crystal socket on the front panel of the main enclosure. The transmitter is mounted to the front panel by the key jack which has one side tied directly to ground. In addition, a barrier terminal strip was provided on the rear panel for connection of an ac power supply for fixed station work. The drive control potentiometer was mounted inside the transmitter since it is not adjusted in normal operation. The final amplifier transistor was mounted on a piece of .040 aluminum about 2 in. square which serves as a heatsink. Since the collector is connected to the transistor case, the heatsink must be insulated from ground.

A twice-size PC board layout for the receiver *rf–first audio* section is shown in Fig. 3. The integrated circuit is soldered directly into the circuit board, although a socket can be used. Don't be too concerned about damaging the integrated circuit. The one here had to be removed from the board once by cutting the leads above the board and remounted by soldering extension leads to it indicating that these little devices are really quite rugged. Without a proper unsoldering tool, it is difficult to remove the IC once it has been soldered in so a socket might be a good idea even though the board is not laid out for one.

To facilitate parts placement and circuit identification, the composite layout/schematic of Fig. 4 is included. This should help to speed your final assembly process.

The audio output amplifier from the discarded broadcast set was mounted on a small piece of Vector board along with the components for the CW monitor. Fortunately, the radio I dismantled had been built in two sections, with the audio stages on a separate small circuit board. Other radios might have to be operated on to get just the audio section to use in the transceiver. The input to the amplifier can be located by tracing down the leads going to the volume control in the original radio. The wiper arm on the control is connected to the input.

The receiver modules are mounted to a small subchassis which was extensively worked on with tin shears. It is made from a standard open-end 5 x 7 x 1½ in. aluminum chassis. The front is cut off to clear the controls on the front panel and the rest notched and cut where necessary to clear the main enclosure. A vertical shield was positioned across the transceiver between the receiver and transmitter, although this is probably not absolutely essential.

The subchassis provides adequate mounting space at the center (between receiver and transmitter) for the transmitter batteries and the two 9V batteries are mounted to the rear enclosure wall with a homemade bracket of sheet aluminum. The 1.5V penlight cell is mounted on the swr bridge channel as shown in Fig. 2.

The swr bridge is mounted near the rear of the transmitter and the bridge channel is insulated from the main enclosure as mentioned above. Connection to the bridge should be made with small coaxial cable going to the transmitter *pi* network and the transmit–receive switch. Similar coax is used to connect the receiver and antenna connector to the switch.

The photographs show the general construction and panel layout used, but other builders may find other arrangements more desirable. The general layout is suggested as a logical one.

Adjustment and Tuneup

When the receiver is operating, the oscillator frequency is adjusted to cover 40 meters by alternately padding the tuned circuit with fixed capacitance and varying the trimmer in series with the main tuning capacitor to achieve the desired bandspread. By setting the trimmer, the entire band can be tuned or the bandspread limited to just the CW portion. Other arrangements might give different bandspread, but the transceiver described covered the first 100 kHz of 40m over about half the dial and the other 200 kHz over the second half, which was felt to be about right and allowed for monitoring sideband as well as CHU just above the top band edge for time checks as an added bonus.

Some trouble with the local oscillator was experienced at first: Spurious responses were obtained, with 7 MHz ap-

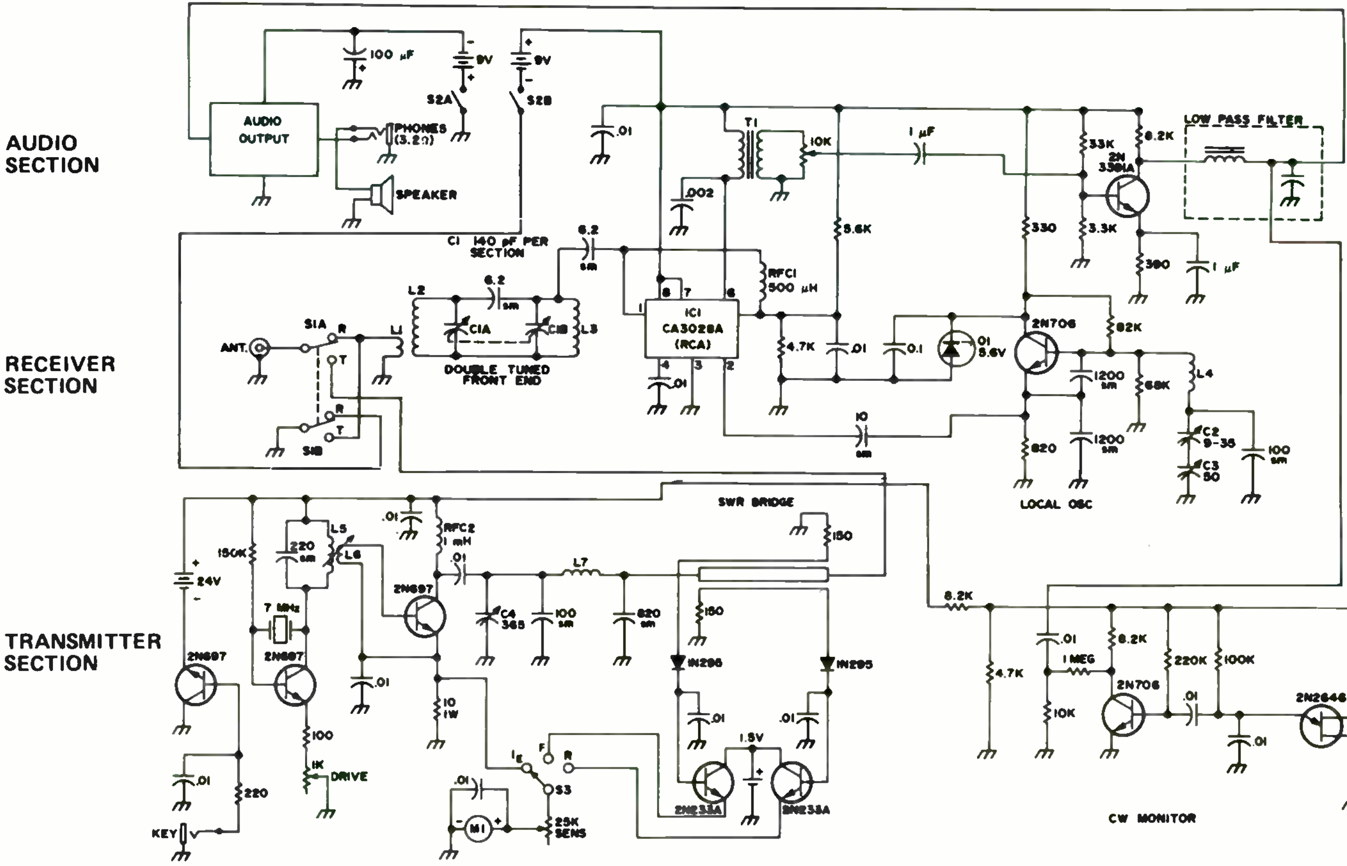


Fig. 1. Schematic diagram of solid-state 7 MHz transceiver.

pearing at several places on the dial. This indicated that the oscillator was operating at too high a level, which generated unwanted outputs. The two capacitors in series from base to ground on the oscillator determine the amount of feedback and the values were set as shown so that the oscillator provides just enough signal to beat against the incoming signal.

The audio output section and the CW monitor can be checked by connecting the monitor to the transmitter supply and making sure that the audio tone is of the desired level and pitch.

When first tuning up the transmitter, it is a good idea to disconnect the final amplifier until the crystal oscillator is operating properly. The oscillator collector coil slug should be tuned up for proper oscillation and keying. With the final connected turn the drive control pot for minimum drive (maximum resistance) and switch the meter to read final emitter current. (The transmitter should have a dummy load connected during all tests!)

With the oscillator working, increase the drive to the final and adjust the oscillator coil slug and the drive control for around 100 mA of emitter current with 24V collector supply. This gives a little over 2W input to the final and is the normal level for CW operation. A 51Ω (1W) resistor makes an ideal dummy load and it should get warm to the touch after several minutes dissipating the output from the transmitter. A No. 47 pilot light can also be used as a load for indications of maximum output and should light to about full brilliance when the rig is properly loaded up and tuned.

When proper operation into a dummy load is verified, the transmitter can be connected to an antenna and the matching network adjusted in the usual way for proper loading. It is possible to QSY over 100 kHz of the CW portion of 40m without retuning either the crystal oscillator or final amplifier, with only a slight readjustment of the antenna coupler.

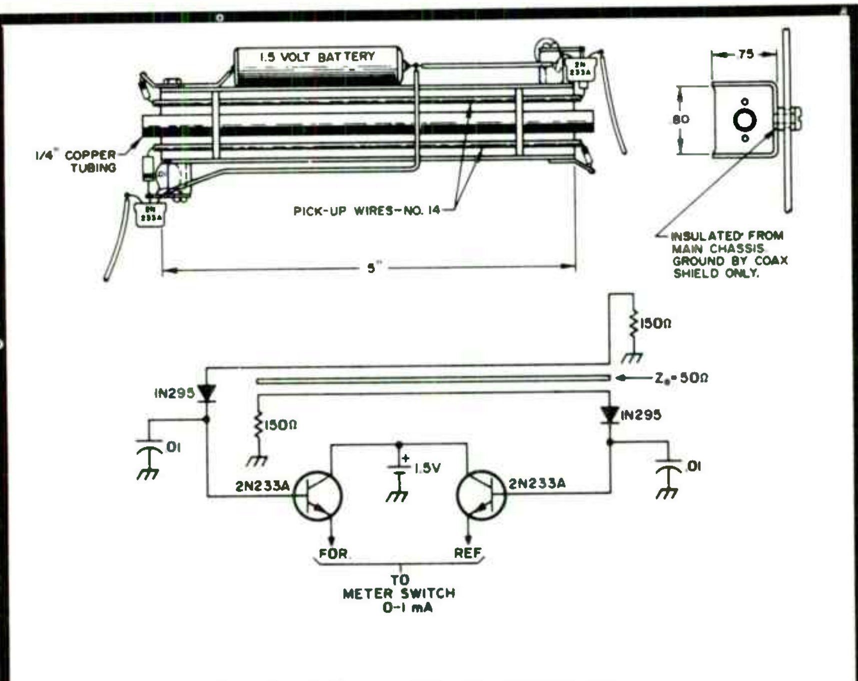


Fig. 2. Swr bridge assembly and placement data.

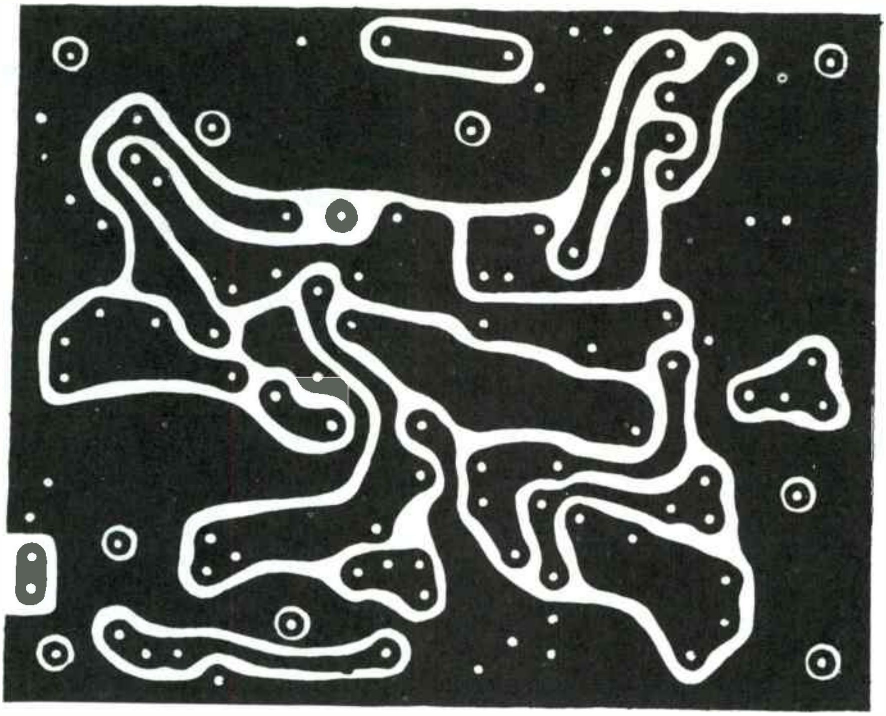


Fig. 3. Double-size PC board layout for the rf/af section of the receiver.

Results and Afterthoughts

Results with this little transceiver have been very good. A watt of power into the antenna may not seem like a lot to the kilowatt operator, but don't underestimate the punch of the signal. I have had stations at 400 miles during daylight operation insist on giving me a 599 report. In general, stations come back to the first call (unless you're covered up by a higher power station calling) and your contact probably won't know you are QRP until you tell him. Crystal control does limit the operating convenience somewhat with low power.

Since the receiver has an oscillator right on the operating frequency some will probably wonder why it is not used as a transmitting vfo also. The answer is that it can and if I were building this unit again I would probably include vfo operation. In fact, some checks were made using the receiver oscillator as a rfo with moderate success. However, the output of the local oscillator is quite low and it was necessary to provide a comple of stages of rather high gain to properly drive the transmitter. These tended to be somewhat unstable in operation and the vfo idea was abandoned, without too much work done in that direction. A conversion-type receiver would probably be better in this respect with the vfo operating at a frequency different than that of the transmitter.

With a suitable antenna coupler or matching network the rig will load into just about any piece of wire for portable work. Operation from the home station has been with a Windom antenna cut for 40m and fed with single wire. This still requires an L-network between the transmitter and antenna for proper matching. A coax fed dipole should work connected directly to the transceiver antenna connector.

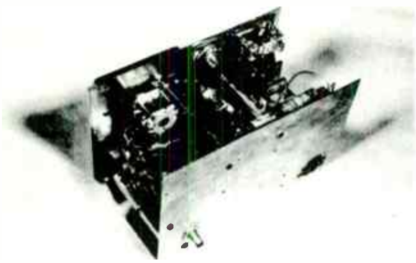


Photo showing shielding and construction of 40m transmitter and receiver in main chassis housing.

Receiver performance has been exceptional and sure beats using a regenerative set with its limited audio output and instability. Drift from a cold start with this receiver is practically nil and it can be used to copy sideband signals with no trouble at all. For CW work, a more selective audio bandpass filter can be used either in place of the low-pass filter inside the transceiver or in the headphone line. A couple of 88 mH toroids and some 0.5 μF capacitors will increase the selectivity markedly when connected as an audio filter. For general operation the filter is not necessary and it does prevent good copy on sideband signals.

With the addition of a balanced modulator. this setup could provide the basis for a simple little sideband rig using the audio section for the speech amplifier, generating a double sideband signal at 7 MHz.

Using currently available transistors, this little transceiver can be used to drive a class C stage to 15 or 20W for a little more signal, but at these power levels a battery of rather monstrous proportions is required; the 2W power level is just about

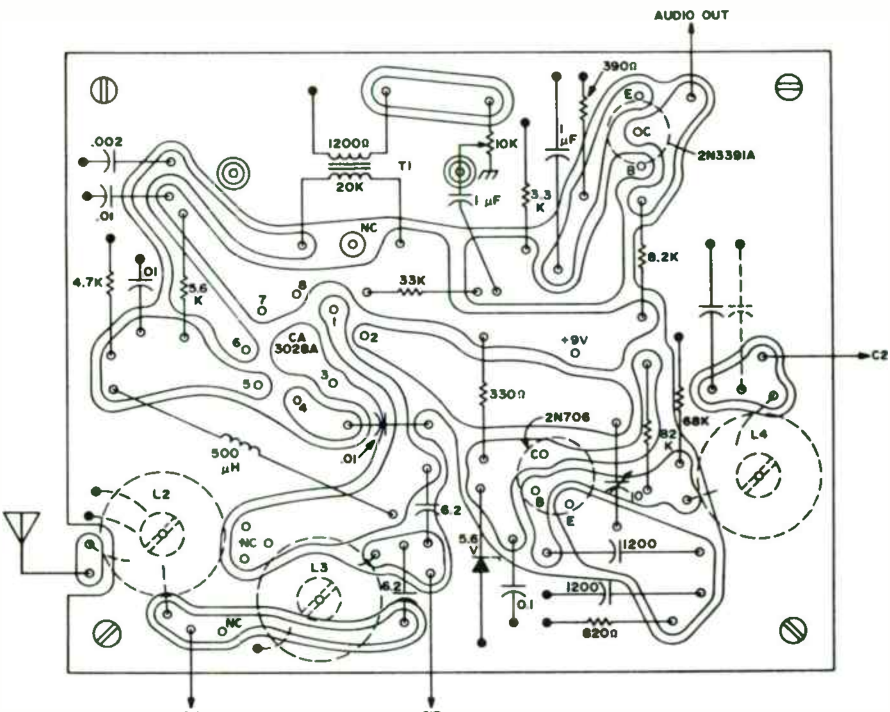


Fig. 4. This composite sketch shows positioning of components on the etched board.

right for the mercury batteries used. And 20W isn't really QRP anyway.

Table of Parts

C1 – 140 pF peri section, dual variable capacitor
C2 – 9–35 pF ceramic trimmer
C3 – Main tuning 50 pF variable capacitor
C4 – 365 pF broadcast variable
CR1 – 5.6V zener, 1N708 or equivalent
T1 – Interstage audio transformer 1.2–20 kΩ 6-T-12 PC (Allied Radio)
BA1, BA2 – 9V batteries
BA3 – 5V cell
BA4 – Two 12V batteries in series, Mallory TR289 or equivalent
IC1 – RCA CA3028A
L1 – 5 turns 28 AWG spaced over L2
L2, L3, L4 – 36 turns 32 AWG wound on .380 diameter toroid core (Arnold A4-380-125-SF)
L5 – 12 turns 28 AWG close-wound on 3/8 in. diameter slug-tuned form
L6 – 3 turns 22 AWG over B+ end of L5
L7 – B&W Miniductor ½ in. 16 TPI, 1 in. long
M1 – 0-1 mA meter (Emico Model 13)
Main tuning dial – Millen Type 10039 midget panel dial

. . .W9ZTK■

Practical IC Regulator Circuits for Hams

Don Nelson WB2EGZ

The experimenter has broad horizons with the many low-cost ICs and semiconductors available today. While many of the ICs have been designed for a specific service, such as an audio amplifier or a logic switch, there are types which contain several transistors to be used in almost any application. The RCA CA3018 exemplifies the second category. Let's examine the device as an introduction to integrated circuitry.

Four silicon transistors are formed on a common monolithic substrate within the CA3018. Two of these transistors are interconnected by having the emitter of one tied to the base of the other. Either of these transistors may be used separately since an external lead connects to the emitter-base link. The intent of the interconnection is to use the transistors in a Darlington circuit. The other two transistors are isolated as shown in Fig. 1. Notice the substrate is important enough to have its own lead. In this IC, as in most others, there is diode action between the substrate and some elements of the circuit. You can see this characteristic using an ohmmeter between the collectors and the substrate. These diodes are reverse-biased by connecting the substrate to the most negative point in the circuit, thus isolating the transistors. If the substrate is not connected in such a fashion, you may not get transistor action from the circuit.

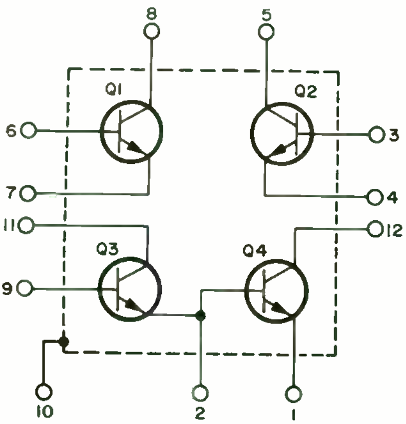


Fig. 1. Schematic diagram of RCA CA3018. Numbers refer to pins of IC.

The transistors in the CA3018 are useful from dc to 120 MHz. One of the big advantages of integrated circuitry is the matched characteristics of the transistors within. Gain, for example, is matched better than 10% and the base–emitter voltage match is better than 2 mV over a wide temperature range. Because of these characteristics, this and other integrated circuits are excellent for temperature-compensated circuitry.

Another plus for most ICs is the excellent low-frequency noise figure. Transistors in the CA3018 array boast 3.2 dB of noise at 1 kHz. At 100 MHz, the noise figure is typically 7 dB, so this device is favored for operation below VHF.

Learning to use the best characteristics of a device and learning to design around its limitations are good engineering practices. I once heard a ham brag, "The 4X150 is a great little tube – mine is dripping solder from the radiator, but my power output hasn't dropped." Well, those good old days are gone when you use semiconductors – so learn what the ratings are all about. Maximum and minimum values for the CA3018, which must be observed, are listed in Table I.

Those readers who are familiar with transistors will find only two new ratings; a total-package power rating, and a collector-to-substrate voltage rating. The first is a

Table I. IC Parametric Limitations at 77°F.

Parameter	CA3018	CA3018A
Maximum power dissipation:		
any single transistor	300 mW	300 mW
total package	450 mW	450 mW
Derate 5 mW/°C for TA >85°C		
Maximum collector-to-emitter voltage	15V	15V
Maximum collector-to-base voltage	20V	30V
Maximum collector-to-substrate voltage	20V	40V
Maximum emitter-to-base voltage	5V	5V
Maximum collector current	50 mA	50 mA

limit for the sum of the power dissipation of the individual transistors. For example, if one transistor operates at the 300 mW level, the average power of the remaining three may not exceed 50 mW each (300 + 3 x 50 = 450). The second new rating refers to the breakdown voltage of the collector-to-substrate diodes. These diodes are always reverse-biased because the substrate has been connected to the most negative part of the circuit. It is easy to determine the exact voltage in a circuit because the collectors are frequently at the most positive voltage.

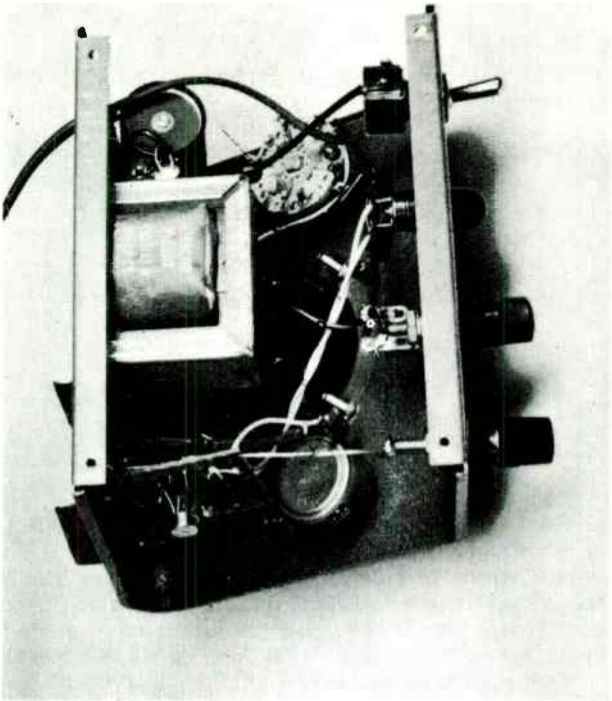
Many ICs have low breakdown voltage ratings, which can be a serious disadvantage. In some cases, you may be able to design around the problem. Another problem at the experimenter's level is the fact that if one transistor is destroyed, the whole IC has to be replaced. This disadvantage is somewhat offset by the low $1.62 cost of the CA3018, but it is hard to rationalize soldering 24 leads for each mishap. A discrete transistor could be used in a crisis with the CA3018. Better yet, use a socket for breadboarding.

Applications

The CA3018 is a natural choice for a power supply regulator amplifier. Such an application, I feel, is also interesting for analysis. In Fig. 2, I have drawn a very basic regulated power supply using three of the transistors. Let's examine its characteristics from the ratings of the IC.

Nearly all the current to a load will flow through Q4 (pins 12 and 1); therefore, we are limited to 50 mA, which is the

Exterior view of the 500 mA power supply. The case used is a Bud CMA 1930.



Placement of parts is only critical with respect to squeezing everything into the box. The 2N5295 transistor is outside the rear apron and insulated from the main chassis. A piece of copper 2-1/2 x 2 x 1/32 in. forms a heatsink for the power transistor.

maximum collector current of any single transistor. If you think about it, this is all the current required by many circuits. The power supply could service almost any circuit now operated on small 9V batteries. An FM portable receiver draws peaks of

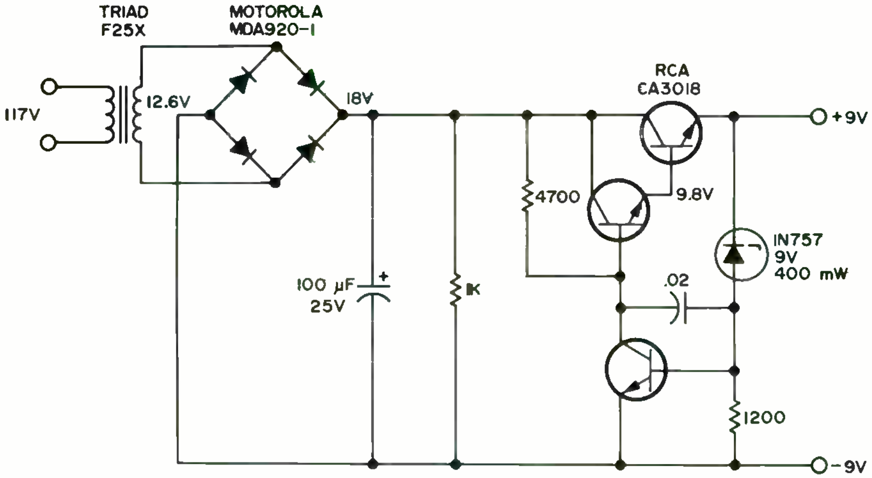


Fig. 2. Regulated 9V 30 mA power supply using CA3018.

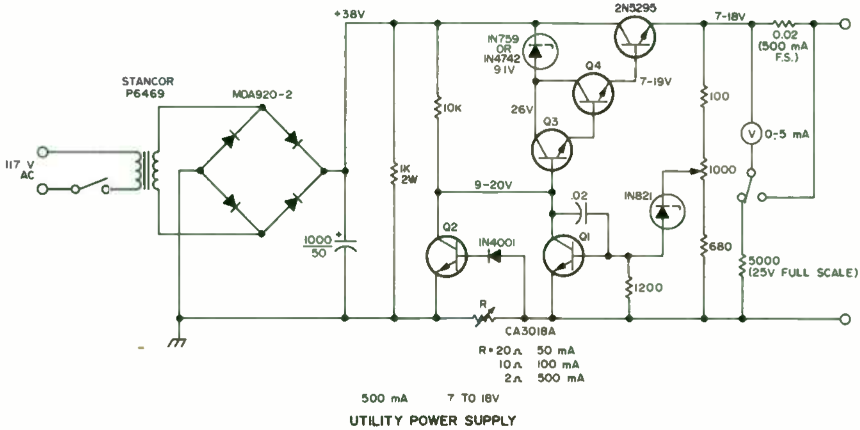


Fig. 3. "Utility" power supply with 7–18V output. Resistor R determined by current requirements: 20Ω–50 mA; 10Ω–100 mA; 2Ω–500 mA.

less than 20 mA; AM radios draw even less.

The second rating to consider is the collector-to-emitter voltage. The greatest permissible value is 10.5V, which is well under the 15V rating of the transistor. The collector–base breakdown voltage is not exceeded and the emitter–base junction is never reversed-biased, so the design is clean in those respects. Don't forget the collector-to-substrate rating. In this case, the maximum voltage seen is 18V: okay for this circuit, but not for the next one.

When considering the power rating, it is safe to assume we will not exceed 85°C (no derating necessary), and the power dissipation of Q4 will be much greater than the total power of Q1 and Q3. At 50 mA of current, the power (P=IE) would be .050 (18–9)=0.450W, which is high for a single transistor. A reduction in output current or lowering the input voltage on pin 12 is necessary to operate within ratings. In any case, it would be wise to heatsink the IC for this circuit. You may expect it to get hot above 30 mA.

The shortcomings of the CA3018 become apparent as we attempt to design a more versatile supply such as the one in Fig. 3. A 2N5295 power transistor was added to the basic circuitrv in order to operate to supply to 18V and 500 mA. Other changes include the use of a temperature-compensated zener diode, variable voltage output, and current limiting by use of the fourth transistor on the IC. Incidentally, you might just want to build a power supply like this if you are newly acquainted with semiconductors.

The 2N5295 will handle 500 mA and higher voltage with ease. If we assume the beta of that transistor is 50 at full load, the base drive required would be 500/50, or 10 mA. That is no problem for Q4 of the CA3018. On the other hand, it soon becomes apparent that we need higher than 20V breakkdown. Some relief is obtained by using the CA3018A, but not quite enough to handle the collector–emitter voltage drop across the Darlington pair. By

The completed 500 mA regulated low-voltage power supply.

using a zener diode between the high-voltage point and pins 11 and 12 of the IC, the collector-to-emitter voltage requirements are reduced, as is the collector-to-substrate voltage.

It is not necessary to use a temperature-compensated zener here. Collector-to-emitter voltages are the only ones above ratings now. Three CA3018s tested had collector-to-emitter breakdown voltage ratings in excess of 23V; however, this is a weak point in the design.

Power requirements are lower with the addition of the 2N5295. The maximum power through Q4 is 23V x 0.010A= 0.230W, enough less than the maximum rating that no heatsink is necessary.

A word about the operation of the current limiter is in order for any prospective builder. Transistor Q2 is normally turned off until an overload occurs. At this time, the voltage drop across R is enough to turn on the 1N4001 and the emitter–base junction of Q2. As the transistor is turned on, its collector is pulled toward ground, thus turning off Q3, Q4, and the 2N5295. As soon as the overload is removed, the supply recovers to its former voltage. This limiter may be made variable for limiting at lower currents. Approximate limiting values are shown in the schematic. Under extreme current limiting conditions, the collector-to-base voltage of Q3 and Q4 will approach 26V. This is acceptable for the CA3018A.

While the CA3018 is not truly a circuit, it provides an interesting introduction to the technology and characteristics of integrated circuitry. Its versatility invites experimentation. For those who would like to duplicate the regulated power supply shown in the photos, I include a full-size

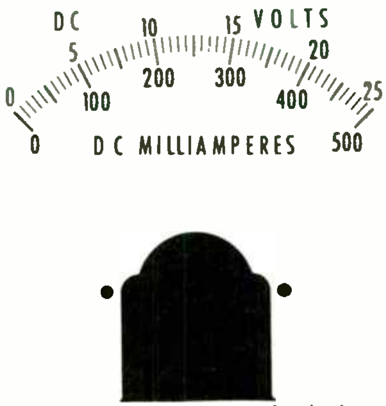


Fig. 4. This meter faceplate can be glued to any appropriate sized meter with a 5 mA full-scale rating.

copy of my meter panel as Fig. 4. This can be copies or cut from the magazine and glued to the face of an appropriate 0–5 mA meter.

Reference:

Several other interesting circuits for the CA3018 are discussed in the RCA integrated circuits manual, where complete ratings and typical characteristics are listed.

. . .WB2EGZ ■

Low Cost Function Generator for the Experimenter

Richard Factor WA2IKL

Most experimenters have audio frequency oscillators, yet their use is severely limited by the lower frequency limit of about 20 Hz in most low-cost units. This limitation exists because of the extremely high impedance necessary to avoid loading the frequency determining network. Until the advent of the field-effect transistor (FET) it was almost impossible to build a low-distortion, low-frequency transistor signal generator. It is still difficult to lower the output frequency significantly without introducing severe distortion.

The unit described in this article has none of the low-frequency limitations of bridge-type oscillators. It generates signals in the range of 4 kHz down to about 10 mHz (that's 10 millihertz, or 0.01 cycle per second!). Such frequencies have many uses, such as generating slow-sweep displays, checking the response of low-frequency filters, and voltage-control applications in electronic music generation, instrumentation, etc. The circuitry is similar to that of a high-priced laboratory instrument known as a "function generator," but by taking advantage of low-cost ICs and nonprecision resistors, it can be built for substantially less than most kit-type audio generators.

Circuitry

Most audio oscillators directly generate a sine wave. Sine-wave oscillators usually are much more critical than other types. The oscillator of Fig. 1 generates a triangular wave which is then shaped into a sine wave using a FET instead of a reactive filter. This renders it insensitive to fre-

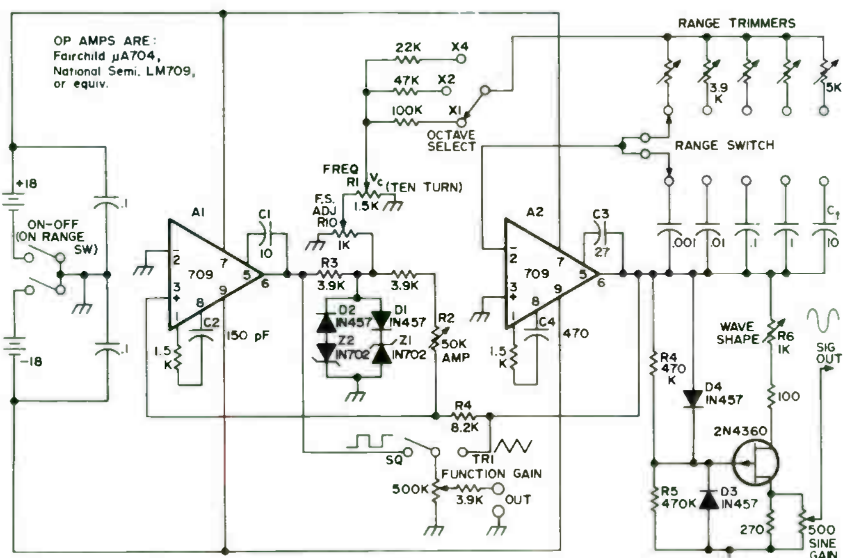


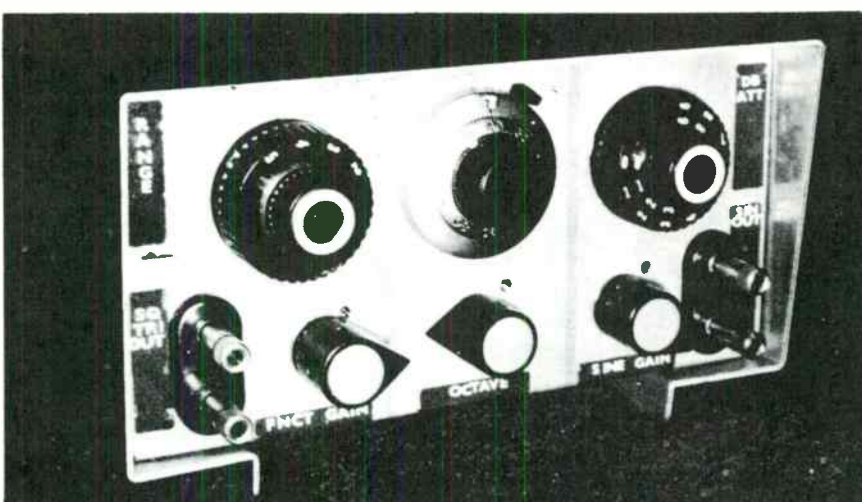
Fig. 1. Function generator schematic diagram.

quency, enabling the distortion to be constant over the entire range. Due to difficulties in obtaining precise symmetry in FET characteristics, plus nonideal characteristics of the transfer function itself, it is difficult to get the distortion lower than about 1.2%. This is adequate for almost all applications.

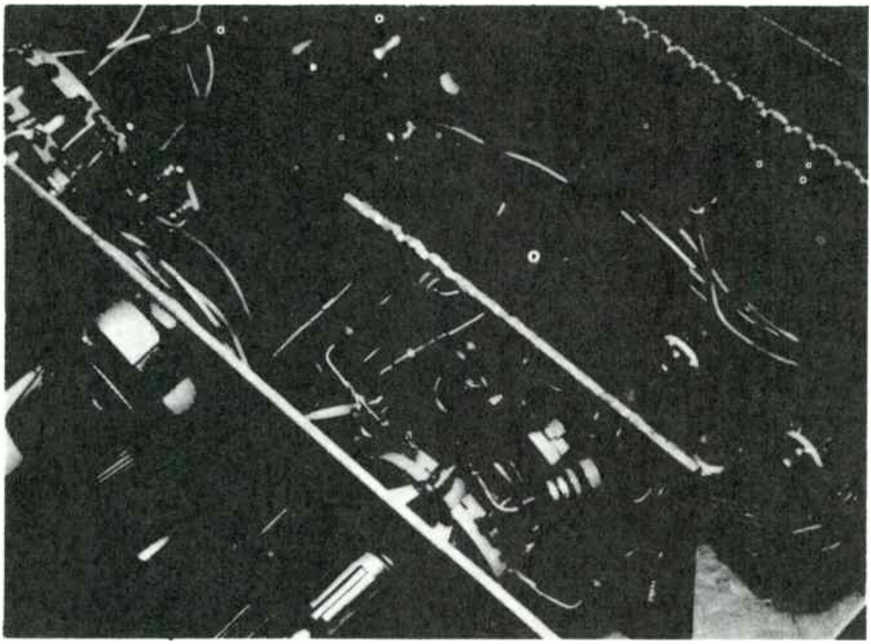
Integrated-circuit operational amplifiers (opamps) are used to generate the triangular wave. Unit A2 is connected in a circuit arrangement known as an integrator. Time-determining capacitor C_t is charged at a rate directly proportional to the current fed into pin 2 (the inverting input) of the opamp. Pin 2 is kept at ground potential by feedback through the capacitor, so the current equals the voltage at point V_c divided by the total series resistance, consisting of the resistor on the octave switch and the range trimmers. Thus, it is possible to control the charging rate with R1, and the high current capacity vs low leakage current of the opamp allows the charging rate, and hence the frequency, to be varied over a 500-to-1 range, compared to the normal 10-to-1 range for normal capacitance-controlled oscillators.

If a current were just fed into the opamp, the capacitor would continue charging until the limit imposed by the opamp and the power supply voltage was reached. However, A1 is connected to the frequency control pot, and supplies the voltage that is converted into the charging rate by the various resistors. A1 is connected in a positive feedback loop with high hysteresis. The opamp is sensitive to the voltage between the two input terminals. If the difference is only a millivolt or so, the amplifier goes to either its fully positive or fully negative limit, depending on the input polarity.

Feedback to the input is through R2 and R4. Assume the output of A1 is fully negative. This voltage is clamped to about 4V by D2 and Z2 (D1 and Z1 when positive). This 4V is connected back to the input through R2 (and the 3.9 kΩ protective resistor). As the voltage at the output of A2 gets increasingly more positive, a point is reached where the current through R4 becomes opposite to and slightly greater than that through R2. When this happens, the amplifier instantly (in a few microseconds) changes state, aided by the positive feedback through R2. Of course, when the opamp changes state, the integrator starts to charge in the opposite direction, thus creating the tri-



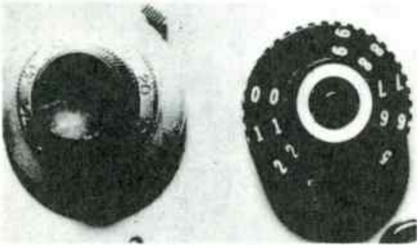
Front panel of function generator.



Interior view of unit with step attenuator resistors in front.

angular wave. The zener diodes insure that the wave is highly symmetrical. If your zeners are out of tolerance, or distortion is critical, connecting a small resistor in one leg or the other might help a little.

The sine wave, square wave, and triangle wave are all synchronous and available simultaneously. To conserve panel space on my unit, I used a switch to connect either the square or the triangle to the binding post, though this is not necessary. Whatever arrangement you use, avoid loading the output on the opamp by more than



Front panel view of frequency control dial and step attenuator dial.

about 2 kΩ. It won't harm anything, but the output frequency may change slightly when connecting and disconnecting loads if this precaution isn't observed. It is normally very difficult to measure the frequency of VLF sine waves. Connecting the square wave output to a counter while using the sine output is a great convenience.

The sine wave shaper takes advantage of the symmetry of the source and drain with respect to the gate of the 2N4360 J-FET. (This unit was chosen primarily for its very low cost, as well as its relatively high pinchoff voltage which gives a greater output voltage.) The output of A2 varies symmetrically about ground, and as the absolute value of the voltage increases, the FET resistance also increases due to the gate bias. Incidentally, note that the power to drive the FET circuit is supplied directly by the opamp output. Proper bias polarity is insured by D3 and D4.

Since the output is symmetrical about ground, no coupling capacitors are necessary. It would be hopelessly unwieldy to use one for 0.01 Hz anyway. The output amplifier is designed for unity gain with no voltage offset. Simplicity is achieved through the use of complementary transistors. The output impedance is under 100Ω. Resistors R6, R7, R8, R9, and diodes D5 and D6 are the bias network, which eliminates crossover distortion in the amplifier.

If you plan to use the equipment for checking professional audio equipment, you might desire to add the attenuator shown in Fig. 2. It provides 89 dB of attenuation into a 600Ω load. Two concentric single-pole, 10-position switches break the 89 dB into 1 dB steps.

To simplify the attenuator design and allow the use of standard resistors with 5% tolerance, an unususal circuit was used. The 10 dB steps are provided by a constant-impedance (200Ω) "T" attenuator, and the single dB steps are provided by switching in series resistors. Thus, the attenuator is not of a constant-impedance design, although the voltage into a 600Ω load conforms to the proper setting.

If you are using a high-impedance load, the output should be bridged with a 600Ω resistor. A low-impedance load will exaggerate the effect of the 1 dB step switch.

I emphasize that this whole assembly, while desirable, is purely optional.

Adjustment

As you can see from the photograph, the frequency control is a 100 turn precision potentiometer. These are readily available for a couple of dollars in the surplus market. Using one, it is possible to calibrate the unit so that for the top 900 divisions, the output frequency is correct to within 1%. As the octave select switch is switched to a lower series resistance, the linearity of the frequency control is decreased. The full-scale adjustment pot sets the output frequency to 100 Hz on the range with the .01 μF capacitor when the octave switch is in the X1 position (100 kΩ series resistance). Then the various range adjust pots are set until each range has the appropriate full-scale value: 1 kHz, 100 Hz, 10 Hz, 1 Hz, 0.1 Hz.

Potentiometer R2 is adjusted so that the triangle amplitude is proper for conversion to a sine. This control interacts with the full-scale adjustment pot, so they both may have to be readjusted several times. Waveshape pot R6 is adjusted for symmetry in the top and bottom halves of the sine wave. A scope or harmonic distortion meter is advisable for the above two adjustments, although a reasonable null can be achieved by a musically trained ear, especially if there is an undistorted signal available for comparison.

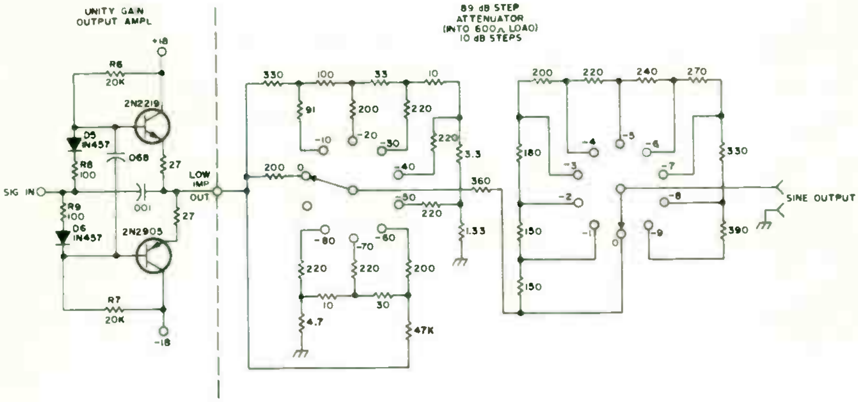
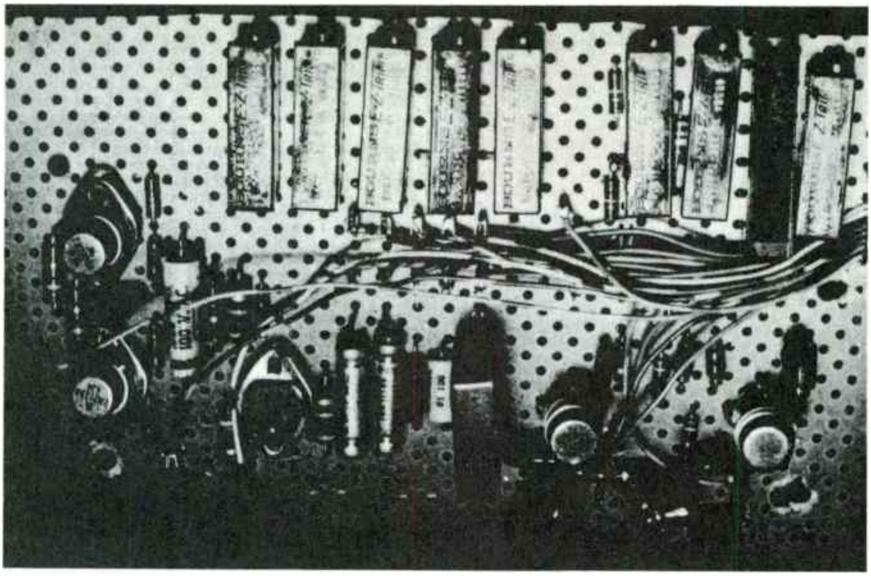


Fig. 2. Output amplifier and attenuator schematic.



Closeup view of circuit board.

Construction

The circuitry is noncritical and any convenient layout may be used. Power is provided by four 9V batteries. Do not exceed this voltage as the operational amplifiers are rated for a maximum power supply potential of ±18V. The batteries should be bypassed with 0.1 μF capacitors. If any trouble is encountered with the opamps oscillating at high frequency, it will probably be eliminated by bypassing the power supply pins to ground as close to the IC as possible. If this doesn't work, add capacitance to C1, C2, C3, and C4 a few picofarads at a time until oscillation stops. The values given in the diagram are calculated to work with almost all production opamps, but occasionally one will be near the edge of the specifications and require the extra capacitance.

If you don't need precise frequency calibration or attenuation, substantial cost can be saved by eliminating these features. The range trimmers can also be eliminated (just replace with a short circuit). The octave switch can be replaced with a fixed resistor down to 15 kΩ or so. The output attenuator can be replaced with an ordinary pot; and substitute the 500Ω pot and 470Ω opamp resistor before the amplifier with a 240Ω resistor; and connect the opamp input directly to the FET. The pot must be connected to the amplifier output to attenuate the amplifier noise as well as the signal. Use any convenient pot for the frequency control, and calibrate the dial yourself.

Following these procedures can reduce the cost to under $10 and yet provide you with an oscillator that, in the lower frequency ranges, is better than anything selling for under the price of a laboratory function generator.

The IC-mitter

Stephen Goldstein

This transmitter was designed with the QRP fiend in mind. It uses inexpensive ($3.80) RCA type CA3000 integrated circuits. As crystal oscillators they will work to about 10 MHz, and if you're lucky you just might be able to use them at 20 meters. This transmitter will work on 160, 80, 40, and maybe 20. This contraption will put out AM or CW at the flick of a switch.

The schematic should be self-explanatory. If you build on Vector board I would recommend using sockets. As a matter of interest, the 2N3904 is for agc. Capacitor C1 should be chosen for good agc action – try 0.05 or 0.1 μF. Transformer T1 is a swamped 10 to 5 kΩ matching transformer. The values shown for it give a fairly good match. The C2-L1 network should resonate at the crystal frequency. Tap L1 wherever you get the best match. The AM-CW function is switched by a 3-pole double-throw wafer or toggle switch. For AM work, use a 50Ω mike (or something close)

These circuits are by no means completely original. Only partly. They were lifted from RCA publication ICAN5030. Also from File 121. Both are available from RCA for the asking. These papers provide much useful data on the CA3000.

Have fun!

Stephan Goldstein ■

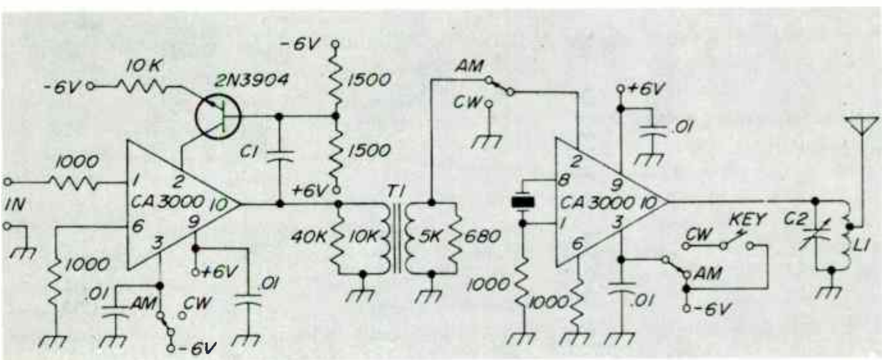


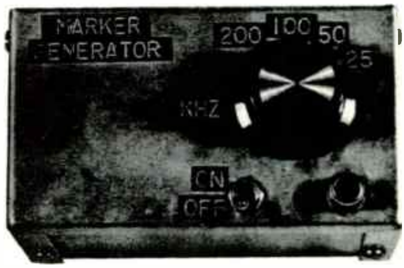
Fig. 1. See text for C1, C2, L1, T1. For CW you can also switch out the transistor if you want. Instead of keying the voltage, you could key the output. If you don't want agc, simply ground pin 2 of the first IC.

An IC Marker Generator

D. A. Poole K4BBC

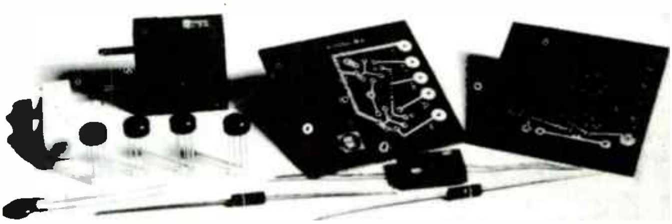
The incentive licensing regulations have created a need for identifying restricted band segments. The easiest method of doing this is by using a crystal-controlled oscillator to provide marker signals at the edges of the subbands (getting the Extra license is another good approach).

The marker generator described here provides calibration signals at 200, 100, 50, and 25 kHz intervals, usable up through the 6 meter band. Although the basic components shown were purchased



in a kit (R&R Electronics, December, 1969. 73 Magazine) containing the two circuit boards, the generator could be

R&R (311 E. South St., Indianapolis IN 46225) supplies all components and PC boards in kit form.



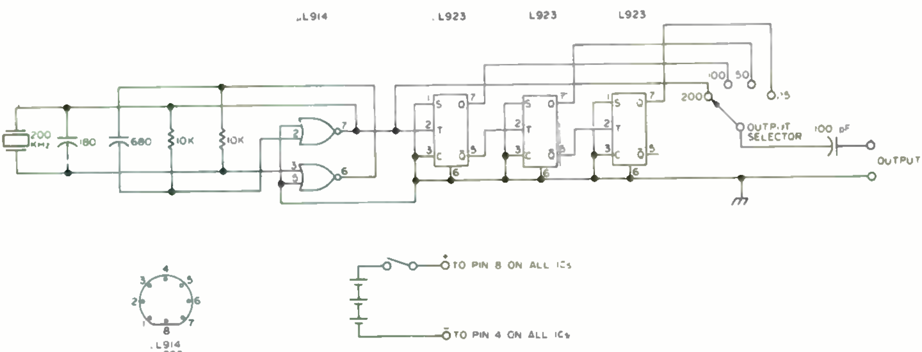


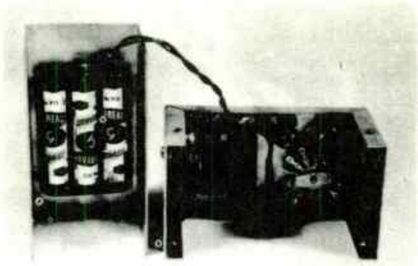
Fig. 1. IC interconnection and base diagram for the band-edge marker generator.

built on perforated board from individually purchased components.

The circuitry consists of a 200 kHz oscillator and three divide-by-two stages. In the oscillator, a Fairchild μL 914 integrated circuit is connected as a crystal-controlled multivibrator. A 200 kHz crystal is used because it is cheaper and more stable than the 100 kHz crystals usually used in marker generators. With the values shown the frequency is within a few cycles of zero-beat with WWV.

The output from the 200 kHz oscillator is successively fed to three μL 923 integrated circuits, each connected to divide the frequency by two. A rotary switch selects the desired output. Three penlight cells supply the 4.5V required by the integrated circuits.

The generator in the photo was built in a 4 x 2¼ x 2¼ in. aluminum box. The printed circuit boards with integrated circuits are attached to the box with screws and small spacers. The small rotary



The finished marker generator. Note that penlight cells mount handily in cover section of minibox.

switch, battery holder, and output jack are from Lafayette.

The generator has proved to be reliable and stable, providing subband identification at a flick of the switch. Because the harmonic output of the divider is quite strong, tight coupling to the receiver or transceiver is not necessary. Good results can usually be obtained by slipping the end of the output wire down into the shield of the first rf amplifier tube.

. . . K4BBC ■

A Simple Integrated Circuit Q-Generator

John J. Schultz W2EEY

Almost any integrated circuit operational amplifier can be used to build this Q-multiplier. Its advantages are extreme circuit simplicity and a useful frequency range that extends from audio frequencies to almost all i-f frequencies. Both the peaking frequency and Q can be made variable.

Both vacuum tube and transistor Q-multiplier circuits find wide application in improving the selectivity of transceivers and receivers, particularly on CW when the Q-multiplier is used to peak the i-f response following a steep-skirted crystal or mechanical SSB filter. The Q-multiplier provides a very narrow bandpass but if used alone does not provide steep skirt selectivity. When used in conjunction with an SSB filter, however, the latter provides the necessary skirt selectivity (see Fig. 1).

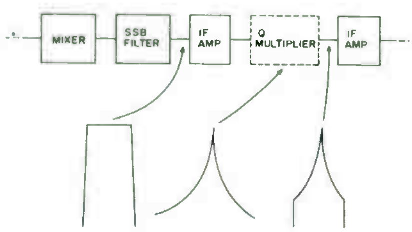


Fig. 1. Placement of Q-multiplier after SSB filter provides effective narrowband response for CW reception.

The Q-multiplier described in this article is meant to be inserted between any i-f stage in a transceiver or receiver following an SSB filter. It can be easily switched for broadband operation so it does not affect normal SSB operation. Since it is adjusted for unity gain, it does not upset any gain relationship in the i-f stages. As compared to vacuum tube and transistor-type Q-multipliers, the circuitry of the unit is extremely simple due to the use of an integrated circuit operational amplifier. The unit can be successfully used on frequencies far higher than those normally used with vacuum tube or transistor Q-multipliers up to 5 MHz or more, depending upon the integrated circuit used. The circuit can also be used at audio frequencies, if desired. One can also build an audio selectivity unit for outboard use when it is not desired to make internal modifications to a transceiver or receiver.

Circuit Description

The Q-multiplier is constructed around an integrated circuit operational amplifier. Many such amplifiers are available on the market at prices starting at a few dollars. The main requirements for choosing a suitable unit are that it have a differential input (inverting and noninverting inputs), and a single-ended output and a bandwidth sufficient for the frequency of operation. For example, Fairchild 709T amplifiers are avail-

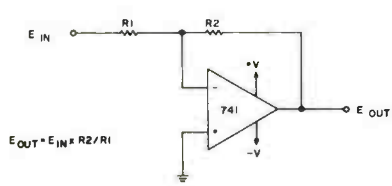


Fig. 2. Basic type of integrated circuit operational amplifier with a differential input and single-ended output that is needed for the Q-multiplier circuit.

able for about $3 and are usable up to at least 500 kHz. I built a unit using a Fairchild 741 which is usable up to 1-2 MHz. Other amplifiers such as a Motorola MC1530 can be used up to 10 MHz.

Figure 2 shows the schematic of the type of operational amplifier which is used and the formula for the output voltage of an ideal amplifier. The Fairchild 741 amplifier which I used requires no external frequency compensation components. Other amplifiers may require a few external components for this purpose as specified on their data sheet. The frequency rolloff components should be chosen such that the amplifier gain starts to decrease just above the frequency where it is used as a Q-multiplier. There is no advantage to having the gain "rolloff" at any higher frequency and would just make the amplifier more susceptible to oscillation due to a stray feedback path via external components. As noted from the gain formula, the gain of the amplifier depends upon the ratio of R2 and R1. If R2 is made equal to R1, the gain is unity. The Q-multiplier effect is based upon replacing R2 with a parallel resonant circuit which will present a very high impedance at one frequency and, therefore, maximize the overall gain at that frequency. Positive feedback is also used to enhance the Q-multiplying effect of the circuit.

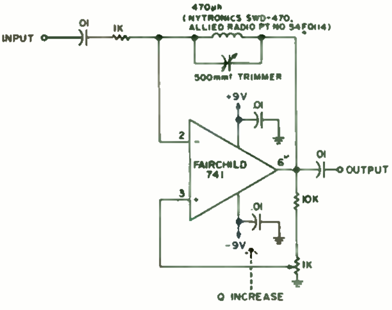


Fig. 3. Circuit of the Q-multiplier as constructed for a 455 kHz i-f

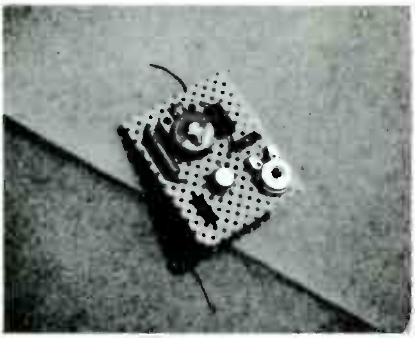
Figure 3 shows the schematic of a practical Q-multiplier circuit with the LC circuit resonant at 455 kHz for use in a 455 kHz i-f chain. Positive feedback is supplied via a 1 kΩ potentiometer. As with other Q-multiplier circuits, the Q can be increased by regulating the feedback until a point is reached when the unit will break into oscillation. The Q-multiplying effect is most effective when components are used for the resonant circuit which in themselves have a good Q. The inductor used for the 455 kHz Q-multiplier is a molded type which provides a Q of about 55 at 455 kHz by itself. The circuit can multiply this value by about 50 times or more. Another suitable inductor can be obtained by using only one or two sections from a regular 1 mH rf choke. The trimmer capacitor is used to set the frequency of the Q-multiplier in the middle of the i-f passband. It can, of course, be used as a variable tuning control for the peaking frequency of the unit. The input resistor, although shown with a nominal value of 1 kΩ, should be chosen so that the gain of the overall circuit is approximately unity.

Construction and Adjustment

There is nothing critical about the construction of a unit utilizing the circuit described as long as the various lead lengths are kept short. The photograph, for instance, shows how the various components can be directly wired together on a small piece of Vector board. The board itself can be directly mounted near the i-f chain in a transceiver, receiver, or together with the potentiometer used for feedback control if the latter is panel mounted.

If the Q-multiplier is not made tunable, the initial adjustment consists of peaking up the LC circuit. This can be done with the unit connected in an i-f strip and using any test signal centered in the i-f bandpass. The adjustment is best done with the feedback control set for minimum Q and with a barely audible CW test signal. The input resistor value should be then chosen for approximately unity gain. The adjustment is not difficult and need not be made exactly. The output level produced by a test signal without the Q-multiplier connected is noted. Then when the Q-multiplier is used, the input resistor is chosen so that the output level remains approximately the same. With most transistor i-f stages, the value of the resistor needed will be about 1-5Ω.

Bypassing of the Q-multiplier can be done in a number of ways in order to allow for normal SSB reception. If the Q-multiplier is made frequency tunable, it can simply be tuned outside the i-f bandpass. In order not to have the i-f gain decrease too far when doing this, the input resistor must be chosen for unity gain to take place when the Q-multiplier is just tuned outside the i-f bandpass. This will result in some increase in gain when the Q-multiplier is tuned to the center of the i-f bandpass but normally the result will not be objectionable. Another way to bypass the unit is to replace the LC circuit with a simple resistor equal in value to the input resistor (as in Fig. 2). The switching action can be accomplished by using a 1 kΩ potentiometer for the feedback control which also incorporates a SPDT switch. The switch must be wired such that the resistor replaces the tuned circuit when the wiper arm of the potentiometer is at ground potential.



The Q-multiplier components can be directly wired together using Vector board mounting. In this case, a PC board potentiometer is used to set the operating Q rather than making it continuously variable.

Summary

The simple Q-multiplier circuit described can be used for a variety of purposes besides that of improving i-f selectivity. It is useful for improving the Q and selectivity of a variety of tuned circuits as they might be used in FSK converters, audio filters for distortion tests, etc. Stagger-tuned circuits used in series can be formulated to provide a variety of bandpass shapes, often replacing more expensive components where bandpass shape factor is not important.

The permission of Dick Gerdee of Optical Electronics to present this circuit, which he originally developed, is gratefully acknowledged. . . . W2EEY ■

3 Versitile IC Testers

Richard Factor WA2IKL

As integrated circuits become widely used in industry, large numbers of them find their way into the surplus market, often at prices of a few cents each. The cheaper assortments contain many rejects and are frequently unmarked. With transistor assortments, this is not a problem, since transistors generally have three leads, and only a few trials are necessary to find out if they still "transist."

ICs compound the problem, since they not only have many more leads, but they don't all do the same thing. Figure 1 shows just a few of the many configurations available. With three simple IC testers, you can test and identify virtually all of the ICs a ham is likely to come across (including those of Fig. 1).

After the sweeping claims above, I should clarify some of the things the testers won't do: They will not rapidly and automatically test all the static and dynamic parameters of ICs. They will not automatically check such complex function units as 256-bit read-only memories and other MOS circuits. They will not test the vast assortment of rf and video amplifiers available. They will give no measurements good to 0.1%. But all three can be built for under \$20; so if you don't expect the impossible, you can make up a very handy item to have around.

What the testers will do, and do quite nicely, is test RTL, DTL, and TTL digital logic circuits, decade counters, Nixie drivers, and assorted operational amplifiers and comparators, and tell you whether or not they work. And believe me, that's all you really care about.

Testing Philosophy

To test an IC, the simplest thing to do is to regard it as a black box. You apply power to it, connect an input to the proper pin, and observe whether the output is as expected. The procedure for doing this to an unmarked IC is as much an art as a science. I will describe my procedure for going through a batch of ICs. Bear in mind that this is only a representative method – you might hit on one that works better for you.

Assume you have latched onto a handful of unmarked TTLs known to belong to the SN7400N series. This family has gates, flip-flops, and assorted complex functions. Since gates are the most common function, it is best to start by looking for them. Connect the patch cords so that pin 7 is grounded, and pin 14 goes to the +5V terminal. One advantage of looking for the gates first is that these power supply connections cannot damage any flip-flops or counters.

It is a very good idea to have the manufacturer's literature handy so that once you identify an IC, you can label it with a number rather than writing a truth table for it, and so you can avoid pitfalls like improper power supply connections. Insert the ICs into the socket and watch the output current of the supply. When it increases (about 5–10 mA), you have reason to believe that the IC is a gate.

The next step is to look for outputs. The easiest way to do this is to connect the scope successively to each pin while holding the tip of the patch cord. Holding the cord puts 60 Hz on the scope input. Inputs have very high impedance, and thus the signal will not disappear when the scope is connected to one. Outputs have either saturated transistor outputs in either state (TTL), or low-resistance pullups (DTL or RTL), and will effectively short out the 60 Hz current picked up by the body. Once an

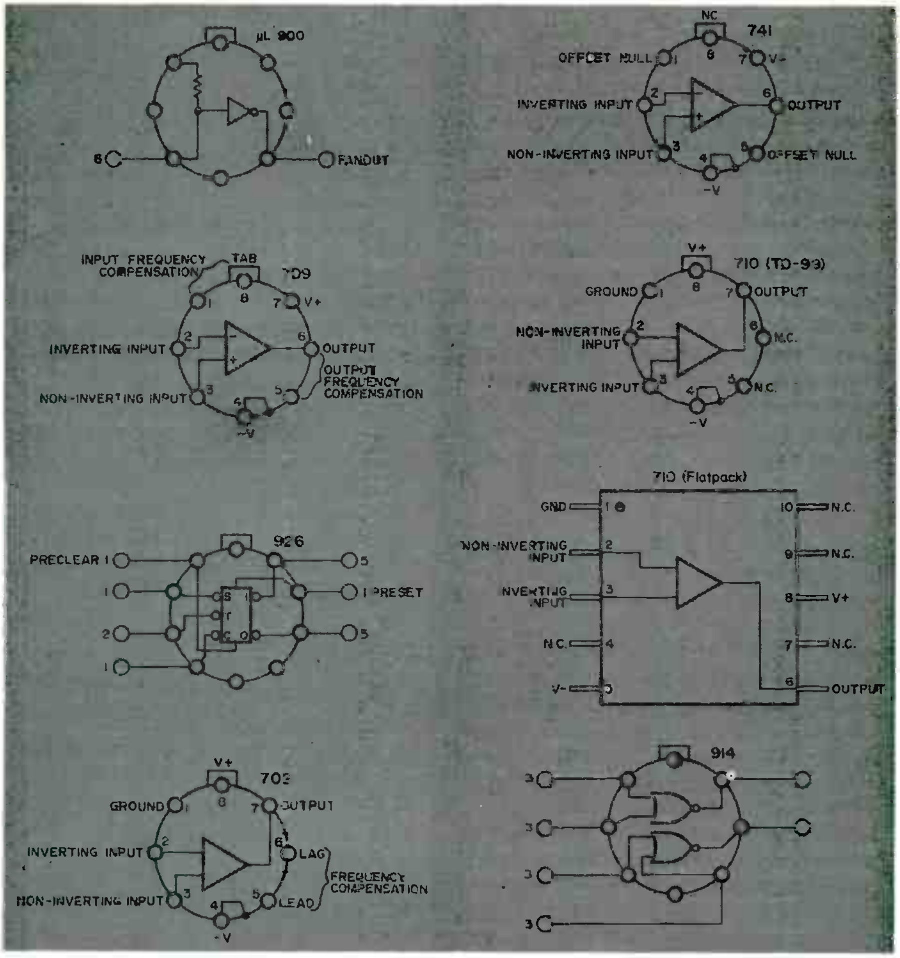


Fig. 1. Connection diagrams for some common ICs. All are shown from top view.

output is identified, connect a pulse input to the other pins and find out which ones give an inverted output. This is now sufficient to identify the IC.

Frequently an IC will be perfectly good except for one input. If it has only a minor defect (one of four gates bad, a defective reset on a flip-flop, or such), you can break off the pin corresponding to the defect. This is much more economical than discarding the unit.

Now that we have identified the gates, let's try some other possibilities. Connecting pin 11 to ground and pin 4 to +5V is appropriate for the SN7473N flip-flop. Again, identify the outputs as above and connect the scope to one of them. Connect the pulse generator to the other pins. If you truly have a flip-flop, you should see the pulse waveform divided down to a square wave at half its frequency.

Similar procedures can be followed for almost any digital IC. Some of the newer functions are sufficiently complex to make a data sheet necessary for testing, or you

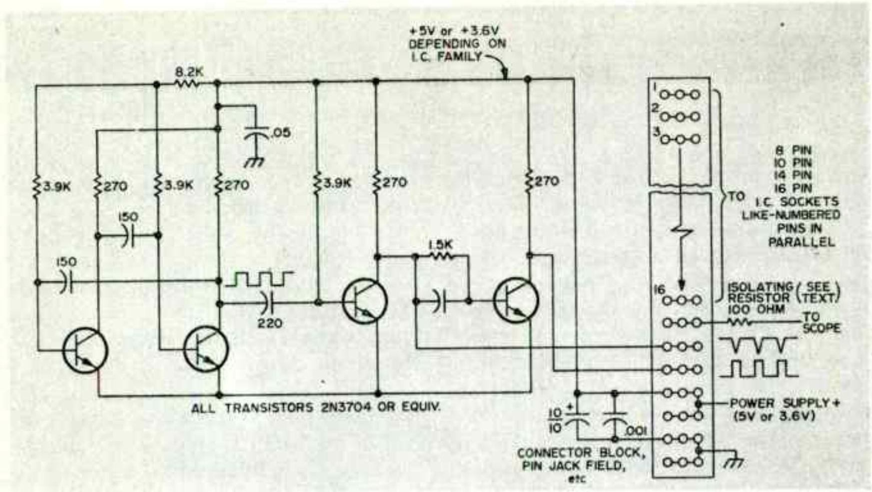


Fig. 2. General purpose digital IC tester.

may never hit on an appropriate combination of inputs and outputs. In such a case the digital tester of Fig. 2 can still make valid operational tests, although it will be of little use in identification.

Decade Counters, Nixie Drivers

Counting circuits and readout tube drivers are becoming more popular as their prices decrease. While the SN7490N decade counter can be tested by the above method, it is rather tedious, and the driver cannot be since it is designed to interface with a high-voltage device. To simplify testing of the counter and SN7441N Nixie driver, a special unit was built (Fig. 3) which simulates decade counter operation. Half of an SN7400N (or any other 2-, 3-, or 4-input dual, triple, or quad gate) is used

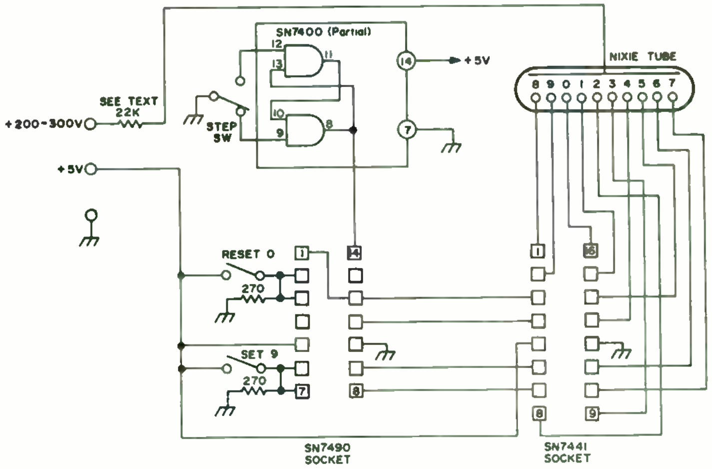


Fig. 3. Decade counter and nixie driver tester.

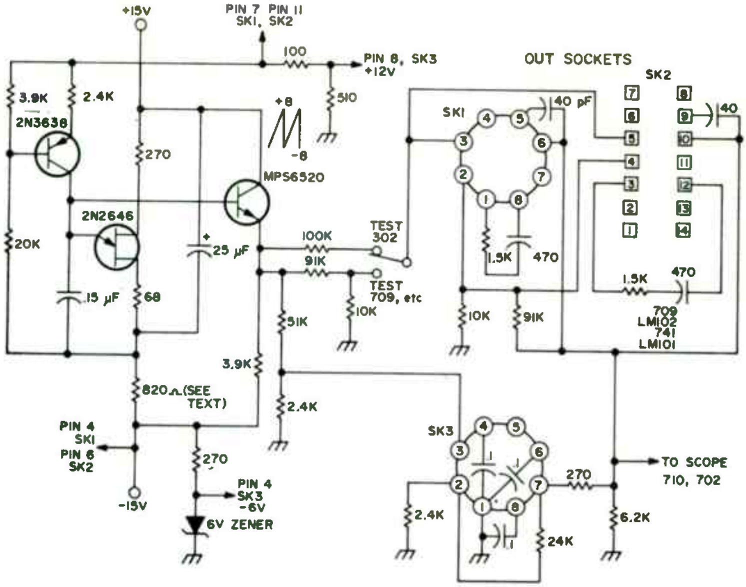


Fig. 4. Linear IC tester.

to produce a single pulse each time the SPDT switch is closed. This steps the decade counter, and the Nixie tube is observed for proper operation. Obviously, it is necessary to have a working Nixie driver in the socket when testing the counters, and vice versa.

The 22 kΩ resistor in series with the Nixie is for current limiting. This resistor value depends on the B+ supply and the particular tube used. A rule of thumb is to select a resistor which insures that all of each digit is lit up when the corresponding cathode is grounded. The value is not particularly critical.

Linear ICs

Compared to digital ICs, linears are a horse from a different stable. If you have no idea what type you have, you are quite likely to destroy it during the testing procedure. They are designed for a wide variety of positive and negative voltages, and pin connections are quite nonuniform. The most useful type of linear IC is the operational amplifier, or opamp, and the most popular of these is the type 709. They are so popular that the price has plummeted from over \$50 to under \$2 in reasonable quantities, and are available from numerous sources. The linear tester (Fig. 4) was designed primarily for 709 and the 710 comparator. It will also test other opamps, such as the LM101, the 702, the µA 741, and the LM102 voltage follower. The diagram showing the test-circuit equivalents (Fig. 5) gives the appropriate output waveforms to look for. The tester doesn't test for parameters such as dc offset and open-loop gain, but does provide a go/no-go test.

The unijunction and current-source transistors generate a linear sawtooth of about 16V amplitude. The approximately 820Ω resistor is adjusted so that the voltage swing is symmetrical about ground

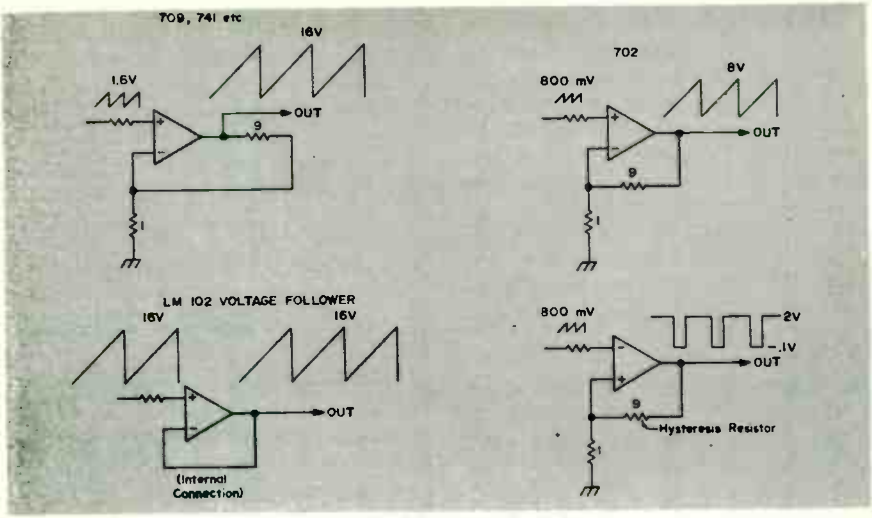


Fig. 5. Simplified test circuit equivalents.

when measured at the emitter of the MPS 6520 buffer. This signal is then appropriately attenuated and applied to the device input. The 709 output should be a symmetrical linear sawtooth of 16V amplitude. The 302 output should be similar, but without attenuation. Testing the 709 without attenuation will show its peak-to-peak output swing, without delivering enough current to damage the input. A tester switch is used to control the attenuation.

Building the Testers

The diagrams are more or less self-explanatory. The only special component used in any of the testers is the patching arrangement for the digital IC tester. I used some sort of connector block into which small conical pins fit. It had been lying around for so long that I forgot where I got it. If you can't find something like this, you might try an arrangement of terminal strips and alligator clips, or perhaps a field of pin jacks. Also, while it is not shown on the diagrams, it is a good idea to bypass the power supply connections to ground as near to the IC sockets as possible, especially in the linear IC tester.

Unless you're an old pro with ICs, testing them is not as boring a job as it may seem. It is impossible to describe a complete test procedure for all digital ICs, since they fail in as many ways as there are internal components. It is highly instructive to test the ICs with both the pin diagrams and the internal schematics in front of you. It will give you insight into IC operation and a better knowledge of logic functions than can be obtained from the literature. And the money you save from salvaging just a few ICs can equal the cost of the testers.

. . .WA2IKL ■

An IC Audio Processor

John J. Schultz W2EEY

To write about another type of speech compressor and still, yet, to call it an "ultimate" type may seem a bit overdone at a time when speech compressor circuits of every variety are commonplace. But the unit to be described is by no means just another garden-variety speech compressor. The phrase "audio processor" as used in the title, may, on the other hand, seem a bit nondefinitive, but it has been used to indicate that the unit provides a far more useful function than just speech compression alone. In fact, it is meant to convey the idea that it has processed a speech waveform on an af basis to such a degree that the waveform is the best possible to be fed to a transmitter for full modulation.

The unit makes use of two ICs – an SL630 amplifier and an SL620* automatic gain control unit. The two ICs were designed to mate together for a speech processing function and, therefore, should be used together even if one is tempted to replace the relatively simple amplifier IC by another type of IC.

The complete schematic of the unit is shown in Fig. 1. The unit is designed for both a low impedance input and output and a matching transformer is necessary for use with a high-impedance microphone. Usually, no transformer is needed on the output even if it is connected to a high-impedance audio input on a transmitter because of the amplitude of the output voltage. The input provides either for a balanced or an unbalanced type of microphone input. The former can be quite useful since by having both microphone leads ungrounded, many problems with noise pickup, rf pickup, etc. on the microp-

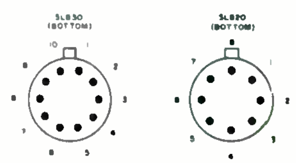
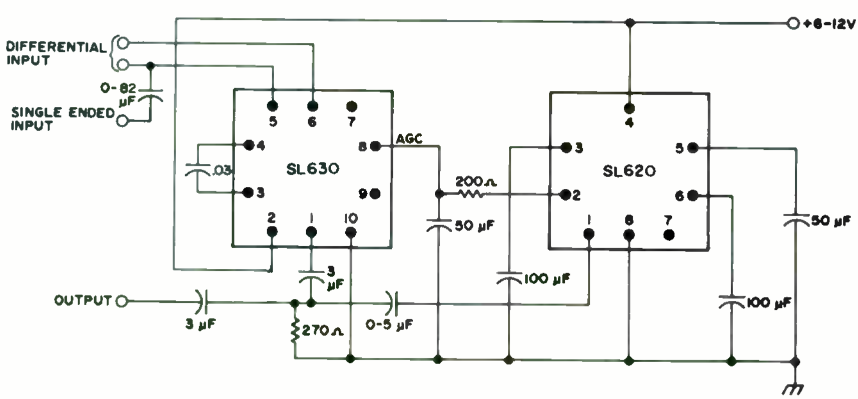


Fig. 1. Complete schematic of the IC audio processor.



hone leads are automatically avoided. If the usual type of unbalanced input is used, however, it is connected to pin 5 of the SL630 via a 1 μF capacitor. Pin 6 is then left unconnected. The capacitor between pins 3 and 4 of the SL630 provides a high-frequency rolloff characteristic. The values shown provide a rolloff starting at 3 kHz, but this can be changed, if desired, by experimenting with the capacitor value.

The output is taken from pin 1 of the SL630 via the two 3 μF coupling capacitors. Part of this output is coupled to pin 1 of the SL620 IC via the 0.5 μF capacitor. The SL620 IC uses this voltage to generate an agc voltage which is eventually available at pin 2 of the SL620. From there it is coupled to pin 8 of the SL630 to control the gain of the latter IC. What these two little ICs can accomplish is shown nicely by Fig. 2. The upper curve shows a varying speech input, while the second curve shows how the audio output appears after processing. Notice that when the speech input either rises rapdily or falls rapidly the output remains essentially constant. Noise bursts, because of their much shorter time duration, are recognized separately by the unit. The noise burst shown occurring during speech, although much higher in amplitude than the speech level, produces practically no increase in output. An automatic squelch feature is also provided. When there is a pause in speech, the output is disabled to prevent the background noise buildup common to most simple compressors. The pause time before the output is disabled is about 1 second and can be changed, if desired, by varying the value of the capacitor from pin 6 of the SL620 to ground.

The control range of the unit is illustrated by Fig. 3. Only a very slight change in agc voltage is necessary to control the output over a 60 dB range. In practice, the input can change over a 35 dB range and the output level will remain between 70 and 87 mV.

Construction

The photograph shows how the author constructed a unit on perforated board stock. The parts layout is in no way critical and is just a matter of convenience. The photograph of the underside of the board is just shown to illustrate how easily the unit could be adapted for etched PC board construction as a club project since only one wire crossover is necessary and even that can probably be eliminated by experimenting with the parts layout.

The units will operate equally well with a 6 or 12V dc supply and draw up to about 15 mA. The operating voltage can be borrowed from a well filtered point in a transmitter or a battery supply used. In the latter case, a usual 9V battery is ideal to use. There are no controls to the unit and so it may be housed easily inside a transmitter, avoiding only any location near high rf fields.

Operating Results

I compared operation of the unit to several types of conventional audio compressors. In every case, the unit described exhibited a much smoother compression action without pops, clicks, etc. It simply

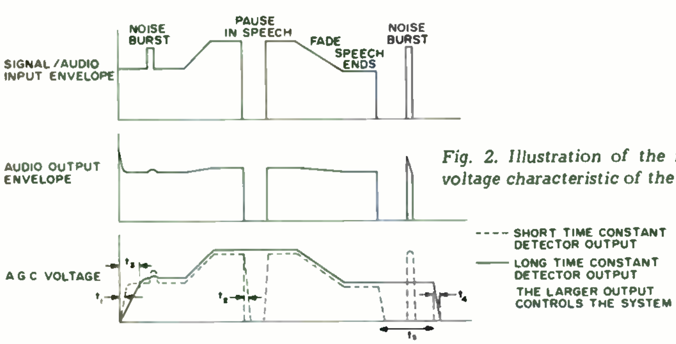


Fig. 2. Illustration of the input/output and agc voltage characteristic of the processor.

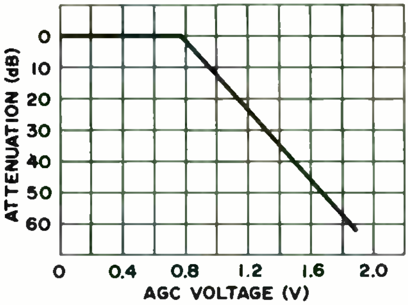


Fig. 3. Control range of the unit excees 60 dB. The graph illustrates the control of the input signal by the SL630 with agc voltage supplied by the SL620.

sounded more like the type of quality speech processing found on commercial circuits rather than the usual harsh, noisy compressor type action, which usually has to be disabled for local contacts because the inherent distortion then becomes so noticeable. The squelch and noise immunity features added also a great deal to the cleanliness of the speech output and should be particularly useful in a mobile type situation where a great deal of extraneous background noise can exist.

The IC Internals

The circuitry of the SL630 audio amplifier is not too different from that of many IC audio amplifiers, except that it includes provision for agc control of its output over a wide range. The input is coupled to a differential amplifier directly without the use of coupling capacitors for a balanced input. The agc voltage (pin 8) controls the emitter current flow return for the differential amplifier via the transistor whose base goes to pin 8 and a 3.6 kΩ resistor. The 750Ω resistor between base and collector acts as a "linearizing" element to give the smooth control range shown in Fig. 3. The rest of the unit continues to provide amplification (40 dB overall), ending in a series-connected output pair, which can provide an output up to 250 mW if the unit were used as solely a power amplifier device. Unlike many ICS, however, the necessary bias resistors and capacitors to suppress parasitic oscillations are all included in the unit, thus saving many external components. Pin 7, marked "muting," can be switched to ground to disable the audio output. If manual control of the gain of the unit were desired (instead of by the SL620 unit), a potentiometer can be connected between pins 9 and 2 with the wiper arm going to pin 8.

The circuitry of the SL620 unit, on the other hand, is quite different than most ICs because of its specialized functions. T1 through T4 are the input af amplifiers. The af output is coupleo to a dc output amplifier (T16-T19) by means of two detectors (T14 and T15). T14 in conjunction with C1 has a short rise and fall time constant. T15 in conjunction with C2 has a long rise and fall time constant. Thus, any input signal will rapidly initiate agc action via T14 (in 20 ms), but after a longer time (200 ms) T15 takes over to control the agc. The effect is rapid initial agc response but not false agc response to sudden peaks after the speech input has started. T6–T8 form a trigger circuit which detects sudden peak inputs above 4 mV, such as noise bursts. When such a burst occurs during a pause in input it prevents via T10 and T13 the output from being turned on.

T9 in conjunction with C3 forms a sort of memory circuit having a time constant of about 1 second. So long as a speech input is present, it does not act but during a pause exceeding one second, C3 discharges to turn on T12 via T11 and turn off the audio output of the SL630/SL620 combination. The capacitors mentioned above for the various time constants, C1, C2, and C3, are external to the IC and their value can be experimentally changed to suit individual preferences. W2EEY■

**If units are not available locally, write to Plessey Microelectronics, 170 Finn Street, Farmingdale, L.I., NY for location of nearest distributor.*

An Integrated Circuit CWID Generator

P. J. Ferrell, W7PU9

A project to develop an automatic, solid-state CW ID generator was recently initiated by members of the Seattle repeater group. Although there have been a number of recent articles concerning such devices,[1–2–3–4] our starting point was the FM Magazine article by Woore.[3] The outcome of this project must be classified as an engineering overkill. The resulting CW identification generator features a clocked character generator (for flawless CW with variable speed), inexpensive RTL integrated circuitry,[3] and a computer-designed diode read-only memory matrix. Also included are "pulse" starting, a discrete "hold" voltage available during ID execution, and a continuously adjustable keying speed (from far too slow to far too fast). Not only are these generators ideal for repeater identification, but they may be used to identify any amateur station such as RTTY, ATV, etc.

The block diagram in Fig. 1 shows the major divisions of the ID generator. Many excellent articles covering RTL logic design have appeared in amateur literature[5–6] and the reader is referred to them for background material.

Program Counter and Start/Stop Flip-Flop

This six-stage ripple counter consists of three dual JK flip-flips. The first five cascaded stages are the program counter and count from 0 to 31. The last stage is employed as the start/stop flip-flop. Each stage of a ripple counter is arranged to toggle (change state) on the output of the preceding stage. A five-stage program counter has 32 distinct stages ($2^5=32$). When arranged as shown in Fig. 2, the program counter advances under the control of gate G1 which derives its input from the character generator. Each dot, dash, or blank character advances the program counter by one count. The last character (number 32) resets or clears the first five stages, but toggles the start/stop flip-flop to the "stop" or set position, thereby halting the character generator.

A positive pulse into the "preclear" input of FF6 clears the halt and allows the

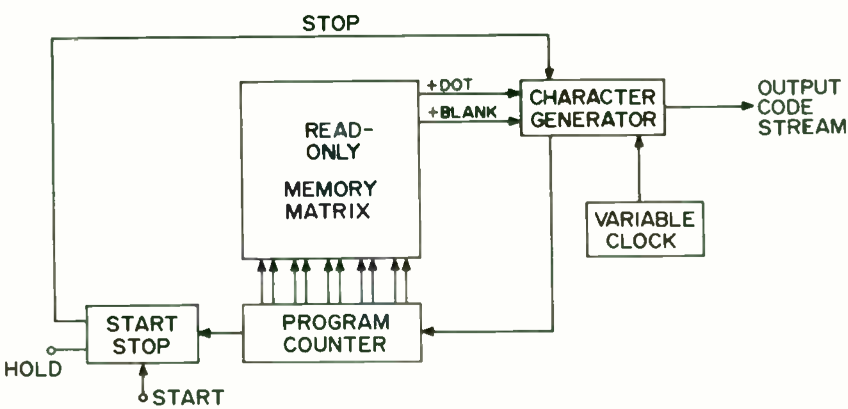


Fig. 1. CW ID generator block diagram.

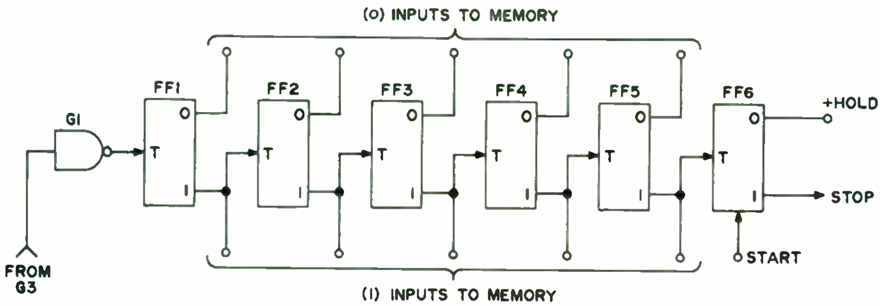


Fig. 2. Program counter.

character generator to run, thereby initiating a cycle of operation. A five-stage program counter was chosen since virtually all amateur calls can be encoded in 32 characters worth of dots, dashes, and blanks. RTL JK flip-flops are adversely affected by capacitive output loads, and will not toggle reliably if the capacitance is too high. This fact precludes the use of silicon diodes in the *and* portion of the diode memory.

If the program counter output lines were buffered (isolated from the flip-flops with gates or inverters), then any type of diode could be used in any memory position; but germanium diodes have very low capacitance and may be connected directly to the flip-flop outputs, thereby saving the cost of 10 buffer stages.

Variable Speed Clock

The clock circuit must deliver a negative-going pulse with leading edge of less than a microsecond duration in order to toggle an RTL JK flip-flop. The pulse repetition rate should be variable to permit choice of code speed. The circuit is shown in Fig. 3 with a PNP silicon transistor paired with gate G2. The net effect is a PNPN switch. Capacitor C charges to about 2V and then discharges through the gate with a leading edge which is very abrupt.

Nearly any of the new PNP silicon transistors will work in this circuit. The minimum value for R is about 33 kΩ else sufficient current is available to hold the switch in conduction (just like a neon relaxation oscillator). For R much above 1 MΩ insufficient current is available to initiate the regenerative "snap" action. Values of R between these limits work well.

Character Generator

The electronic generation of Morse code requires the creation of dots, spaces, dashes, and blanks which have a precisely specified relative length. The dot and space are each of one unit duration, while the dash and blank are each of three units duration.

An extremely clever character generator was borrowed from the Micro-Ultimatic Keyer[7] and forms the heart of the ID generator. The character generator consists of two JK flip-flops (FF7 and FF8) and gates G3 and G4 as shown in Fig. 3. A positive (stop) voltage on terminal C of FF7 holds it in the clear state, thereby stopping the character generator. If the stop voltage is removed, the character generator toggles in such a manner as to produce a string of dashes at the output of gate G4. If a positive (+DOT) voltage on terminal C of FF8 changes the string of dashes into a string of dots. A dash (or blank) requires four clock pulse intervals while a dot requires two. Gates G1 and G4 each invert the output of gate G3.

The output of gate G1 must be either a dot or dash character (never a blank), and is used to advance the program counter at the end of each character. The second input to gate G4 will blank out the output code stream, and is used to produce a blank character. If a blank is required, then

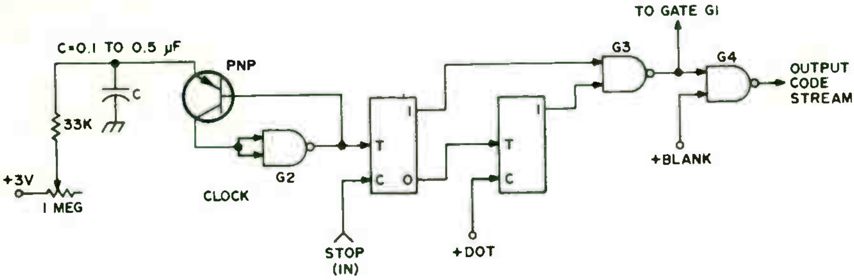


Fig. 3. Clock and character generator.

a positive "+BLANK" input from the diode memory causes the dash generated by the character generator to be blanked

polarity of the (1) output

	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
FF5	o	o	o	o	o	o	o	o	o	o	o	o	o	o	o	o
FF4	o	o	o	o	o	o	o	o	+	+	+	+	+	+	+	+
FF3	o	o	o	o	+	+	+	+	o	o	o	o	+	+	+	+
FF2	o	o	+	+	o	o	+	+	o	o	+	+	o	o	+	+
FF1	o	+	o	+	o	+	o	+	o	+	o	+	o	+	o	+
+DOT	o	+	+	o	+	o	o	+	o	o	o	o	o	+	+	+
+BLANK	o	o	o	+	o	+	+	o	o	o	+	o	o	o	o	o
CHARACTER	—	:	·	x	·	x	x	·	—	—	x	—	—	·	·	·
		D			E				W					7		

Fig. 4. Diode memory inputs and outputs.

out – which results in the transmission of a blank character.

Thus, the role of the diode read-only memory matrix is to provide just that sequence of +DOT and +BLANK inputs to the character generator which results in the transmission of the desired code stream. Gate G4 is the output with a plus representing "key down" and a zero representing "key up."

Diode Read-Only Memory Matrix

This is the hard part! Each desired code stream requires a distinct and different read-only memory design. A +DOT voltage must be produced by the diode memory for each program counter state that corresponds to a dot in the desired code stream, and a +BLANK is required for each blank character. An example is presented in Fig. 4.

Suppose that the first letters of the desired code stream were "DE (blank, blank)W7". The program counter states are shown corresponding to the required outputs from the diode memory matrix. A dash is seen to be the "default" condition; i.e. if neither a +DOT nor a +BLANK occurs, a dash results. Each of the 32 program counter states must be accounted for since they all appear on each ID execute cycle.

Figure 5 shows the method for decoding the program counter. The five terminals marked FF1 through FF5 are connected to either the 1 or 0 side of the respective flip-flop. If *any* of the five connections is low, then the whole common line is low.

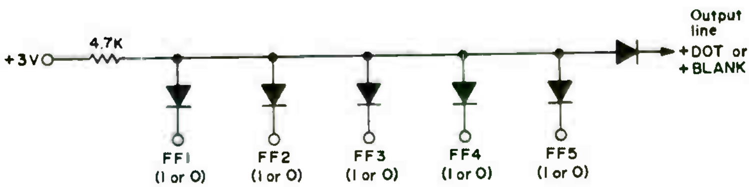


Fig. 5. Diode decoder for program counter.

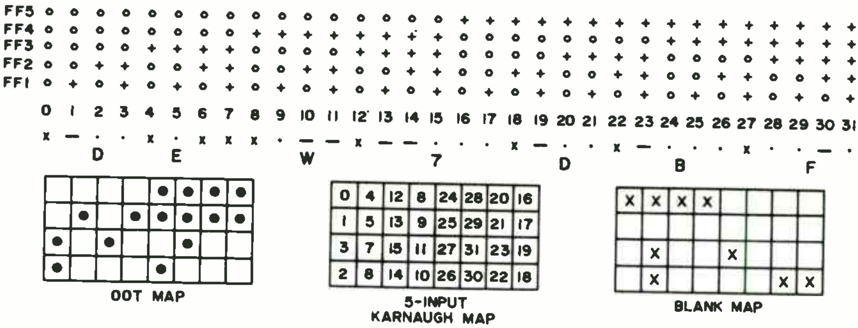


Fig. 6. Using the Karnaugh map.

The only time the common line can be high is when all five input connections are high. For any given connection, this will occur exactly once during each program counter cycle. This type of diode decoder is often called an *and* gate since it has a high output only when all inputs are high.

Because of the diode output from the common line, these decoders may be paralleled to obtain the required +DOT and +BLANK functions. This paralleling is often referred to as an *or* gate since *any* high input results in a high output.

In the example of Fig. 6, the desired code stream has 15 dots and 9 blanks. If we employ a separate diode decoder for each one, then a total of 24 decoders would be required: 15 would be paralleled to give the +DOT signal and the remaining 9 would be paralleled to provide the +BLANK signal. Each decoder requires 6 diodes for a grand total of 144 diodes to build a straightforward read-only memory using this technique.

Fortunately for us, the English philosopher George Boole published his *Investigation of the Laws of Thought*, in which he resolved the ambiguity of the words *and* and *or* by means of a kind of algebra. In 1938, eighty-four years later, Prof. D. E. Shannon (the *Information Theory Shannon*) put Boole's algebra or Boolean Algebra to use in the *Symbolic Analysis of Relay and Switching Circuits*. This classic paper has revolutionized switching design, and has led to the development of minimization techniques which can dramatically reduce the diode count of our ready-only memory. The details of these methods and the underlying theory are beyond the scope of this article, but for those who are fascinated by this stuff, standard texts are available which will quickly dispel the aura of "black magic" that seems to surround this area.[8]

For our purpose, a graphical reduction technique known as a *Karnaugh map* will be employed. Figure 6 illustrates the process for the code stream *DE W7DBF*. The polarity of the 1 output levels of flip-flops FF1 through FF5 are shown as the program counter steps from 0 (all FFs clear) to state 31 (all FFs set). The sample code stream begins with a blank and has three blanks separating the DE from the W

A Karnaugh map organization of program counter states is presented, flanked on the left by the DOT map and on the right by the BLANK map for the desired code stream. Reduction is accomplished by "folding" the map about any of the dividing lines and pairing the marks (dots or blanks) which overlap. For example, folding the DOT map about the vertical centerline pairs dot 5 with dot 21, and dot 9 with dot 25, and dot 15 with dot 31. Successive pairing, then pairing pairs, etc. allows a reduction from the original 144 diodes to a total of 48 diodes arranged as shown in Fig. 7. Note that in Fig. 7, the 0/1 flip-flop lines are reversed for FF2 and FF4. This reversal materially simplifies printed circuit construction.

As an example of pairing, consider the blanks in positions 0, 4, 8, and 12 of the BLANK map of Fig. 6. If the map is folded

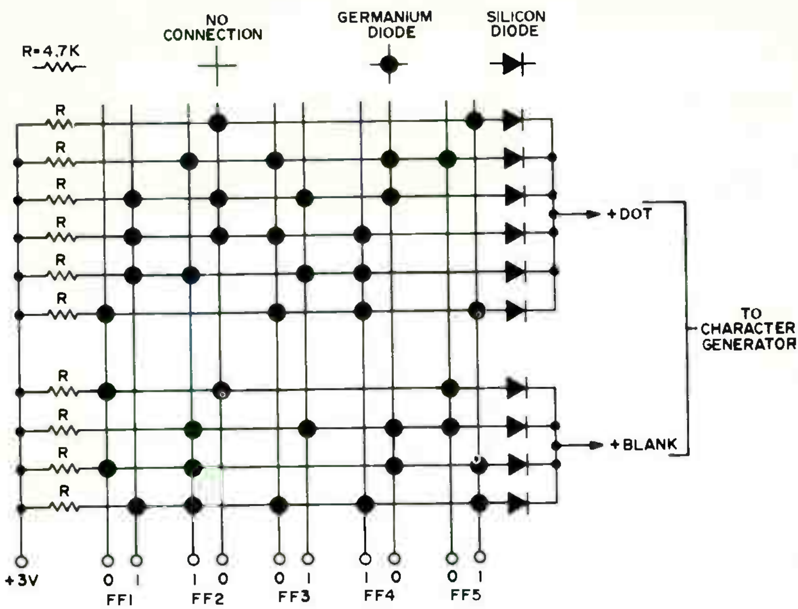


Fig. 7. Diode read-only memory matrix for "DE W7DBF."

about the second vertical line (the one separating positions 4 and 12), then blank 4 pairs with blank 12, and blank 0 pairs with blank 8. If we fold again, then all four blanks coincide with each other. This is illustrated in Fig. 8. In part A, the program counter states for positions 0 and 8 are compared. They differ in exactly one FF position (FF4) as all pairs must. A single diode decoder of the type shown in Fig. 5 could get both blanks simply by neglecting to connect to FF4. It even saves one *and* diode.

Part B of Fig. 8 presents the same comparison for blank 4 and blank 12, and again they differ only in FF4. In part C, a comparison of the 0/8 with the 4/12 pairs shows that these differ only in FF3. Thus, if a three-input diode decoder of the type presented in Fig. 5 were connected to FF1 (0), FF2 (0), and FF5 (0), it would give a +BLANK for all four desired program counter states (0, 4, 8, and 12). Without this reduction, 24 diodes (six for each blank) would have been required rather than the four actually required. This diode decoder may be found in Fig. 7 as the first line in the +BLANK group.

FF5	o o o	o o o	o o o
FF4	o +	o +	
FF3	o o = o	+ + = +	o + =
FF2	o o o	o o o	o o o
FF1	o o o	o o o	o o o
	0 8	4 12	Pairing A and B
	(A)	(B)	(C)

Fig. 8. Actual reduction procedure.

The foregoing example illustrates the use of a Karnaugh map. The actual reduction is performed as in Fig. 8. It should be noted that some states will not pair, such as blank 27 in Fig. 6. To pick up this blank, a full five-input diode decoder is required. From Fig. 6, we see that for blank 27, the program counter is in state "++0++" and the resulting diode decoder can be found in Fig. 7 as the bottom diode line.

The rule when pairing states is that the two program counter states can differ in exactly one flip-flop position. All other

```
EXECUTION BEGINS...


CODE STREAM [ -.. .   .-- --... -.. -... ..-.]
REQUIRES 48 DIODES AND 10 RESISTORS.

PLACE SILICON DIODES (S), GERMANIUM DIODES (G), AND
4.7K RESISTORS (R) IN THE FOLLOWING POSITIONS:

D B  0 1 2 3 4 5 6 7 8 9  +
S          G           G  R
S        G   G     G G    R
S      G   G   G   G      R
S      G   G G   G        R
S      G G     G G        R
S    G       G   G     G  R
  S  G     G         G    R
  S      G     G   G G    R
  S      G         G   G  R
  S    G G   G   G     G  R


EXECUTION COMPLETE...
```

Fig. 9. Computer printout.

positions must agree, including any omitted connections as in Fig. 8, part C. If there is disagreement in more than one FF position, then these two states do not pair.

After completing the reduction process, check to make sure that every necessary state has been covered at least once by one of the final decoders; otherwise, you may be surprised at the resulting code stream. This type of calculation, once understood, is not particularly difficult, but it certainly is tedious and has lots of room for errors. Slight changes in the desired code stream (even position) can have a huge impact on the diode count.

For example, the diode count for the code stream in Fig. 6 is 48. If just the *DE* is slid one count to the right, a new code stream is formed which starts with two blanks, and has two blanks between the *DE* and *W*. The diode count for this new stream is 55. If this new stream is shifted two positions to the left, so that the two leading blanks become trailing blanks, then the diode count becomes 85. These effects are unpredictable, and for complete optimization each code-stream version must be reduced separately, and the results compared. This greatly increases the already great tedium of such calculations.

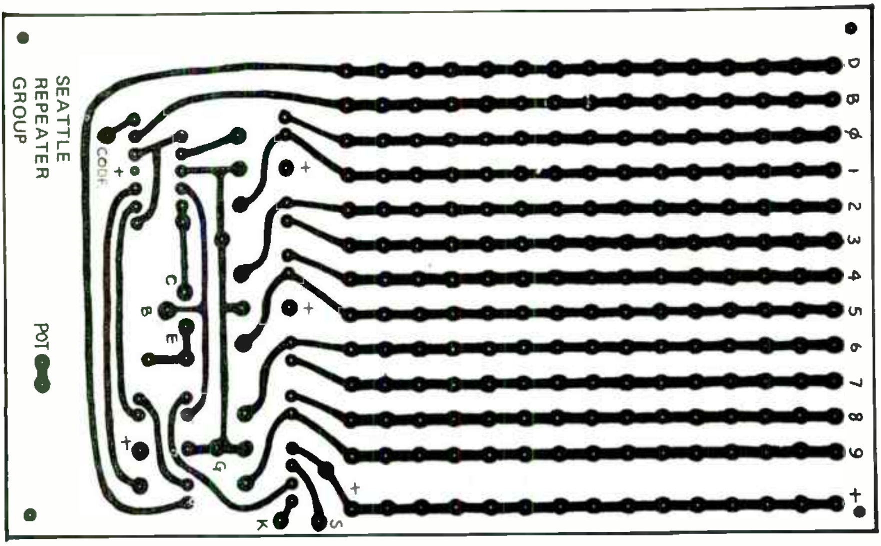
Computer Optimization

In order to minimize the pain of diode memory design, the task was subcontracted to a digital computer. The Seattle repeater group is extremely fortunate in having remote access to the University of Washington Computer Center's Burroughs B5500 computer, one of the nicest hardware/software systems ever put together. The resulting program in extended ALGOL accepts the desired code stream (in dots, dashes and blanks) as an input and performs a complete Boolean reduction for both +DOT and +BLANK diode decoders.

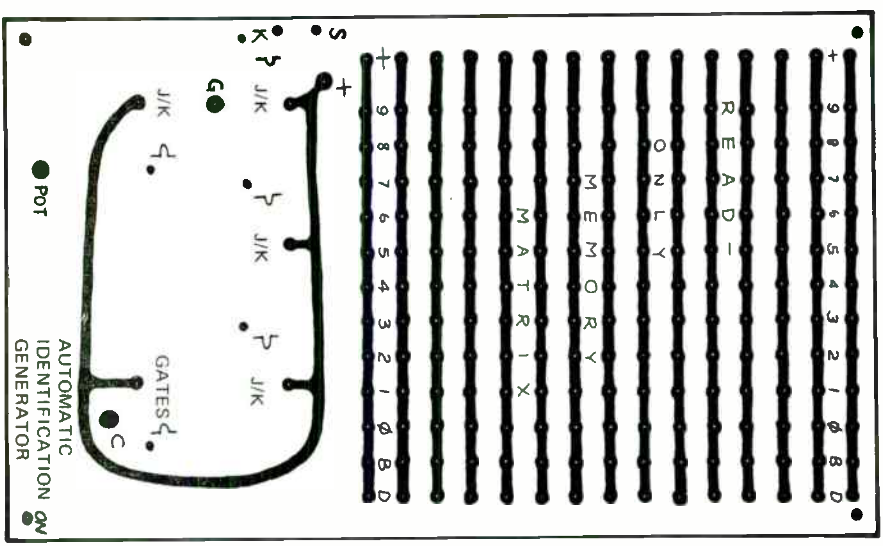
If the specified code stream is less than 32 characters (more than 32 characters are not allowed), then the computer assigns the necessary trailing blanks and performs the required reductions, and repeats the process for each shift of the code stream until all the trailing blanks have become leading blanks. The code stream version having the smallest diode count is printed out along with diode and resistor counts and an actual map of the entire diode read-only memory matrix. Examples of the digital computer printout are shown in Fig. 9.

Construction

The four dual JK flip-flops are Motorola MC790P (or HEP 572) and the quad 2-input gate is a Motorola MC724P (or HEP 570). Virtually any PNP silicon transistors will function in the clock circuit. A HEP 57 is a good choice. Both germanium and silicon diodes are used in the diode read-only memory. Germanium diodes are employed for the *and* function, since their low junction capacitance will not load the

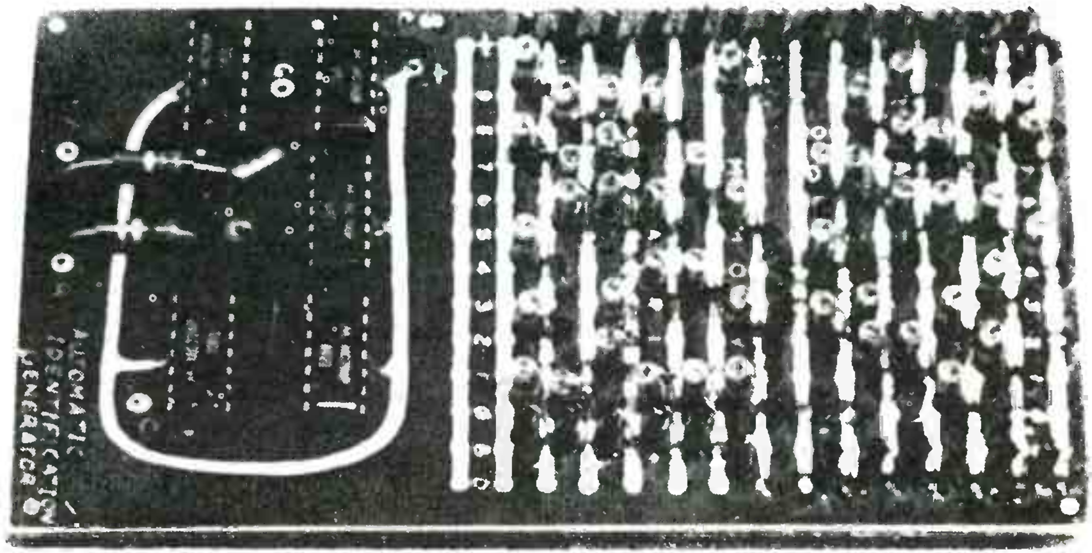


Circuit board (bottom).



Circuit board (top).

Fig. 10. Full-size reproductions of PC board.



Here's what the board looks like when the flatpacks and diode matrix are soldered in. The vertical placement of the diodes helps to keep the size down to this 3 x 5 in. circuit board.

JK flip-flops in the program counter. Either silicon or germanium diodes may be used in the *or* function, with silicon signal diodes preferred since they result in a higher noise margin for the memory. Cheap diodes are available from various solid-state supply houses. Poly Paks features 50 silicon or germanium diodes for $1.

Both sides of 3 x 5 in. double-sided PC board are shown in Fig. 10. A one-sided board was used for the first few models, with a second 3 x 3 in. board used to complete the matrix connections. This "cordwood" construction is a pleasure to look at, but a nightmare to wire up. If a diode goes "west" on a cordwood style generator, it is best to throw it away, since unsoldering about 80 diodes and resistors and then getting things back together is even worse than the initial construction effort. A double-sided epoxy–glass PC board is recommended for the ID generator. An operational generator should be enclosed in a metal box with all leads bypassed for rf. Even VHF/UHF fields have the ability to drive RTL logic circuitry absolutely crazy. Before undertaking the construction of this ID generator, the builder should obtain as many of the referenced ID generator articles[1-2-3-4] as can be found and read them over carefully. The additional background material will amply repay the effort involved.

The Seattle repeater group can supply a moderate number of tinned epoxy–glass circuit boards for this ID generator. The board is not drilled, but assembly instructions and a computer optimized diode map for the circuit board is included. Be certain to specify the desired code stream, keeping in mind the absolute limit of 32 characters (dots, dashes, and blanks). Unit cost is $10, and they may be obtained from the Seattle repeater group, 18235 46th Pl. S., Seattle WA 98188.

Acknowledgement. The author expresses appreciation to K7EVO for art photography, and to K7MWC for the snapshot.

References

1. *A Digital Identification Generator*, Barry Todd K6ZCE, 73 Magazine, March 1967.
2. *IC Repeater Identifier*, Tom Woore WB6BFM, FM Magazine, January 1969.
3. *Automatic CW Identifier* John Connors W6AYZ, Ham Radio Magazine, April 1969.
4. *A Digital Morse Code Generator*, Jerry Hall K1PLP. QST. June 1969.
5. *Digital Logic Devices*, QST, July 1968.
6. *Micro-Logic for Non-logic Users*, Hank Olson W6GXN, 73 Magazine, June 1967.
7. *The Micro-Ultimatic*, Tom Pickering W1CFW, 73 Magazine, June 1966.
8. *Logical Design of Digital Computers*, Montgomery Phister. John Wiley & Sons, 1958.

IC Square-Wave Generator

Charles Jimenez, WA4ZQO

This project was designed primarily for those who wish to acquaint themselves with, and gain experience using, integrated circuits. The square-wave generator described is a rather easy construction project. This is not to say, however, that it sacrifices performance for the sake of simplicity. In fact, several shortcomings of usual square wave generator circuits have been overcome in this design. The construction of this unit will result in a fine piece of test equipment which will be handy around any ham shack.

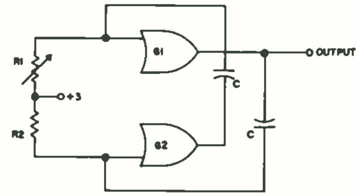


Fig. 1. Simple astable multivibrator circuits.

Circuit details

The circuit which generates the basic square wave form is shown in Fig. 1. This is called an astable multivibrator. The gates G_1 and G_2 are from a Fairchild µL 914 integrated circuit. It's a dual two-input gate and should be familiar to many *73* readers.

Each gate is cross-coupled to the other through a resistor-capacitor network which determines the operating frequency. Different capacitors are switched in for changing frequency bands. There are five bands: band A—10 Hz to 150 Hz; band B—100 Hz to 1.5 kHz; band C—1 kHz to 15 kHz; band D—10 kHz to 150 kHz; and band E—70 kHz to 1 MHz. In order to vary the frequency within these bands, normally you have to vary both R_1 and R_2 simultaneously. But by varying only R_1 we can obtain the same bandspread and save the cost of a ganged pot. Unfortunately this will destroy the symmetry or squareness of the output waveform. This can be remedied and, as you will see later, the remedy brings along a couple of extra advantages of its own.

The simple astable multivibrator of Fig. 1 would work nicely if it were not for one big shortcoming. It may cease oscillating when switching frequency ranges or it may fail to start up when turned on. This happens when both gates saturate at the same time. In normal operation, G_1 and G_2 conduct on alternate cycles; that is when G1 conducts, G_2 is cut off. This process is insured by the capacitors which drive the gates by charging and discharging alternately.

However, suppose now that you are changing bands. As the arm of the bandswitch moves from one capacitor to the next there will be a time interval where there is no capacitor in the circuit at all. Both gates will now see a positive voltage at their inputs through R_1 and R_2 and will conduct heavily. The multivibrator will now be locked and cannot be started up again unless you first turn off the power. Obviously, it would be very frustrating to have to turn off the power whenever you wanted to change bands. I ought to know since it kept happening to me in my early stages of experimenting.

The seemingly insurmountable problem was easily overcome by using a couple of diodes. Fig. 2 shows the circuit, known as a self-starting circuit. By referring to Fig. 1 and 6 you'll be able to see how this circuit works. The two diodes are connected to each output and to the junction marked (X). The +3 V for R_1 and R_2 is now supplied through D_1 and D_2 from the collector of either gate. Remember that when a gate is cut off the collector goes positive and +3 V appears at junction (X). The circuit will operate properly as long as at least one gate is cut off. Now if both gates should happen to saturate at the same time when switching capacitors, the positive voltage at (X) will disappear, tending to cut off the gates immediately. In other words, the diodes, which form the OR gate, will not allow the multivibrator to lock in a saturated condition. Proper operation will begin when the next capacitor is switched in. We now have a reliable astable multivibrator circuit which produces rectangularly shaped waves.

As stated earlier, the method used for varying the frequency destroys the output wave's symmetry. When R_1 is varied, the output may change from a square wave to a rectangular wave or pulse, for instance. Of course, this change in wave shape has no effect upon the frequency as it is varied. In order to correct the wave shape, the output of the multivibrator is fed into a Fairchild μL 923 J-K flip-flop. The action of this flip-flop is shown in Fig. 3. The μL 923's output *changes* only when the input signal goes *negative*. Notice that the output is always a perfectly symmetrical square wave, regardless of the shape of the input waveform. The input can be spikes, pulses, rectangular waves or any other waveform which has a fast negative going portion. It can also be seen from the diagram that

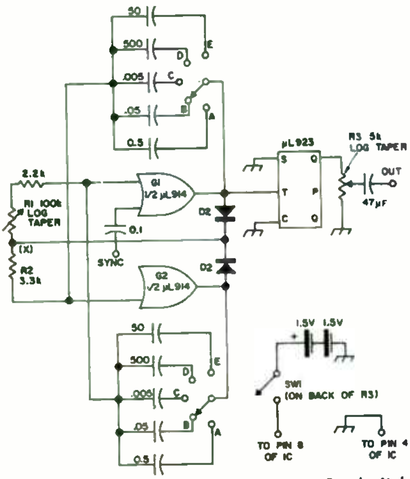


Fig. 2. Complete square wave generator. Bandswitching capacitors are 10% or better tolerance. Resistors are 1/4 watt.

the output frequency is one-half the input frequency. This means that the multivibrator is actually operating at twice the frequency indicated on the front panel dial. The generator puts out a beautiful square wave to 1 MHz and beyond. A slight amount of overshoot on the rising portion of the square wave is normal at high frequencies. The S (set) and C (clear) inputs are both grounded, and the P (preset) input and the $\overline{Q}$ output are disregarded.

Synchronization pulses are fed into gate G_1 for locking the generator's frequency to some external source or oscillator. For in-

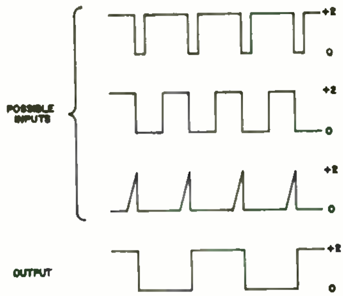


Fig. 3. Operation of the μL 923 flip-flop. Note that regardless of the shape of the input waveform, the output is always a perfectly symmetrical square wave.

stance, by feeding a 100 kHz signal from your receiver's crystal calibrator, you can lock the generator at 100 kHz, 50 kHz, 33 kHz, 25 kHz, etc. Of course, this will result in excellent frequency stability and accuracy. Be careful not to feed too much signal into the sync terminals, as you might cause erratic operation.

The output level is controlled by a 5-k pot at the output of the flip-flop. The actual value isn't too important as long as it isn't too low. Otherwise you might load the flip-flop too much. Don't go below 1k. The use of a log-taper pot will permit adjustment down into the millivolt region for low-level audio work. The output voltage is about two volts into a high-impedance load.

The supply voltage for the unit is taken from two 1.5 volt D cells in series. Current drain is less than 40 mA. Remember that pin 8 of both IC's is connected to the +3 V and pin 4 of both is connected to ground or minus. A colored line or flat portion on the edge of the IC's body identifies pin 8. Diodes D_1 and D_2 can be almost any signal diode. Parts values should be followed rather closely to insure adequate band coverage.

Construction

The printed circuit layout is given for those who want to make their own PC boards. You can get an idea of the front panel arrangement from the photo. Actually there is nothing critical about layout or construction so you can arrange things inside to your liking. I used a 4 x 5 x 6 minibox for my unit, which is just right if you use a Millen 10039 vernier dial as I did. This is a compact unit, and using a larger dial will mean using a larger cabinet. The Millen dial is rather expensive and maybe you'll

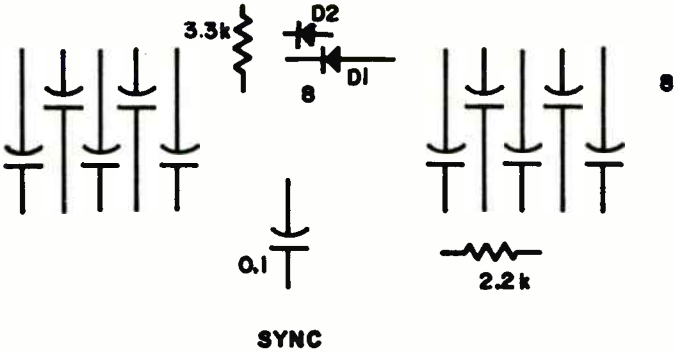


Fig. 4. Full size layout of PC board. This is a bottom view with components mounted on top.

want to use one of the imports and make your own scale. Since I'm on the subject of cost, I might as well say that the whole project will come to about $20.00 with all new parts, including the Millen dial. With an imported dial, you can probably knock $5.00 off that figure.

Calibration

You might have noticed by now that the scale on my dial is not linear. This is because I used a linear taper pot for R_1, since it was available. I'm not particular about such things but if you prefer a more linear scale, I would suggest trying a log- or semi-log taper pot. Keep in mind the fact that most vernier dials turn only 180° as opposed to the normal 270° turn of a pot. You might have to adjust the position of the pot in the dial to insure proper bandspread.

Calibration can be achieved only by the use of a scope or frequency counter. If you don't own one maybe you can gain access to one for about fifteen minutes or so. By using 10%, or better, tolerance capacitors you'll only have to calibrate the lowest frequency band. On each succeeding band, the frequency is ten times the frequency at the same point on the previous band. The simplest method of calibration is with a 60-Hz sine wave, which can be supplied internally on most scopes.

To calibrate the lowest band (band A) the following procedure can be used. First, allow the scope to warm up for a few minutes until it becomes stable. Turn off the internal sync of the scope. Apply the 60-Hz sine wave to the vertical input of the scope and adjust the sweep frequency until you obtain six full cycles on the screen. Since you are not using the internal sync, you'll have to adjust it very carefully to stabilize the pattern. With six full cycles on the screen, the sweep frequency is now set at 10 Hz. Next, feed the square wave from the generator to the scope and tune the generator's frequency until you obtain one full cycle of a square wave. Be careful not to move the sweep frequency of the scope. The square wave generator is now set at 10 Hz and can be marked on the dial. Tune the generator again until two full cycles are visible on the screen. The scale can now be marked at 20 Hz. This process can be continued on up to 100 Hz. Afterwards, go back and repeat it all over again to make sure you have the proper calibration. Once you have made certain that there are no errors, you can mark the rest of the bands as outlined previously. On band E, you can listen to the signal on a broadcast receiver to see if it checks out. The bands on my unit did not exactly come out in multiples of ten because I used 20% tolerances capacitors from my junk box. Even so, they came out very close.

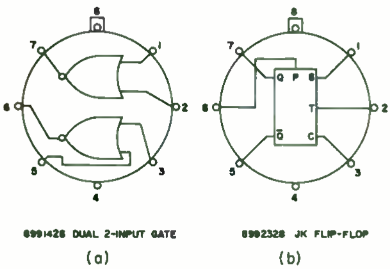


Fig. 5. Basic diagram for the IC's.

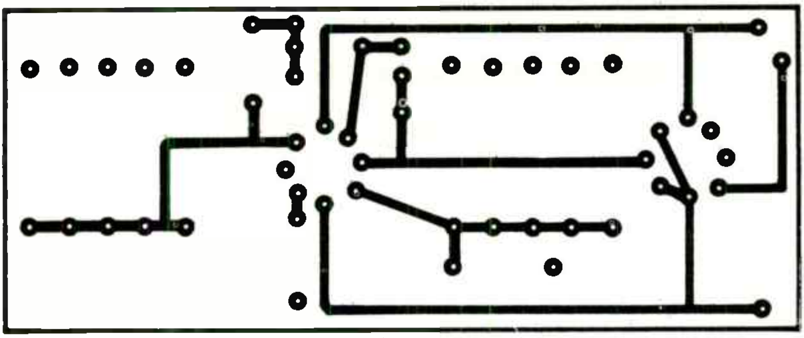


Fig. 5. Full size printed-circuit board used in the IC square-wave generator.

Operation

Square waves are very handy for testing amplifiers of all sorts in conjunction with an oscilloscope. In audio work they will reveal poor high or low frequency response, ringing and other ailments. Of course, you don't need a scope just for general testing of audio amplifires and such. A simple signal tracer will do.

Speaking of oscilloscopes, you can use 500-kHz square waves for adjusting compensating capacitors in scope probes and step, or decade, attenuators. Usually, the instruction manual of your scope will outline the proper procedure. Since this unit will supply a signal at up to 1 MHz in frequency, it can be used to fix or test amateur or broadcast receivers. However, a detailed discussion of testing with square waves is beyond the scope of this article.

I'm sure that if you build this square wave generator, you'll be very pleased with its performance and reliability.

. . . WA4ZQO

An Audio Sinusoid Generator

Howard Phillips, W6FOO

Do you need an audio generator which delivers a sine wave over the frequency range of 500 hz to 5 khz? The oscillator described here will do the job while maintaining a nearly constant output voltage over the range of 1 khz to 5 khz.

An audio oscillator which generates a low-harmonic-content sinusoid has many uses. Examples include two-tone generators for use in amplifier linearity tests and AFSK oscillators for encoding Teletype signals for radio transmission. The oscillator described here is well suited for the latter application because switching different capacitors into the circuit allows generating two equal-amplitude non-simultaneous tones using the same oscillator.

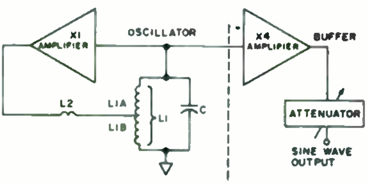


Fig. 1. Sine wave oscillator and buffer.

Fig. 1 illustrates the complementary audio oscillator circuit. The npn-pnp amplifier has a unity voltage gain and is active over a 10 volt dynamic range. The complementary amplifier provides drive through L2 to the tuned circuit composed of L1 and C. The single-ended buffer amplifier prevents heavily loading the tuned circuit. The circuit uses two 88 mh telephone loading coils. These toroid inductors have two windings

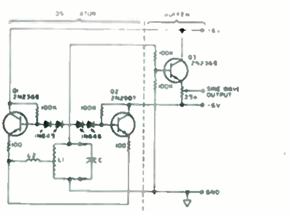


Fig. 2. Functional diagram of oscillator and buffer.

which can be connected in series for L1 = 88 mh or only one of the windings can be used for L2 = 22 mh.

The operation of the oscillator is better understood by considering the functional diagram illustrated by Fig. 2. The current which flows through L2 is in phase with the voltage developed across the tuned circuit. The output voltage from the X1 complementary amplifier is divided between L2 and L1B. The drive voltage to L1B does not vary with frequency between 1 khz and 5 khz.

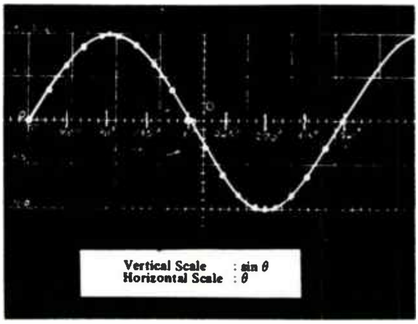


Fig. 3. Plotted points of sin θ superimposed upon measured output waveform.

Consequently, the output voltage is constant over this frequency range. The buffer amplifier has a voltage gain of 0.4 and drives a potentiometer which can be adjusted to provide a sine wave output of zero to 9 volts peak-to-peak.

Fig. 3 shows an oscilloscope photograph recorded to show the quality of the sine wave as measured at the output of the buffer amplifier. Points are plotted on the oscillograph to show the theoretical waveform of a pure sinusoid. The oscillograph in Fig. 4 shows the waveform quality at four operating frequencies in the range of 400 hz to 4 khz. The waveform distortion decreases with frequency. The results of distortion analyzer measurements made on the signal output from the buffer show a waveform

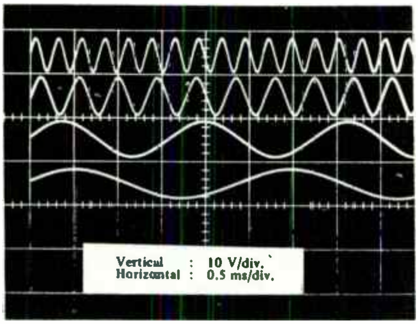


Fig. 4. Measured output waveforms.

distortion of 4% at 3.5 khz and a 1.4% waveform distortion at 500 hz. Filters can be used to decrease the waveform distortion to less than 0.1% if required.

The output voltage is plotted as a function of frequency in Fig. 5. The decrease in output voltage at lower frequencies is caused by: (a) The Q of the resonant circuit decreases at lower frequencies; (b) the inductive reactances of L2 and L1B decrease at lower frequencies, and the voltage drops across the 100 emitter resistors become appreciable. The oscillator can be made to operate at lower frequencies by increasing the values of L1 and L2.

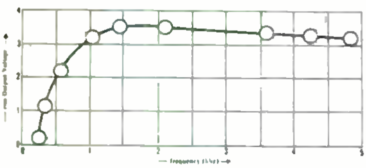


Fig. 5. Measured output voltage from buffer.

Fig. 6 shows a graph which can be used to select a value of C for any desired oscillator frequency. The error bars above and below the plotted data points reflect the ± 20% tolerance of the capacitors used to obtain the data.

The choice of semiconductors is not critical except that silicon transistors and silicon diodes should be used. The bias conditions are designed to make use of the forward voltage drop (0.7 volt) across a silicon diode. The power supply voltage values are not critical. If the semiconductor

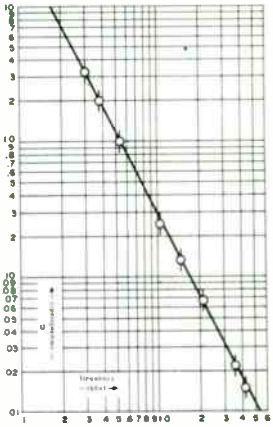


Fig. 6. Capacitance vs oscillator frequency.

breakdown voltage specifications are not exceeded, higher output voltage can be obtained by using higher power supply voltages. The physical layout of the components is not critical, and special construction techniques are not required.

. . .W6FOO

A Six Meter IC Converter

G. W. Reynolds, K2ZEL

This converter is one module of a solid state mobile receiver and transmitter designed, constructed and tested by the author. The advent of solid state devices make it possible to design highly stable receivers and transmitters which will operate directly from the car battery. The modular system was selected to insure adequate shielding between sections of the receiver and the transmitter and to facilitate changes in the system at a future date if desired.

The converter is constructed of two "ICs" and associated components. The one is an *rf* amplifier of 30 db gain and the second is a crystal oscillator, mixer, amplifier stage of 40 db gain. Evaluating insertion losses, the overall gain of the converter is 55 db. The factors considered when selecting the particular "IC" used were cost, availability and suitability. The Fairchild 703 operates as a differential amplifier of 20 to 30 db gain at frequencies up to 200 mhz and is inexpensive and readily available. The RCA CA-3028 is more expensive and can be used as an *rf* differential amplifier with nearly the same gain figure, but its advantage in this converter is that it can be used as a mixer, local (crystal or tuned circuit controlled) oscillator and amplifier of 40 db gain. The circuit employs a third overtone crystal in the series resonant feed back position of the oscillator circuit.

The shield placed across IC is necessary to prevent inductive feedback and self-oscillation of this "IC." The tuned circuit L1C1 is provided to match the antenna input to the converter and link coupled through a shield to reduce cross-talk and birdies being fed through the converter to the tunable *if* module (or receiver).

Construction

Construction was made on a vector board using vector board clips as mounting terminals. This form of construction permits substitution of electrically similar components without detriment to the circuit configuration. The vector board was cut to fit the standard 5"x2¼"x2¼" minibox and mounted with 5/8" nylon (plastic) standoff insulators. The metallic shields were made of sheet copper and soldered to supporting vector board clips. All resistors used are ¼ watt dissipation and the capacitors are either disc ceramic or mylar printed circuit type. The jacks for coupling the converter to the antenna and to the tunable *if* module (or receiver) are standard BNC jacks. Coil forms are ¼" nylon rod. Component layout is shown in Fig. 1.

Alignment Procedure

After completing construction use a grid dip meter or tunnel dipper to adjust the frequency of the LC combinations to the appropriate frequencies per chart.

Tuned Circuit	Frequency
L1C1	50 mhz
L4C2	50 mhz
L5C3	50 mhz
L6C4	Xtal Frequency
L7C5	IF out.

The *if* output will be the difference between the crystal frequency and the 50 mhz.

Connect the converter to a receiver set and apply power to the converter. Using the grid dip oscillator in the wavemeter condition adjust C4 for maximum *rf* signal when the dipper is near L6.

Next apply a 50 mhz modulated signal from a signal generator to pin 2 of IC2 and adjust the receiver for the signal at the *if* output. Adjust C5 for maximum output. Disconnect the signal generator and apply signal to J1 and adjust first C3, then C2, and then C1 for maximum signal output. Repeat ad-

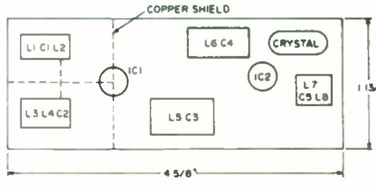


Fig. 1. Component layout.

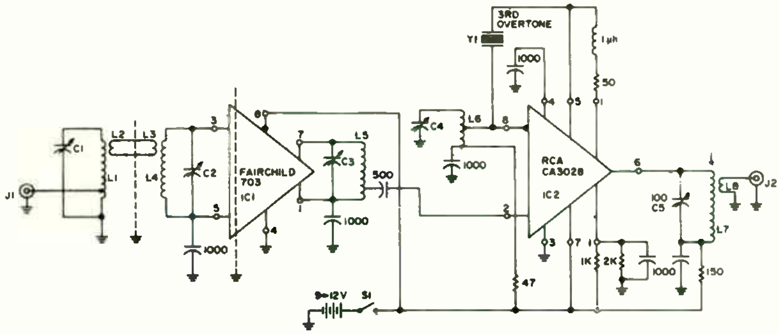


Fig. 2.

L_1 6T ¼" dia. closewound	L_7 30T ¼" dia. "	Y_1 = Xtal 43.5 mhz for 6.5 ic.5 mhz *if*
L_2 2T ¼" dia. "	L_8 3T ¼" dia. "	
L_3 2T ¼" dia. "		
L_4 6T ¼" dia. "	C_1 50 pf	R = ¼ watts
L_5 6T ¼" dia. " Tap 2T	C_2 50 pf	C = 50 vdc
	C_3 50 pf	C1-4 = 7-45 pf trimmers
L_6 6T ¼" dia. " Tap 2T	C_4 50 pf	
	C_5 100 pf	

justments C5, C3, C2, C1 with the generator at antenna input (J1) because of slight interaction between adjustments. Remove the signal generator and connect the antenna. Happy listening to six meters.

...K2ZEL

An IC Audio Notch Filter

John J. Schultz, W2EEY/1

A bridged-T audio notch filter combined with an IC amplifier produces a highly versatile, wide-range audio rejection filter with both variable frequency and variable "Q" controls.

Audio filters are certainly nothing new. They have long been used to improve the selectivity of a receiver or transceiver when it was not desired to "dig" into the *if* circuitry and improve the selectivity at the *rf* level. The disadvantage of such a method of selectivity improvement is that the selectivity takes place late in the receiver processing chain. Therefore, when one is listening to a weak station, a strong station near in frequency can control the avc or overload the receiver stages.

Nonetheless, audio selectivity is easy to apply and can take the form of either an audio frequency peaking or notching type function. Audio peaking can easily be provided by a number of fixed frequency filter designs and numerous inexpensive units are available from surplus outlets. The disadvantage of the peaking approach is that most filters which are of any real use produce a "ringing" effect. The sound is unnatural and definitely very tiring if the filter is constantly left in the circuit without any provision for disabling it. Stations can also be lost when scanning a band unless tuning is done very slowly when using the filter on CW.

The notching type filter, on the other hand, is usable on both CW and phone. It does not cause any ringing effect and does not mask any signals when quickly scanning a band. A single notch filter can only eliminate the one frequency to which it is set, but when a receiver already has a good phone filter – such as the multiple crystal or mechanical types found in SSB transceivers – one notch frequency possibility seems to suffice in most qrm situations. Audio notch filters built around passive components only are fairly old, but their use has disappointed many operators because to achieve reasonable narrowness and high attenuation at the notch frequency, expensive capacitors were necessary and the filter could only be used in very high impedance circuits. The use of an integrated circuit amplifier with a notch filter in a feedback arrangement, however, produces a notch filter of very high Q using inexpensive components. It is neither critical as to circuit impedance nor does it introduce any overall circuit insertion loss.

Basic Circuit

The circuit of the IC notch filter is shown in Fig. 1. The actual notch filter consists of the bridged-T network – the ganged 50 K potentiometer and the two .05 mf capacitors. The circuit presents a very high attenuation at one frequency which is related to the time constants of the circuit. The frequency of maximum attenuation can be changed by either changing the value of the resistive or capacitive legs. As was mentioned, however, unless special precautions are taken, the bridged-T network alone will tend to produce a very broad rejection notch which is particularly unsuited to CW work. The integrated circuit, however, corrects this situation in the following manner: the input signal passes through the bridged-T network to one input of a differential operational amplifier (a Motorola MC1533 in this case). Feedback through R4 is coupled from the amplifier output back to the signal input point. The other amplifier input (-) then receives a combination of the original input and feedback signals. Since the nature of a differential amplifier is such that when the (+) and (-) inputs receive equal level signals, the output is zero, R2 and R3 are chosen such that this condition exists and infinite attenuation takes place at the notch frequency.

There may be situations where a somewhat broader rejection notch is desired with correspondingly less maximum attenuation at the notch frequency. This adjustment is provided by making R1 variable, as shown in Fig. 1. As the wiper arm on R1 moves from right to left, the feedback voltage around the operational amplifier decreases and the effective Q of the bridged-T network is reduced. Thus, if desired, R1 can be made variable and functions as a "Q" control. Otherwise, R1 can be a fixed value resistor and R2 is connected to the junction of R1 and R4 to produce a single frequency, deep notch audio filter. In this case, the only variable control would be the dual 50K ohm potentiometers which vary the time constant of the bridged-T network and hence the notch frequency. The dual potentiometer is capable of varying the notch frequency over about a 10:1 frequency range – 300 to 3,000 cycles approximately. This range should certainly suffice for most applications but, if desired, the range can be changed by using different (but equal) values of capacitance for C1 and C2.

The differential operational amplifier used may seem an unusual component to many readers. Although the basics of integrated circuits cannot be explored in this article, it should be realized that the integrated circuit used is only a multi-stage transistor amplifier packaged into a housing the size of the usual single transistor. The main feature that separates the differential amplifier from a conventional amplifier is its input circuitry. The "differential" input has a non-inverting (+) and inverting (-) input. A positive-going voltage applied to the non-inverting input produces a positive-going output voltage. The same voltage applied to the inverting output produces a negative-going output. Thus, the same value and polarity voltage applied to both inputs will produce no output.

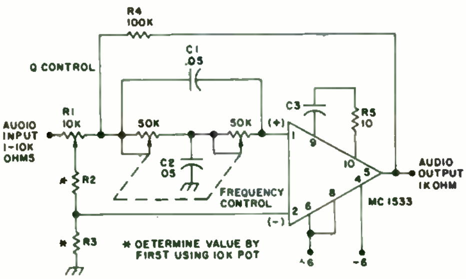


Fig. 1. Schematic diagram of the variable frequency and variable bridge-T notch filter. Various other IC units may be used besides the unit shown.

Fig. 1 shows the use of a Motorola MC1533 operational amplifier, but almost any similar unit will suffice. A number of inexpensive "surplus" IC operational amplifiers are available from such suppliers as Poly Paks, Lynnfield, Massachusetts 01940. Other units may differ in their voltage requirements and roll-off compensation needs (the RC network between pins 9 and 10 on the MC1533), but this information is usually supplied with the unit. It should be noted that a simple integrated circuit audio amplifier cannot be used; such units do not have differential input circuits.

Construction

There are very few precautions to be observed in constructing the unit because of the nature of its operation. One possible method of construction which the author explored is shown in Fig. 2. All of the circuit components are mounted on a piece of vectorboard which in turn is mounted on the rear potentiometer of the dual 50K ohm potentiometer. A piece of foam plastic material is glued between the underside of the vectorboard and the rear potentiometer to achieve the mounting.

The parts layout shown for the vectorboard in Fig. 2 need not be followed exactly, although it was the simplest which the author could devise. If an integrated circuit packaged in a dual-inline container is used (rectangular with 5-7 connections on each long side), the IC can be mounted on the left side of the vectorboard and R5 and C3 both placed under C1. No "Q" control is provided for in the parts layout shown, and R1 is a fixed value resistor. R1 need only be made variable, as shown in Fig. 1, if this feature is desired. The values of R2 and R3 can only be properly chosen by first using a 10K ohm potentiometer in their place (with the wiper arm going to terminal 2 on the IC) due to component value variation. The procedure is fairly simple. The wiper arm on R1 (if a variable unit is used) is first set towards the junction of R1 and R4. Then, using an input signal containing a frequency which the notch filter can reject, the dual 50K ohm potentiometer is adjusted for maximum rejection. Leaving this control set, the temporary potentiometer used in place of R2 and R3 is adjusted for complete signal rejection. At this point, the arms of the 10K ohm potentiometer are measured and replaced by equivalent value fixed resistors.

The resistors used in the notch filter can either be ¼ of ½ watt sizes. The capacitors need be rated no higher than the maximum value of the supply voltage used for a particular IC. The capacitors used in the bridged-T network should be of good quality to achieve the sharpest notch selectivity. Disc ceramic types are acceptable, although, if possible, low-loss types such as the Aerovox P123ZN series are preferred. The current demand from the power supply is in

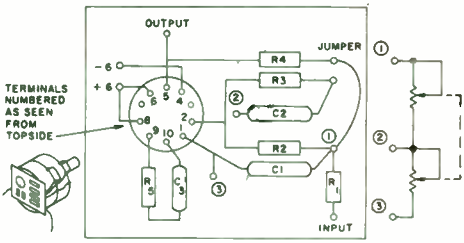


Fig. 2. No particular circuit layout is necessary but the components can so compactly be grouped on vectorboard that the entire circuit mounts on the back of the frequency control potentiometer.

the order of a few milliamperes, but the operating voltages should be obtained from a well-filtered source to avoid any possible hum problems.

The dual 50K ohm potentiometer used in the bridged-T circuit is a standard linear taper type. Although one can purchase such potentiometers from various supply houses, particular attention should be paid to being "cost-conscious" about this item. The author has found such potentiometers available for as low as 35¢ as compared to prices of $3 for similar units at regular catalog prices.

Mounting and Operation

The mounting or placement of the notch filter in a receiver or transceiver is fairly flexible. The unit can be inserted between almost any two audio stages. DC blocking capacitors must, of course, be used to prevent other than audio frequencies passing through the filter. Alternatively, the unit can be used completely external to a receiver or transceiver. The audio output from the receiver or transceiver can be taken from a headphone jack or from the loudspeaker terminals through a transformer (4-10 ohms to 1K ohm or more). The gain of the operational amplifier, in the latter installation, is more than sufficient to drive any pair of medium to high-impedance headphones.

Operation of the notch filter is extremely simple. When not in use, the filter is adjusted for maximum low-frequency attenuation. As qrm develops, the filter is used (in conjunction with tuning of the receiver bandpass) to eliminate the most severe interfering beat (on SSB or CW). The result is an almost complete attenuation of the interfering signal while still retaining the full fidelity of the desired signal. The difference between this method of qrm elimination and that which depends upon a severe reduction in bandpass to accomodate only the desired signal is quite startling in terms of fidelity and ease of tuning.

Summary

Many thanks are due to Herman Gelback, W7JPU, of the Boeing Company, who completely developed the original circuitry of the IC notch filter and who allowed the author to present this description of its operation. Herman has also suggested that besides the MC1533 IC, another suitable off-the-shelf IC would be the General Electric PA-230, which sells for just over $3

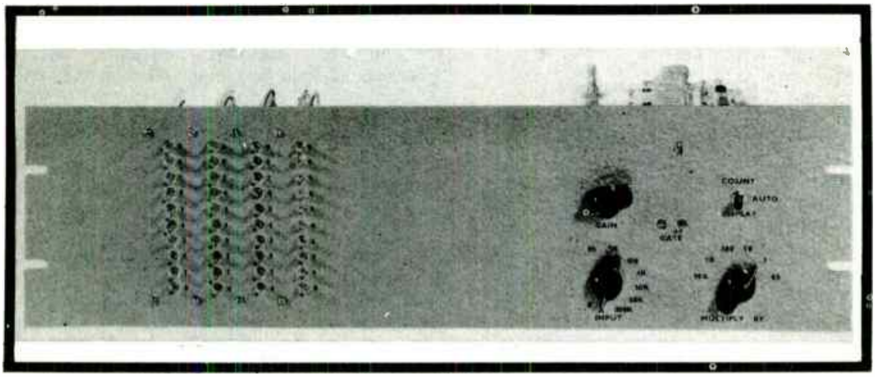
An Integrated Circuit Electronic Counter

Geo. W. Jones, W1PLJ

A digital frequency counter is a useful, though not common, piece of equipment in the ham shack. The writer built a counter many years ago using old fashioned vacuum tubes in order to place high in the ARRL Frequency Measuring Tests. The unit only worked up to 100 kHz, but was adequate for the intended purpose. The recent reduction in the prices of plastic encapsulated integrated circuits prompted the writer to see if a better unit could be built with transistors and integrated circuits. The result is a counter which will go up to 10 MHz and has every feature a ham could want, including direct decimal readout and completely automatic operation. The unit shown is useful not only during the ARRL FMT but also in everyday ham operation. During normal operation it is connected to the VFO of my transmitter-receiver setup and is set on the 100 Hz range, thereby acting as a digital "tuning dial" with 100 Hz divisions; a feature not found on any ordinary receiver or VFO. Later, when I go on RTTY, it will be useful for setting the transmitter frequency shift and aligning the receiver converter.

Principles of operation

This counter displays the frequency in decimal numbers so that the operator doesn't have to convert from binary to decimal. On the one-hertz multiplier range, the cycles of the input signal are counted for precisely one second, and the progress of the count can be watched on the neon lamps. The final count is then displayed for one second. The count period can be extended to any multiple of one second if greater than one-hertz accuracy is needed and, likewise, the display can be held for as long as desired. At the end of the display period, the counters are reset to zero and the process starts over again. On the 10-hertz multiplier range the same process is repeated five times a second, on the 100-hertz range, fifty times a second, etc. To avoid confusion on the ten-hertz and higher ranges, the neon lamps are not lighted during the counting period and are, therefore, seen only displaying the final count. On the 10-hertz range, the display blinks five times a second, but on the 100 Hz and higher ranges, it appears continuous and appears to change immediately if the input frequency



Front view of the integrated-circuit frequency counter. The neon counting decades are on the left, count controls are on the right.

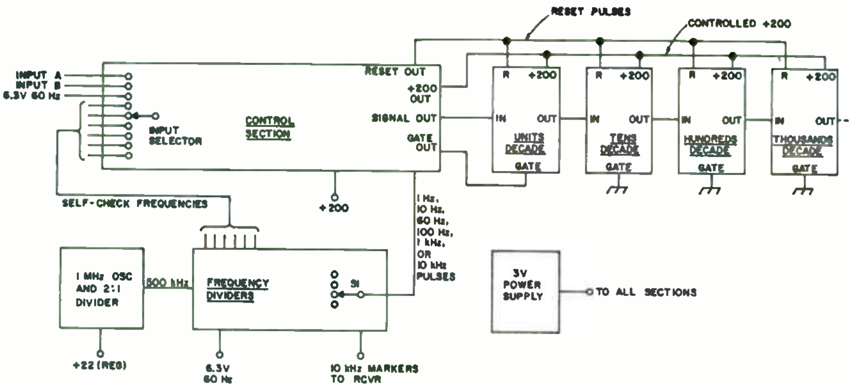


Fig. 1. Block diagram of the complete integrated circuit frequency counter. Any number of decades may be used, but for proper display, the units decade should be to the right, the tens decade to its left, etc.

changes. Therefore, it combines the convenience of an analog display with the accuracy of a digital display. The last digit in this case usually vacillates between two adjacent numbers because of the one hertz per gating period error inherent in a digital count.

The counter consists of three main sections. First, a frequency divider divides the signal from a 1 MHz standard down to 10 kHz, 1000 Hz, 100 Hz, 10 Hz, or 1 Hz, as required. A time base derived from the 60 Hz line could have been used but this would have limited the accuracy to 0.1% and would only have permitted the 10 Hz and 1 Hz ranges. This section also applies 10 kHz markers to the remainder of the frequency measuring setup. The 10 kHz pulses are rectangular in shape and have strong harmonics above 30 MHz. Therefore, they might as well be used as markers.

Second, a control section takes the desired time base frequency and turns on the units counter for the correct length of time. It also shapes the input signal, so that the units counter will accept it, turns on the high voltage supply for the neon lights during the display period, and supplies a reset pulse to all counting decades at the end of the display period.

Third, the counter proper consists of as many counting decades as the builder desires, one for each digit to be displayed. The units decade is gated by the control section and only counts pulses when the control section wants it to. For each ten pulses the units decade is allowed to count, one is passed on to the tens decade, likewise for each ten pulses the ten decades receives it passes one on to the hundreds decade, etc. The decade counters, after the units decade, are not gated since they only receive pulses if the units decade is supplying them. Although the decades count by binary flip-flops, suitable feedback circuits make them count in decimal instead of binary. A decoding network and ten transistors allow one of ten neon lamps on the decade to be turned on to display one digit of the measured frequency. Each decade can also be reset to zero by a reset pulse from the control section.

Digital integrated circuits

The counter uses RTL integrated circuits because of their low cost. These have been described in 73 magazine both in articles and integrated circuits, and in two excellent articles about IC electronic keyers; therefore, they will only be described briefly here. The reader who is not familiar with RTL circuits should review these references before trying to understand the counter in detail. He might also find it advisable to build the "Kindly Keyer", before he builds the counter, as the writer did. Although the counter could probably be built and made to work by just following the diagrams, a previous knowledge of RTL circuits, gained by building a simpler device, will help in trouble shooting.

Oscillator and frequency dividers

A 1-MHz crystal oscillator is used as the main frequency standard at W1PLJ. One MHz is used instead of the usual 100 kHz because a 1-MHz crystal gives better stability than a 100-kHz crystal if one wants to pay a reasonable price for the crystal. This is probably because the 1-MHz crystal can be AT cut. The oscillator and a divider to 500 kHz are mounted in a separate box so that the oscillator can be kept on all the time for better stability. Also, 500 kHz can be used for other purposes including future plans to use it to synchronize a phase-locked oscillator for the first conversion of the receiver. If the builder already has a frequency standard, it is not necessary to build another crystal oscillator for the counter—the existing one can be worked in easily. Conversely, if the builder is interested in frequency measurement but does not yet want to build the counter, he can build the oscillator and the dividers down to 10 kHz and at least have markers for his receiver. The oscillator and first divider are shown in **Fig. 2.**

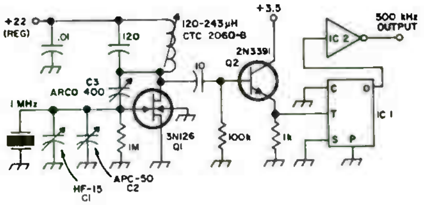


Fig. 2. The oscillator and 2:1 frequency divider used with the decimal counter. This unit was built into a separate box and may be used for obtaining markers as described in the text. Although the FET is a Motorola 3N126, an MPF-105 is less expensive and would probably work. IC1 is a Fairchild 923 or one-half a Motorola MC-790-P; IC2 is a Fairchild 900 or one-half of a Motorola MC-799-P.

The remainder of the frequency divider section, Fig. 3, consists of 2:1 and 5:1 dividers. The 2:1 dividers are simply J-K flip-flops; the 5:1 dividers are J-K flip-flops with an RC network and inverter on the set input which only allows every fifth input pulse to produce an output pulse.

The 5:1 divider can be best understood from the diagram and waveforms of **Fig. 4.** Without an input signal, the inverter input is held high by the connection to positive voltage thru R_1. The inverter output is, therefore, low so that low appears on the set input of the flip-flop. If the O output of the flip-flop is initially high, the first negative going transition on the toggle input will make it go low. This change will be passed on to the inverter through C_1 and this will make the set input go high so that the O output cannot go low again when more input pulses come in. C_1 will charge through R_1 and, after a delay, the inverter output and the set input will go low again so that the flip-flop can respond to an input pulse. If the divider is adjusted correctly, it will pass every fifth input pulse. Other division ratios can be obtained, and maybe it would work with a division ratio of ten, but the ratio of five makes the division ratio very stable. In fact, it does not go out of adjustment for a change in the supply voltage from 3.0 to 4.0 volts.

The first three 5:1 dividers are identical except for time-constant values. The output of the 10-kHz divider is fed through a buffer to the station receiver and frequency measuring equipment. The markers are very strong through 30 MHz, the limit of the author's receiver. If the receiver calibration cannot be trusted to 10 kHz, the 50-kHz test button, shown in dotted lines in **Fig. 2** and not used by the author, can be provided. Pushing this button makes the 50-kHz markers louder and the other 10-kHz markers turn into 25-kHz markers. The counter proper does not read correctly while this is being done, but this doesn't matter since identifying the markers is done separately from making the final count.

The divider form 500 Hz to 100 Hz uses a discreet high-beta transistor instead of a gate, so that a higher resistor value and, therefore, a smaller capacitor value can be used. The dividers to 10 Hz and 1 Hz use decade dividers, with four J-K flip-flops in order to avoid even larger capacitors. This type of circuit could have been used for all the dividers and would have eliminated the need to adjust the dividers. The circuit of these dividers will be described in the section on the counting decades which use the same circuit.

The switch, S_1, selects the divider frequency whose period is equal to the desired gate time and is calibrated in factors, by which the counter reading must be multiplied, rather than in gate time. The X60 position takes the time base from the ac line instead of the dividers, and is useful in adjusting the dividers. For example, to adjust the divider whose output frequency is 50 kHz the input switch is set to 50 kHz, the multiplier

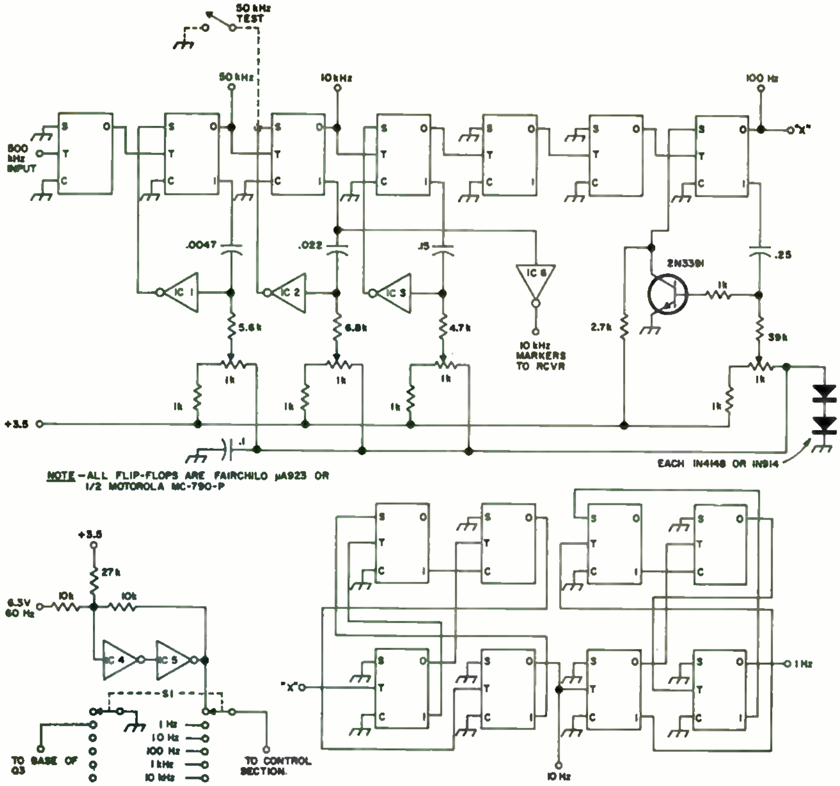


Fig. 3. The frequency dividers used in the IC counter. Integrated circuits IC1 through IC5 are one-half of Fairchild 914's or part of Motorola MC-789-P or MC-724-P; IC6 is a Fairchild 900 or one-half a Motorola MC-799-P.

switch to 60 and the counter should read 833. This reading will jump around a bit, due to instability in the ac line frequency, but the reading for the 10-kHz divider will only vacillate between 166 and 167.

Control section

The input selector switch, S_2 (Fig. 5), selects the desired input which can be either a signal input for measurement, or one of the divider outputs for self checking. IC_8 and IC_9 can be regarded either as an amplifier with positive feedback, or as a flip-flop. They make the signal into a rectangular wave with sharp edges and reject noise which may appear on the input signal. At any instant of time, either IC_8 or IC_9 will conduct, but not both at once, because the one that is conducting turns the other one off. The positive half cycle of the input signal will make IC_8 conduct and once it is turned on, the high output from IC_9 will supply holding current through R_4 to keep it on.

The negative half cycle will then overcome this holding current and turn off IC_8 whereupon the holding current will be removed and IC_8 will continue not to conduct. A small amount of noise riding on the input signal will not be able to overcome the holding current and will, therefore, not make the circuit change state. The resulting rectangular wave is fed to the units decade at all times and the necessary gating is done in the first J-K flip-flop of the units decade. Provision for gating already exists in the J-K and it is simpler to use it than to do the gating in the control section.

The remainder of the control section can exist in either of two states, count or display. We will discuss these quiescent states before we examine how it gets from one to the other. In the count state the 1 output of IC_7 is high and the O output is low. If S_1 is not in the X1 position, the high 1 out-put of IC_7 turns on Q_3 and turns off Q_4, thereby turning off the neon lamps. The low O output of IC_7 goes to the gate input of the units decade and allows it to count. It also turns off Q_1 so that the "gate on" light will be illuminated and it per its the "gate pulse" light to be turned on if a gate pulse is present. The opposite conditions exist in the display state. Power is applied to the neon lamps through Q_4, and a high output is supplied to the gate so that further counting cannot occur, and both Q_1 and Q_2 are turned on so that the two gate lamps are shorted and not illuminated.

To understand how we change state, assume we are on display and S_3 is in the automatic position. IC_1 and IC_2 form a monostable multivibrator which supplies the reset pulse and the trigger for IC_7. The positive-going edge of the rectangular wave from the frequency dividers turns on IC_1 momentarily and this makes the output of IC_2 go high. Furthermore, this holds IC_1 on until R_2 charges up C_2 again, whereupon the output of IC_2 goes low again. The result is a short pulse which occurs once every timing period. Since we are on display and automatic, this pulse will be passed by IC_3, IC_4, IC_5, and IC_6, inverted each time, and appears as a high pulse to reset the counters. The trailing edge of the pulse from IC_2 will toggle IC_7, putting us in the count mode. The next pulse from IC_2 will not reset the counters because IC_4 has a high input from IC_7 and, reset can only occur if all three inputs to IC_4 are low. The trailing edge of the pulse still toggles IC_7, however, and we are in the display mode; displaying the number of input pulses that occurred between two timing pulses.

The switch, S_3, is used if you want to count, or display, for a multiple of the basic timing period. The switch itself does not switch the counter to display or count, since only the timing pulses can be allowed to do this; rather, it prevents the counter from going into the other state. The "display" position of this switch is useful if you have just made a critical count and want to hold it a few seconds to make sure of writing it down correctly. It is also useful if the circuit for blanking the neon lamps isn't working or isn't yet built and you want to make a reading on the higher ranges. In this case it is difficult to read the display on the automatic position because you will see both the counting and the display, but placing S_3 on "display" will hold the last count and allow you to read it. The switch can be thrown to "display" either during count or during display. In either case, a timing pulse will still switch IC_7 from count to display at the right time, but the next timing pulse will not put it back on count due to the high level on the clear input. Also, the counter will not be reset due to the high input of IC_4 which will hold its output low.

The count position of S_3 is normally used only on the X1 position of S_1, and is used when you want a gate time of several seconds for an error of less than one Hz. This is useful in the ARRL Frequency Measuring Tests where it is desirable to use a 10-second gate time in order to obtain an accuracy of 0.1Hz. With this arrangement,

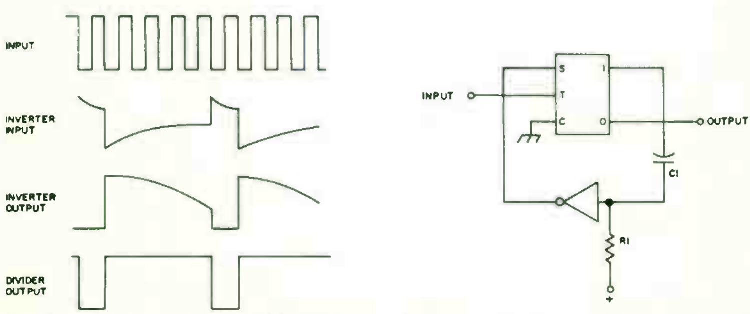


Fig. 4. The basic 5:1 frequency divider using a J-K flip-flop, and RC circuit and an inverter, along with the waveforms.

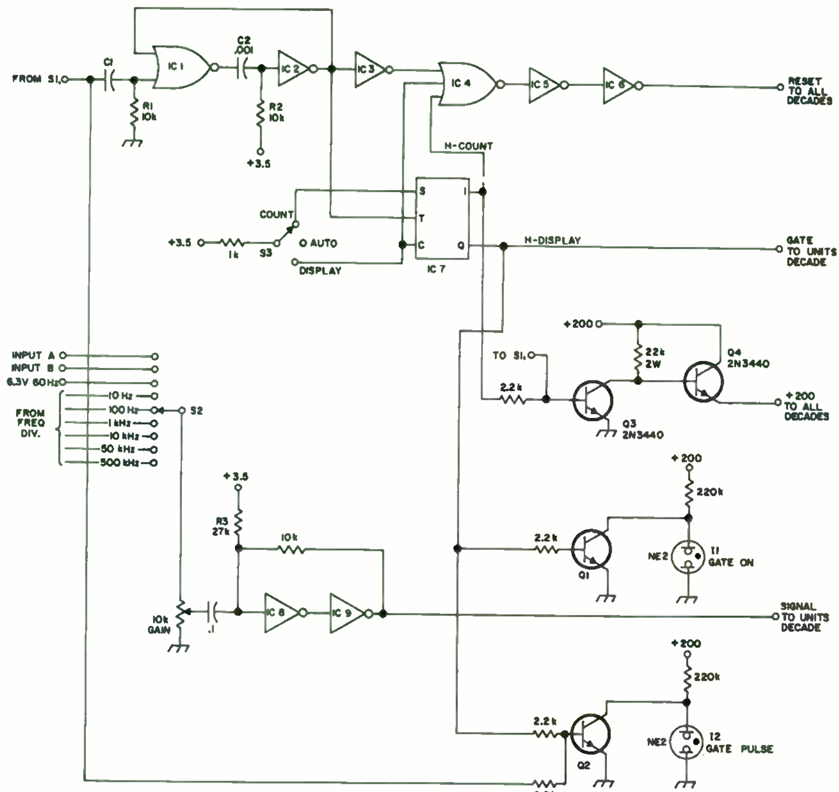


Fig. 5. The control section of the digital frequency counter. IC1 is a one-half a Fairchild 914, one-fourth a Motorola MC-724-P or one-third a Motorola MC-792-P. IC4 is one-third a Motorola MC-792-P. IC2, IC3, IC5, and IC9 are one-sixth of Motorola MC-789-P, one-fourth of Motorola MC-724-P, or one-half of Fairchild 914. IC6 is a Fairchild 800 or one-half a Motorola MC-799-P. IC7 is a Fairchild 923 or one-half a Motorola MC-790-P. Q1 and Q2 are 2N3877's or Poly Pak 2N1893's.

if you start a ten second run and the signal starts to fade, you can stop the test at the next timing pulse by throwing the switch to display and still obtain a meaningful reading. To make a ten second run, you start with S_3 on display, and throw it to count when everything is ready. The next timing pulse will put you in the count mode, but the next one will not put you back on display.

Each timing pulse will flash the "gate pulse" lamp once, and after it has flashed ten times, you put S_3 back on display. The next pulse will put the counter on display and you will be able to read the frequency in tenths of hertz. With a little practice, you will find that running a multiple second count is much easier than reading about it.

In wiring the counter it should be remem-ered that the supply to the neon bulbs is a 200-volt square wave because of the lamp blanking circuit, and also, the collectors of Q_1 and Q_2 (Fig. 5) have 60-volt pulses on them since they turn on neon lamps. Both of these must be kept away from the inputs to the IC's; otherwise, erratic operation will result. In particular, the 200-volt lead to the counters must not be cabled with the signal and gate inputs to the counters and the leads to I_1 and I_2 must be kept at least an inch away from the leads of S_3. If the counter shows any erratic operation which cannot be easily explained, the blanking circuit should be disabled by grounding the base of Q_3 so that the lamps are on continuously, and I_1 and I_2 should be shorted to ground. A test can then be made to see if the trouble still exists. Except for these precautions, no other difficulties should be encountered with the unit.

Counting decades

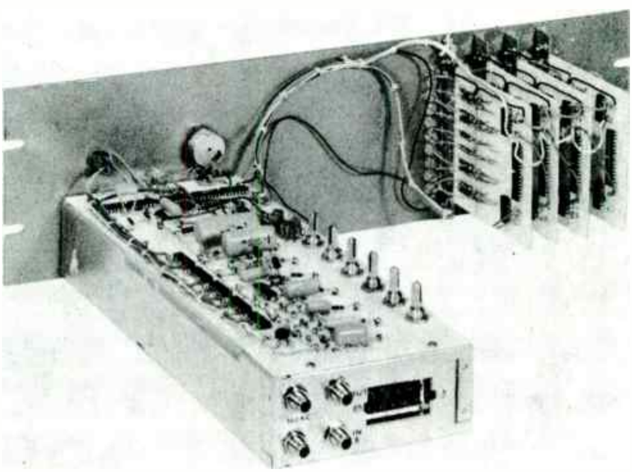
Fig. 6 shows the circuit on one counting decade, including neon lamp drivers. The gate input on the units decade is connected to the gate output of the control section, but the gate inputs on the other decades must be grounded since each must accept any pulses put out by the preceeding decade. The actual counting is done by four J-K flip-flops and, with the help of the table shown, the reader can follow the count as an interesting exercise. The input pulse following the ninth count makes the decade go back to zero and passes a negative transition on to the next decade making it count once.

IC_5 through IC_{12} are needed to amplify the voltage output of the J-K flip-flops. The J-K's give only one-volt output with light external loading due to the fact that they internally load their own outputs. This was not found sufficient to drive the resistor gates used for the neon lamp drivers. An inverter, however, gives almost full supply voltage when lightly loaded and drove the resistor matrix satisfactorily.

It is necessary to use discrete transistors to drive the neon lamps at the present state of the art, but these are not expensive, expecially if Poly Paks* 2N1893s are used. The transistors are used as shunts scross the lamps. This makes gating simpler and also limits the voltage across each transistor. For a given count one lamp must be on and the other nine off. The driver for the desired lamp must have low level on all its inputs so that the transistor will not conduct, allowing the lamp to light. The other nine drivers must have high level applied to at least one input; this will be sufficient to extinguish the lamp, regardless of what appears on the other inputs. The gating of the lamps could have been done entirely with IC's but this method was found to be simpler and cheaper, at least at the present state of the art.

Count	A B	B A	C D	D C	E F	F E	G H	H G
0	H	L	H	L	H	L	H	L
1	L	H	L	H	H	L	H	L
2	H	L	L	H	L	H	H	L
3	L	H	L	H	L	H	H	L
4	H	L	L	H	H	L	L	H
5	L	H	H	L	H	L	L	H
6	H	L	H	L	H	L	L	H
7	L	H	L	H	H	L	L	H
8	H	L	L	H	L	H	L	H
9	L	H	H	L	L	H	L	H

Table 1. Truth table showing the proper levels on each of the logic lines of the decade counter in Fig. 7.



All of the count-control circuitry is mounted on the large chassis to the left. The small perforated boards on the right each contain one decade counter.

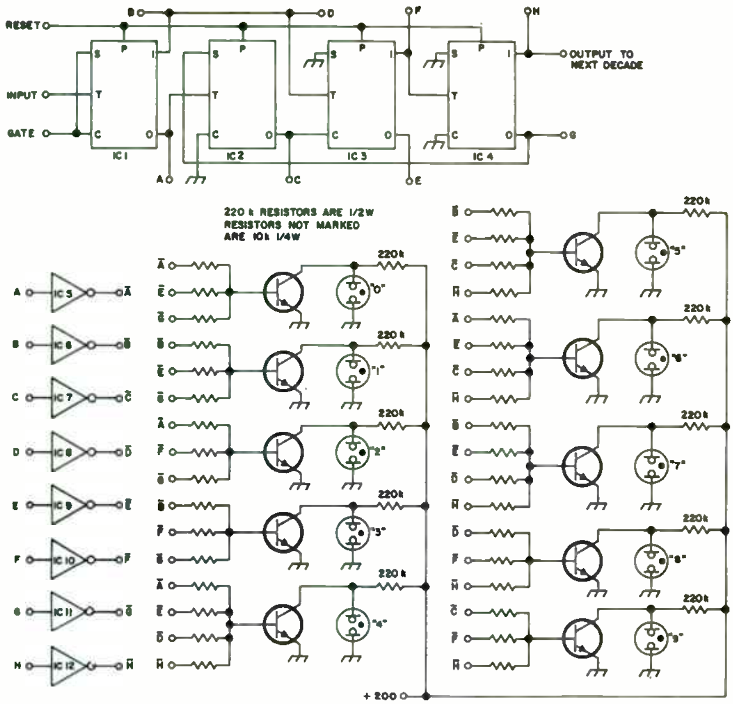


Fig. 6. A typical counting decade. In this circuit integrated circuits IC1 through IC4 are one-half Motorola MC-790-P's or Fairchild 923's. IC5 through IC9 are one-sixth Motorola MC-789-P's, one-fourth Motorola MC-724-P's or one-half Fairchild 914's. All transistors are 2N3877's or Poly Paks 2N1893's. All neon lamps are NE-2's.

In testing the decades, +200 volts must not be applied unless all transistors, which are in place, have neon lamps across them. Otherwise, if a transistor is not conducting, the collector voltage rating will be exceeded since there is no neon lamp limiting the voltage. Also, if +200 volts is applied to a decade but +3.5 is not, all lamps should light since the logic circuitry only acts to short out the undesired lamps. No harm is done by this and it is a quick way to check the lamps and driver transistors. If a lamp does not light under this condition, its driver transistor should be suspected first.

Power supply

The counter, as shown in Fig. 6, requires about one ampere at 3.5 volts and 40 mA at 200 volts. Neither supply needs to be regulated and the IC's will work on any voltage from 3.0 to 4.5 volts, although 3.6 ±10% is recommended by the manufacturer. The power supply used by the author is shown in Fig. 7. An 8-amp transformer was used because it didn't cost much more than a 2-amp one in the same series. The 2-amp unit would probably work and would save space and weight. For the 200-volt supply, anything from 150 volts on up would work, although with anything much over 200 volts, the 220K collector resistors must be increased or a dropping resistor must be provided. If this voltage is taken from a supply powering other equipment, it must be remembered that the current drawn will be a 40 mA peak square wave at 5, 50, 500, or 5000 Hz which may cause a buzz to be heard on the other equipment.

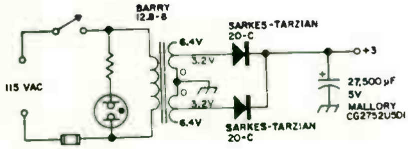


Fig 7. Three-volt power supply for use with the integrated circuit frequency counter. A truth table showing the proper levels on each logic line are shown in Table I.

Construction

The individual counting decades are built on See-Zak MM-492 boards and the remainder of the unit on a See-Zak MM-512 board mounted on See-Zak R-25 and R-212 rails. See-Zak M-25 terminals are used for the larger components, including the Fairchild IC's. The hole spacing on these boards is 0.2" whereas the Motorola IC's require 0.1" spacing; therefore, seven extra 1/16 inch holes must be drilled for each Motorola IC inbetween existing holes. Connections to the Motorola IC's are made with #26 bare wire covered with Teflon spaghetti. No other construction details are given since the writer is more interested in circuitry than packaging and other builders will probably have ideas of their own. The use of printed circuits would be ideal.

. . . W1PLJ

*Poly Packs, Post Office Box 942A, Lynnfield, Massachusetts 01940.

References

1. Skeen, "Low Cost Precision Frequency Measurement", *QST*, January, 1965, p. 32.
2. Brassine, "An Electronic Counter for Amateur Use," *73*, December, 1966, p. 20.
3. Pickering, "The Micro-Ultimatic", *73*, June, 1966, p. 6.
4. Daughters, "The Kindly Keyer", *73*, July, 1966, p. 46.
5. Lancaster, "Using New Low Cost Integrated Circuits", *Electronics World*, March, 1966, p. 50.
6. Olson, "Micro-Logic for Non-Logical Users", *73*, June, 1967, p. 58.
7. Suding, "An Integrated-Circuit Frequency Counter," *73*, November, 1967, page 9.

100 KHZ Thin-line Pulse Generator

Jim Ashe, W2DXH

Digital integrated circuits are an entirely new kind of electronic component. These finished, ready-to-go devices contain complex transistor circuits in tiny, convenient packages. Until recently they were too expensive for one-off projects, but a burgeoning market and competition between manufacturers have brought some prices to the dollar apiece level. In fact, very good digital IC's are now available on the surplus market.

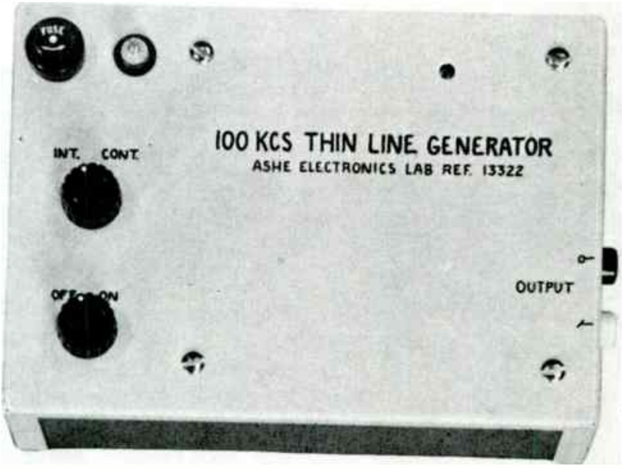
Perhaps because they are so new, it is hard to see applications for digital IC's outside the computer and industrial control scene. It takes a little while to adjust mental perspective, too, before their input-output characteristics begin to seem natural. Yet, there are applications for them which are not very difficult. For instance, how about a frequency standard?

Ordinary 100-kHz frequency standards are usually audible up to a few tens of megahertz. A good one might be usable at 50 MHz. The circuit described here uses a dual NAND gate to generate a 100 kHz signal whose harmonics are usable to 432 MHz or higher. And it can be built without benefit of special instruments and knowledge.

The thin line pulse

One rather surprising result of higher mathematics is that all repetitive signals are composed of harmonically related sine and cosine waves. For example, the familiar square wave is composed of a fundamental frequency, which sets its basic repetition rate, and of odd harmonics only of its fundamental, which contribute to its square corners. If the harmonics' amplitude or phase relationship is upset, the square wave is distorted. This feature makes the square wave very useful for amplifier testing, but its harmonic content is not very good for frequency standard applications.

Now suppose that we start adding up signals of F, 2F, 3F, and so on, phased in so that they all reinforce each other once per cycle. Let's say thay are all the same amplitude. What would we get? See Fig. 1A.



External view of the 100 Hz thin-line generator.

The five equal amplitude sine waves peak simultaneously at the beginning of the fundamental's cycle. Everywhere else, until near the end of the cycle, they are more or less out of phase. Trying to see what will happen, we try adding the first two frequencies. **Fig. 1B**, the result, might suggest something to a mathematician.

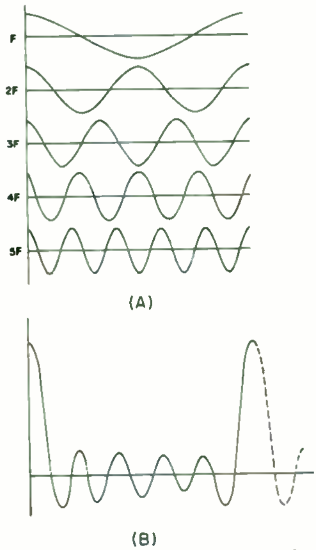


Fig. 1. Five sine waves (A) and the waveform as a result of point-by-point addition (B).

As the number of frequencies is increased, their amplitudes tend to average to zero everywhere except at the beginning of the cycle. Here, they all add up to a short, sharp pulse. It follows that a short, repetitive, one-sided pulse should contain odd and even multiples of the fundamental frequency.

An ideal thin line pulse has infinite frequency content.* No real signal could meet this spec, but a fast digital IC can produce a very workable approximation. **Fig. 2** shows a Tektronix 545A view of the generator output and tests with other scopes indicate the real pulse has better rise time and sharper corners than shown here.

If this pulse is viewed on a low-performance service variety scope, its appearance will be greatly changed. There will be an apparent loss in amplitude, since the pulse occurs and terminates before the slow circuitry can properly respond. The apparent duration is increased, also because of the slower viewing circuitry. And the fast pulse may excite circuit resonances, so that the thin line pulse appears as a damped oscillation. But these problems do not interfere with constructing the generator, because the very simple NAND gate circuitry contains no critical elements or adjustments.

How it works

There are four circuit sections, shown in **Fig. 3**. A 100-kHz crystal-stabilized oscillator sets the basic frequency, and a dual NAND gate circuit converts the oscillator output to a thin line pulse. A 1-Hz astable generates the output marking signal. A 6 volt dc power source is provided by a voltage doubler zener-regulated supply.

Multivibrator oscillators are not ordinarily very stable frequency sources. But if the oscillator is designed to run slightly below required frequency, and an appropriate crystal is connected between transistor base terminals, oscillations are stabilized at the crystal frequency.

The crystal does not change the multivibrator's style of operation. It synchronizes the astable to its own frequency, by triggering the **OFF** transistor into conduction shortly before normal RC turn-on. The output is a squarish wave with good fall time, but a long rise time as shown in **Fig. 4A**.

In passing through the first NAND gate the pulse is squared up and becomes slightly unsymmetrical. See **Fig. 4B**. A differentiating network, C7 and R11, converts the square wave into the pulses shown in **Fig. 4C**. These pulses, applied to the second NAND gate, reappear as the thin line pulses shown in **Fig. 4D**.

Since one CW signal sounds just like another and there may be several in the vicinity of a check point, a marker feature is required. This is provided by the 1-Hz astable, which paralyzes the second NAND gate part of the time. Its base bias resistors are unequal, giving a distinctive duty cycle to the output signal. A switch disables the astable if a continuous signal is required. **Fig. 5** shows the output when the second astable is operating: the output is locked in the up condition during half of each 1-Hz astable cycle.

Sometimes an astable oscillator will refuse to start oscillating when it is turned on. It does not start because both transistors are in saturation. This reduces loop gain so that available noise cannot be amplified around the loop. It would never start without some strong, outside interference.

*Smith, *Applied Mathematics for Radio and Communications Engineers*, Dover Publications, 1961.

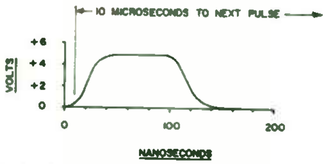


Fig. 2. Real circuit output as seen by a Tektronix 545A oscilloscope. A faster scope shows shorter rise-time and sharper corners.

A pair of diodes, D1 and D2, provide a reliable remedy. The diodes are arranged so that base bias must come from whichever collector is at the higher voltage. If both transistors are in saturation, their collectors are at perhaps 1 volt, which cannot provide enough base current to keep the transistors in saturation. This contradictory situation does not arise in the real circuit, which starts reliably.

Additional diodes, D5 through D8, appear in the base circuit of the 1-Hz astable. These are protective diodes. The collector swing at turnoff of about 5 volts is conveyed powerfully to the opposite base through the large coupling capacitors C5 and C6. The reverse B-E breakdown voltage of these transistors is not known, so the diodes are provided to prevent the turnoff voltage exceeding 2 volts or so.

DC power for the Generator circuitry comes from a voltage doubler supply based on a low-current filament transformer. Its design is conventional, but a large capacitor, C12, is provided across its output to minimize noise on the supply line. The supply could be replaced with some batteries, shunted by a 50 μF or larger capacitor to absorb transients. The original breadboard ran very well, powered by four flashlight batteries.

Construction

The generator is built in a Premier #PMC 1008 3x5x7 inch heavy aluminum box. Its top cover was refinished in light green enamel, and four ⅜ inch grommets in the bottom piece serve as protective feet.

Inside the box, the 6.3-volt transformer and cheater cord connector are mounted on

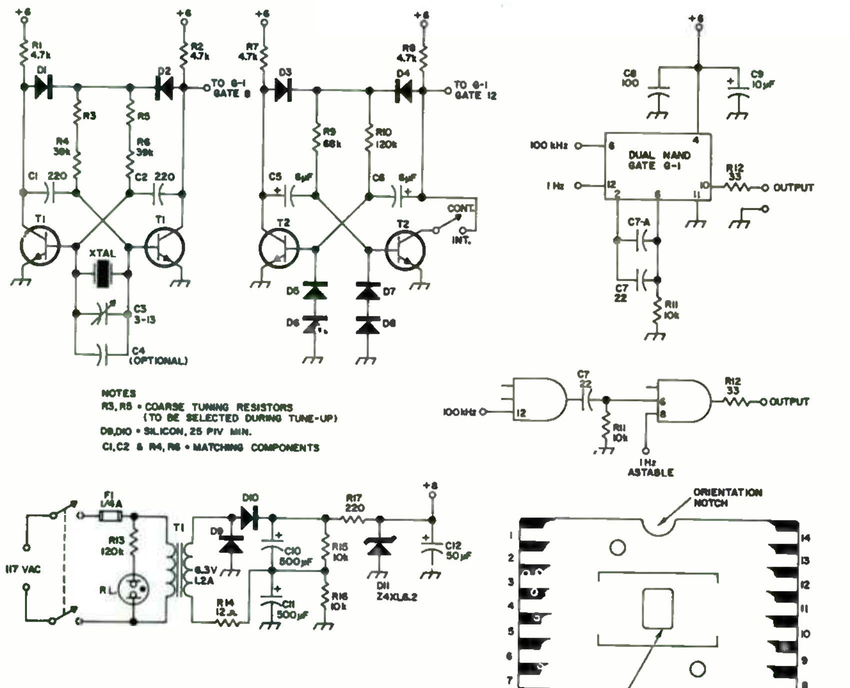


Fig. 3. Schematic of the 100 Hz thin-line generator.

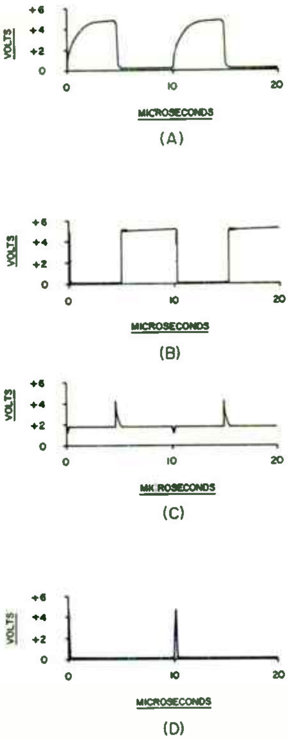


Fig. 4. Signals at four critical points in the generator, as displayed on a Tektronix 545A oscilloscope. They are shown in time coincidence.

the left-hand wall. A pilot lamp, fuse, and two switches are mounted on the horizontal panel, at the extreme left. This leaves just enough open space for the two circuit boards which occupy most of the box. Two banana jack output connectors are placed on the right-hand side, just below the panel.

The circuit boards are cut to 4⅛ x 5 inches, from Vector 5/32 inch pattern A stock and mounted parallel to the panel. The upper board is spaced an inch from the panel, and carries both astable oscillator circuits. The other board is mounted one half inch below, and carries the digital IC and the power supply circuitry. Assembled, the two boards make a sandwich with wiring sides together.

Both boards are mounted on the same four centers. These are through the second hole diagonally inward from each corner. The 1 inch 6-32 internally threaded spacers are modified by adding a short length of 6-32 threaded shaft to one end of each, simplifying assembly.

Component assembly on the boards is largely a matter of plugging in Vector T9.4 lugs. The finished product looks much better if some thought is given to facing the lugs in one of two directions. Mounting and transistor holes should be drilled and reamed to size before installing lugs.

The general arrangement puts all wiring on one side of the board and practically all components on the other side. This approach seems a little inflexible but is straightforward and looks good.

Possible board orientation problems may be overcome by working out a handling and wiring procedure that doesn't require constant reference to actual components. A good approach assumes that the board is only turned over an imaginary hinge at its bottom edge, so that top down when one side is up becomes bottom up when the other side is down. This preserves left-right relationships. Another useful convention is that all supply wiring goes to left-hand end of components.

Wiring is carried out one network (plus supply lines; ground lines; interstage lines, etc.) at a time, with prearranged color coding. Bare wire goes for short runs and where there is no chance of a short. Solder each lug when convenient. #22 solid wire fits the T9.4 lugs well, but flexible stranded wire is used for the four lines from one board to the other.

Transistors precede other components into the board, because they are convenient position markers. They are placed in their

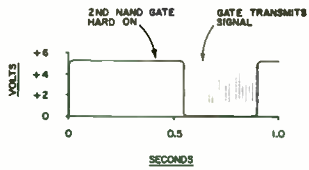
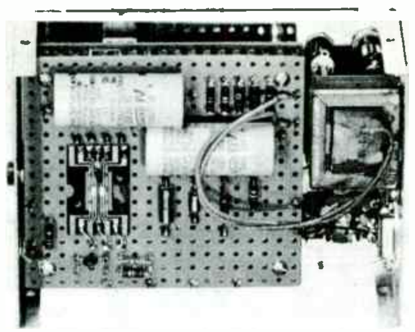


Fig. 5. The second NAND gate locks in its up position part of the time to produce an intermittent output.

mounting holes in the board from the component side, and their leads brought to the T9.4 lugs.

Then the other components are mounted on the boards.

product. Diode and electrolytic capacitor mounting polarity should be double checked. The T9.4 lugs may need a little bending before they will take a good grip on the



Inside the assembled thin-line generator showing the component side of the power supply and IC board.

components, but no component soldering is done until everything is installed.

Trimmer capacitor C3 is mounted on its tabs just under the top panel. Then a small screwdriver access hole is drilled over it in the panel, before painting, for vernier frequency adjustment after final assembly.

Certain components are matched before installation. An ohmmeter and a capacitor checker will do a satisfactory job of selecting C1 and C2, and R4 and R6, for equal values. These components are chosen alike for best symmetry of the 100-kHz oscillator operation. It might be good planning to leave these components unsoldered until tuning is completed, but everything else can be soldered to the board at this point. Note that the R3 and R5 sites do not get resistors until later.

Two optional capacitor sites are included. These are for C4, an additional and probably unnecessary padder across the crystal; and C7A, which can be added to increase the width of the thin line pulse.

Apparently, the digital IC comes in a specially designed package for testing before use. To mount the IC, solder a ⅜ inch piece of #22 wire in each of the T9.4 lugs carrying supply and signal voltages to the IC. Place the IC between the two rows of lugs, bend the wires against the proper terminals, and solder. No other mounting is required.

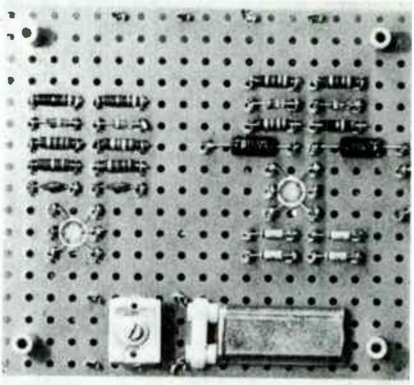
The original breadboard showed a lot of transient noise in its supply circuit. This originated from the IC, which was trying to get big chunks of current to manufacture pulses. Since the IC cannot deliver frequencies not available from the supply lines, very careful bypassing is indicated.

High-frequency bypassing consists of C9, a .01μF disc ceramic capacitor across the IC supply terminals on the wiring side of the board, and C10, a 100 picofarad capacitor soldered directly between supply terminals on the IC. The capacitor leads are provided with spaghetti insulation and placed for minimum open space between the capacitor leads and the IC's supply leads.

Testing before final assembly is very easy, because the odd appearing board layouts go together giving a structure that opens out like a book. The hinge is the four leads between boards. Leave transformer leads long, so that the circuit may be tested well free of its cabinet.

The upper half of the Premier box is prepared by a powerful cleaner which removes its original paint. After thorough removal of the cleaner, the metal is roughened with wet sandpaper, rinsed in vinegar solution and then clear water, leaving a very good surface that does not require priming for excellent paint adhesion. Watch out for greasy fingerprints.

Rustoleum #868 Green applied from a convenient spray can gives a fine finish.



View of the component side of the astable oscillators board.

Follow instructions on the can. After drying, the fresh, clean enamel will take waterproof India ink, applied with a Leroy drafting pen. When the ink is thoroughly dry, a final coat of Rustoleum #717 Clear finishes the job. The enamel is soft at first, but hardens into a coat durable in normal lab use.

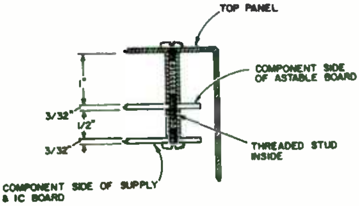


Fig. 6. Mounting dimensions and spacer assembly diagram.

Table of special parts

Crystal: 100 kHz parallel resonant 32 pF. shunt capacitance normally designed quartz crystal.

The following parts were obtained from Solid State Sales, P. O. Box 74, Somerville, Mass. 02143.

- T1 & T2: 2N2060 type dual NPN transistor
- D1, D2, D3, D4: fast point-contact Germanium diodes coded 1N59
- D5, D6, D7, D8: fast point-contact Silicon diodes marked S284GM
- G1: surplus digital integrated circuit Solid State Sales type G1. (comes with data sheet)

Tuning up

The generator should be zeroed to frequency before installation in its case. This is a two-step process. First, the 100-kHz astable base resistances are adjusted by choosing resistors for R3 and R5 to bring the oscillator frequency within trimmer range of 100 kHz, perhaps a few hundred cycles high at 15 MHz. Then the trimming capacitor brings the frequency to accurate coincidence with WWV.

To roughly zero the generator, set the trimmer capacitor, C3, at minimum capacitance. Identify WWV on a short-wave receiver, and tune around a bit to familiarize yourself with what's happening in the vicinity. It would be nice if things are fairly quiet.

Then put 4.7k resistors into the astable board at the R3 and R5 sites, turn on the generator, and look around for the signal. Depending upon actual values of C1 and C2, the signal may be on either side of WWV but is likely to be on the high side. If so, try again with resistors one size larger, which will lower the frequency. You should shortly find resistors that bring the frequency near enough to WWV for final zeroing with the capacitor. Verify tuning range on both sides of WWV.

Correct values for R3 and R5 may be approximated very quickly if a good triggered scope is available. Try selecting resistors for a period of 11.4 microseconds with the crystal removed.

Using the thin line generator

A breadboard test showed that (as might have been expected) there should be some way to distinguish generator signals from other CW signals. The continuous/intermittent feature provides the marking, and once the correct signal is located the generator can be switched to "continuous" for accurate work.

At low frequencies, the generator output and behavior resembles a conventional 100-kHz standard. The signal simply is not as strong. A greater difference appears at higher frequencies: the original model yields an audible beat note at 80-MHz from a diode mixer through an inexpensive audio amplifier. And another test shows a usable signal at 432 MHz: the 4,320th harmonic.

Some connection to the receiver or other detector is required. This is a natural consequence of a circuit design that puts the signal where it belongs, rather than spraying it all over the lab. A few picofarads coupling capacitance is sufficient at all frequencies.

Perhaps this circuit can be used for purposes other than a frequency standard. Its moderate amplitude but wideband output should be ideal for detecting changes in receiver sensitivity over a broad tuning range. In fact, with a little decoupling of the input leads and provision of a coax output connector the generator should do well as a stable, reliable small-signal source. A piece of adjustable waveguide-below-cutoff would make an excellent attenuator for work not requiring exact measurements. Another thought that occurs is possible further development by provision of some arrangement for detecting which harmonic is actually being heard.

Using the First Ham Integrated Circuit

Rbt. A. Hirschfield, W6DN6

For several years, 73 has been a leader in publishing articles on ham applications for currently available digital and analog IC's. Until now, however, no IC's had been produced specifically for two-way-radio use. At least one IC manufacturer, recognizing the need for such circuits, in the potentially large commercial, military, and amateur market, is now aiming a major development effort toward communication "subsystems on a chip", and it is expected that other manufacturers will follow. Besides the direct benefits of improved performance, decreased size, and lower component costs, which will reach amateur radio through commercially-built rigs, the new specialized chips will enable even the casual homebrew artist to construct sophisticated, complex equipment he might previously have considered beyond his reach.

The first Communication IC now available is the National Semiconductor LM270 Audio AGC/Squelch Amplifier. It is basically an operational amplifier, whose gain is controlled by a dc voltage, plus a built-in sensitive squelch threshold detector. The ten pin circuit replaces entire sections of today's transmitters, receivers, or transceivers, and makes speech compression, VOX, receiver squelch, and other functions practical in even the simplest homebrew rigs. While the chip contains 36 junction devices (transistors and diodes), and 20 resistors, it is size, rather than complexity which determines an IC's cost, so that the LM270, which is about the size of a single medium power transistor, is already cost competitive with the less complex discrete-component circuits it replaces. As volume commercial use of the circuit increases, the circuit is likely to be available at even more attractive prices.

Inside the can

The LM270 consists of several separate functions, designed to work together in a self-contained system, to produce control voltages for external use, or to respond to applied control signals. Heart of the circuit is a balanced series-shunt variable attenuator, formed by the four transistors in Fig. 1, which allows a large gain control range, with low distortion (for inputs less than 100 mV p-p), and which can be directly coupled to other parts of the system, eliminating the transformer or capacitor coupling necessary with all other variable arrangements. From a twelve volt supply, the gain vs. control voltage relationship is a smooth curve, as in Fig. 2, which gives a constant gain of +40 db for control voltages between zero and +2 volts, and is effectively "shut off" above +2.6 volts.

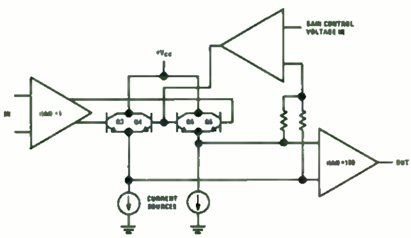


Fig. 1. Block diagram of variable gain amplifier.

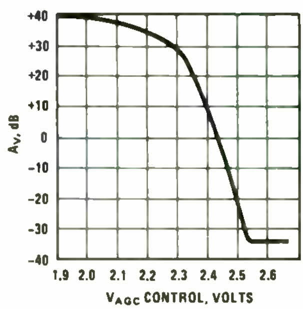


Fig. 2. Typical voltage gain vs. control voltage applied at pin 4.

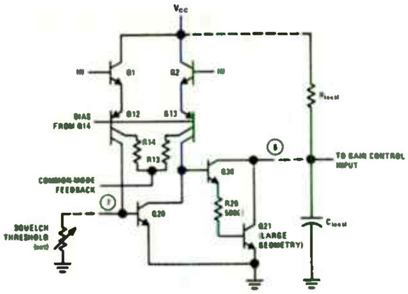


Fig. 3. Differential input circuit and squelch detector.

A separate subsystem within the LM270 is the squelch detector, **Fig. 3.** Using the same input differential amplifier as the variable gain circuit, the high gain peak detector formed by Q20, Q36 and Q21 responds to very small inputs (as little as a millivolt, depending on setting of the external threshold pot), by rapidly discharging an external capacitor. In the absence of input signal, C(ext) charges above +2.6 V, which, when tied to the gain control input, keeps the output amplifier "off". A momentary input peak above the threshold causes Q21 to rapidly discharge C(ext) below +2 volts, turning the amplifier fully "on". This arrangement gives a fast attack, slow release squelch, which catches first speech syllables, and waits long enough to avoid "cutting out" between words.

The complete circuit appears in **Fig. 4.** A detailed explanation of each part, too lengthy for inclusion here, may be found in the references.

Practical ham applications

Before going into specific circuits, a few general remarks are in order. Those familiar with operational amplifiers will easily recognize the LM270 configuration. Differential inputs allow inverting or non-inverting gain, or drive from a "floating" signal source. If single-ended drive is needed, the unused input is simply tied to the same reference voltage as is the actual input. All that is required is that both inputs be at equal dc potential, somewhere between +4.5 volts and the positive supply. Like an "op amp", the LM270's dc output voltage stays at approximately half of the positive supply voltage, for all supplies between +4.5 and +24 volts, so that symmetrical output clipping occurs.

Two identical gain control inputs, pins 3 and 4, are provided, which allows control by two independent sources at the same time, such as simultaneous AGC and squelch. By bypassing pin 2, the gain control inputs become emitter-follower positive peak detectors. The control inputs are protected by 6.5 volt zeners (Q33 and Q34). If the control input is expected to rise above +6.5 volts, a 10K series resistor at that input should be used to prevent excessive dissipation in the zeners.

Remote gain-controlled audio amplifier

A simple application is a preamplifier, **Fig. 5,** whose gain is manually controlled, noiselessly, by a dc voltage from a remote location, rather than running long, capacitive coax signal lines to and from that location. Pin 4 is bypassed by an external capacitor, to eliminate noise pickup. Since the gain-control curve, **Fig. 2,** is approximately logarithmic, a linear pot will give a desirable logarithmic audio attenuation characteristic.

For illustration, the second control input is shown connected to an IC logic gate, of the DTL, RTL or TTL varieties now available at low cost. This gate, operating from a five volt supply, can be part of a logic arrangement to override the remote control, and shut off the amplifier under present conditions. The resistors and capacitors shown biasing the single-ended input are used to illustrate one way of operating the inputs at a fixed dc voltage; subsequent examples will show simpler schemes.

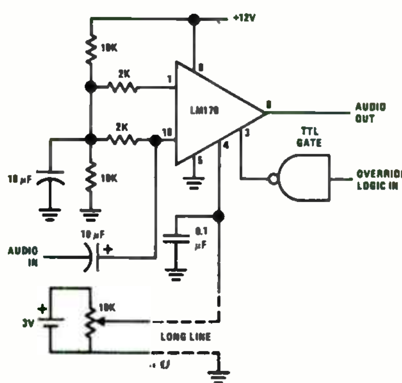


Fig. 5. Remote or digital control amplifier.

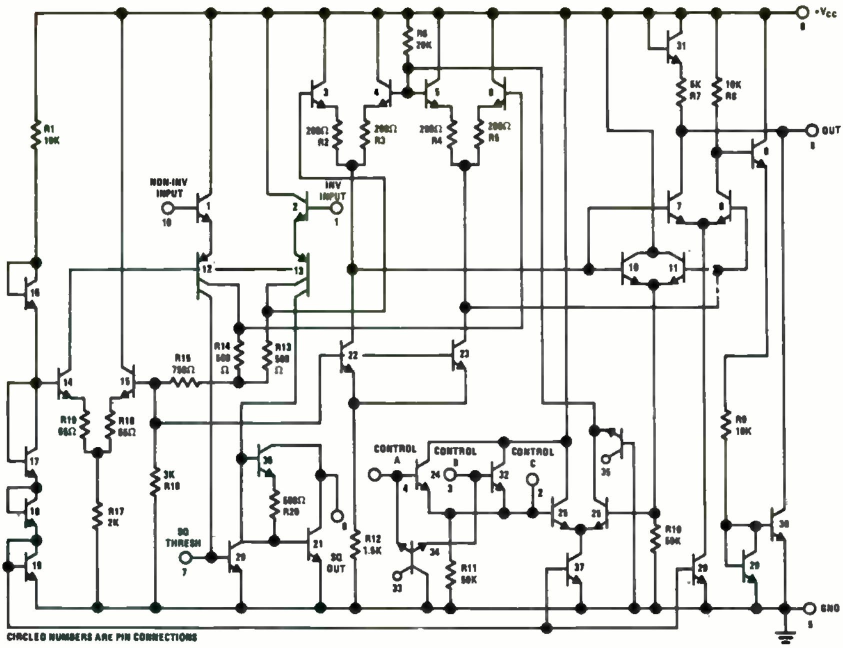


Fig. 4. Complete LM270 Schematic.

Speech compressors

Fig. 6 and 7 are basically audio AGC systems, which respond to peak speech levels above a set threshold by quickly reducing gain to a level which keeps succeeding similar peaks below the threshold. This differs from the usual "speech clipper", as it causes no distortion, but simply keeps the output level at an approximately constant level. In a modulator (any type), such AGC keeps modulation always near, but never in excess of, 100 percent. In Fig. 6, a PNP transistor (almost any type will do) adds enough gain to the control loop to operate over a large range of input levels. In Fig. 7, the additional gain of the receiver or modulator is used for this purpose. Varying load impedances can cause the gain of these stages to vary; taking the control signal from the system's audio output automatically compensates for load variations, in much the same way as an ALC system operates. The scope photo, Fig. 8, shows how the output (vertical axis) remains nearly constant while the input (horizontal axis) varies over a wide range. Note that both AGC circuits use the internal emitter-follower detectors, and that both inputs are biased from the positive supply through equal resistors, although other biasing works equally well.

Squelch preamplifier with hysteresis

Audio squelch is useful in both receiving and transmitting systems, to cut out background noises. The sensitive circuit of Fig. 9 includes a number of refinements, which make it smooth-acting, and easy on QRM weary ears. The threshold pot at pin 7 can be a front-panel control, to cut in at any desired level. Attack time is on the order of a millisecond for nearly any capacitor value at pin 6, but release time is determined by the external RC time constant. The fixed 100 k resistor may be replaced by a 100 k pot, in series with a 10 k resistor, to give operator-adjusted release time.

Part of the voltage at pin 6 is fed back to the threshold pot; since there is an "on" and an "off" voltage at pin 6, this creates a controlled amount of threshold hysteresis, which greatly enhances the circuit's im-

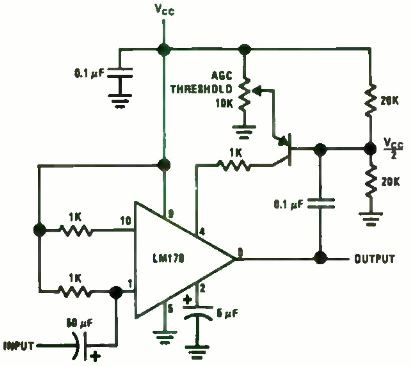


Fig. 6. Speech compressor.

munity to rapid fading or erratic speech patterns. A typical threshold control setting might be one at which amplification cuts in above 20 mV p-p inputs. With the feedback values shown, the input level must drop below 12 mV p-p for a time equal to the RC time constant, before gain is cut off. Shorting across the 200 ohm resistor defeats the hysteresis.

Unlike most squelch systems, which are just switches, the LM270 provides a gradual fadeout of background noise, when releasing. This is because the RC combination charges slowly along an exponential curve, and passes through the variable gain region on its way to complete cutoff. Fig. 10 shows the squelch action with a 25 μF capacitor and 100 k charging resistor. In the upper trace, a constant 1 kHz signal just below the squelch threshold keeps the output, in the lower trace, off. Abruptly increasing the input above the threshold immediately turns the amplifier on. Reduc-

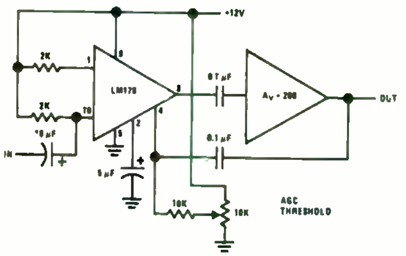


Fig. 7. Speech compressor using subsequent gain for better control.

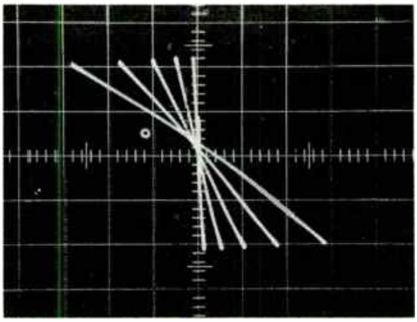


Fig. 8. AGC transfer characteristics, input vs. output, for varying input.

ing the input does not turn off the output, but merely reduces it proportionally, during the release period. Finally, after about one second, the output tapers off to zero again.

In this example, another input biasing scheme is illustrated; the LM270 can be driven directly from a high impedance dynamic microphone, such as the Shure 401A, with dc bias for both inputs derived from the positive supply, and no other external components required. In receiver squelch, one of the previously illustrated input arrangements might be used. The high frequency response of the squelch may be rolled off with a .05μF capacitor from pin 7 to ground, to reduce squelch triggering from high frequency noise above the speech spectrum.

A simple VOX mike preamp

Using a small power transistor driving a relay, the LM270 makes a combination VOX

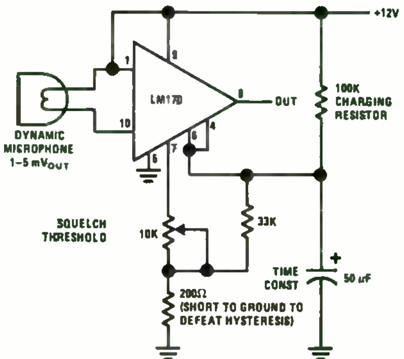


Fig. 9. Squelched preamplifier with hysteresis.

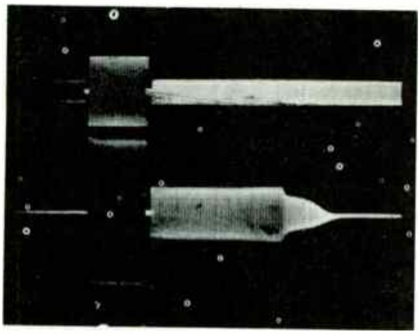


Fig. 10. Fast attack, slow release squelch action.

and microphone preamp small enough to build into a mobile-type communications mike. With the relay contacts wired across the push-to-talk switch, such a microphone can add VOX to existing transmitters with minimum disturbance of wiring. The basic circuit of Fig. 11 can be improved, as in Fig. 12, by driving one amplifier input from the microphone, and the other from an attenuated part of the receiver's loudspeaker output. (Correct phase must be determined experimentally, by reversing either loudspeaker or microphone leads for best performance.) This takes advantage of the differential inputs provided on the LM270, to cancel ambient speaker signals reaching the mike (anti-trip VOX). A diode shunts the relay coil to protect the PNP power transistor. Any relay drawing less than 100 mA from a +12 volt supply may be used, small model-airplane types being suited for inclusion inside the mike case. In Figs. 11 and 12, amplifier gain is not cut off by the squelch detector; however, the VOX circuit may combine with any of the preceding applications to give, for example, a preamp containing both VOX and speech compression.

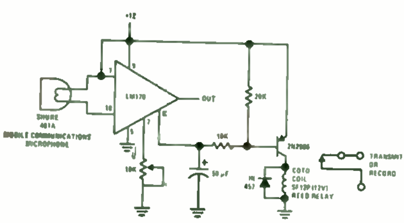


Fig. 11. VOX/mike preamp.

Twin-tee constant amplitude audio oscillator with remote level control

Oscillation occurs in a twin-tee, op-amp type circuit, when total feedback gain equals unity (including filter losses). Conventional methods of regulating oscillator amplitude usually rely on nonlinear loading of the gain stage. With the LM270, however, gain may be set by detecting the output, and using this to force the gain to exactly the minimum value required to sustain low distortion oscillation. The "AGC Oscillator" circuit, Fig. 13, automatically compensates for changes in oscillator load impedance. The exact amplitude at which this action occurs is set by an external pot, and may

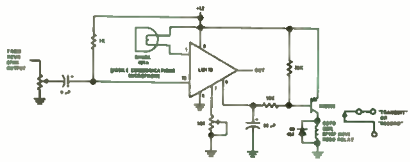


Fig. 12. VOX/mike preamp with anti-trip.

be set at any value below the maximum undistorted output of the amplifier itself. The "twin-tee" values shown give a 1 kHz output; other frequencies can be calculated from the formula:

$$f = \frac{1}{2\pi RC}$$

A modulated 455 kHz signal generator

An inexpensive, high "Q", 455 kHz ceramic filter can be substituted for the twin-tee

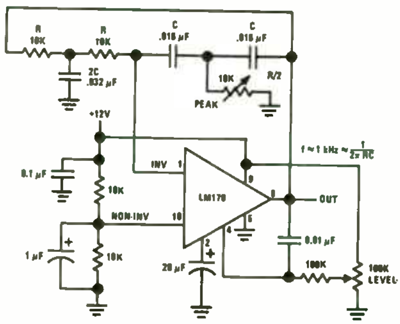


Fig. 13. Twin-Tee constant amplitude audio oscillator.

feedback network in the preceding example, to make a regulated-output AM *if* alignment generator, Fig. 14. If the AGC threshold voltage, which determines the amplitude of stabilized output, is varied at a slow (audio) rate, the output *rf* amplitude will be forced, by the AGC feedback, to track the audio modulation.

Conclusion

The LM 270 is a very versatile ham IC, which can make your next homebrew rig more advanced than many commercial jobs, with a minimum of the usual headaches. A little thought will reveal many applications, not covered in this article, in speech processing, RTTY, mountaintop repeater control, and others requiring either a variable gain amplifier or a sensitive squelch detector.

Future developments in the communications IC area are going to raise a few more eyebrows; it is expected that nearly all low power level sections of both receivers and transmitters will be built in integrated form in the near future, but these developments must wait for subsequent articles. Meanwhile, whet your appetite with the LM270, the first ham IC. . . . W6DNS

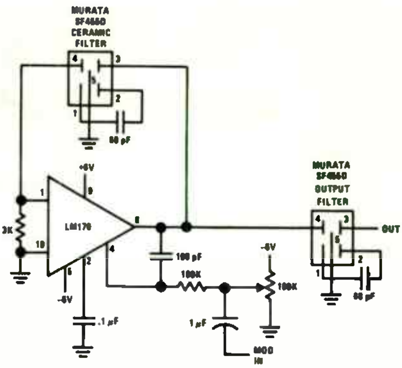


Fig. 14. 455 kHz modulated, regulated output signal generator.

REFERENCES

1. *R. Hirschfeld*, "A Unique Monolithic AGC/Squelch Amplifier", Application Note AN-11, National Semiconductor, Santa Clara, Calif. July, 1968
2. *R. Hirschfeld*, "Linear Integrated Circuits in Communication Systems", WESCON 1968 Proceedings, Session 1, Paper 3
3. "Ceramic Filter Data Sheet", Murata Corp. of America, 160 Broadway, N.Y., N.Y. 10038

Integrated Circuit TV Sync Generator

Claude Wiatrowski, K9AAC/7

The prices of both digital and linear integrated circuits have fallen so quickly in the past few years that it now costs more to build certain electronic systems from discrete components than to buy the equivalent integrated circuit. This article describes just such a case: the use of digital integrated circuits in a synchronizing generator for an amateur television station. This is not the usual sync generator probably found in the majority of amateur television stations. It generates the same sort of synchronizing, blanking, and drive signals that commercial television stations generate. Equivalent sync generators using vacuum tubes are found in commercial stations to this day. They are notoriously unstable, difficult to adjust, consume upwards of 450 watts of power, and may fill one or two large relay racks. The integrated circuit version described in this article is extremely stable, has only five easily set independent adjustments, consumes a little over one watt of power, and will fit behind a standard rack panel. More remarkable is the fact that the total cost of the unit is under fifty dollars.

Now, if you are interested in contructing a sync generator, I will assume that you have some experience in television systems. For those of you interested in the theory of the unit, I will also assume you have a familiarity with basic digital circuits and logic.

The blanking signal generated by this unit is identical to the signal all television stations, both amateur and commercial, use for this purpose. The sync signal generated, used by all commercial stations and very few amateur stations, may not be familiar to the amateur television operator so a short explanation is in order.

The Synchronizing Signal

Many simple camera designs that have been published in amateur journals rely on the blanking signal to synchronize the scanning oscillation in the receiving television. A more sophisticated approach is to add separate sync pulses on top of the blanking pulses while establishing a definite time relationship between the line and field frequencies, a truly interlaced system, as opposed to random interlace. Note, however, that in both systems the horizontal scanning oscillator of the receiving television will not be synchronized during the vertical sync pulse.

In order to achieve accurate interlace, it is imperative that the horizontal oscillator remain synchronized at all times. To accomplish this, the vertical sync pulse is serrated at twice the horizontal rate. The sync separator in a television receiver consists of a differentiator and an integrator. The output of the differentiator is a series of very short pulses corresponding to the edges of the synchronizing pulses. Since the widths of the pulses have no effect on the differentiator output, its output remains the same through the vertical sync interval due to the extra edges on the serrated vertical. It is this output that keeps the horizontal scanning oscillator synchronized at all times.

The integrator, on the other hand, responds to the widths of the pulses. The output of the integrator is roughly the average over an interval of time of the area of the pulse train. Indeed, then the output of the integrator increases as the pulses become wider and decreases as the pulses become narrower.

Since the vertical sync pulses are wide compared to the horizontal pulses, the output of the integrator corresponds to the vertical sync pulse interval. It is this output that synchronizes the vertical scanning oscillator.

Because every other field ends on a half line, the last horizontal sync pulse will vary

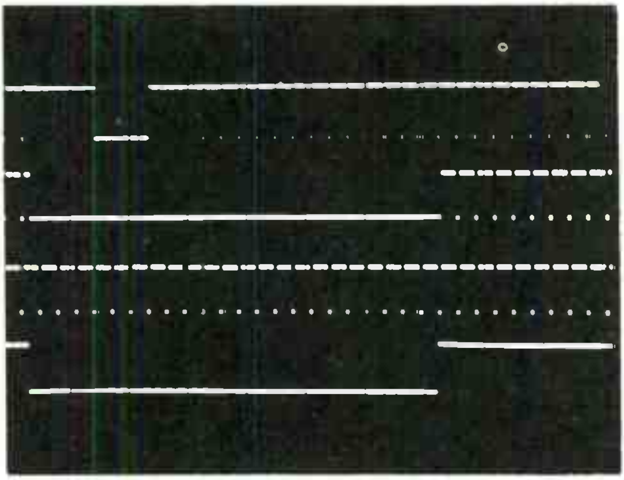


Fig. 1. Scope trace photo of sync and blanking signals and horizontal and vertical drive.

in position relative to the vertical pulse by 31.75 μs. This will cause the output of the integrator to be shifted by a small amount on alternate fields. This shift is enough to cause the lines on alternate fields to pair rather than be spaced equally. The group of pulses that remedy this situation are called equalizing pulses.

Six pulses, each of half the area and twice the repetition rate of the horizontal pulses, are placed before and after the vertical sync pulses. Being the same on every field, they serve to buffer the integrator output against shifts of the horizontal sync pulses which now occur farther away in time.

The complete sync, blanking, and drive signals are shown in Fig. 1. This has not been intended as a rigorous treatment of the television synchronizing system. For the interested reader, the subject is covered in many texts.

All gates are negative logic *nand* or positive logic *nor*. As this unit was designed with negative logic in mind, the analysis proceeds most easily from this viewpoint.

The flip-flops toggle on the negative-going edge of a trigger (T) pulse provided the set (S) and clear (C) terminals are left floating or are grounded. A signal of +3V on the set and 0V on the clear will cause the flip-flop to set on a negative-going trigger edge. (Set is interpreted as meaning terminal Q in the drawings will go to +3V and terminal Q will go the 0V.) An analogous situation exists for the clear input and the reset condition of the flip-flop. These flip-flops also have a dc clear which resets the flip-flop on a 3V pulse or level which overrides the inputs on all other terminals.

The Clock and Frequency Dividers

Basic components of any accurately interlaced scanning system are frequency dividers. One frequency divider divides the input frequency of 31.5 kHz by 525 to obtain a signal at the field rate of 60 Hz. Another unit divides by two to obtain 15.75 kHz, the line rate. In the sync generator described in this article, the dividers are actually electronic binary counters requiring no adjustments. The simplest of all these circuits will divide only by powers of two. An example is the 4:1 counter of Fig. 2 used to obtain a 31.5 kHz signal from a 126 kHz oscillator. It is necessary to apply feedback to count by numbers other than a power of two. The 15:1, 7:1, and 5:1 counters that make up the 525:1 divider in Fig. 2 are examples of this type of counter. Briefly stated, the output of the *nand* gate associated with each counter resets the entire

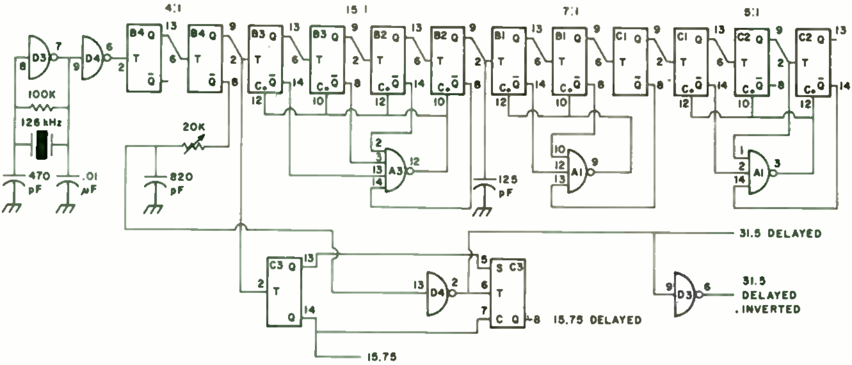


Fig. 2. Master oscillator frequency dividers.

counter to zero when the required count is reached.

The 7:1 counter has some minor variations. The last stage is reset by the action of the previous stage resetting. A capacitor is placed across the input of this counter. These two modifications are necessary to prevent extraneous pulses from resetting or triggering the counter stages. This is fast logic and the offending pulses are less than 1 V in magnitude and 50 ns in length.

There are two separate 2:1 counters. One counter is triggered by a signal delayed by an inverter and RC circuit connected to its input. The connections from the outputs of the nondelayed counter to the set and clear inputs of the delayed counter keep the two counters in phase. Without these connections it would be possible for the sync pulses to appear between rather than on top of the blanking pulses The delayed outputs are used to trigger all basic pulses except the horizontal blanking pulses. This delay which is adjusted to 1.27 µs accounts for the space between the leading edge of the blanking pulse and the leading edge of the sync pulse (sometimes called the front porch).

The delay is accomplished as follows. Assume that the input of inverter D4 is initially at 0V and its output is at 3V. If the input voltage to the potentiometer suddenly rises to 3V, the capacitor will only slowly change to 3V. The rate of change is adjusted by the potentiometer so the input to the inverter rises to the point necessary to saturate the transistor in the inverter after 1.27 µs. Thus, 1.27 µs after the input transition takes place, the output transition will take place from 3V to 0V.

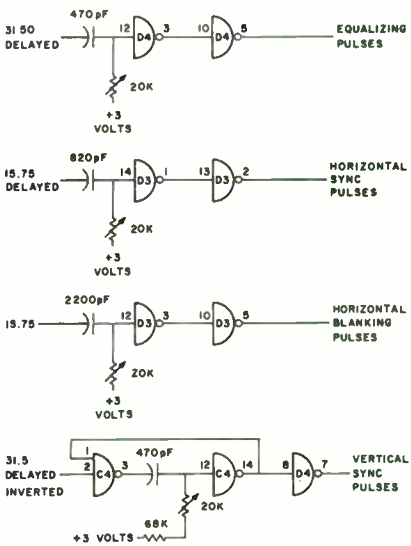


Fig. 3. Pulse generation.

The master oscillator operates at 126 kHz. This frequency was chosen because crystals for this frequency are less expensive and more easily obtained than lower frequency crystals. The circuit is a Pierce oscillator constructed from a digital inverter. A second inverter is used to shape the output to trigger the 4:1 divider to 31:5 kHz. The 0.01 μF phase-shifting capacitor may have to be increased for a lower Q crystal. It is also possible for this circuit to oscillate at a harmonic of 126 kHz. The cure here also is to increase the value of the capacitor.

Pulse Generation

Four monostable multivibrators are used to generate the following pulses: equalizing pulses, 2.54 μs; horizontal sync pulses, 5.08 μs; horizontal blanking pulses, 10.16 μs; and vertical sync pulses, 27.31 μs.

The 27.31 μs monostable is a standard design contructed from two gates. The other monostables are constructed from one inverter each. They are not true monostables in the sense that they do not have feedback.

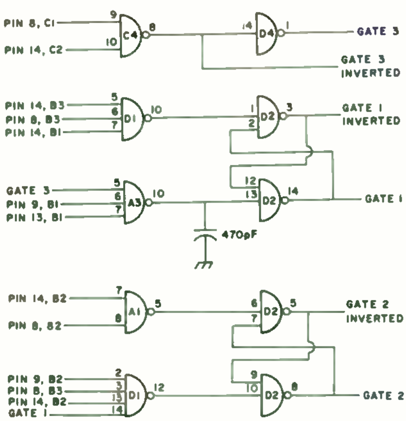


Fig. 4. Gate generation.

They will work in this application because the half period of the driving waveform is longer than the resquired pulse width. The circuitry is shown in Fig. 3. All outputs are inverted.

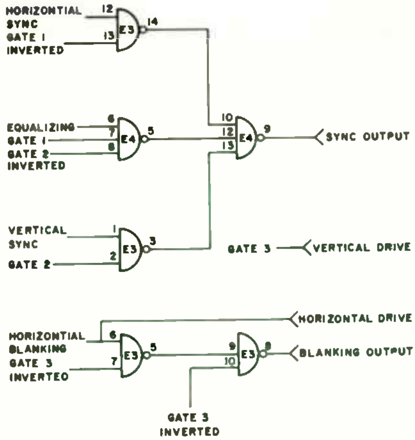


Fig. 5. Output gating.

Generating the Gating Signals

The main counter has 525 unique states of each of its flip-flops. These states can be decoded to provide gating signals.

Gate 3 is the vertical blanking signal. It is generated directly as a combination of the states of two flip-flops of the counter. This gate is about 3 horizontal lines longer than the recommended width. This is virtually unnoticeable and the additional logic necessary to correct it hardly seems worth the additional complication and expense.

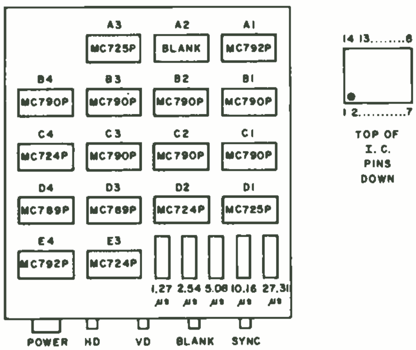
Gate 1 gates off the horizontal sync pulses and gates on the equalizing pulses. Gate 2 gates off the equalizing pulses and gates on the vertical sync pulses. Gates 1 and 2 are both produced by gating pulses from the main counter at the beginning and end of the gate times and using these pulses to set and clear flip-flops made of two cross connected *nand* gates. The circuitry is shown in Fig. 4.

Output Gating

Gates 1, 2, and 3 are used to turn off and on the basic pulses at the required times. Gate 3 (Fig. 5) gates off the horizontal blanking pulses during the vertical blanking interval and it itself is inserted as the vertical blanking signal.

Possible Changes

It is possible to generate "on" and "off" pulses and use these to set and clear a flip-flop to generate gate 3 analogous to the circuits for generating gates 1 and 2. In amateur service this additional circuitry hardly seems worth the additional cost and complication. If the unit is intended for commercial service, this modification will be necessary. Other modifications that would normally be required for commercial service would be separate width and position flip-flops and associated gating for the vertical and horizontal drive signals. Another "must" is the inclusion of alternative methods of master frequency control, such as a method to lock the generator to the power line or some external source (such as network or remote site sync generator).



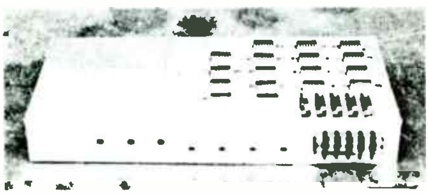


Fig. 6. The completed sync generator.

Construction

The unit is constructed on a 6 x 12 in. chassis of which only 6 x 7 in. is used for the sync generator (Fig. 6). The primary consideration to bear in mind is ease of wiring. Each integrated circuit has a possible fourteen or more connections to it and space for wiring can be rapidly depleted. This unit could ideally be constructed on a single printed circuit card, although the layout of such a large printed circuit is not a simple task.

On all integrated circuit sockets, pin 4 is ground and pin 11 is connected to +3.6V. The crystal socket was mounted on a bracket on the underside of the chassis. All discrete components were mounted on a directly on the IC sockets. I do not recommend this procedure as it makes an already crowded situation worse. A possible improvement would be to mount all discrete components on a piece of Vector board mounted on the underside of the chassis.

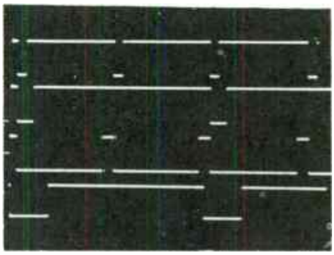
The multiturn potentiometers are a convenience but satisfactory operation should be possible with ordinary carbon potentiometers. Four phone sockets are used for sync, blanking, and vertical and horizontal drive connections.

The large empty space on the chassis, the extra output connectors, and the two spare terminals on the power block are for possible expansion to color. The IC socket is a 16-pin type that works as well for these 14-pin packages.

Adjustment Procedure

Assuming that you prefer to wire and test the generator in sections, the first section to be wired should be the clock and frequency dividers. Observe the input and output of the 15:1 divider on a dual-trace oscilloscope. There should be 15 input pulses for every output pulse. If you do not have dual-trace facilities, you can mix the two signals in a spare gate. Although the output does not look the same as with a dual-trace scope, the number of pulses that should occur *during the output pulse* for each counter can be determined from Fig. 7, which shows scope trace photos of the gating and counting circuits. The other dividers may be checked in the same way.

See that the 15.75 kHz and delayed 15.75 kHz signals are in phase. They will be if you have not made a wiring error. Adjust the delay so that there is a 1.27 μs between the leading edges of the two signals. Again, without a dual-trace oscilloscope, mix the

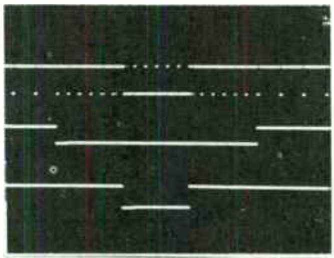


Equalizing pulses
2.54 us

Horizontal sync
5.08 us

Vertical sync
27.31 us

Horizontal blanking
10.16 us

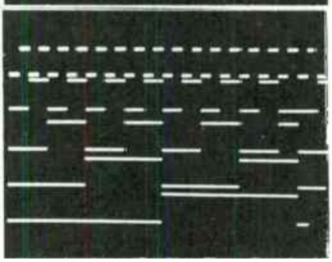


Sync

Gate 1

Gate 2

15:1 Counter



Pin 2, B3(Input)

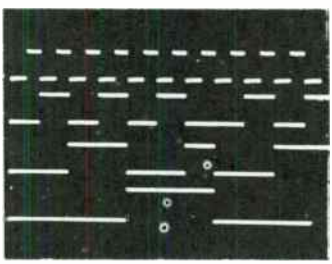
Pin 13, B3

Pin 9, B3

Pin 13, B2

Pin 9, B2 (output)

7:1 Counter

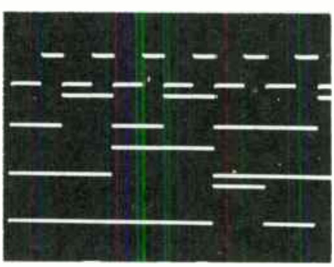


Pin 2, B1

Pin 13, B1

Pin 9, B1

Pin 9, C1



5:1 Counter

Pin 2, C1
Pin 13, C1
Pin 9, C2
Pin 13, C2

Fig. 7. Relative-time photos of gating functions. In (A), notice the leading edge of blanking pulse precedes other pulses by 1.27 μs. Photo (B) shows the action of gates 1 and 2 in producing the sync signal. Photos (C), (D), and (E) show the output signals from the counters.

inverted nondelayed output with the normal delayed output. The pulse on the oscilloscope from the output of your mixing gate has a width equal to the delay time.

Wire the monostables and their inverters next. Look at the output of the inverters and adjust the pulse widths to the correct values. Remember the pulses are negative going at the outputs of the inverters. Wire gate 3 first, then gate 1, and finally gate 2. The outputs of the three gates may be compared with the scope pictures in Fig. 7. Their widths should be: gate 1, 571.5 μs; gate 2, 190.5 μs; gate 3, 2857.5 μs. Lastly, wire the output gating and connect it all together. Sync, blanking, and drive should now appear at their respective output terminals. Check the appearance of the blanking and drive with an oscilloscope. The sync waveform may only be viewed reliably using an oscilloscope with delaying sweep. The delay has to be longer than two fields, as the horizontal sync pulses alternate their position every other field. Without these precautions you can not be certain that your scope trace is an accurate indication of the output waveform. Note, however, that *no* adjustments *have* to be made while looking at this waveform so if you trust your wiring and previous adjustment procedure, you can be confident of having a correct sync signal. If you still insist on seeing the sync waveform, it is possible to add a single flip-flop as a 2:1 counter connected to the output of this 2:1 divider to the external sync on your scope and you should be able, after some adjustment, to see the sync signal.

Since the outputs of this unit are the outputs of gates, they *will not* drive a low impedance load. All inputs to other equipment should be high impedance and the line should *not* be terminated with a resistor. This situation is ideal in my station as most equipment is of IC contruction and the output of the sync generator simply drives other gates. If you absolutely require low impedance signals, the outputs of the gates may be buffered with simple emitter followers.

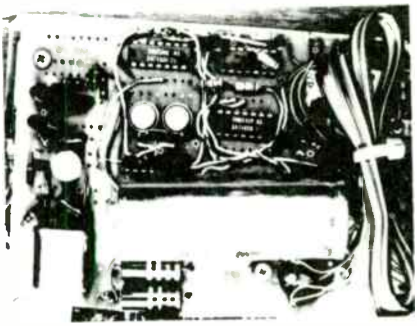
A Digital Readout for your VFO

R. Factor

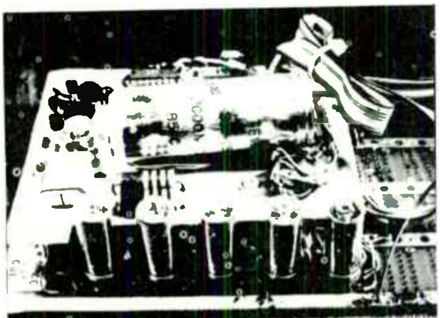
Are you on frequency? Are you within the band? Incentive licensing subbands have made this an increasingly difficult question to answer. Even if you have an Extra license, it is still nice to know where you are for net operation, OO work, etc. The declining price of integrated circuits and readout devices has made it unnecessary to continue to drool over the advertisements for that $1600 transceiver with the digital dial. A large percentage of currently manufactured gear can be fitted with a digital dial with greater accuracy, and very reasonable cost. Little modification of your gear is required, and what is required will not affect the resale value, since it is invisible and is easily removed.

As an extra added plus for the homebrewer, the digital dial makes unnecessary the greatest hate object of the electronic purist; the tuning dial; with its attendant impossibilities of getting linearity, accuracy, and above all, of inscribing, decaling, engraving, or calligraphing the dial in a neat, readable manner.

The device to be described is an adaptation of the basic electronic counter circuit. It takes the vfo signal of your receiver or transmitter and tells you what it is. Accurately. Not to the nearest kHz (with luck, at 57.8% relative humidity, with the rig on a cast iron bench weighted with two tons of sand and suspended in a mercury pool) but rather down to the nearest 10 Hz, almost always, with no precautions or requirements other than a periodic check to zero-beat the oscillator with WWV. If it is installed in a transmitter, it can measure received signal frequency by zero-beating the transmitter with the receiver. It can also measure frequency shift of an RTTY signal by zero-beating the mark and space frequencies and subtracting the two readings.



Top view of power supply and timing section.



Front view of digital dial showing sandwich construction.

You know frequency counters are expensive and you're scared away by the price. Right? You shouldn't be – I mentioned that IC and readout prices are declining. Below is a comparison based on the advertised prices of the major components involved. I haven't included resistors and small components because you no doubt already have them. Even if you don't, the total should be less than $10. It is assumed that your rig has high voltage and filament supplies so that no power transformer must be bought.

Type	Quan	Total
SN7490	10	$20
SN7475	5	$10
SN7441AN	4	$12
B5750 (Nixie)	5	$25
1MHz Xtal	1	$ 5
		$72

This is effectively for the "worst case." i.e., no external oscillator, using 1 MHz crystal, 5-digit readout with storage. By eliminating the 10 Hz resolution, using an existing 100 KHz calibrator, and deleting the most significant digit, the price becomes drastically lower.

Type	Quan	Total
SN7490	7	$14
SN7441AN	3	$ 9
B5750	3	$15
		$38

Since dial drive mechanisms with less than one tenth the accuracy cost over $30, the price is not at all out of line. Incidentally, the prices quoted above are the highest you should have to pay. Due to a current oversupply in the IC industry, substantial discounts may be obtainable, and there is little doubt that prices will drop substantially between the time these words were written and the time the article appears in print.

Now that you have decided to build the digital dial, let's see if its circuitry makes it compatible with your rig. The rig should have crystal-controlled front end or conversion oscillator and a vfo covering a reasonable frequency range without odd number kilohertz tacked on. Ideally, a range of from, say 5.000 to 5.500 MHz would be covered. The unit can be used with a unit whose vfo covers, for example, 5.300 to 5.800 MHz with only minor changes. Input frequency is unimportant – anything up to 15 or 20 MHz is okay with the ICs specified. What is important is that there be no odd numbers at the end. A vfo covering 5.455 to 5.955 would be unacceptable. Of course, the counter will measure such frequencies as well as any other, but they will be impossible to mentally relate to the operating frequency. Another

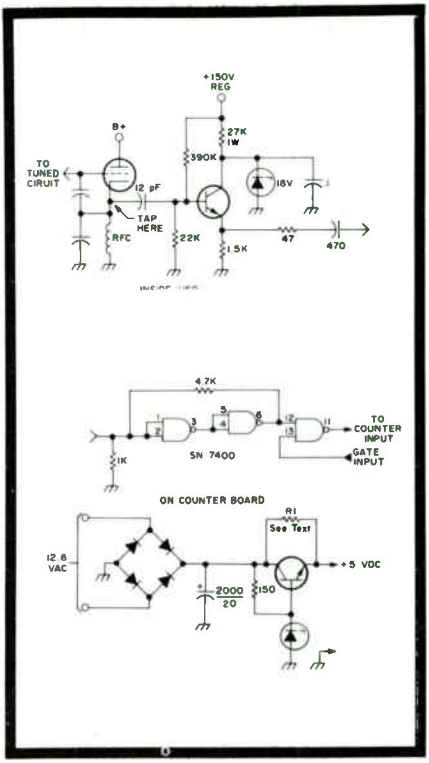


Fig. 1. Vfo isolation, input circuit, and power supply.

requirement is that the frequency mixing scheme be either additive or subtractive, but not a combination of the two. A quick glance at your instruction manual will tell you the exact vfo frequency and mixing scheme. If your rig tunes in the same direction on all bands, you should have no problem with the additive/subtractive question. If your rig does not fit the above criteria, I'll have a few comments on possible remedies, untried but theoretically sound, at the conclusion of the article.

The digital dial described herein is being used with my HX500 transmitter, whose vfo is 3.9–4.4 MHz, and which employs subtractive mixing. However, the principles and circuitry can be used with only minor modification in any transmitter or receiver fitting the above criteria.

The Circuit

The first step in going digital is to modify your rig. The digital dial needs +5V dc at about 500 mA, B+ at about 15 mA, and 1V of signal at the vfo frequency. To get the 5V, tap the filament supply (6.3 or, preferably, 12.6V and connect it to the bridge rectifier shown in Fig. 1. If your filament supply is 6.3V, it may be necessary to connect another 6.3V, 500 mA transformer in series with the filament supply. Nixie tubes like 170V at about 3 mA across them, but since they tend to act as voltage regulators, current limiting is necessary. Measure your B+ supply, subtract 190V, and calculate a resistor that will give 3 mA per Nixie with this voltage across it. Be sure to make a power calculation also, as a 2W resistor may be required. Small series resistors are connected to each Nixie to equalize current; the resistor just calculated goes to the junction of the series resistors.

The purpose of the vfo buffer is to make sure that the vfo in the rig is not disturbed. An emitter follower is ideal for this application. The component values shown are for a supply of 150V, as this is the most commonly used. No trouble should be encountered if the pickoff capacitor (12 pF in the schematic of Fig. 2) is at least an order of magnitude smaller than the capacitor from the cathode to ground. The actual capacity will be much smaller, since the pickoff capacitor is in series with the transistor base, but it doesn't hurt to be safe. This completes the necessary modification of the rig. No damage to the front panel – see?

Circuit Description

The theory behind the electronic counter is that the number of cycles of the signal

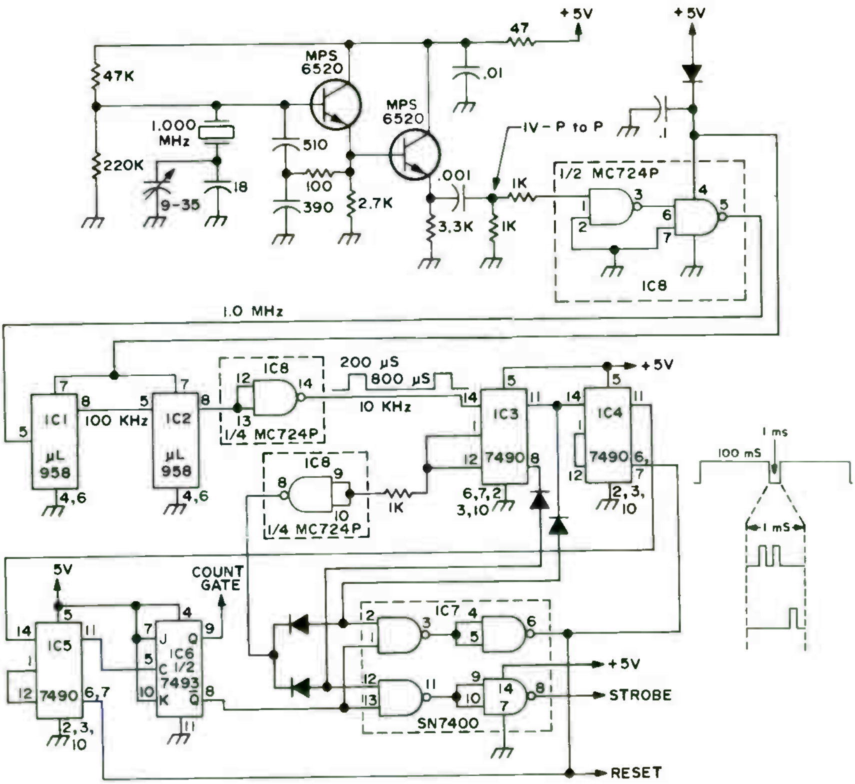


Fig. 2. Frequency-standard oscillator and counter control circuitry with timing diagrams.

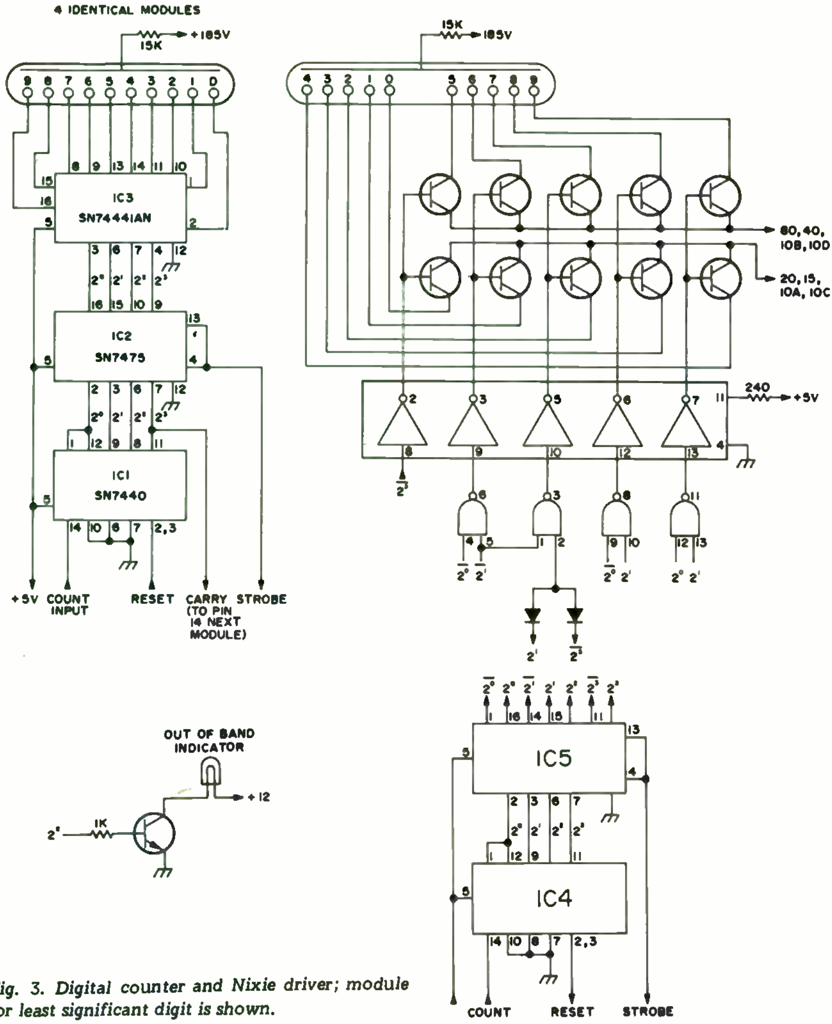


Fig. 3. Digital counter and Nixie driver; module for least significant digit is shown.

in a given period can be counted. If the period, for example, is 1 second, the number counted comes out to be cycles per second. Deriving the period precisely is done by counting down an accurate reference oscillator and then using the counted down signal to gate the signal to be measured.

The circuit works as follows (see Figs. 3 and 4). The input signal goes into a divider chain, the first stage of which is the least significant digit. After 10 pulses, the first stage is reset and a carry pulse generated. Since the counter has five stages and the counted frequency is in the MHz range, you can see that the counter will overflow several times during each count interval. However, we know what the first digit will be, so there is no point wasting a counting stage on it. All the stages are identical except for the most significant digit, which uses discrete components since special decoding is required.

After the counters have counted, the SN7441AN decoder/drivers ground the appropriate Nixie cathodes and the corresponding numbers light up. Between the counters and the decoder/drivers, one additional stage is necessary – a buffer storage register. The reason for this is that the gate time is 0.1 second, which will cause blurring during counting. If you wish to read out once per second, this isn't too bad. However, the dial is more useful when it's responsive, and you can zero in a frequency much more quickly when you don't have to wait a full second to see where the dial turn took you. By using storage registers and a few logic gates, it is possible to make ten 100 ms measurements in 10.1 seconds with no digit blurring. How this is done brings us to the timing circuitry.

First, an accurate frequency must be generated. The circuit shown in Fig. 2 is a stable oscillator designed for 32 pF crystals. A trimmer is included for fine frequency adjustment. There is no reason why the 1 MHz output of another oscillator can't be used. Approximately 1V (peak-to-peak) is required. You can also use a 100 kHz oscillator and delete one divider stage. The MC724P quad 2-input gate shapes the signal into a square wave suitable for triggering the dividers.

The first two dividers I used in my unit are Fairchild μL958s. I used them because I had them lying around, and it is okay to use SN7490s here. A section of IC8 is used as a buffer to make the two IC families compatible. If you use 7490s, delete this gate.

ICs 3, 4, and 5 are the remaining dividers. At time zero, let us suppose that IC6, the gate flip-flop, is reset (gate at logic 0). The carry pulse from the last divider brings the gate high, starting the count. One carry pulse later (100 ms), IC6 is again reset, stopping the count. (The abbreviation "ms" means milliseconds; μs" means microseconds.

Here the fun begins. IC6 Q, being high, sets half of the *and* condition on the first two sections of IC7. IC3 is still counting from binary 0 to binary 9 at a rate of 1000

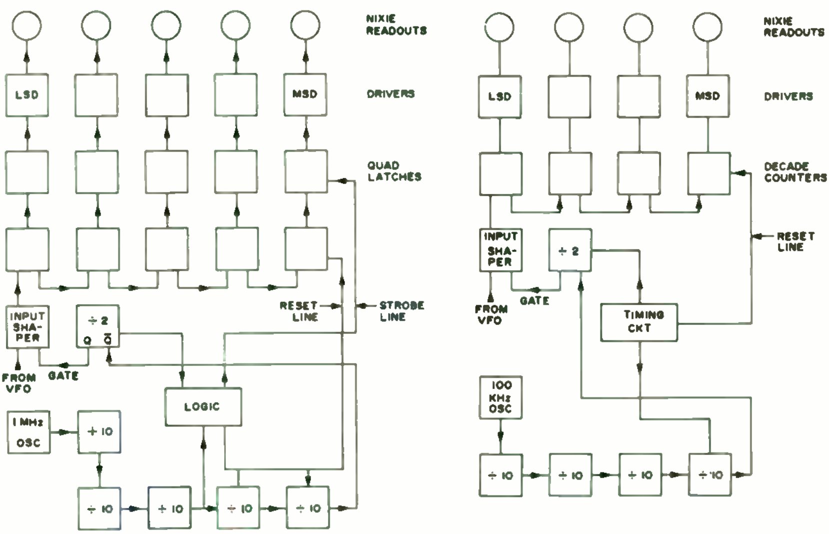


Fig. 4. Logic block diagram. The left block gives readout in 10 Hz increments at 101 ms intervals without blurring. The right block reads in 100 Hz increments; the update is adjustable and blurring occurs during 100 ms count.

times per second. Its output states are decoded by the remaining inputs to IC7 in the following sequence: At binary 4 and 6, pins 12 and 13 are both high, causing pin 11 to go low, pin 8 to go high, and thus strobing data into the buffer register. The strobe is performed twice because doing so eliminates the need for an additional decode – the same data is strobed each time.

At binary 8, pins 2 and 1 of IC7 are both high, causing pin 3 to go low and pin 6 to go high, thus resetting the counter and resetting the last two 7490s in the dividing chain. However, these two dividers are reset to 9 instead of to 0. This means that just 0.1 millisecond later, when IC3 carries, IC4 and IC5 will also, starting the count again. Note that the total time between the end of the first count and the beginning of the next is just 1 ms, allowing almost 10 measurements per second.

Backwards or Forwards?

If your rig is of the subtractive mixing type, you are probably wondering how the readout frequency corresponds to the output frequency of the rig. It doesn't – everything is upside down. Of course, it will be accurate on one frequency in the middle of the band, but that's little comfort. Despair not. Instead, connect the Nixies to the counter backwards. (Switch 0 for 9, 1 for 8, 2 for 7, 3 for 6, and 4 for 5.) Through the magic of mathematics, for every 10 Hz increment of the vfo frequency, there is a 10 Hz decrement of the readout frequency. You can now see why I said that the rig can be additive or subtractive, but not both . . . unless you happen to have a two-position, 50-pole switch around.

Most Significant Digit

The one remaining readout problem is that of the most significant digit. If you want the digital dial to be accurate on all bands it must read: (3)745.92 kHz on 80 meters, and (7)245.92 kHz on 40 meters. One way this can be done is to have different drivers for the Nixie, and ground the appropriate one with an extra wafer on the band switch (see Fig. 5). By driving the bases in parallel and grounding the emitters of the appropriate group, a cheap and dirty "adder" is constructed.

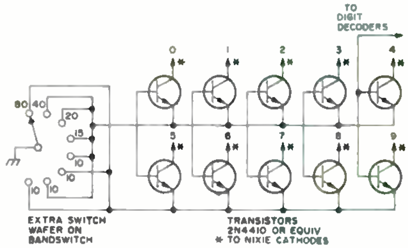


Fig. 5. Using a wafer switch to ground the appropriate Nixie drive for "adder" function.

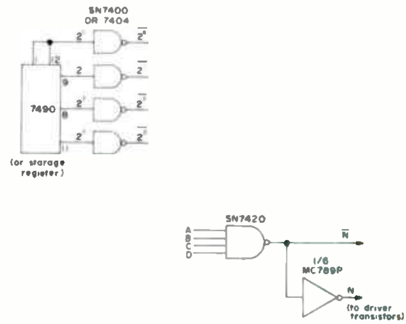
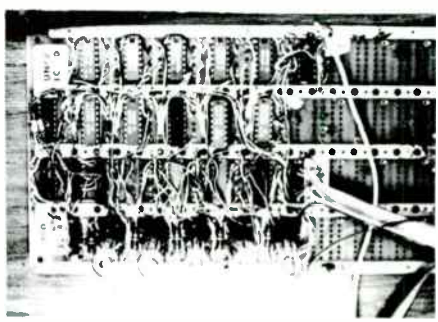


Fig. 6. General decoding scheme for all BCD states.

If your vfo starts from some frequency other than an even megahertz, the decoder must be designed to match. If the vfo is, say 3.9 to 4.4 MHz (as is the HX500), where 3.9 corresponds to the top of the band and 4.4 to the bottom, then the decoding is as follows: 4.399.99 to 4.300.00 is 000.00 to 099.99 kHz on the



Top view of counting circuit.

dial. Thus, 3 must be decoded as 0; 2 as 1; 1 as 2; 0 as 3; and 9 as 4. To do this the gate that decodes a 3 binary state should drive the 0 (and 5) digit line, etc. The decoding section is fairly simple since only 5 states must be decoded, compared to the 10 states decoded by the SN7441AN. Figure 6 and Table I give the *general* decoding logic for all possible binary-coded-decimal states. Considerable simplification is possible if you have a good understanding of digital logic. For instance, the schematic of my unit shows the decoding function performed by one 3-input gate equivalent, three 2-input gates, and one 1-input gate.

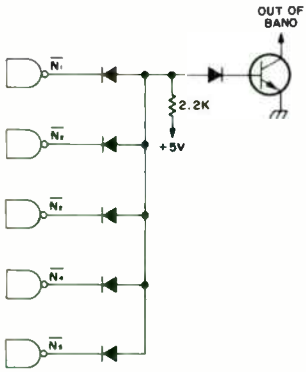


Fig. 7. An out-of-band indicator is constructed by anding the unused N states and feeding a simple driver.

If you wish, the unused states of the counter can be used to give an out-of-band indication (Fig. 7). None of the N outputs will be low if none of the proper states is decoded. This allows the transistor to be saturated. You can use it to light a light, ring a bell, disable the VOX circuit, or start a tape recording with hosannas to the FCC.

El Cheapo

Thus far, my comments have been about a "deluxe" counter with storage and 10 Hz readout. If you wish to build a more austere model (read "cheap"), there are ways to cut down on the cost. The best is to make a four-digit model with a 10 ms time base. Since the count interval is so short, you can sample once every tenth of a second, count for a hundredth of a second, perform housekeeping functions in 2.0 ms, and have a display visible for almost 90% of the time without using storage registers. One Nixie stage is eliminated, and IC6 is unnecessary.

To make the stripped-down version, substitute the circuitry of Fig. 8 for the bottom half of Fig. 2. It works as follows: As the dividers count up from 0, the diodes decode a count of 88 and generate a 1 ms long reset pulse. At the transition to 90, the count gate decodes a binary 9 on the most significant digit and goes high for 10 ms, forming the gate interval. Thus, the SN7490s essentially perform the timing functions.

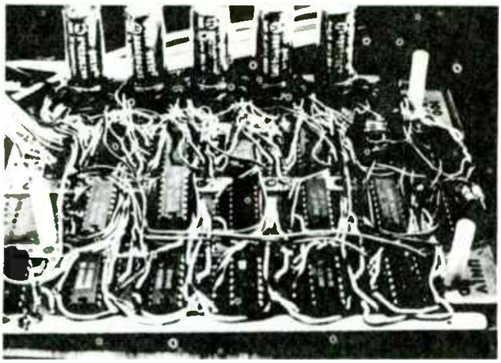
An additional Nixie stage can be eliminated by deleting the most significant digit from the count chain. This also saves the trouble of figuring out the decoding circuitry. Since any vfo will be accurate to the nearest 100 kHz, a second of mental work will compute the correct number.

Construction

Unless you want to make a PC board, the easiest way to build the unit is to use a "universal" IC card. The ICs and the Nixies can be mounted on it, and no drilling is required. Since most of the wiring is repetitive, it can be accomplished quite rapidly. The counter board will take about 4 hours, the timing board about 2. Looking at the photographs you may disbelieve me, but it goes very fast.

One suggestion to speed things up is to use the wire that can be found in multiconductor telephone cable. A four foot scrap of this stuff has 200 ft of hookup wire in it, and the wire has a soft plastic insulation which is very easy to strip. It melts at a low temperature, but since you're using a low power soldering iron, there should be no problem.

The crystal oscillator is quite stable, but it can obviously be improved by installing the crystal in an oven. The TTL circuits used in this project are much better than RTL with regard to noise immunity. A significant amount of rf can be floating



Rear view of counting circuit.

around before erratic operation occurs. However, it is recommended that the unit be carefully bypassed and shielded; first, because often there *is* a considerable amount of rf floating around, and, secondly, because the pulses in the digital circuitry are very fast and can radiate noise to the receiver.

Mechanically, my unit was constructed on two boards which were then sandwiched together. This saved about 10 sq in. There is no reason why the unit can't be built on one board. Final dimensions exclusive of case were 5 by 4-1/2 by 2-1/4 in. Something this size could probably be installed inside the cabinet of the rig. If you wish, the Nixies and their drivers can be placed on a separate board which should easily fit in the space taken by the typical dial drum or scale assembly.

The unit requires 5V and some convenient high voltage. I described the B+ "supply" earlier. The 5V supply is worth a few words. Regulation is not critical, but if the supply has lots of hum or noise, filter it out before it gets to the oscillator. The oscillator only takes 5 mA or so and an RC filter should be sufficient.

I used some nondescript power transistor for the series regulator. Anything that can handle an amp and 5W of dissipation should be sufficient.

The resistor shown bridging the transistor bypasses some of the current without, theoretically, greatly affecting the regulation. Measure the voltage across the transistor – there should be a large ac component. Pick a value for the resistor so that when the voltage is at an instantaneous minimum, the current through the resistor is about 100 mA. If all this is too much trouble, don't bother. Almost any transistor you choose can handle the power even without the resistor.

Final Thoughts

I have described a fairly simple and relatively cheap way of obtaining accurate frequency calibration. There are some inherent limitations. The main limitation is that the counter doesn't actually measure the transmitter output frequency. Since the assorted conversion oscillators are inevitably crystal-controlled, there is little cause to worry about significant errors. When you install the unit, it might be a good idea to set the vfo at 000.00 and zero-beat the transmitter output with your calibrator by adjusting the trimmer capacitors on the conversion crystals.

If you can't build the counter described because your rig doesn't fit the criteria, there is still hope. For instance, if it is additive and subtractive, you might consider using the bidirectional counters that are now available. Direction of count can

TABLE I.

Connection Chart for all Binary Combinations

		CONNECT			
N=	BINARY	A	B	C	D
0	0000	$\overline{2^0}$	$\overline{2^1}$	$\overline{2^2}$	$\overline{2^3}$
1	1000	2^0	$\overline{2^1}$	$\overline{2^2}$	$\overline{2^3}$
2	0100	$\overline{2^0}$	2^1	$\overline{2^2}$	$\overline{2^3}$
3	1100	2^0	2^1	$\overline{2^2}$	$\overline{2^3}$
4	0010	$\overline{2^0}$	$\overline{2^1}$	2^2	$\overline{2^3}$
5	1010	2^0	$\overline{2^1}$	2^2	$\overline{2^3}$
6	0110	$\overline{2^0}$	2^1	2^2	$\overline{2^3}$
7	1110	2^0	2^1	2^2	$\overline{2^3}$
8	0001	$\overline{2^0}$	$\overline{2^1}$	$\overline{2^2}$	2^3
9	1001	2^0	$\overline{2^1}$	$\overline{2^2}$	2^3

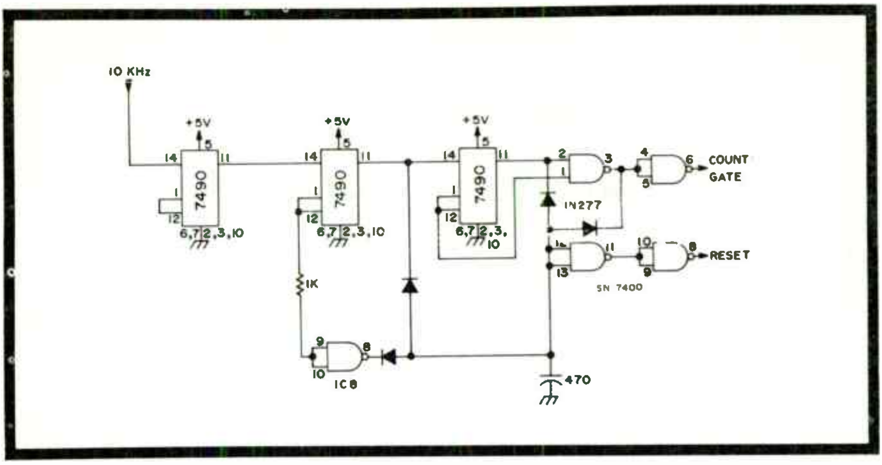


Fig. 8. By substituting this circuit for the bottom half of Fig. 2, a cheaper but less accurate digital counter results.

Ref. Macleish, Kenneth, "A Frequency Counter For the Amateur Station," QST, October 1970, pp.15.

be controlled by the band switch. These units are presently more expensive than the 7490s.

If your rig uses some odd vfo frequency, perhaps with a 455 kHz added or subtracted, you can try several things. One is to mix the vfo with a signal that will subtract the odd number. Another is to preset the counter with the reciprocal of the odd number and set to 9 all digits more significant than the odd number. I don't know how well these ideas will work, since it was not necessary to try them, but they seem to be theoretically sound. The reference gives a design for a more elaborate counter that might solve your problem.

If you feel as I do, that electronic equipment should have uses comprehensible to the layman, witness an additional characteristic of the digital dial. Take vfo knob firmly in hand and give it a spin. The last two digits of the display will be random numbers! Not wishing to become an accessory before the fact, I won't attempt to suggest what you might use those random numbers for, except perhaps compiling a random number table. But that shouldn't stop you from thinking . . .

...WA2IKL■

The Mod 2 Digital ID Unit

Tom Woore, KB6BFM

The reliability of most electromechanical repeater identifiers leaves something to be desired. In many cases an amateur repeater will operate as many as 500 times a day. According to a recent FCC ruling, each repeater must be identified at least every three minutes of operation. This means that the repeater may be identified just as many times as the repeater is operated. Most electromechanical devices such as relays, code-wheel devices, and tape decks cannot withstand the constant on/off operation of repeaters for any great length of time. Remember, those devices were for intermittent use – the code wheel for distress signals; the tape deck for listening pleasure, and so on. None of these devices were made to take the constant on/off use that is needed, let alone the environmental conditions.

One only has to climb to the mountaintop site after the first snow of the season because of an identifier failure to realize there ought to be a better way! Why not make the identifier solid state and eliminate those moving parts that wear out? Better still, why not use integrated circuits to accomplish the task? With a parts cost of less than $20, the Morse code digital identification unit (DIU) described herein does just that and it will outlast anything mechanical that you might otherwise put on top of a mountain.

The System

The DIU is unique in that it uses a simplified computer address principle for selecting the information it is programmed to send. There are four basic units in the DIU:

- Counter
- Matrix (memory)
- Signal logic
- Oscillator

The counter establishes which sequence is next. The matrix determines what instruction is next by the sequence. The signal logic converts the instruction information into the actual signal to be sent. The tone oscillator sends the requested signal. The whole system is based on a closed loop and therefore no standard clock is employed in the logic.

Logic Terms

To understand how the DIU works we must first become familiar with some of the simple logic terms that the system is based on.

High: Maximum output of logic unit (at least +1.5V)

Low: Minimum output of logic unit (less than +0.5V)

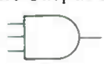
Inverter: Device used to produce opposite logic state of what is applied to it. Example: +2V into an inverter would produce a 0V output while a 0V input would produce a +2V output.

Symbol: 

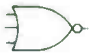
Or gate: Device used to give a high output when any of its input lines are high. Example: 3 input lines; one at +2V, the other two at 0V produces a +2V output on the output line of the gate.

Symbol: 

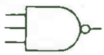
And gate: Device used to give a high output when *all* input lines are high. Example: 3 input lines; +2V on all 3 input lines of gate produces a +2V output on the output line of the gate.

Symbol: 

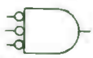
Nor gate: An inverted *or* gate; device used to give a low output when any of its input lines are high. Example: 3 input lines, one at +2V, the other two at 0V, produces a 0V output on the output line of the gate.

Symbol: 

Nand gate: An inverted *and* gate; device used to give a low output when *all* input lines are high. Example: 3 input lines, +2V on all 3 input lines of gate, produces a 0V output on the output line of the gate.

Symbol: 

For this article, *nor* gate logic was used to implement the *nand* functions; therefore, the definition for our purpose of a *nand* gate is a device used to give a high output when all of its input lines are low. Example: 3 input lines, 0V on all three input lines of a gate, produces +2V output on the output line of the gate.

Symbol: 

Note that the zero placed before or after the inverter, *nor*, and *nand* logic gates defines the expected state of the input or the output for the function to occur.

Flip-flop: A device used to store information a bit at a time. In the DIU application, a string of flip-flops is used as a counter. The purpose of the counter being to sequentially address the required instructions for the DIU.

Symbol: 

Unit: Smallest bit of information sent by the DIU (dih, dah, or blank).

The DIU uses the MC700 series of Motorola integrated circuits due to their inexpensiveness and availability.

System Operation

A 0V signal through the start network (see Fig. 1) from the transmitter keying circuit resets all the flip-flops in the counter to the zero state. All Q' lines become high and all Q lines become low. Approximately 2V and 0V are fed into the diode matrix, which decodes the counter number into an instruction for the oscillator keying logic. In the DIU there are four basic instructions: (1) send a dit, (2) send a dah, (3) send neither dit nor dah (blank), and (4) stop.

If the diode matrix decodes the first sequence count (0) to be instruction number 1 (send dit), the dit signal line from the matrix will be high. This will cause the dit inverter to have a low output and one-half of the "dit enable" gate will be enabled. Since the space line is also at "low" level at this time, a trigger pulse will be sent through capacitor C7 to the "dit" one-shot. (A one-shot is a monostable device used to generate a predetermined pulse-width.) The dit time pulse determined by the one-shot is sent through the "dit or dah" gate and the "dit or dah/blank" gate to enable the "dit or dah send" gate. The *nand* gate keys the oscillator circuit to produce the dit signal.

At the same time the dit is being sent by the one-shot to the oscillator, the "space" one-shot logic is being reset via the "dit, dah, or blank" gate, inverter, and "space enable" gate.

Upon completion of the dit signal, the "dit, dah, or blank" gate becomes high, making the inverter output low. Since the stop instruction has not been called for by the matrix, the "space enable" gate produces a high output. The high output in turn sends a pulse through capacitor C3 to trigger the "space" one-shot. (The space time period is used to separate the units of a letter. Example: D = dah-space-dit-space-dit.

The space period is the same as the period for the dit. The space signal, besides allowing for the time to distinguish the units of a letter, advances the counter through an inverter to the next unit and resets the "dit" and "dah" one-shots by discharging capacitors C5 and C7.

If the diode matrix decodes the next sequence to be instruction 2 (send dah), the dah signal line from the matrix will

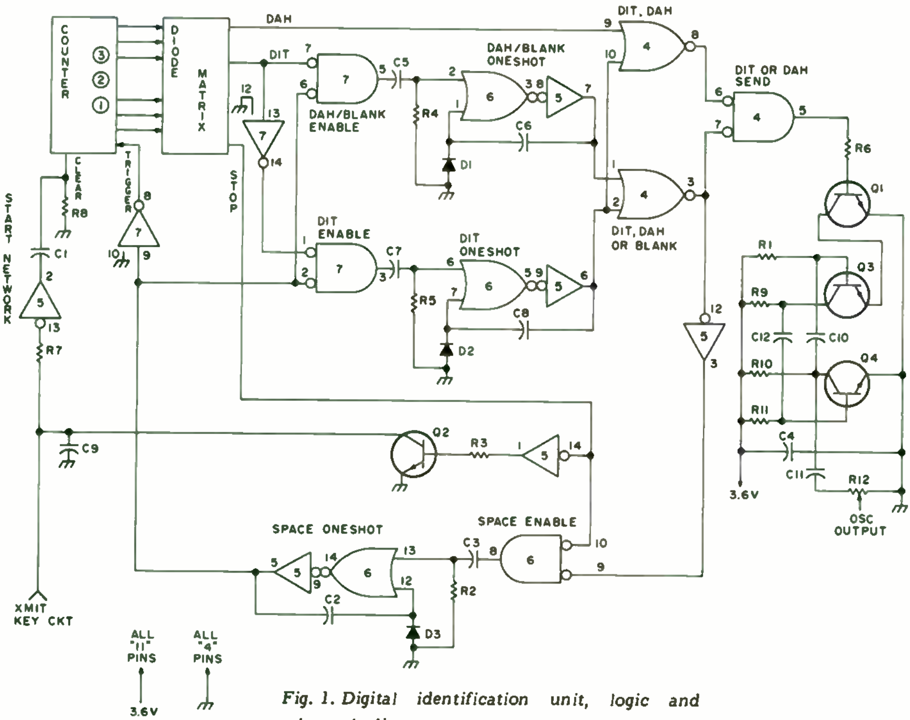


Fig. 1. Digital identification unit, logic and schematic diagram.

become high and the dit signal line will become low. When the space line becomes low, the "dah/blank enable" gate will send a pulse through capacitor C5, triggering the "dah/blank" one-shot. The dah/blank pulse would then go through the "dit or dah/blank" gate while the dah signal from the matrix would go through the "dit or dah" gate. These two gates would then enable the "dit or dah send" gate to trigger the oscillator for the dash period. The "space" one-shot is again triggered to advance the counter to the next unit.

If the diode matrix decodes the next sequence to be instruction 3 (send a blank), neither the dah nor dit line will become high. The same will occur as above for the dah except that when the signal reaches the coincidence gates the "dit or dah" gate will not be enabled. Thus the oscillator will not be keyed. This generates the blank period which is put between letters. (Example: DE = dah-space-dit-space-dit-blank-dit-blank) Again the loop through the "space" one-shot is triggered and the counter is advanced to the next unit of information.

The counter is advanced each time a unit of information is sent until it is advanced to the "stop" instruction. This instruction causes a blank to be automatically sent and stops the "space enable" gate from triggering the space one-shot. The DIU remains in the stop state until a reset pulse is sent to the counter from the transmitter keying circuit and the whole process starts over again.

Of course the DIU works much faster than it can be described. Depending on the component values selected for C2, C6, and C8, the DIU can function at any reasonable speed. The particular values used in the prototype and listed for Fig. 1 (see Table I) causes the unit to identify at the rate of 42 wpm (2 seconds for DE W6FNO). If a faster or slower rate is desired, capacitor values should be changed accordingly. It should be noted, however, that C8 must be three times as large as C2 and C6 to give the proper character formation. This is a critical relationship and follows from the fact that a space and dit are identical in time length while a dah or blank is three times the length of a dit.

Transistor Q2 is used to lock on the transmitter keying circuit while the digital identification unit is sending its identification code. If a timer is used in conjunction with the identifier, the transmitter will be

Table I. Parts list for DIU Logic Board and Diode Matrix

DIU Logic Parts	
R2-8	10 K ¼W
R9, R10	3.3K ¼W
	R1, R11r33K ¼W
Ç1, C3, C5, C7, C9	.05 µfd disc 25V
C10, C12	.008 µfd disc 25V (1000 Hz)
C2*, C1, C8*	10 µfd/15V
C4, C6*	30 µfd/15V
	*Most change in direct ratio
IC1-3	Motorola 791P
IC4,6,7	Motorola 724P
IC5	Motorola 789P
Q1-4	2M3415 or equivalent
D1-3	1N34 or equivalent
IC sockets wire-wrap type	Vector R-714
22 pin PC socket	Vector R-644
Matrix Parts	
10-20 resistors	3.2K ¼W
60-100 diodes	ge or si (all same type)
22-pin PC socket	Vector R-6⁄

0	1	2	3	4	5	6	7	8	9	10	11	12	13
X	—	•	•	X	•	X	X	•	—	—	X	—	•

14	15	16	17	18	19	20	21	22	23
•	•	•	X	•	•	—	•	X	—

24	25	26	27	28	29
•	X	—	—	—	Stop

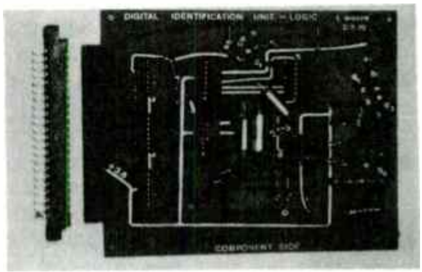
Fig. 2. Unit breakdown diagram.

keyed for the duration of the identification every time the DIU is reset. This allows a complete sending of the identification regardless of whether the transmitter remains keyed by an external circuit such as a COR (carrier-operated relay) or not. If this feature is not desired, Q2 should not be installed.

Transistor Q1 is used to key the oscillator, while Q3 and Q4 – along with the feedback and bias networks – make up the oscillator. The oscillator was designed to be fed directly to the grid of the modulator in the transmitter.

The Diode Matrix

Up until now very little has been said about the diode matrix other than the fact that it determines what instruction to give the keying logic. The actual construction of the matrix can be considerably simplified and consequently cheaper. Up to 70% of the diodes necessary for the diode matrix can be eliminated by using mathematics. A much more sophisticated, economical, and space-saving layout can be achieved using Boolean algebra. Thanks to Mr. Karnaugh, it is not necessary to give a complete discussion on Boolean algebra.



The IC logic board is shown here from the component side. Note the use of IC receptacles, which simplifies test, checkout, and replacement.

The Karnaugh map is a device for mechanically determining the mathematical equivalent of the diode matrix. For the purpose of this discussion the MCW message will be "DE W6FNO." Of course, any other message can be developed by this method and consequently this discussion may be used for developing any matrix logic.

The first step in determining the diode matrix for the message is to break up the message into the units to be sent: • = dit, – = dah, x = blank. This is shown in the breakdown diagram Fig. 2.

It is seen that 30 units of message will be sent (0 is actually used for a blank). To convert units 0 to 29 into a diode matrix, the Karnaugh map is used (see Fig. 3).

The numbers in the boxes correspond to the decimal equivalent to units on the output of the counter. The numbers across the top and along the side of the chart correspond to the binary output of the flip-flops – 1 for true or 0 for false. The letters written diagonally in the top left corner refer to the six flip-flops. Example: Box 17 has flip-flop A tue, B false, C false, D false, E true, and F false. Written in Boolean form, 17 would be represented by AB'C'D'EF', where the apostrophe after the letter indicates that the flip-flop is false and, conversely, a letter without an apostrophe is true.

To simplify the matrix, a Karnaugh map is constructed separately (Figs. 3 and 4) for the dits and dahs to be sent. From Fig. 2, units 2, 3, 5, 8, 13, 14, 15, 16, 18, 19, 21, and 24 represent the dits to be sent in the message. In the dit Karnaugh map (Fig. 4) a 1 is placed in each box corresponding to the number. An X (not the X which represents a blank) is placed in all boxes after the stop code number. These are "don't care" conditions because the counter will not count to these codes.

From the dit Karnaugh map (Fig. 5) it can be seen that the third unit of information is a dit and that flip-flop A is true, B is true, C is false, D is false, E is false, and F is false, or ABC'D'E'F'. To put this in matrix form, the Boolean algebra tells us that this dit would be represented by a diode connected to Q_a lead (the true lead of flip-flop A), another to Q_b, another to

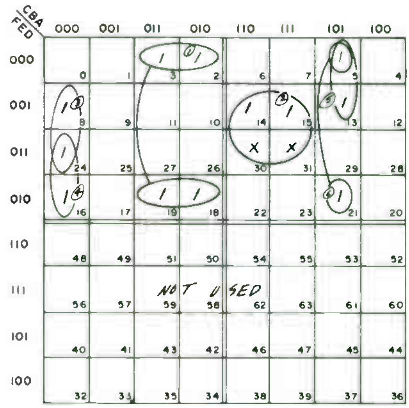


Fig. 4. Karnaugh map of dits to be generated in DE W6FNO.

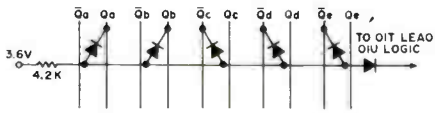


Fig. 5. Unit 3 information – dit.

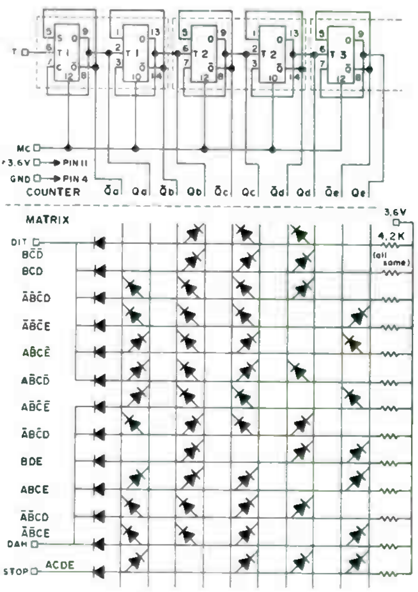


Fig. 6. DIU matrix and counter

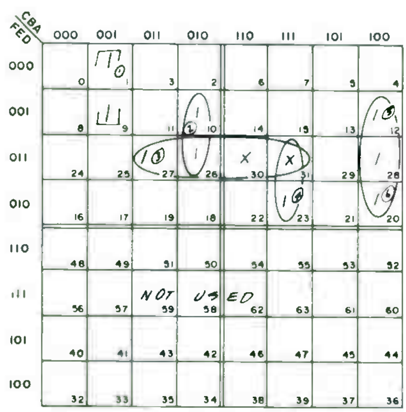


Fig. 3. Karnaugh map of dahs to be generated in DE W6FNO.

Q_c' (the false lead of flip-flop C), another to Q_d', another to Q_e'. Since there are only 30 units of information, flip-flop F is not used. A line may be used over any of the symbols to indicate the same thing as an apostrophe. It would normally take six diodes (seven when the F flip-flop is used) to send this unit of information. See Fig. 6.

Actually, it would take six diodes (seven when the F flip-flop is used) for each unit of information in the message or 29 x 6 = 174 diodes. This includes the diodes

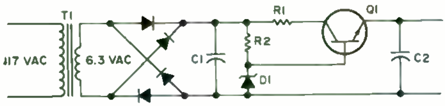


Fig. 7. DIU regulated power supply.

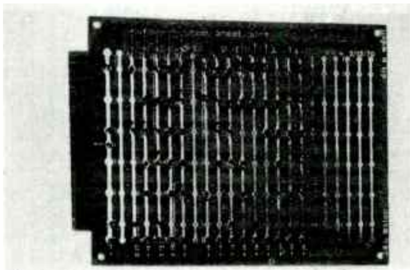
needed to *or* the dahs together and the dits together. This is where the Karnaugh map saves diodes. Again on the amp in Fig. 4 any adjacent box or any box that changes just one variable from another box eliminated that variable. Boxes 8 and 24 simplify to A'B'C'D, eliminating the E flip-flop altogether. Boxes 3, 2, 19, and 18 also simplify since they change one variable at a time or BC'D'. Note that not only is 3 represented by BC'D' but also 2, 19, and 18, resulting in a savings of 20 diodes – 4 numbers x (5+1 or) –(3+1 or used) = 20. 14 and 15 combine with "don't cares" 30 and 31 to equal BCD. The final expression though not the only expression that will work) for the dits is BC'D' + BCD + A'B'C'D + A'B'C'E + AB'CE' + AB'CD'.

Figure 3 was used to develop the dah equation which is AB'C'E' + A'BC'D + BDE + ABCE + A'B'CD + A'B'CE. 28 diodes were used to develop the dit matrix, 29 were used for the dah matrix, and 5 were used for the stop code, giving a total of 62 diodes for the entire matrix. Quite a few less than 174!

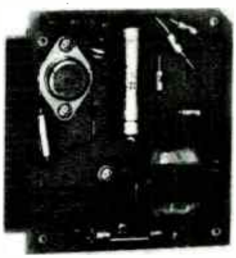
The final matrix appears in Fig. 6 for the message DE W6FNO. Note that any matrix of this magnitude can be determined by the above method. To expand to 64 units of information the mirror image of the first 32 units is used in the Karnaugh map. In Figs. 3 and 4 the "not used" portion would be used. The upper portion of Fig. 6 illustrates the wiring of the counter; note that it mates to the leads of the matrix.

Construction of the DIU

Since the publication of the first article on the DIU numerous people have sought printed circuit boards for the unit. In order to obtain the DIU logic, matrix, or power supply boards, write to Keith Whitehurst, Box 538, Claremont, Calif. 91711.



This photo of the call letter matrix board shows the layout of the diodes for W6FNO. The same board is used for other calls, though the diode placement will vary.



Regulated power supply module provides +3.6V to digital, identifies logic and matrix boards.

Table II. DIU Regulated Power Supply Parts List.

T_1	**Stancor P6465 117 to 6.3V, 600 mA.**
D_1	**3.9V Zener 1N748 Motorola**
D	**1A 12V diodes or HEP Bridge**
C_1	**1000 μF at 12V**
C_2	**200 μF at 12V**
R_1	**10 Ω at 6.3V 5 Ω at 3.2V**
R_2	**220 Ω at 6.3V input, 110 Ω at 3.2V input***
Q_1	**2N4921 or HEP 245**

Power Supply

Figure 7 illustrates the schematic of the DIU power supply. Table II lists parts required. Any power supply, however, may be used if the power output is 3.6 V with less than 5% ripple (including voltage spikes).

Installation

The signals normally received and sent to and from the DIU should meet the following criteria:

1. From power supply – 3.6V dc, well filtered and regulated.
2. From transmitter keying circuit – 0V, transmitter keyed; approximately 6V transmitter unkeyed (filtered).
3. To transmitter keying circuit – identifier off: 10 MΩ; identifier keyed: 10Ω.
4. To modulator circuit – high impedance DIU oscillator output.

Note that all dc input lines to the DIU logic should be filtered. In some relay circuits the output of a bridge rectifier is used to directly key the transmitter relay. So the pulsating dc does not key the digital identification unit, a 60 μF capacitor (or greater) should be placed across the relay supply.

It is worthwhile to note that the original prototype, after two and a half years, is still operational atop Johnstone Peak in San Dimas, California, sending out for all to hear – DE W6FNO.

An IC Pulser for the Amateur Experimenter

Hank Olson, W6KXN

In designing an amateur pulse generator, the first consideration has to be compatibility with ICs. From the standpoint of convenience, economy, or versatility, it has become unattractive to build *anything* digital without IC logic. What this means to pulser design is that the pulser should be compatible with the types of digital ICs one finds himself using most.

Although RTL is the most widely used IC logic family in amateur circles, it is extremely doubtful whether this situation will last. The two types of "current-sinking" logic (DTL and TTL) are far out in front of RTL in industry usage and gaining daily. The reasons for industry preference of DTL and TTL over RTL are several: better noise immunity, higher speed, larger fanout, and a larger selection of devices around which to design.

This pulser was designed so that its output would be compatible with DTL or TTL and also with ECL. The implementation of these two outputs is simplified by use of ICs that are designed for *buffer* and *clock driver* service. (The pulse generator section is shown in Fig. 1, and the output section in Fig. 2.)

The free-running multivibrator which determines the basic "rep" rate, the *delay* one-shot, and the *pulse length* one-shot use HEP versions of MECL ICs. The entire basic pulse-forming system is similar to that of my previous article (*Pulse Generator for the Amateur*, 73, Nov 1967). A number of improvements have been made, however.

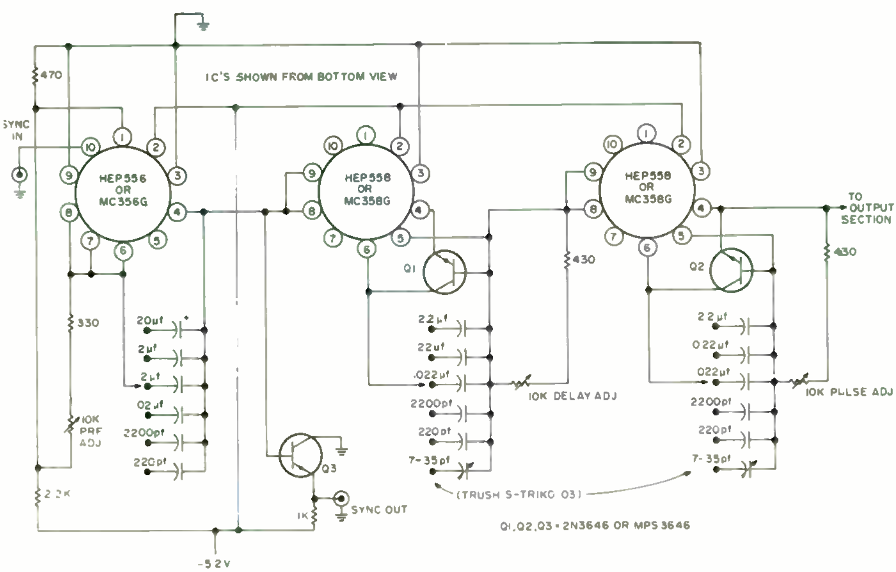


Fig. 1. Pulse generator portion of IC pulser.

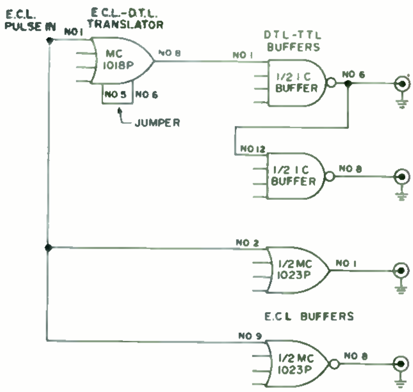


Fig. 2. Output section of pulser.

By using an etched circuit board, it is possible to reduce lead lengths and get shorter pulses. Longer pulses are also provided by extending the range at the other end of the range switches. An additional transistor has been added to both the *delay* and the *pulse length* one-shot. These transistors allow the one-shots to recover more quickly, providing more stable operation for pulse lengths approaching the period of the basic rep rate.

Since one of the two outputs is to be ECL-compatible, the basic pulse-forming section is powered from −5.2V. This means that V_{ee} (terminal 2 of the HEP 556 and 558s) is connected to −5.2V, and V_{cc} (terminal 3) is connected to ground.

The DTL–TTL portion of the pulser, of course, requires +5V, which is also provided. By carefully arranging the pin numbers of the 14-pin socket, any one of three different families of logic may be used in the DTL–TTL buffer position. The least expensive is the DTL buffer (MC832P). The SN7440N is a TTL buffer that is compatible pin-for-pin. And it can also be used at a slightly higher cost. The MC3025P, member of a different TTL line (MTTL III), can also be plugged in. The cost of the MC3025P is a bit higher than the SN7440N; each step up in cost corresponds to an increase in speed. For two of the types of buffers mentioned above, there are numerous replacements made by different semiconductor manufacturers. These are listed in Table I; they are *not* different families, but rather second-source items.

Table I.

Specified Type	Manufacturer	Equivalent	
SN7440N	Texas Inst.	DM 8040N	National Semiconductor
SN7440N	Texas Inst.	USN 7440A	Sprague
SN7440N	Texas Inst.	N 7440A	Signetics
SN7440N	Texas Inst.	FJH 141	Amperex
SN7440N	Texas Inst.	MC7440P	Motorola
MC 832P	Motorola	DT μL 932	Fairchild
MC 832P	Motorola	SW932-2	Stewart Warner
MC 832P	Motorola	DTL 932	Sperry
MC 832P	Motorola	PD 9932	Philco
MC 832P	Motorola	SN15 832N	Texas Inst.
MC 832P	Motorola	S 9323	Sylvania
MC 832P	Motorola	MIC 932	ITT
MC 832P	Motorola	HSC 932	Hughes
MC 832P	Motorola	CD2306E/832	RCA
MC 832P	Motorola	RM 932	Raytheon

Interfacing *between* the ECL section and the DTL–TTL buffer is an MC1018P *translator* IC. This IC requires ground, +5V, and −5.2V for supply connections. A simplified circuit of the MC1018P is shown in Fig. 3 as used in the pulser. Of course, such a circuit *could* be built of discrete components, but not as simply and inexpensively as using the MC1018P. If the MC1018P is hard to find, the MC1018L may be used. It is the same "chip" in a *ceramic* case, and is being sold at a slightly higher price than the plastic unit.

All the components are mounted on the side of the board away from the panel.

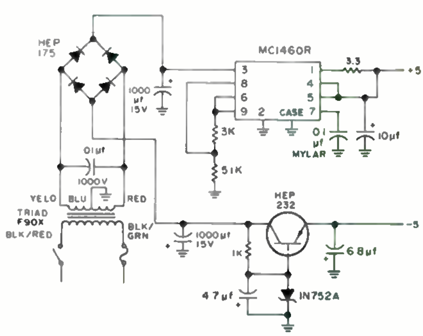


Fig. 4. Power supply for new ECL pulser.

except the three HEP ICs. These three ICs were reverse-mounted to ease layout, allowing shorter trace lengths.

The 7-7/8 x 7-7/8 in. panel is made to fit a Bud CD-1480 cabinet which has enough panel space for all the controls and jacks. The power transformer is mounted (off the board) inside this cabinet.

As to the DTL-TTL output capability of the pulser, it is dependent on the exact type of output stage. Each of the ECL outputs will drive 24 ECL gates (has an ECL fanout of 24). But the fanout of each of the two current-sinking logic outputs is as follows: MC832P – 24 DTL load units, SN7440N – 29 TTL (SN7400N series) load units, MC3025P – 19 MTTL III load units. The load units are not the same for these three current-sinking families, so it is best to use the type of output IC for the sort of family *you* use most. It is quite all right to use any of the types of current-sinking logic ICs with the pulser (no matter which IC is used in the output stage), but some reduction of fanout may be experienced with certain combinations.

Performance

The pulser will produce pulses (and delays) from about 50 ns to 30 ms. DELAY and PULSE switches each have six positions, and the variable control associated with each continuously varies each over at least ten-to-one. Pulse repetition rate is adjustable from about 0.5 Hz to 1 MHz with a six-position switch and variable control. The labeling of the three panel switches was minimal, with 200 kHz to 2 Hz (in that order, so that the period steps in the same direction as delay and pulse width) for the rep rate. These labeled frequencies correspond to frequencies *within* each switch position, not the center or either end point. The DELAY and PULSE switch positions were similarly labeled from 0.1 μs to 10 ms for simplicity.

The regulated +5V and –5.2V are provided by a common power transformer and rectifier. Since the centertap of the transformer is grounded, the circuit may be considered as two full-wave rectifiers (one positive and one negative) across the same

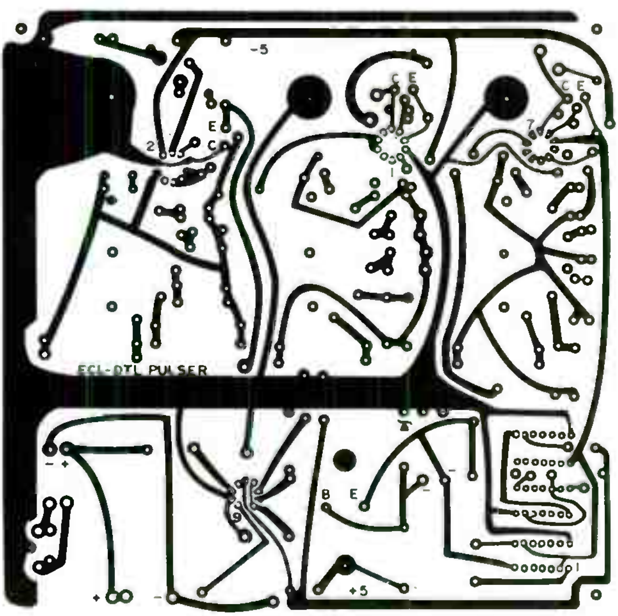
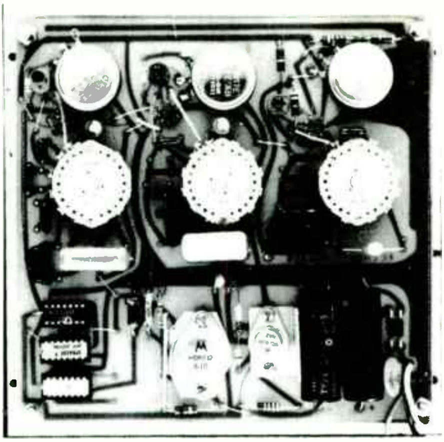


Fig. 5. PC layout of IC pulser



Circuit board, parts side.

transformer. Of course, the diode configuration comes out the same as a full-wave bridge; so an IC bridge (HEP 175) is used.

The positive regulator is an MC1460R, an inexpensive IC that provides excellent regulation with few external components. The negative regulator is an emitter follower with a 5.6V zener and a germanium PNP transistor. Since the base-to-emitter drop of germanium transistors is about 0.3V, 5.6V minus 0.3V gives us close to the −5.2V required for the ECL−ICs.

The output section for the ECL compatible pulses is provided by an MC1023P. Although this IC is billed as a *clock driver*, it makes the best output stage of any of the MECL II series. This is true because of its exceptional ability to drive capacitive loads.

Both the MC1460R and the HEP 232 (PNP power transistor) are diamond-shaped and meant to be fastened to a heatsink. They each have an aluminum bracket attached to them to fulfill this requirement for a dissipator.

The entire circuit of the pulser is built on an etched circuit board. In fact, the wafer switches are assembled so that the board is clamped into the switch assembly. The shafts of the switches are cut to 7/8 in. to extend out the same length as the pot shafts. The entire board is then mounted to a panel using 1-1/8 in. spacers. Some care must be exercised not to allow the spacers to short any of the traces of the circuit board to ground. This can be insured by using fiber washers between board and spacers. The one spacer in the power supply corner is intended to connect the board ground to the panel, so no fiber washer should be used at that corner.

Construction

The entire pulser is built on an etched circuit board, including the power supply, whose circuit appears in Fig. 4. Figure 5 is a half-scale copy of the board.

Since the etched board method has been used, some specialization in components is necessary. The Cornell-Dubilier BR1000-15

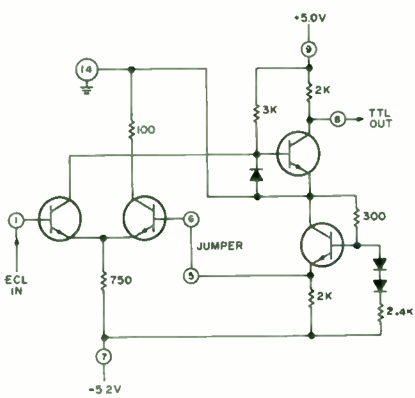
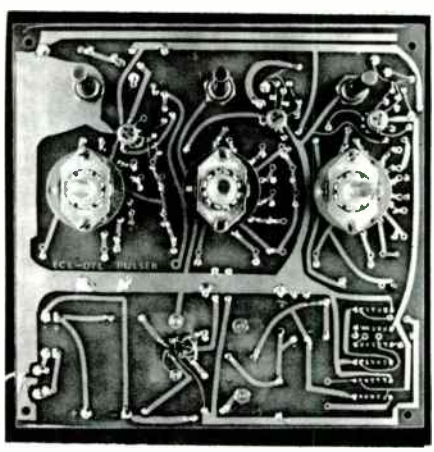


Fig. 3. Simplified ECL−TTL translator IC circuit.



Circuit board, trace side.

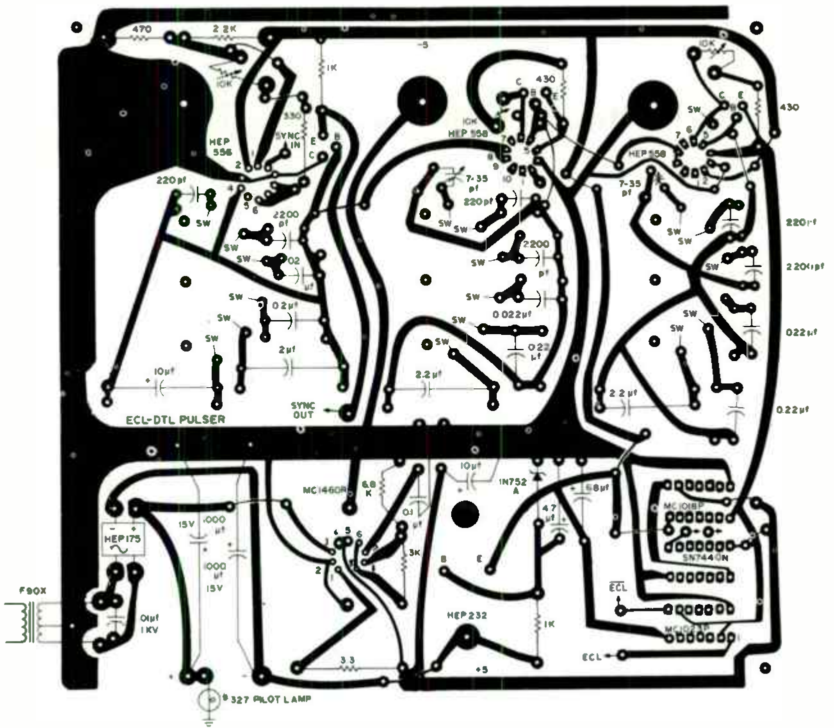


Fig. 6. Component placement (etched side of board shown).

filter capacitors are physically smaller than most other brands of the same capacity and voltage rating, and so are best used. Similarly, the switches are Centralab PA-1 and PA-300 combinations, and the two variable capacitors are Trush S-Tri ko 03 types. The IC sockets were HEP 451 for the round-can types and Methode M1141 for the dual-inline types. Figure 6 is the layout of all components on the board.

Both circuit board and a kit of parts are available from Project Supply Co., Box 555, Tempe AZ 85281.

Bibliography

1. Olson, H., *A Pulse Generator for the Amateur*, 73, Nov 67.
2. Renschler, E., *Design of Monostable Multivibrators Using MECL Integrated Circuits*, Motorola Application Note AN233.
3. Texas Instruments, *Integrated Circuits Catalog 1967–68*.

. . . W6GXN ■

IC/Photocell Compressor/AGC Unit

John J. Schultz, W2EEY

One problem with many compressors or audio agc units is that they cannot be conveniently built into an existing transmitter or receiver, since the amplifier and control sections of the unit cannot be readily separated. Consequently, such units are usually placed in separate enclosures and mounted in the microphone lead to a transmitter or in the loudspeaker or headset audio output leads of a receiver.

Most such conventional compressor/ audio agc circuits use transistor stages for both amplification and control functions. It is difficult to separate the stages physically unless additional coupling stages are added, so that the amplifying and control functions can each be located where each can function best and where power and space in a transmitter or receiver are most readily available (see Fig. 1A).

The IC/photocell unit to be described overcomes most of these limitations. The amplifying and control functions can be separated as desired (Fig. 1B) through the isolation medium of a photocell-lamp module. Although described as an audio compressor/agc unit, the photocell-lamp module allows control to be achieved of an rf stage as well through its biasing network. Only the derivation of the control function is restricted to an audio frequency point in a receiver or transmitter, since the IC amplifier used operates at audio frequencies. In an rf amplifier, an rf actuated and rf control compressor could be achieved. Although it has not yet been tried, it would seem that this latter approach might produce a highly effective SSB rf level compressor without the need for two sideband filters as is required with SSB rf level dipping circuits.

Besides separation of the amplifying and control stages, the use of a photocell-lamp module also overcomes the noise build-up problem associated with conventional audio compressors. With such compressors, noise build-up occurs at the output of the compressor during speech pauses because, without speech input, the gain of the compressor rises to a high value and amplifies the self-noise of the first stage in the compressor to a high value. Choosing a long time constant in the gain control stage of the compressor will act to suppress such noise build-up, but one is quickly limited as to how far this approach can be used. If the time constant is made too long, low level speech inputs to the compressor

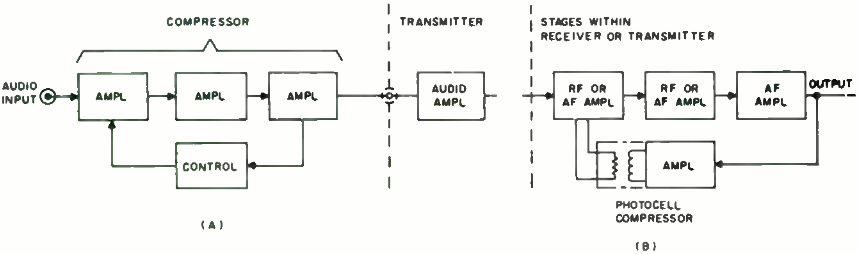


Fig. 1. Amplifier stages of usual compressor are in series with transmitter or audio chain (A). Photocell compressor (B) works parallel to controlled stages and the photocell module provides both noise isolation from compressor amplifier and feedback isolation between controlled amplifier stages.

following a higher intensity input will not be amplified sufficiently. Since the amplifier stages of the photocell compressor need not be in series with the audio chain in a receiver or transmitter, its noise output is not reflected in the controlled stages. Also, the thermal-photoelectric interface within the photocell-lamp module prevents the coupling of noise or spurious frequencies from the audio amplifier of the compressor. In fact, the audio amplifier portion of the photocell compressor can be rather simple and produce considerable distortion without affecting the units' performance. The only real requirement is that it produce a power output sufficient to drive the photocell module which is directly proportional to the audio level at the sampling point within a receiver or transmitter.

The use of a photocell also provides feedback isolation between the sampling point and the controlled point in a receiver or transmitter. Since no direct electrical connection is involved between the two points (except for the minor capacitance between the lamp and photocell in the module), one does not have to worry that the feedback loop has a time constant greater than the lowest frequency at which the gain of the controlled stages is greater than unity – a criterion for stability in an electrically coupled compressor feedback loop.

Circuit Description

Figure 2 shows the schematic diagram of the compressor/agc unit. The integrated circuit used is an RCA CA3020 which can produce about 500 mW output. Various other audio amplifier integrated circuits can be used such as the GE PA234 and also various surplus operational amplifiers can be used. A module type audio amplifier or discrete stage transistor amplifier can also be used. The prime criterion is that the amplifier produce enough power output to properly drive the photocell – from 150 to 250 mW.

The external components used with the IC are chosen primarily to give sufficient power output rather than maximum undis-

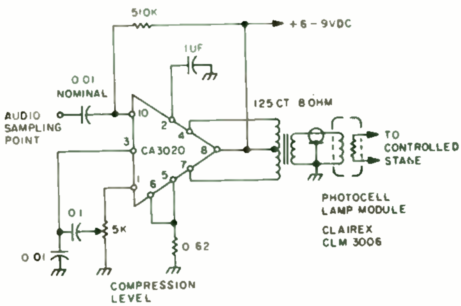
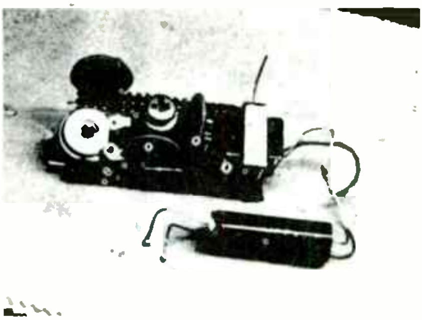


Fig. 2. Photocell compressor/agc circuit schematic. Voltage rating of capacitor to terminal 10 must be chosen to protect unit from voltage found at sampling point. DC operating voltage need not be supplied from an extremely well filtered source since audio quality of amplifier is not significant.

torted power output, as would be the case if the IC were used for strictly audio reproduction. A 5 kΩ potentiometer between stages in the IC acts as a compressor level control. No input potentiometer is used due to the fact that even if some slight overdrive of the input should occur, it would not be significant in this application. The output transformer secondary drives the lamp of the photocell module. The module may be placed any reasonable distance from the amplifier and connected to it by shielded audio cable. It is not necessary to rectify the output of the amplifier, since the thermal inertia of the lamp in the photocell module will "wash out" instantaneous level variations. The photocell module itself can be any one of a number of Clairex or General Electric units which sell for $3–$4. The Clairex CLM3006 unit works well for a general variety of applications. Its lamp drive requirements are 6 volts at 40 mA maximum, and the resistive element in it will vary from a value of over 100 kΩ when the lamp is not excited to about 200Ω when the lamp is fully driven. The photocell module itself also provides a sort of automatic delay action, since the change in resistance is not linear with lamp drive but generally is slower at low lamp drive levels. Thus, depending upon how the resistive element is placed in a circuit, compression action increases with higher signal levels.



Assembly of compressor/agc units on Vectorboard. Photocell unit is shown next to output transformer. Compressor level pot is at other end of Vectorboard, and IC unit is in middle with circuit components grouped around it.

Construction

The photograph shows how a typical photocell audio amplifier driver circuit can be assembled on a piece of Vectorboard. The photocell module would, of course, be located remotely from the amplifier circuit. A fin-type heat dissipation device should be used on the IC, depending upon the manufacturer's recommendation for the unit used. A PC board type trim potentiometer, located to the right of the IC, acts as the compression level control. Since the control does not have to be readjusted normally after initial setup, it can be left as a component on the Vectorboard. If it is desired to have some means to continuously control the compression level as well as turn off the compression action, the potentiometer – brought out as a panel control – will perform both functions. The output transformer on the left is a conventional miniature transistor type with an 8–16Ω secondary.

Placement

The objective in placing the resistive element of the photocell module in a receiver or transmitter circuit is to make maximum use of its wide resistance change. At the same time, the resistive element cannot be used such that it must dissipate over 1/10 watt. One possible placement for the resistive element is shown in Fig. 3A, where it is used as part of a voltage divider network in a fairly high impedance circuit. In some units the microphone input itself or the interstage coupling point after the first audio amplifier might be used. Theoretically, the large resistance range change of the photocell resistive unit could produce voltage output changes in a high impedance divider network of over 40 dB. Another possible placement of the resistive element which also prevents dc from flowing through it is shown in Fig. 3B. As the photocell module is driven harder, the resistive element increasingly shorts to ground the audio bypass capacitor. Such a capacitor would be placed between audio stages in a receiver or transmitter to shunt the audio signal to ground as compressor action takes place. There are various other placements possible for the resistive element, such as in the bias line to the stages it is desired to control or even as a shunt element across a low level audio transformer. The placement that will achieve the best control can be quickly determined by experiment.

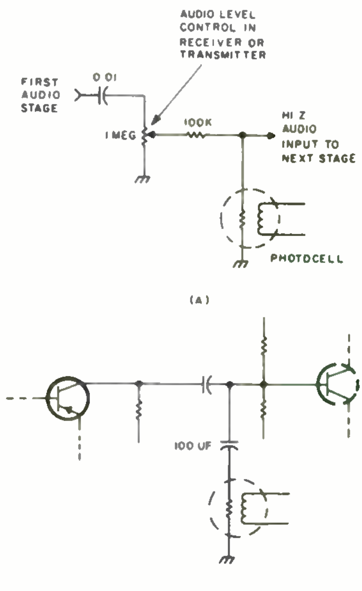


Fig. 3. Possibilities for using the resistive section of the photocell for control of the signal amplitude in an audio amplifier. Voltage divider method (A) and capacitance bypass method (B).

The sampling point for the compressor's audio amplifier input is usually taken at some interstage point towards the high level end of the audio chain in a unit. The sampling connection should have no effect upon the normal operation of the audio amplifier in a unit. If possible, the sampling point should be chosen such that the audio level control in a unit is located earlier in the audio chain. The controlled point can be either before or after the audio level control.

Operation

With the compressor control initially set for minimum gain, the audio level control in a unit is set for the highest desired audio level. The compressor control is then advanced until the audio level decreases significantly. The audio level control is again advanced until the desired output level is achieved. Basically, the circuit should then be set, although one will have to initially adjust back and forth between the audio level and compressor controls until the optimum compression range is obtained.

Summary

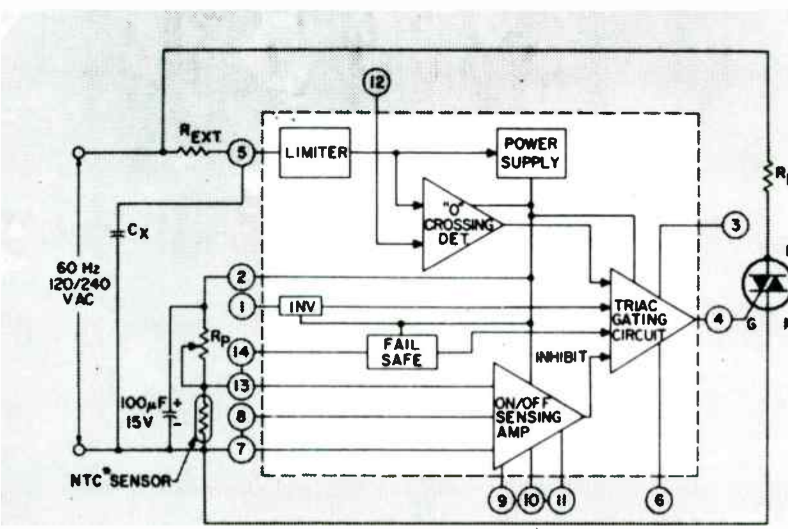
The photocell module is a very versatile unit for compressor and agc control circuits. Many more circuits are possible with it than the one illustrated here. For instance, if one wanted to use a varying dc voltage as the control source for the module this can be done by using a dc amplifier and direct output coupling in place of the audio IC amplifier shown. With some low power vacuum tube circuits it is also possible to place the lamp of the photocell module in series with a cathode resistor and achieve direct control without any amplifier at all.

AC Switching with Self-powered IC'S

Eugene L. Klein Jr., W2FBW

A cure-all to ac switching problems? Not so! But RCA's little CA3059 integrated circuit device goes a long way in eliminating many of the bugs we inherited with Tesla's genius. Formerly known as a monolithic silicon zero-voltage switch, it is now more handily identified by RCA's "CA" number. It is no bigger than a 2W resistor, yet it contains its own power supply and all the other functions shown in Fig. 1. When used with a half-dozen or so other common components, it does many wondrous things. Several are described in this article.

Basically, we would like to accomplish switching when there is no incoming voltage; that is, when the power line voltage crosses zero. This happens twice each cycle, or 120 times per second at 60 Hz. If we switch at this precise moment, no current is flowing through the switch; thus, radio-frequency interference is eliminated, as is contact wear. Incandescent devices, such as tube heaters and pilot lights, which have very low cold-resistance, are heated up "gradually" during a half-cycle if voltage is applied at one of the "zero" crossing times. They are, therefore, not subjected to a destructive high-current surge as before. In the same manner, we reduce the high surge current and minimize the peak inverse voltage imposed on our rectifiers



* NTC = NEGATIVE TEMPERATURE COEFFICIENT

Fig. 1. Functional block diagram of CA3059. Few external components are required in this basic circuit when used to trigger the gate of a triac. The sensor may be a temperature-sensitive thermistor, a photoelectric cell, or a simple off-on switch.

when initially turning on a power supply. The zero-crossing detector in the CA3059 synchronizes the output pulses of its circuit with the time of zero-voltage in the ac cycle. Figure 2 shows this relationship.

But to be useful in switching normal loads, an external power device is needed. Thyristors such as SCRs (half-wave) or triacs (full-wave) are ideal for this purpose. Either one is readily triggered by the 100 μs pulses from the CA3059. The zero-voltage switch is designed primarily to trigger a thyristor that switches a resistive load.

Because the output pulse supplied by the CA3059 is of short duration, the latching current of the thyristor becomes a significant factor in determining whether other types of loads can be switched. (The latching-current value determines whether the thyristor will remain in conduction after the trigger pulse is removed.) Provisions are included in the CA3059 to also accommodate inductive loads.

For example, for load currents that are less than approximately 4A rms (or that are slightly inductive), it is possible to retard the output pulse with respect to the zero-voltage crossing by insertion of capacitor C_x from terminal 5 to terminal 7 as shown in Fig. 1. The insertion of capacitor C_x permits switching of loads that have a slight inductive component and that are greater power than approximately 200W. For loads less than 200W, it is recommended that the user employ the RCA-40526 sensitive-gate triac with the CA3059 because of the low latching-current requirement of this triac.

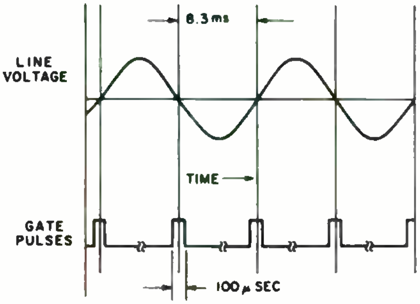


Fig. 2. Timing relationship between output pulses and the ac line voltage. Notice that the gate pulses are present to trigger the triac at zero line voltage.

Figure 3 shows an application which is useful in switching noninductive loads. Lamps and electric heaters can be controlled in this circuit by a simple switch, photoelectric cell, or temperature-sensing thermistor.

Inductive loads up to 200W can be controlled by the configuration shown in Fig. 4. In this case, terminal 12 is connected to terminal 7, and the zero-crossing detector is inhibited. Whether a "high" or "low" voltage is produced at terminal 4 is then dependent only upon the state of the differential comparator within the CA3059 integrated circuit, and not upon the zero crossing of the incoming line voltage. Of course, in this mode of operation, the CA3059 no longer operates as a zero-voltage switch. However, for many applications that involve the switching of low-current inductive loads, the amount of RFI generated can frequently be tolerated. This curcuitry is particularly useful as a differential comparator. Such comparators have found widespread use as limit detectors which compare two analog input signals and provide a go/no-go output, depending upon the relative magnitudes of these signals. In many industrial control applications, a high-resolution, fast switching unit is not essential. The CA3059 is ideally suited for use in such applications.

The chart below compares some of the operating characteristics of the CA3059, when used as a comparator, with a typical high-performance commercially available IC differential comparator.

PARAMETERS	CA3059	Typical IC Comparator
1. Sensitivity	50 mV	2 mV
2. Switching speed	>20 μs	90 ns
3. Output drive capability	4.5V at ≤4 mA	3.2V at ≤5 mA

The CA3059 can be used as a simple solid-state switching device that permits ac currents to be turned on or off with a minimum of electrical transients and circuit noise.

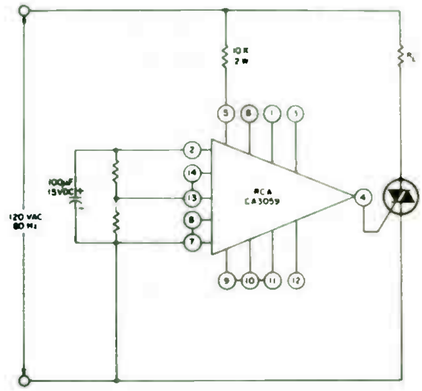


Fig. 3. Controller for resistive loads. The RCA 2N5444 triac can be used for load currents up to 40A. The RCA 40668 triac will switch intermediate loads and the 40526 will handle lighter loads and those which are somewhat inductive.

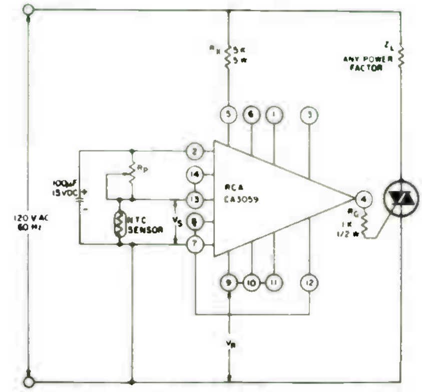


Fig. 4. Differential comparator circuit. The load is switched on when the voltage difference between V_S and V_R becomes less than 50 μV. Note the jumper between terminals 7 and 12, which deactivates the anti-RFI feature.

The circuit shown in Fig. 5 is connected so that after switch S1 is closed, electronic logic waits until the power-line voltage reaches zero before power is applied to the load. Conversely, when the switch is opened, the load current continues until it reaches zero.

This circuit can thus switch a load at zero current regardless of whether it is resistive or inductive. The maximum load current that can be switched depends on the rating of the second triac. If it is an RCA-2N5444, an rms current of 40A can be switched.

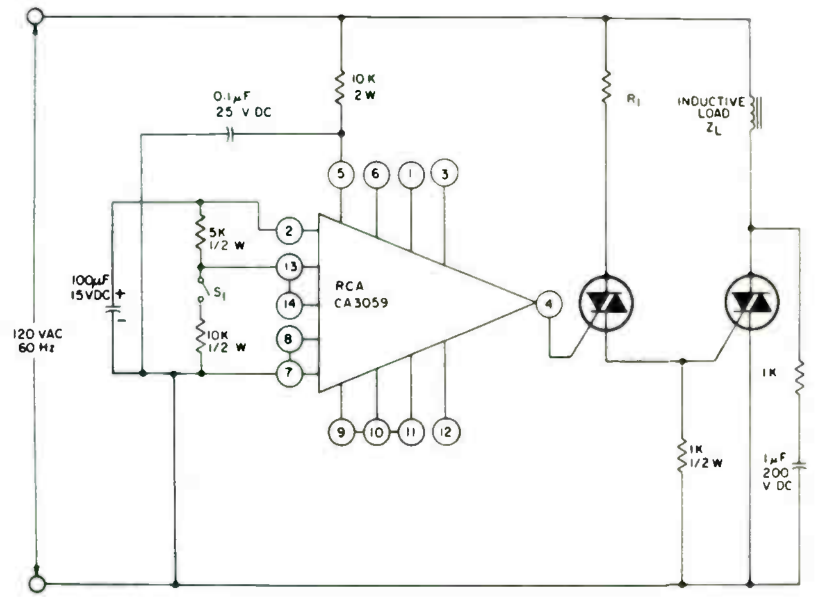


Fig. 5. Practical power switch. Heavy inductive loads can be handled by adding the 0.1 μF capacitor and by cascading the two triacs (RCA 40526 and2N5444). With the latter heavy-duty triac, R1 should be the highest suitable value less than 10 kΩ.

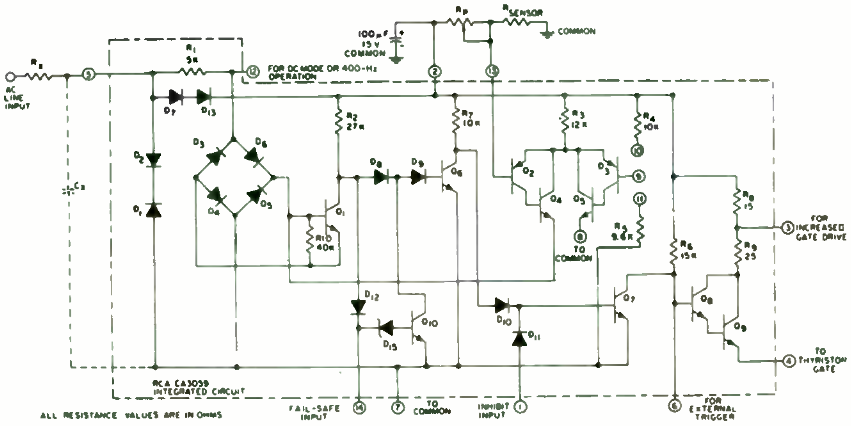


Fig. 6. Schematic diagram of CA3059 zero-voltage switch. This miniature device contains the equivalent of 10 transistors, 14 diodes, and 10 resistors.

Figure 6 is a schematic diagram of the CA3059 zero-voltage switch. This should prove helpful to the more advanced experimenter who wishes to design his own circuits for specific applications. Some of these may include:

- Relay control
- Heater control
- Valve control
- Lamp control
- Photosensitive control
- Synchronous switching of flashing lights
- Power one-shot control
- On-off motor switching
- Differential comparator

By referring to the schematic and to the functional block diagram of Fig. 1, we can identify the following circuits within the zero-voltage switch:

1. Limiter, power supply – permits operation directly from an ac line.
2. Differential on/off sensing amplifier – tests the condition of external sensors or command signals. Hysteresis or proportional-control capability may easily be implemented in this section.
3. Zero-crossing detector – synchronizes the output pulses of the circuit with the time of zero-voltage in the ac cycle to eliminate RFI when used with resistive loads.
4. Triac gating circuit – provides high-current pulses to the gate of the power controlling thyristor.

In addition, the CA3059 also provides a built-in protection circuit which may be connected to remove drive from the triac if the sensor opens or shorts. And thyristor firing may be inhibited through the action of an internal diode gate connected to terminal 1. In addition, high-power dc comparator operation is provided by overriding the action of the zero-crossing detector.

Additional information is available in RCA specification sheet File 397 and companion application note ICAN-4158.

...W2FBW ■

2 Watt, 6 Meter Transmitter Using the Heterodyne VFO

Bill Hoisington, K1CLL

This article describes the breadboard design, tuneup, and results of an rf power stage on 6 meters using the $2.95 Motorola 3-watt HEP-75 transistor (similar to the famous 2N3866). This rf stage is designed to work from an input power of 120 mW such as furnished by the 6 meter crystal-heterodyne-vfo circuit described in a previous issue of 73 Magazine. Two types of inputs are detailed, one using a single capacitor, and the other a matching network for use with any length of cable. At a dc input power of 2W using a 12V battery, a good clean watt of rf output power is obtained.

Circuit and Design Theory

Figure 1 shows the schematic, using the input matching network. The design of a VHF input circuit can take various forms, depending on whom you read, what you read, and the proposed use of the rig. In this case, we want sure-fire operation, an easy-to-build circuit, freedom from self-oscillation, smooth tuning, 12V operation, and "carrying" type portability for hill-topping emergency use, and mobile work, as well for use at home.

The circuit in Fig. 2 uses a capacitor from J1 to the base. It works, and should be sufficient in the case of an excess of

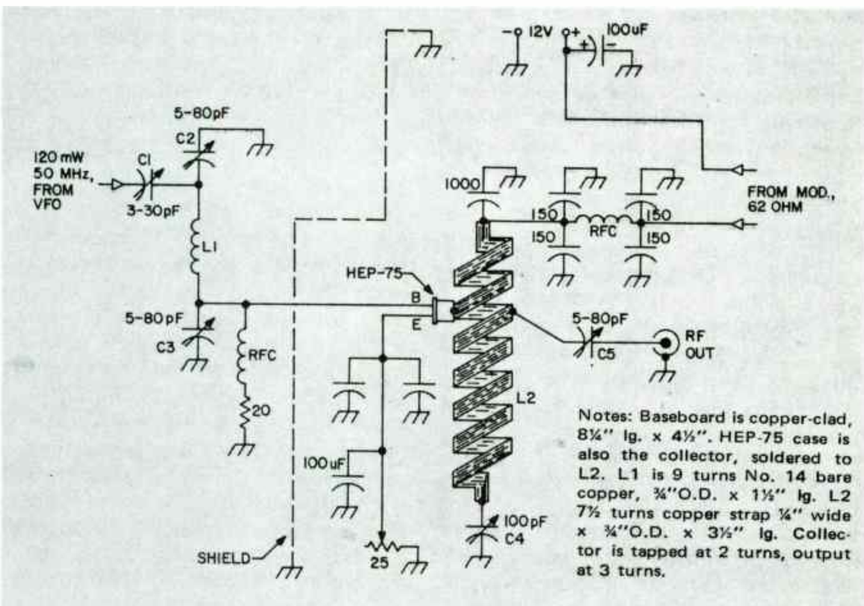


Fig. 1. 2 watt, crystal-heterodyne-VFO 6 meter rig

input power (not too likely), and where you don't mind adjusting cable length. However, if you're building up from onlv 120 mW of vfo output, you may want maximum power transfer along with the ability to match various cable lengths, and direct inputs (as in the case of packaging the entire rig in a box).

The circuit of Fig. 2 is reliable and good for a "quickie" any time. But it does not always furnish maximum drive or best input match unless the cable length and output tap on the vfo output inductor are adjusted. These latter are of course indications of mismatch, but for short lengths of cable there is little loss, so you can operate that way if needed.

DC Base Lockup

This nasty little trouble has not been mentioned in the rather large amount of literature persued here through recent years. I can just hear some know-it-all lads saying, "Ho, Ho, he's just discovered the Poniatowsky effect." Well, maybe so, but whatever its name is, it arrived here and I don't like it! Here's what happens: With an rf choke from the base to ground and *no* resistor, and an rf input driving the collector to about 200 mA, cutting off the rf excitation does not cut off the collector current. There is no rf involved and it seems to be a dc type of hangup. Just for fun I'll wait until someone tells me its name. In the meantime, back on the breadboard, the cure was easy. Just include a little resistance in the base circuit. That's why the 20Ω job is there.

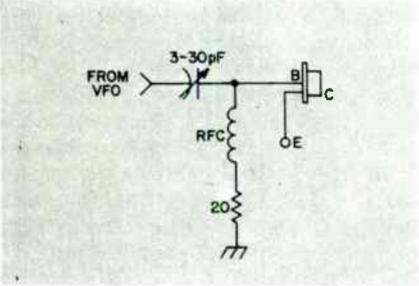


Fig. 2. Simplified input circuit, 2 watt, 6 meters. (Temporary)

Input Matching and Base Circuit

After trying out various combinations of circuits as shown by RCA, Motorola, Fairchild, and others, the circuit of Fig. 1 was judged best. Inductor L1 can also be an air-wound coil 5/8 of an inch O.D., 8 turns per inch, with 5½ turns (not too critical).

Just a word of caution here about overloading receiver inputs while testing transmitters. While checking with the lab receiver, with its antenna only 10 in. away from the 1W rf output of the amplifier, the receiver was completely blocked out and detuned by the rf. The addition of only 1 in. of wire as a test antenna brought things back to normal. A very peculiar effect; expect almost anything when you start to run power. In solid-state VHF, "power" refers to anything over 1W.

Emitter Circuit

No trouble here. Two capacitors were used for bypassing, which are not really 100% in parallel. That is, with the ground leads of the capacitors going to slightly different ground points and the two leads cutting down the lead inductance by a large factor, the emitter is pretty well tied down to the ground plane rf-wise. A 25Ω pot controls power output from about 0.25 to 1.0W with 120 mW of rf input power and 2W of dc power.

Due to the base resistor requirements, no limiting resistor was needed in series with the 25Ω emitter pot which operates nicely as an rf power control (although 2W dc input does not particularly strain the HEP-75). A 100 μF capacitor was later shunted from the emitter to ground for better modulation properties.

Collector Circuit

Several requirements must be met here, some of them not usually compatible, but things worked out quite well as you will see in the results section. A good high-Q inductor is desirable for maximum selectivity when loaded by the antenna, and at the same time good heatsinking is needed. Fortunately, there is a design that will accomplish both of these requirements at the same time. The secret; Plenty of copper. Thus the edgewise-wound copper-strap inductor shown in Fig. 1. The collector is internally connected to the case, and to keep the inductance low and the heat conductance high, the case can be soldered directly to the copper strap. Do not use a large iron, and be sure to tin both the strap and the transistor case first. Use the minimum amount of time and heat to do this, consistent with a good solder joint.

Note that the collector is tapped pretty far down on the inductor (near the cold end); this is done for impedance matching purposes. With direct currents of nearly 200 mA at 12V, you can see that the rf impedance will be low. This gets to be a big problem when you get up into the hundreds of watts, but at 2W matching can be done by tapping the collector on the second turn of a 7½-turn coil as shown.

The copper strap also takes care of the heatsink problem by conducting the heat away from the collector in both directions along the strap. With more powerful transistors, use is made of a beryllium oxide or aluminum oxide insulating stud, which conducts heat quite well but *not* electricity. So far they cost a lot more, but we can always hope they will come down in price.

Collector Dip Notes

With the HEP-75 collector circuit detuned, the current is near 200 mA; with it tuned and unloaded, the current drops to about 50 mA just like in the good old days with tubes. Believe it or not, a spark can be seen when applying the pencil test to the high rf end in the unloaded condition. I generally load the circuit with the antenna so that a slight dip of, say, 10% is obtained. With an rf power indicator in the antenna line a precise adjustment can be made.

The RF Output Match

While not critical, the output connection does require care and testing to obtain a good match and maximum power output at 50 MHz.

Tapped onto L2 and 2½ turns (Fig. 3) is a 5–80 pF Arco compression trimmer (Model 462), which does a good job of matching a 0.9 or 3W bulb to the collector. With the input circuit of L1 tuned correctly and 180 mA showing in the collector circuit while detuned, and rocking C4 slowly through the dip at resonance, the proper adjustment of output coupling capacitor C5 can be obtained.

The adjustment for a pilot light match is not necessarily the same as for a 50Ω cable. Leading to coax requires readjustment of output capacitor C5 and possibly the tap on L2.

I found 160 mA to be about the maximum output. If you load L2 down to where there is less dip, your power and selectivity against harmonics will suffer. It is best to use a shade less coupling, sacrifice a small percentage of your power, and obtain good discrimination against harmonics. (Don't forget that second harmonic of 50 MHz right in the middle of the FM band!)

Homebrew Wattmeter

This is something I'll really have to look deeper into, because there are rf wattmeters on the market, but they start at $49.50 for 2–30 MHz and go up in price along with the frequency. With all the material that the young builder has to purchase, perhaps on an "allowance" budget, he just has to pass up such luxuries.

In the meantime, back at the old bench, turn to page 259 of Lafayette's 1970 catalog and you will find a list of pilot lights with a wide variety of wattage. You can push these along at a lively dc clip to perhaps twice the rated voltage. After all,

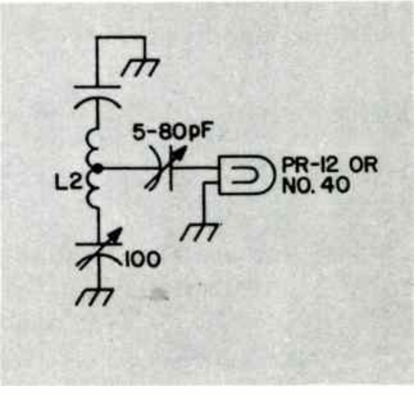


Fig. 3. Output test circuit.

most of them sell for only 15¢ to 90¢. The only thing is that after you get up to a watt and over it begins to be a little tough on the eyes! So use a higher rated lamp than actual rf power. In the Lafayette list you will see the number 48 and 49 for 120 mA (good for oscillators), the 40 and 47 at 1W, the PR13 for 2.34W along with the PR12 for 3W, and the 432 and 433 for 4.5W, etc. A lot depends on what you can find in your local hardware store if you're in a hurry right now to know how many watts output you have. There comes to mind the question of whether or not the pilot lamps light to the same brilliancy on rf wattage as they do on dc. All I can say is that when a lamp is lit by rf to the same brilliance, there must be *at least* that much wattage. If the final is tuning nicely at the proper loading point, as mentioned above for maximum power out, it is reasonable to suppose that most of that rf is going into the filament of that pilot lamp, and will show up as heat.

To make a real handy wattmeter for pennies, set up a little panel or minibox with lamp, battery, pot, and dial knob calibrated in watts. To calibrate, simply read the amps times volts for different positions, using the method of successive approximations to get round numbers for easy reading on the dial.

To solder a connection onto the aluminum base of a bulb, use a good clean tinned iron, scrape the aluminum clear of any oxide, and use a rubbing motion of the iron to tin the aluminum for about 3 or 4 seconds. Friction, plus heat and solder flux, helps solder to adhere adequately to aluminum.

For use with the transmitter, just position the wattmeter bulb alongside the one lit by the rf and adjust for the same brilliance. The human eye is supposed to be quite good at this type of comparison. It works.

The Modulator

There is a ready-made unit for the modulator, which only costs $7.95 and has 3W of audio output. This is the Lafayette "4-TR," 99E91432. It is a chunky little package, uses an RC-coupled input stage, a transformer coupled driver, and two power transistors in push-pull, with an output transformer having 8 and 16Ω outputs.

The unit fully modulates the 2W rig. At this particular time I have yet to decide between the transformerless output circuits and those using a transformer. It may well be that certain systems need transformers, and others, such as hi-fi sets, are better without them. As some engineers have pointed out (mainly those engineers from telephone companies!), the best of FM

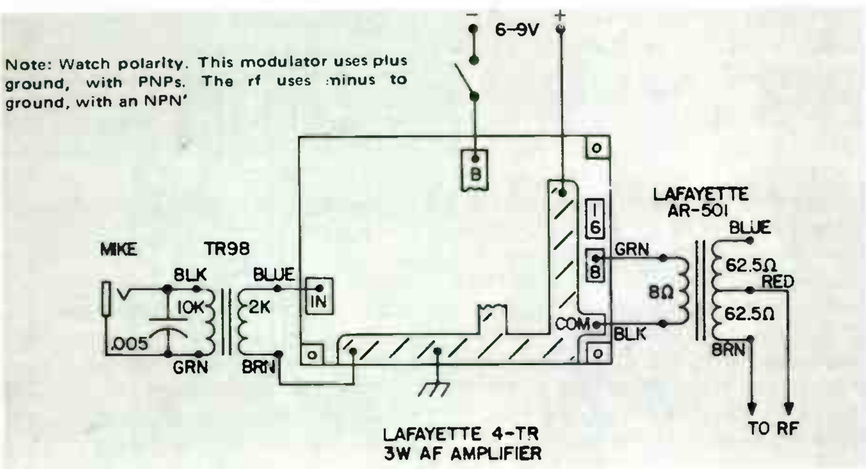


Fig. 4. Modulator circuit, 2 watt, 6 meter rig.

broadcasts reach you through a minimum of ten or a dozen transformers, so why worry about one more?

So, for a modulation transformer, not having found the ideal as yet, we use another, back to back with the 4-TR winding of 8Ω, The nearest to 75Ω I could find was the Lafayette AR-501, which has an 8Ω winding on one side and a 125Ω winding on the other (centertapped). It worked out fine on the air, in the circuit shown in Fig. 4. As the power goes up on this type of rig, with one or more HEP-75s, possibly to 10W or so, the modulation impedance will drop below the 48Ω region. In this area there is the 10W "universal" transformer with taps at 4, 8, and 16 on one side and 8, 12, 16, 24, and 48 on the other side, at only $3.95. A major benefit of low-impedance solid-state devices now becomes evident, as you don't have to pay for much copper.

In my transmitter, only the final is being modulated. Tests were carried out using only a 6V battery on the modulator because the only 9V batteries I had on hand were those little jobs for Jap radios – and they will *not* furnish 500 mA! Excellent modulation reports were obtained on the air, however.

A good idea for checking your own modulation on a dummy load before you put it on the air is to use an Amperex TAA-300 integrated circuit, as shown in Fig. 5. This little gem is actually a miniature hi-fi set all by itself, and it really tells you what your own rig – with *your own voice* modulating it – sounds like to others on the band. To do this you need also a set of earphones, with good padding to keep your voice from reaching your ears through the air. The Lafayette Model 8X stereo headphones ($7.95) are excellent for this, with 8Ω impedance per phone. Just connect the two phones in parallel in a three-circuit jack.

Connect a tuned diode receiver in front of the TAA-300 to pick up the rf from the transmitter, and use the Amperex external circuit as shown in Fig. 5.

The TAA-300 is an excellent example of a modern audio-type IC using eleven transistors and five diodes, with a frequency response of 20 Hz to 25 kHz. Naturally, you don't need all that range for voice communications work so the following modifications were used to cut down the highs and reduce the lows. Replace C1, 0.6 μF, with a 0.01, and install a 0.005 across the input jack. Regular treble and bass tone controls can be made up but they should be put in a minibox to minimize hum because there is a lot of gain at 60 Hz in the little tin can. I also put a pot in place

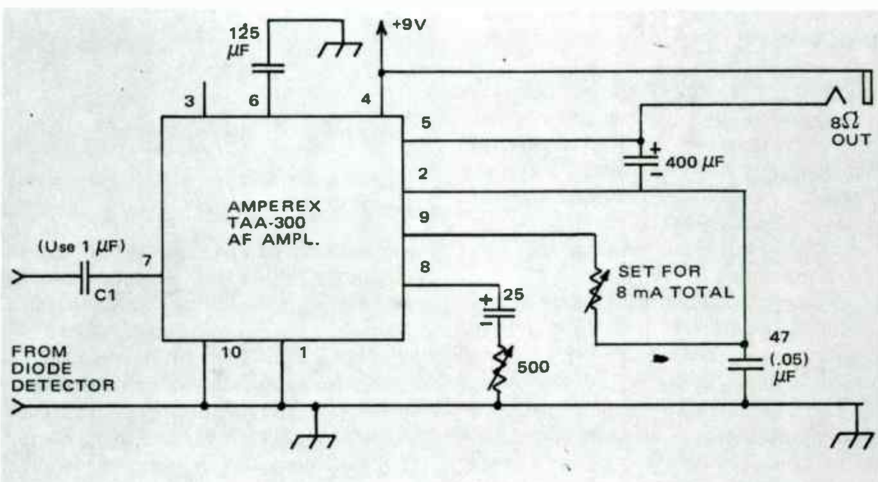


Fig. 5. Modulation monitor, 2 watt, 6 meter transmitter.

of the 48Ω feedback shunt resistor, because making the negative feedback control variable has very interesting possibilities, as you will see when you operate it. The TAA-300 also makes an ideal amplifier for any amateur receiver using transistors.

On The Air

Not waiting to hook up an rf power monitor, I connected the 50Ω cable from my 4-element beam through a ceramic rotary switch for use as in Fig. 6, to change over the antenna and turn the receiver and transmitter on and off, and tuned for the same dip in the collector current that had shown maximum power into a PR-12 3W pilot lamp. This did the job. In October 1969, I tuned the vfo to a good loud signal on the band and he came right back to me. This was W1ZLG, a Massachusetts station 55 miles away. His report: "Modulation very very good."

. . . K1CLL ■

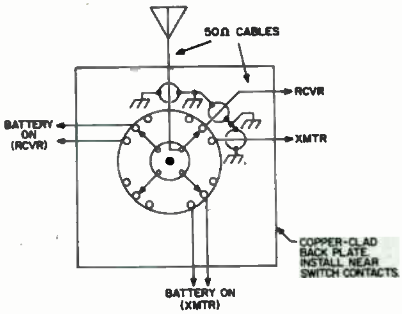


Fig. 6. Send-receive switch, 6 meter, 2 watt transmitter.

Build Your Own 2m FM

Bill Hoisington, K1CLL

All you have to do is look once at the price of commercial walkie-talkies to see how the effort of building your own can be justified. But there's a satisfaction that comes with operating something you've put together yourself, which goes well beyond the knowledge that you've beaten the manufacturers at their own game. I won't be foolish enough to tell you that building up a portable FM transceiver is a snap, because it frankly isn't quite that simple. But it really isn't particularly difficult either; and the time you spend will be amply rewarded by a good solid working knowledge of what's in those little radios you hold in your hand. Besides, it can be great fun!

The construction process is described here in its entirety – first, the receiver, then the transmitter. But since the project is to be a miniaturization job as well as simple construction, there are certain specifics involving components that must also be considered. In the main, these are dealt with individually.

Miniature Components

Capacitors. Bypass units can be the Lafayette thin units (see page 294 of 1970 catalog), where a 0.01 μF job can be found which is only 5/16 in. square by 5/64 in. thick. And the 1000 pF ones are only 11/64 in. square. For the lower values used in coupling and for fixed tuning capacitors, I like the Elmenco dipped silver-mica jobs.

Resistors. Resistors can be the Ohmite ¼ watters, but for the sake of miniaturization, you'll be better off if you get a selection of Allen Bradley 1/8 or 1/10 watt midgets. They're *really* small!

Crystals. The crystals should be the small plug-in kind, about 400 by 175 mils because repeater input channel frequencies do vary across the country. The most prevalent in the U.S. is 146.34, with 146.46 being Canada's prime choice. An absolute must for repeater use is the "Radio Amateur's FM Repeater Handbook," by Ken W. Sessions, Jr., which can be obtained (where else?) at 73.

Coils. The 8 and 24 MHz coils are the 9050 units from J. W. Miller, and are very handy for modifying to suit transistor input impedances, as well as having good, stable, mechanical tuning of the cores.

There isn't much else on the strip that I can see except thin copper-clad, 1/16 in. linen-base Bakelite or fiber glass strips, four Motorola HEP 55s, and two Motorola HEP 75s (2N2866).

Various colors of subminiature wire will help also.

Special Tools

Don't worry about particular tools; they're not too special, as you will see, but you should prepare a little, in order to do a real good job. You must have the usual set of good small tools and it helps to thin down by grinding the already thin needle-nose pliers to get into those really narrow places you will find in back of the mounting strip. Use the same treatment on some small side-cutters also, because you will be cutting off a lot of small wires in even smaller places.

A collection of small low-cost screwdrivers will be handy, too – file them sharp and very small for special places. Sharp-pointed tweezers are handy as well.

For drills I go down to size 65 (35 mils) in a Black and Decker ¼ in. drill with the drill stand for under $10. Depending on how lucky you are, your drill chuck may not take those little drills. Some of them don't. Then you have to lay out another $3 or $4 for a jeweler's chuck, which will take a No. 80 drill (13½ mils).

You do not have to drill the component-lead holes exactly to size but the closer you do the more rigid the parts will be when mounted.

Various fiber TV tuning tools are useful for the trimmer capacitors, and several lengths of ¼ in. Lucite and Bakelite rods make good insulated screwdrivers also.

A slightly unusual aid I employ a lot is a "coffee stick" with an arrowhead-shaped lump of coil wax stuck on the end. When you're winding small coils with small wire it is very handy to put a drop of wax on the coil and let it sink in and cool. You can do this with the tip of your small iron, and it sure helps hold all that tiny wire in place. All the filter chokes shown use this method. Good for a lot of receiver coils to come later, too, and for holding the extra turns wound on the Miller coils for base impedance matching.

Be sure to have plenty of subminiature clip leads with flexible wire of various lengths from 1 in. up to 1 ft.

Have a good selection of Arco midget trimmers on hand also, such as the 400 series, which are just ½ in. long.

No. 48 or 49 bulbs are good for checking rf as you go along the multiplier chain from stage to stage. You should always be able to light one of these, which glow red (dull) at 20 mW. Use a matching series trimmer, as little as 5 pF for 147 MHz and less than 1 pF on 450.

Have a roll of plain masking tape to hold down strips and things while working on them. A small drill vise or the "third hand" bench vise helps, too.

Dos and Don'ts. These hints apply especially to a multiplier chain including a straight-through amplifier used as a phase modulator, which is the circuit being described in this article. It has an 8 MHz crystal oscillator and ends up on 147 MHz, so you must be *sure* of the frequency of each stage as it operates. Do *not*, rely on your receiver or on grid-dipping the inductors first.

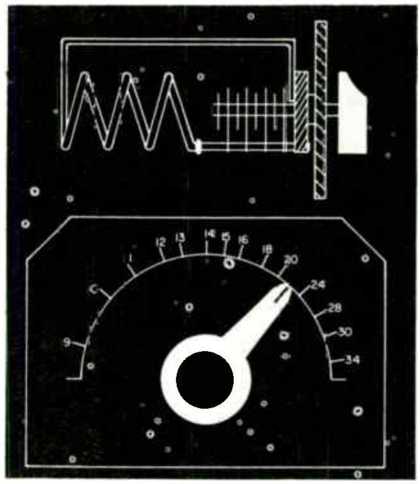


Fig. 1. Typical absorption wavemeter.

These simple and inexpensive test accessories will help you in this work, as I found out – even after 50 years of radio endeavor. I tried rushing this multiplier along without using my homebrew set of absorption wavemeters (see Fig. 1) at every stage and trouble showed up right away.

Getting right down to the point, here is a list of handy items to have on the bench

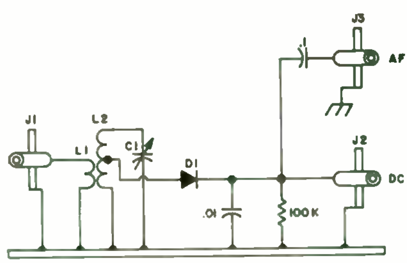


Fig. 2. Tuned diode detector.

while you're building crystal oscillators, phase modulators, multipliers, and amplifiers. Using the absorption wavemeter, any circuit under test can be checked for the real and exact frequency at which it is resonating or oscillating, by lighting a bulb on rf or using a diode detector with a meter. When the absorption meter resonates with the rf in the collector circuit which is lighting the bulb or actuating the meter, a dip in the light or on the meter will show. This indicates the *real* frequency of the main body of the rf present. Some transistor collector circuits not tapped down on coils are especially notorious for this, and may exhibit two frequencies at the same time. For example, there may be energy at 72 and 96 MHz present. This is an indication of mistuning, or overloading, or both. Tap the collector down on the coil, don't load it so heavily to the next stage, check it carefully with the diode meter, and don't worry about a small remnant of off-frequency energy. After all, a multiplier is bound to have some of this present. Just get the *main* amount on frequency and be happy. And be sure the *next* stage also peaks on *its* desired frequency.

A grid-dipper in the *diode* position can also be used for this work. A one-turn link around the low end of the grid-dipper coil and a cable will get you into small places in a rig where you cannot insert the whole dipper.

The diode detector. Figure 2 shows the schematic of one of these useful pieces of equipment which allow you to listen to your transmitter multiplier stage as you build it, and check the actual frequency at the same time. I have a collection of them

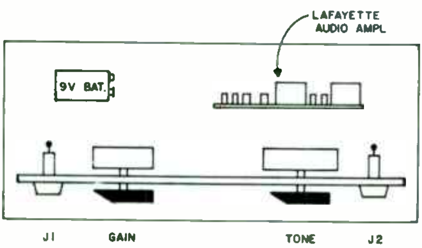


Fig. 3. Handy af assembly, top view

here covering from 125 KHz to 10 GHz. With a good variable capacitor you can generally run over three to one in frequency range – up to the UHF region at least. From there on up things get a little more difficult.

These "receivers," because that's what they really are, although of low sensitivity, are especially helpful in transferring known frequencies on a signal generator to a homemade set of wavemeters.

The meter. This should be as sensitive as possible. Lafayette has good ones down to 50 μA. Use a tap switch to put resistors in series to bring the voltage range up to 10V or so for use with an active portion of the rig such as the 1W final circuit.

The AF Amplifier. This item should not be neglected as it is at times a great aid to getting a trouble-free, *noiseless* carrier, which you can then modulate and be proud of. The valuable RCA handbook, "Transistors, Thyristors, and Diode Manual," has a lot to say about "discontinuous jumps in amplitude or frequency as various levels of drive are encountered." These little termites can be seen on the meter or *heard* on the af amplifier or can show up on both. Figure 3 shows a mounted version of the af amplifier used here for this purpose. It is a worthwhile and handy little piece of equipment to have in a lot of situations, in both receiver tests and transmitter tuneup. Just plug it into J3 of the diode detector in Fig. 2 and hear those unwanted clicks, whistles, rushing noises, squeals, etc., coming from what you may have wishfully thought was good clean rf in your multiplier drive!

Important notice! Overdrive is especially to be avoided in multiplier chains with transistors. Superregeneration is one of the indications. Believe me, it can be a very nasty bug!

Diode detector cable probes. Have a collection of these on hand as in Fig. 4. You can use them also to feed rf into a pilot light, connect up to your lab receiver, etc.

Handy meter jacks. Figure 5 shows an elementary but flexible and useful metering method for checking total or only one stage current.

FOUNDATION RECEIVER

The basic design shown here is for a low-cost single-conversion utility receiver for 2 meter FM; particular attention is given to easy-to-build i-f and discriminator modules for the 10.7 MHz section. The rf is tunable from 144 to 148 MHz, with a switch for AM use. This is a complete portable receiver, *not* tied down to a large ac communications receiver.

Discriminator action, with sample, is shown for easy understanding and home-brewing. Double conversion with crystal control can be added later.

The schematic of Fig. 6 shows how easy it can be. Remember, this is just a basic receiver which, without double-conversion, is a relatively broadband, easy-to-tune job, but it sure pulls in those interesting repeaters!

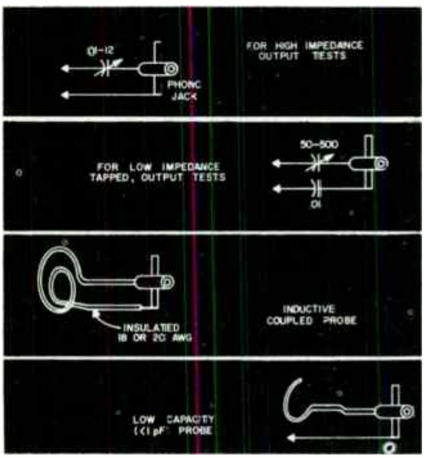


Fig. 4. Coupling methods.

Front End

Simplicity is the word in this module. You can check different transistors for low noise, coils for rf, or add a low-noise stage; and the tunable oscillator is easy to change to a crystal oscillator for repeater operation. All three stages are tunable from 144 to 148 for coil, sensitivity, and selectivity experimentation, and to allow you to check the AM section of the band as well as repeater work in your neighborhood.

The oscillator tuning dial also relaxes preliminary oscillator crystal frequency requirements by allowing you to find out what crystals you will want later, and order them without rushing the deal. The link coupling at low impedance permits easy switching from tuned to crystal control, if you wish to retain the tunable feature.

The rf and mixer stages are tuned by small variable capacitors mounted on the baseboard with small brackets made from copper-clad. Small pointer knobs allow peaking of these circuits. The rf stage has a trimmer capacitor feeding the base which is quite useful, resulting in a welcome balance between gain and self-oscillation. The mixer also has a trimmer for its base input, which permits a selectivity adjustment for this circuit.

The tunable oscillator was mounted on the Miller slide-rule dial for mechanical stability as shown in Fig. 7, and works quite well – with the broadband i-f of course. As I write, the repeater band just below 147 MHz is giving out with various

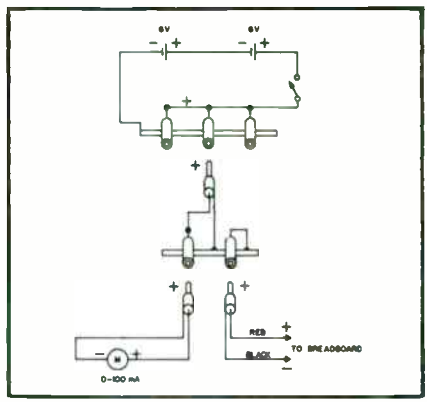


Fig. 5. Handy metering jacks and plugs.

repeater-relayed calls, "W1AHE, anybody around?", K1SHE through W1ABI," "W1JLE through WA1KFY," "K6MVH through W1ALE," and various other calls, through such other repeaters as WA1KFZ, K1ZJH, K1MNS, etc.

The two-meter band can be spread out from 10 to 90 on the dial by trimming L7, increasing C5, and using a smaller C6.

Oscillator coupling can easily be adjusted for maximum conversion efficiency via L4 and L7, and the cable between them is a good place for the crystal–tunable switch mentioned above. To start up, adjust L7, C5, and C6 for the range 133 to 137.5 MHz as a local oscillator for the i-f of 10.7 MHz to be used later. I tuned up the whole front end using the diode detector of Fig. 2 tuned to 10.7 for the i-f section. When there is lots of 10.7 MHz energy out on L6, such as to deliver 5V dc out of the diode, you've got a good front end!

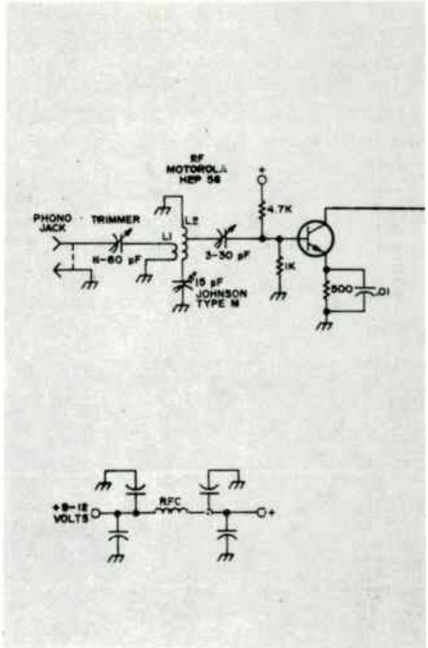
10.7 MHz I-F Stage

The reliable and sure-fire Motorola HEP 590 IC was used here – 25 dB gain, no self-oscillation, what else would you use? Figure 8 shows the circuit, using Miller half-inch shielded coils both on the input and output. Note that the 590 is simply turned leads-up and soldered onto a few resistor supports, with a shield, as in Fig. 9. A gain control is used, which may or may not be kept in later as you wish. With the limiters that can be added, the gain control is *not* needed.

A B+ filter is included in each module, and a 100Ω resistor with a 10 μF capacitor may be needed also to cut out motorboating when more stages are added later, if you go to double conversion. Be sure *not* to return L3 to *ground* dc-wise, as the needed bias is supplied internally through pin 4. Pin 9 is the main B+, along with the cold end of L3. Gain control can be obtained through a pot in the pin 5 lead, where maximum gain is reached with pin 5 at *ground* potential.

For new readers, the internal and external circuit of the Motorola IC 590 is shown in Fig. 10. This IC, which is very useful for frequencies up to at least 6 meters, has extremely interesting features, among which can be noted the absence of internal feedback (even at 50 MHz), the high gain, and the excellence of the gain control at pin 5, either manual or automatic. For this receiver, mainly intended for experimental FM use, no avc is used. Later, if you add double conversion, the limiter section module will eliminate the need for avc.

Trimmers are shown for C1 and C2, but fixed capacitors of the proper value may be used to allow tuning of the i-f coils at 10.7 MHz by the variable tuning slug cores in L2 and L3. Note that these Miller half-inch gems have very good electromagnetic as well as electrostatic shielding, due to the cup-core type of construction, and are available for use from 30 MHz down to 135 kHz. They can also be easily opened



for the addition of primary or secondary low-impedance windings.

Discriminator

After many days on the bench with discriminator circuits I at last hit on one that works like a charm for any frequency tried so far here at least from 10.7 MHz down to about 135 kHz, and at the same time is easy for the homebrewer to build because of the link coupling. Figure 11

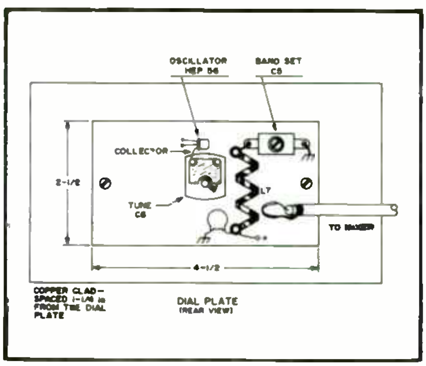


Fig. 7. Dial-mounted local oscillator

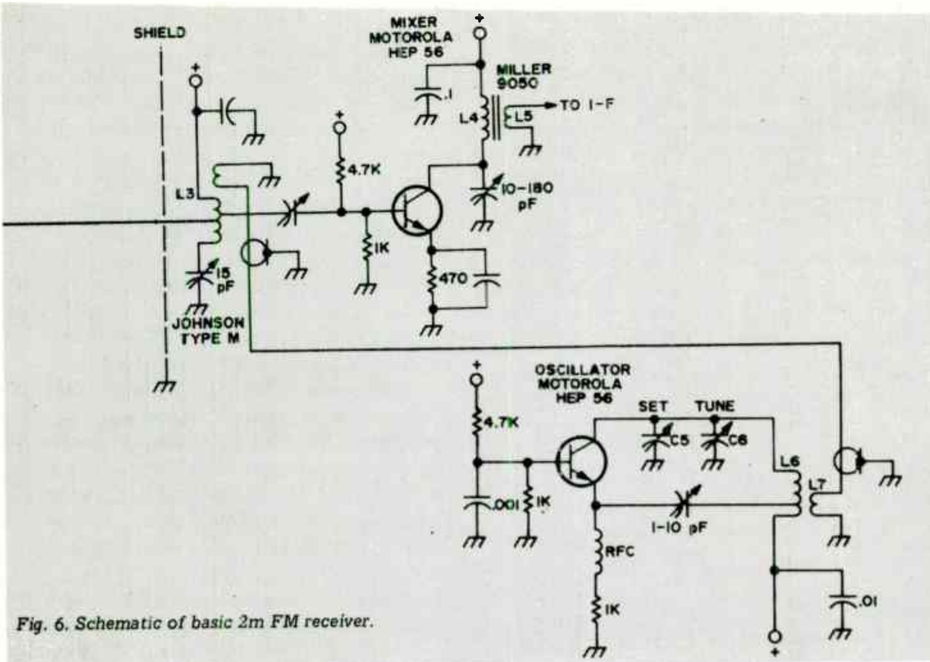


Fig. 6. Schematic of basic 2m FM receiver.

shows the circuit where L1 is a simple tuned coil in the collector circuit, with *no* coupling requirements other than a one- or two-turn link. When the primary of a discriminator transformer has to be coupled just right to the centertapped secondary it is not a job for the usual experimenter at his bench. With the link you can't go wrong. At least I haven't so far. Just tune L1 to 10.7 MHz, put a turn or two around it and another turn on L4 and away you go. Figure 12 shows the discriminator dc output curve, which handles about 25 kHz for 2-0-2V.

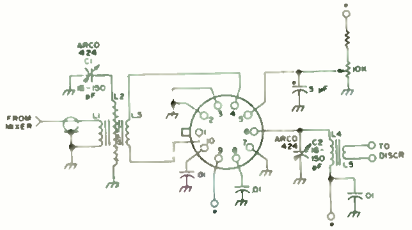


Fig. 8. 10.7 MHz i-f, bottom view of IC.

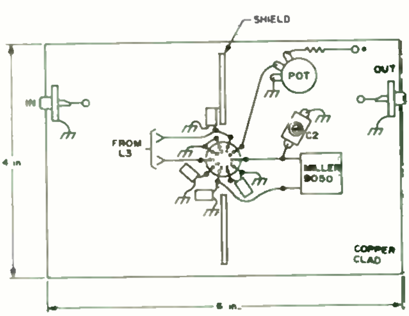


Fig. 9. i-f shielding as seen from top side of board (bottom view of IC).

I got the idea of using 10.7 MHz as an i-f (other than the fact that it is used for many years as the i-f for the FM broadcast sets) from some of the little $20 Nagasaki Hardware Co. sets used to pull in the police bands. It is also used a great deal in the two-way mobile sets, no doubt "because it was there" to begin with. It also can be used later as the first i-f, for narrowband work, by adding a 10.7 MHz converter to 455 kHz, and a 455 kHz i-f and discriminator. So, away we went, and I'm having a great time listening to those repeaters. However, please bear in mind that the i-f selectivity at 10.7 MHz is not sufficient for continued use on 2 meters, except to tune up the front end and to get acquainted with what's going on in your neighborhood, even though it is fun! It will also start you off on discriminator work, which I find has some very fascinating aspects, such as the extreme selectivity of the output curve for

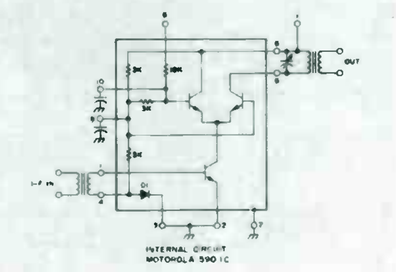


Fig. 10. Internal circuit of Motorola's HEP 590 IC.

one tuned circuit. More later on this if time allows.

Audio Amplification. Just for new readers we'll give a quickie on the Amperex TAA-300 1W "baby-hi-fi" IC which I now use for almost every audio purpose in receivers, and modulators up to a watt. There are about 11 transistors in that one little can, and it is flat to 1 dB from 25 Hz to 25 kHz! Get a halfway decent 8Ω speaker to go with it, because it's worth it. Figure 13 shows how to connect it up.

FM TRANSMITTER STRIP

The transmitter section measures 1 in. wide by 8 in. long, and it puts out over a watt on 146–147 MHz, with low-cost components. This miniaturized transmitter is my logical step toward design and ultimate construction of a "shirt pocket" portable transceiver. The parts for that one jump up a little in cost, because it takes a lot more tools to make subminiatures, such as stereo microscopes, special materials and skills, jewelers tools, and so on.

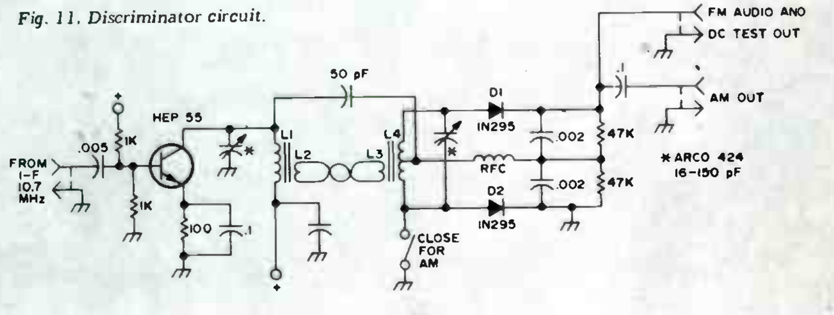
Shape Factor, and Assembly Method.

These are important features, as you will see, allowing the homebrewer to build a complete FM rig in a minibox and still have room enough left over to change components for repairs or design improvements if needed. You can also substitute slightly different components if you have to.

Figure 14 shows the method, using a copper-clad baseboard on which is mounted a drilled ½-in.-high strip of insulating material holding all the components. Bypass capacitor leads to ground are no longer than ¼ in., shielded coils are used, and all tuning is done from one side.

The photos show the happy results of placing the parts to best advantage on such an assembly. Notice that the components are also all on one side, and their leads and connections are on the other side. On the wiring side, every connection is spread out in front of you, with room between each one for good soldering; no resistor supports or other metal tie points are needed.

Fig. 11. Discriminator circuit.



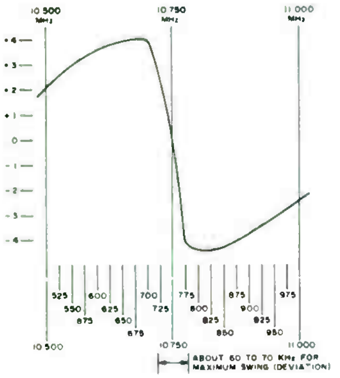


Fig. 12. Discriminator dc output curve.

The B+ lead is red subminiature wire and goes from filter to filter along the strip. The rf lead is green and goes from the coil output tap of each stage to the next base coupling capacitor; the rest of the connections practically fall in place for soldering together. As you can see, there is still room left over!

The detailed planning of the holes to be drilled becomes a large portion of the work. Figure 15, component side and top views, shows how to start this off. The next step is to make a life-size drilling template using the components you have or intend to use. I mention this because most of these are not critical and you may substitute without trouble providing you keep thinking "little." Even here, you can go bigger with the components if you want to, but your overall package size may expand. You can also go smaller if you plan carefully and cram everything together a little tighter. The reason for this will be evident if you study the circuit, where you will see that no critical wires cross over each other, and that the power amplifier is well away from the oscillator.

Ultimate size is actually up to you, and you can judge for yourself after laying out the parts on hand. If you send for a selection of Lafayette Radio very thin and small capacitors, you will have an easier task to get it down in size.

Figure 16 shows two methods of preliminary fastening of the vertical strip to the copper-clad baseboard. This will start off the assembly, and after the wiring you could hardly tear them apart with your fingers. I counted no less than 25 ground connections to the copper on my own 10 in. strip.

You can also make strip modules of any length you want such as modulator af, receiver sections, etc., as shown in the receiver plans. This makes the task of repairs or improvement changes easier later on. These shorter strips can be fastened end on to each other and fastened down to the baseboard as shown.

Miniature Filters.

Do *not* try to make up frequency multipliers *without* rf filters in the dc line to each stage, unless you care to experiment with rf phasing in battery leads – and that isn't good! (Every time I leave out the filters I get into trouble!) You can make up

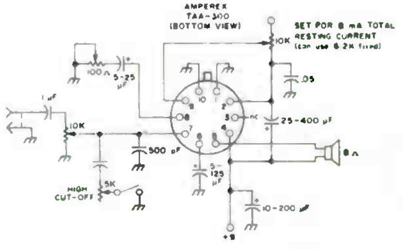


Fig. 13. Audio amplifier IC.

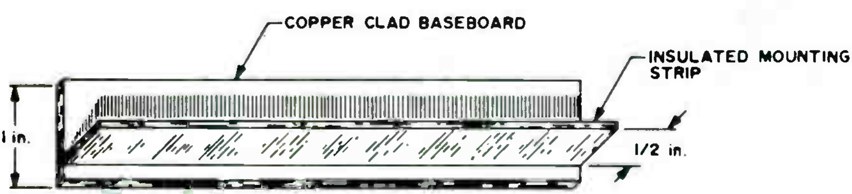


Fig. 14. Baseboard/mounting strip configuration.

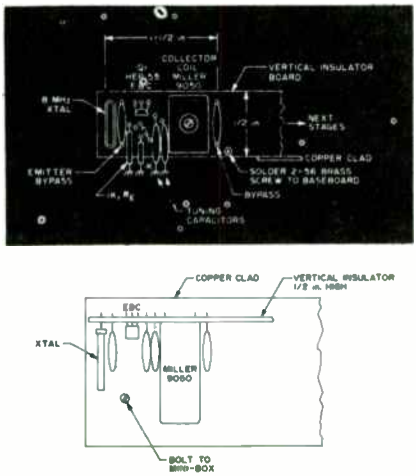


Fig. 15. Component layout, top and side.

"dime" filters without much difficulty if you follow the simple details below. Materials needed are tiny resistors (any value over 1kΩ), some 36 or 38 AWG wire (double silk covered), coil wax (you can use paraffin wax if you can't get coil wax), and small capacitors, such as the Lafayette types. Use 0.01 for HF and 0.001 (1000 pF) for VHF. Figure 17 shows the circuit. The main thing is to interpose an rf trap in the plus lead between each stage and any other.

The series method shown in Fig. 17 is best for high-gain amplifiers because it puts

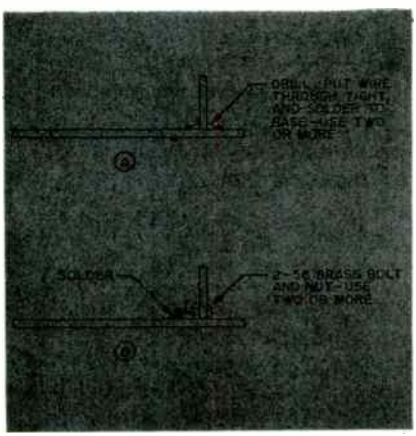


Fig. 16. Methods for fastening insulated strips to baseboard.

several filters between the high-power output stage and the sensitive first stage. However, if the filters are very good you can bring the battery leads from each stage to a common point, but this must be checked carefully if you have to do it.

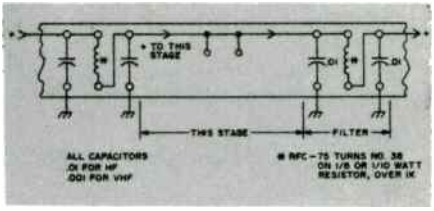


Fig. 17. Miniature filters, interstage.

How to make'em. Clean and tin carefully each resistor form lead close to the body, then melt a thin layer of wax onto the resistor to hold the wire from slipping when you wind it on. Solder one end of the wire onto one lead and then random wind 75 to 100 turns of 38-gage wire onto it, and wrap the end around the other lead ready to solder. Put a drop of wax on the coil before soldering to hold the wire turns in place. The wax should penetrate the whole coil. Most types of insulation on

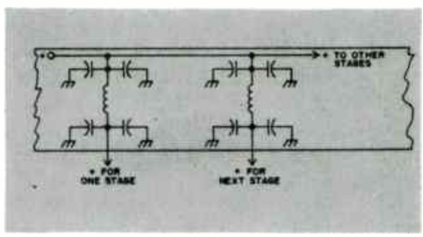


Fig. 18. UHF filters for interstage coupling.

38-gage wire will disappear as soon as solder and heat are applied, so you don't have to bare the wire first. Now you have an rf choke, and if you keep the capacitor leads real short to ground, the filter will do the job for you.

It works fine even up to 450 MHz if you use four capacitors, each to different point on the ground plane of copper-clad, as in Fig. 18.

Phase Modulator

Phase modulation results in a type of frequency modulation of the carrier at the rf output jack which the usual FM receiver cannot distinguish from true FM. Being crystal-controlled it is used by practically all the FM mobile and base stations in the U.S. so it is 100% okay here. And of course with the crystals in there, you *will* be on the amateur FM channels, providing you buy them right. You have to pay around $7 for these but it seems well worth it.

Certain designs of the af section of the phase modulator, its tuneup, and the connections to the phase modulator can be troublesome for the homebrewer, so considerable time was spent to make it as simplified and easy to adjust as possible. It also can be used in the receiver section as the af amplifier because the frequency correction is done *outside*. The use of an 8 or 16Ω output connection into the phase modulator emitter circuit helps to stiffen the af drive and keep it clean.

Phase modulation af sections in commercial rigs are often qualified as "audio conditioning," or "processing" circuits, which they are of course, but don't let that bother you. Excellent FM quality can be

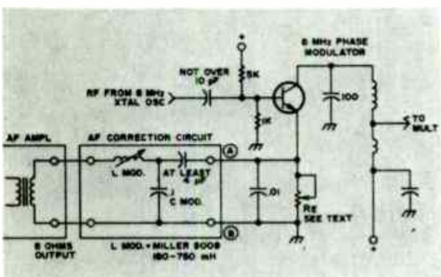


Fig. 19. Phase modulator interconnect circuitry.

obtained by the use of an inductance of large value, placed outside of the af amplifier, in the noncritical low-impedance output circuit. The inductance cuts down the extra high audio modulating frequencies caused by the phase modulator's tendency to make the FM deviation directly proportional to the modulating frequency, which emphasizes the highs too much unless corrected. Being outside of the af amplifier, you can now use almost any good low-cost job and use it in the receiver also.

Figure 19 shows the simplicity of the method used. Having a four-transistor amplifer from Lafayette, at $4.95 on hand, that's what was installed, with a slight adjustment of the feedback resistor. This had nothing to do with the FM unit, it just happened that the Lafayette amplifier sounded and acted awful funny at first. And no wonder –, it was oscillating up in the 100 kHz range! After trying to bypass and decouple almost everything in the little brat my eyes began to focus on that printed lead going from the 8Ω output connection over to near the input, and sure enough that was it: too much feedback! An additional 50 kΩ resistor in series with the one already there did the trick and from then on nothing but good af came out. I mention this because it could happen to you too. My 3W job, also Lafayette, is suspect, possibly the same simple trouble.

The af output needed to drive the phase modulator emitter is several hundred millivolts, and the low impedance allows the usual rf bypass capacitor of 10,000 pF to act simply as an additional af filter, which it does.

As a result, the entire tuneup is done by adjusting the value of the emitter resistor and the phase modulator tank tuning coil. Neither are actually critical but should be adjusted while listening to the 146 MHz carrier on a good amateur narrowband FM receiver. The emitter resistor will be heard to kill the modulation when going much below 2 kΩ and to bring in distortion on large amounts of audio when going a lot more than 2kΩ. This latter condition also causes a drop in the rf output. You may hit it right the first time with the 2kΩ value; I'm just pointing out that this resistor is worth checking up on for a final value when adjusting for best modulation.

The actual phase modulation resulting from varying the emitter voltage with

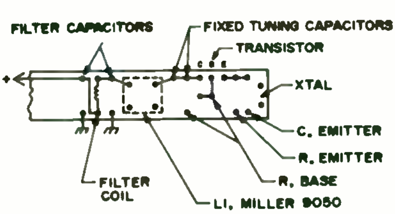


Fig. 20. Drilling layout (wiring side).

audio is adjusted by tuning, which is also smooth and noncritical. I used the tried-and-true method of listening to my own voice with plenty of audio on the receiver and a set of well-padded earphones (you can get a very useful set for under $10 at Lafayette) which keep your voice from reaching your ears directly through the air. It also cuts down audio feedback.

Tuning with af going into the phase modulator as per Fig. 19, you will notice good strong clean FM on either side of the peak tuning. These points occur *before* the 146 MHz carrier output starts to drop from detuning the phase modulator tank, so don't worry about that part. In any case, you are supposed to be following the phase modulator with enough saturated class C multipliers and amplifiers to prevent any variation in amplitude (otherwise known as AM!) I say "supposed to" because you don't automatically get this condition. You may have noticed an unduly large number of tubes showing in ads for surplus commercial FM sets. This large number is due to the designer's wish to get *all* the benefits of FM into his package. In one box if possible. You have to watch *very* carefully when using ICs for modulators, they tend to pick up rf and generate feedback with their wideband audio circuits and sometimes as many as 11 or even more transistors in one little can. Just a word of what to look out for. It's hit me more than once. Also, don't put more than the specified voltage on IC amps. Your can easily drop down with a resistor and a *large* bypass capacitor.

I used my favorite mike on the input, the Astatic 150, my favorite because it only weighs 3 oz, has the most output,

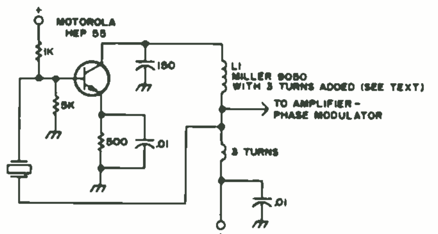


Fig. 21. 8 MHz oscillator schematic.

−44 dB, costs only $3.82 amateur net, and sounds good!

Almost any desired amount of highs and lows can be obtained or surpressed by the manipulation of the LC values in the modulator. If you use a Miller .9009 wide-range adjustable inductor, 180–750 mH you can hear the difference as you adjust the core in and out.

I started out with a large-scale layout for the parts, but you may wish to skip that and go right to a life-size layout as in Fig. 20. To make the life-size drilling template, lay out the components one after the other, "standing up" on a ½ in. strip of good-grade white cardboard and mark the component lead holes, which should result in something similar to Fig. 20. A nice feature of the cardboard method is the easy punching of the holes and the way it holds the drill as you go through the strip. Tape the template in place onto the insulating vertical strip. Do not use anything that melts under heat, though. Even if you ruin part of the strip, or want to make a large change of one stage you can just saw that out and make up another section and go ahead.

The 8 MHz Oscillator

Figure 21 shows the schematic of the crystal oscillator stage. Note the apparent use of negative feedback with the base return through the crystal to a tap on the inductance. It is only apparent though, as the crystal reverses the feedback phase, making it positive. It is a very powerful, sure-fire circuit.

The tap on the coil also provides a good low-impedance match for the next base input. The coil itself is made from a Miller 9050 shielded coil which has magnetic as well as electrostatic shielding, and a good adjustable core that works good mechanically (which is more than you can say for some of those types of cores).

Remove the aluminum can by bending back the four holding tabs and wind on three turns of 30- or 32-gage silk-covered wire onto the existing winding of the coil. Be sure and wind them in the same direction as the turns that are already there. The oscillator coil will then look like Fig. 22, and is ready to mount on the strip.

The wiring on the lead side of the strip is shown in Fig. 23, where most of the leads are seen to fall in place quite well.

Insert the component leads through the strip and bend them slightly in the direction they will go, such as the two base resistor leads which are bent towards the base lead, as shown clearly in Fig. 23. When all the leads to be soldered in one place are all touching each other, a final dressing can be done followed by soldering. In the example mentioned, the base lead has three other wires soldered to it, a wire from the crystal, the 1 kΩ resistor, and the 5 kΩ resistor.

The can of the 9050 coil has a tab which should be soldered to ground. The ground lead of some resistors (or all of them) is not routed through the strip but is soldered to the baseboard on the component side of the strip.

When the oscillator is assembled and wired, B+ can be brought in and the unit tested for rf. Some 5–10 mA of current

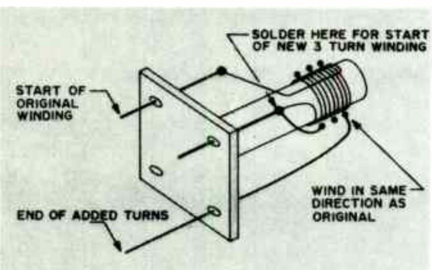


Fig. 22. Miller 9050 coil with added turns.

should register and as soon as the oscillator coil is resonated to the crystal frequency, the oscillator should show rf output to the 8 MHz tuned diode test set connected to the output tap lead.

Check the oscillator carefully on a sharp receiver for its frequency-holding ability while tuning the slug in and out of resonance at 8 MHz. Actually this will be near 8.130 MHz. (With a multiplication of X18, this should land on whatever 2 meter FM channel you're aiming for). It should come into resonance on one side with a good "plop" and gradually build up on the other side as you tune.

I always start with a large calibrated variable capacity at C3 (some 500 or 1000 pF, made from an old BC set three-ganger) and then put in fixed values so that the iron core tuning slug in the 9050 coil tunes properly about ½ in. under the winding of the coil.

Power can be adjusted by the emitter resistor, and feedback by the number of turns between ground and the oscillator-coil tap. (These are of course the number of turns added to the Miller 9050.)

A 48 or 49 bulb, rated at 2V and 60mA, should light up with about 50 to 100 mW worth of rf with a 50–300 pF trimmer in series, as in Fig. 24. When the oscillator is properly tuned and under good power control via the test pot (in Fig. 24) and the plus voltage is checked for the voltages you expect to see, the next stage can be assembled. Of course if you wish, you can mark out the whole strip template, drill all the holes, and mount and solder all components except the coupling capacitor and B+ to the next stage. This allows you to test the oscillator by itself.

The 8 MHz Amplifier—Phase Modulator

This stage (Fig. 25) is not critical, other than to keep the input base coupling capacitor at a low value to avoid self-oscillation. The only requirement is that the tuning should be correct for phase modulation.

Use the same methods of assembly, wiring, and tuneup for power output as with the oscillator stage. You do not need much gain, if any, in this stage.

The 8–24 MHz Tripler Stage

A frequency multiplier has the advantage that generally (though not always) it is free from self-oscillation, due mainly to the output and input circuits being on different frequencies. The bias requirements are different in a tripler from those of a doubler or class C straight-through amplifier, but this can be adjusted simply by varying the emitter resistor during tuneup for maximum output on the desired frequency. Figure 26 shows the schematic of this stage, where the base input coupling

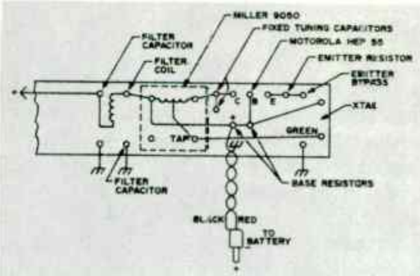


Fig. 23. Oscillator wiring diagram (lead side of strip).

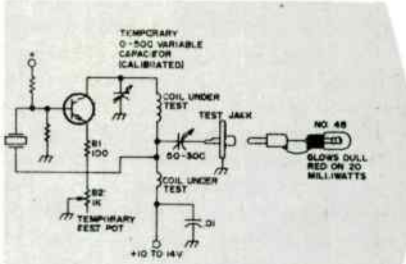


Fig. 24. Oscillator test setup.

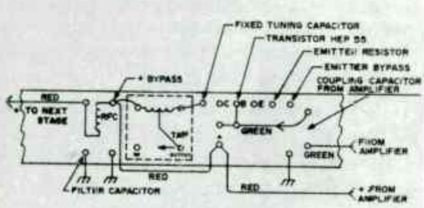


Fig. 27. Wiring diagram, 8 MHz tripler.

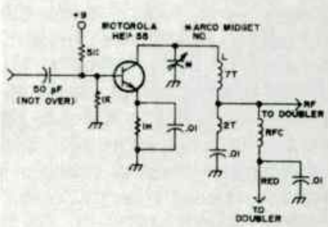


Fig. 28. 24 MHz tripler schematic.

capacitor is seen to be much larger than in the preceding stage. However, in spite of a small tap winding and low impedance in the preceding stage it is easy to cause superregeneration in the base circuit if the coupling capacitor is too large. I found that 150 pF or slightly less is a good value.

The wiring side layout for this stage, which is typical of the multiplier circuits, is shown in Fig. 27. A logical wiring system is seen to prevail, especially as regards the emitter, base, and collector wiring and their components. Two extra wires are used, one red for the B+ and one green for the base input rf circuits, with a filter coil separating the plus of each stage.

A 24 MHz diode detector is clipped onto the rf output tap on the inductor (Fig. 27), as was done in the preceding stages. Be very *sure* you're on 24 MHz, and not on 16 or 32. Here again you should be able to light a 48 bulb with rf with a 5–180 pF trimmer in series for matching. The collector tuning and power output curve with emitter resistor lowering should be clean and smooth.

As mentioned in the test equipment section, it is a real *must* to *listen* to the carrier as you build it up in frequency. I do this with a little af amplifier continually connected to the diode detector output because the carrier has to be free of all spurious noise, squeals, frequency and power jumps, etc.

The Tripler to 73 MHz

This one proceeds in a similar fashion to the previous stage, except that now we begin to use capacitor tuning of the collector coils. The iron-core coils of the Miller 9050 series do not do a good job here, and so far I have not found good ones at reasonable cost, so you have to wind your own but that is very easy, as you will see.

Figure 28 shows the circuit and values obtained by tests here. Do not exceed the value of 50 pF for the base input capacitor. In case of any spurious noise, this is the first place to look; in fact, I always start off with a trimmer at that point to make sure and get the maximum drive possible *without* noise.

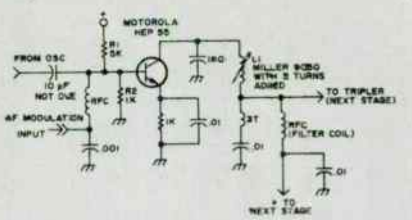


Fig. 25. Schematic of phase modulator/amplifier.

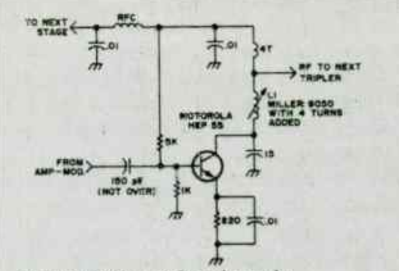


Fig. 26. 8 MHz tripler schematic.

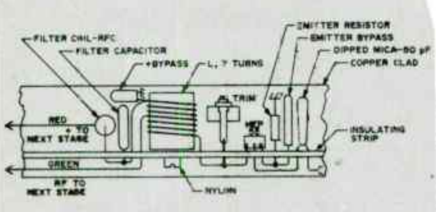


Fig. 29. 24 MHz tripler wiring diagram.

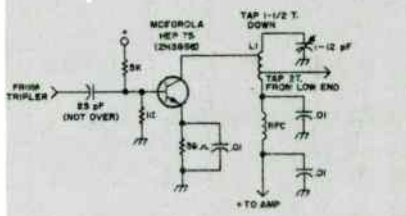


Fig. 30. 73 MHz doubler schematic

When the stage is assembled and wired and under test as done with the previous stages, once again, look out for those *undesired* harmonics, especially the 64 MHz one in this case. It'll sneak up on you if you're not real careful!

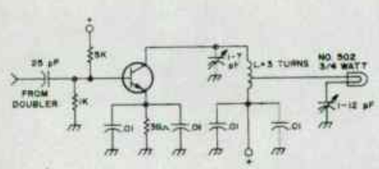


Fig. 31. 146 MHz amplifier schematic.

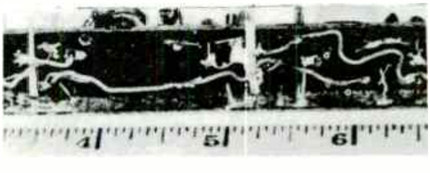
The inductor may be fastened to the mounting strip with a nylon screw (Fig. 29) for ruggedness. The variable capacitor does all right standing up on end with the fixed plates soldered to the baseboard, and the movable plates brought out through the strip with a piece of 16- or 18-gage wire, where it is joined up with the collector, as can be plainly seen in Fig. 29.

Clip on your diode detector for power checks and frequency. You can't check this latter too often, believe me. After testing for power control and noise, you are ready for the next stage, a doubler.

The Doubler to 146 MHz

This stage uses a Motorola HEP 75 (2N3866), always a lively powerful one for VHF. The schematic, shown in Fig. 30, is quite similar to the others except for the different transistor and another coil tap. The base input capacitor worked out at 25 pF maximum, with a 39Ω emitter resistor to keep the power up for maximum input into the final stage. The collector lead is cut off and the collector connection is made by soldering a 1/8-in.-wide soft and thin copper strap, which increases the heat-sinking as well as rf conduction, directly to the HEP 75 case.

Clean the case well by scraping at the place to be soldered. Use small solder, a



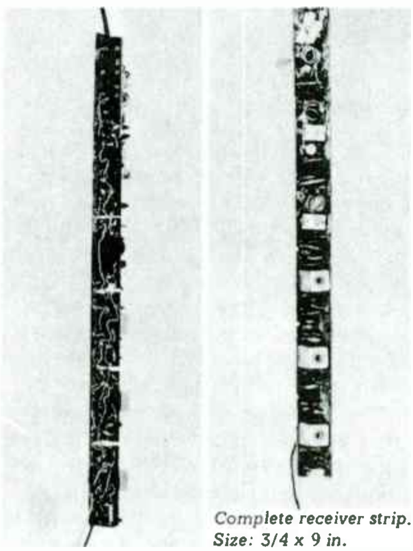
Closeup of wiring side of insulated strip.

small iron, rub the iron gently on the case two or three times for about one second only to effect a good joint for the collector strap over to the coil. The inductance is not actually critical but should be correctly tuned up and tapped for the collector as well as the output tap. After all you *do* need all the power you can get into that final.

You could use a little larger emitter resistor, for a little less current, but here again power is a point to watch. I find about 50 to 120 mW of rf at 146 MHz at the output tap, depending on the plus voltage also. Of course, you can play around with up to 18V if you want to push out a bigger signal. Before buttoning up this stage, check it once more for frequency, please.

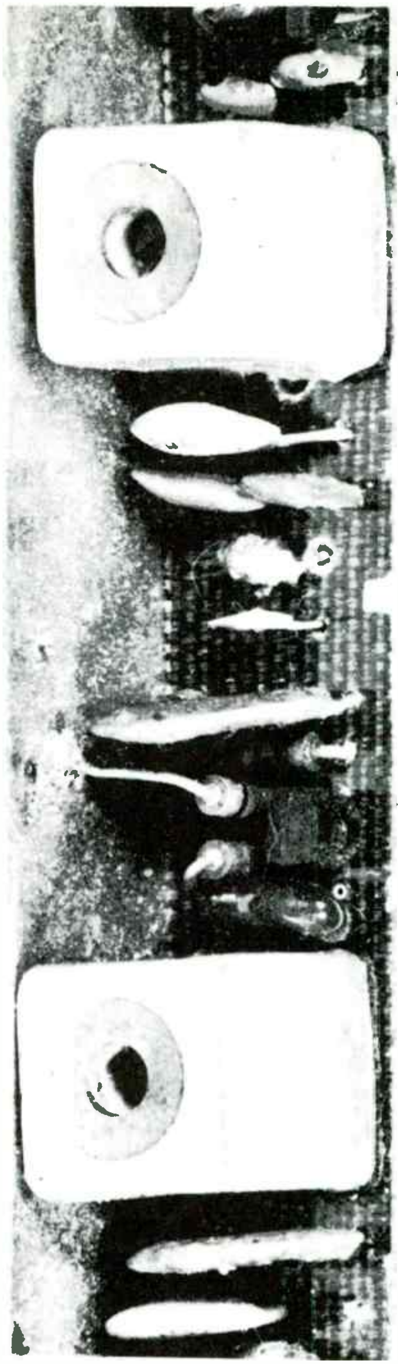
Power Amplifier

Refer to Fig. 31. Everything went along nicely with this one also, as in the previous



Complete receiver strip. Size: 3/4 x 9 in.

stage. You will note a nice feature, true of most electronic circuits that are good and foolproof, that when everything is tuned up correctly and matched properly the whole stage becomes less critical all around. That is, the tuning is not touchy, the power goes up and down nicely, and even the output tap is not too critical.



Blowup view shows degree of miniaturization

Note that with small emitter resistors of under 50Ω the collector current can get pretty high, so always keep at least 10Ω in series, as you test for the best emitter resistor value. Don't forget that 1W is 100 mA at 10V. And for a watt out you will need more than 100 mW even at 15V. "Big" transmitters (50 watters) use as low as 0.1Ω at times in this place, and in some of the new heat-sunk jobs (out of our price range) the emitter goes directly to ground.

I suggest currents of not over 100 mA for this stage. The arrangement shown uses about 50 to 60 mA, depending on drive from the tripler, B+ voltage, and output loading.

Two bypass capacitors are used in the collector circuit. A test bulb (5V at 150 mA), in series with a 1–12 pF trimmer to ground, indicates rf output and loads the collector circuit. Without the test bulb or any antenna loading you can expect self-oscillation as the HEP 75 gives plenty of action on 146 MHz. The output tap can be led into the small but good and very useful 50Ω cable (RG-174/U). This cable will then go to your changeover switch or relay for the final assembly.

As a final note on the transmitter, each tuned circuit of the multipliers and finals should also be adjusted while listening to the carrier modulation. I didn't find any of these at all critical but "the books" say to do this to assure absence of phase-shift distortion in circuits after the phase modulator.

So, good luck with PM, it doesn't seem too tough to build if you avoid the fancy stuff to start with.

The Phase-Locked Loop Comes of Age

Jim Kyle, K5JKX

It's been a little more than 13 years now since the idea of "synchronous detection" entered the ham radio world – and the odds are, unless you're a dedicated VHF and DSB (yes, we said *D*SB) nut like I am, you still haven't heard very much about it.

Not that some of us haven't done our part. The initial publication of an article of the subject back in 1957 (three years before *73's* birth) was accomplished under Wayne Green's guiding hand in CQ Magazine, and the next synchronous-detection bombshell (conservatively titled *50 dB Under the Noise – A Breakthrough*) saw the light of day in the short-lived pages of 6-Up, *73*'s subsidiary VHF magazine of the mid-60s.

But all the way through, the synchronous detection idea has had a tough obstacle to battle: While it does give all the performance claimed for it, it requires several times as many components as does a conventional detection circuit. The 1957 version used eight double-section tubes to give 16 stages; the 1964 edition (actually a completely separate implementation of the same basic idea but giving other features) was solid-state and required 24 transistors.

Quite obviously, this is a larger stage count than that of many complete receiver designs. So long as so many devices were necessary, the synchronous-detection idea just couldn't make it in the face of its much simpler competition.

But all that was changed in the last year when a major manufacturer of integrated circuits put together a single-chip circuit which does just about everything required for the synchronous detector, in a single standard-sized 16-lead dual-inline IC package. Although it contains approximately 50 transistors, the whole thing takes only about 10 mA from an 18V supply, and is hardly as large as a commemorative postage stamp.

And to top it off, the price of the device is surprisingly low, considering what you get. There are several different versions with different features, but the least costly of them is still a complete FM receiver which accepts a minimum 300 μV signal (across 3 kΩ) at any frequency from 100 Hz up to 60 MHz or so, with any deviation up to +20% of center frequency, and produces 60 mV of audio output across 8 kΩ. This one costs $30 in single-lot quantities, according to the most recent price list, but that's all you need for a complete receiver within its frequency range and sensitivity.

On top of this, remember that these are current prices, and the unit is not yet in wide use. If it catches on as it apparently should, the cost is bound to come down. Remember when the CK-722 transistor sold for $7.50 each? And now they're down to 4 for $1 from the mail-order houses?

What Is Synchronous Detection?

The words "synchronous detection" have been applied to many different detection schemes, but all of them share the idea of using a stable local oscillator to mix with the incoming signal, and developing an error signal should the local oscillator get off frequency. You might call it a sort of superhet receiver, with automatic frequency control and a zero-frequency i-f, if that brings any reasonable picture to mind.

The most common type of synchronous reception currently used (yes, it *is* being used – by radio astronomers, space communications systems, long-range radar, and any place else that the ultimate in radio communications is required, thus justifying the complexity and cost of the older approaches) involves a "phase-locked loop."

The phase-locked loop includes a phase detector, a low-pass filter, and a voltage-controlled oscillator. The phase detector is a circuit which accepts two different rf input signals, and produces a dc-to-audio output signal which reflects the phase differences between the two inputs. That

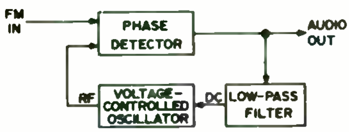


Fig. 1. Basic phase-locked loop or synchronous detection hookup for FM reception uses voltage-controlled oscillator in a servo loop. Output of VCO is continually compared with input signal in phase detector, which produces an output consisting of dc voltage which varies as input frequency changes (which amounts to superimposed audio). Low-pass filter wipes off audio and dc goes to VCO to keep it locked to input signal. Unfiltered error voltage from phase detector is the audio modulation of the signal.

is, so long as both inputs are in the same phase relation to each other (usually 90 degrees), output of the phase detector is zero. If the phase of input A begins to lag, output goes positive, and if input A begins to lead the other, output goes negative.

If we could keep an oscillator absolutely stable on the center frequency of an FM transmission, we could feed its output to one input of the phase detector, and feed the other input with the FM signal. The output would then faithfully reproduce the phase differences, which would reproduce the audio signal envelope.

If we could keep an oscillator absolutely stable on the center frequency of an FM transmission, we could feed its output to one input of the phase detector, and feed the other input with the FM signal. The output would then faithfully reproduce the phase differences, which would reproduce the audio signal envelope.

And that's the way the phase-locked loop (abbreviated PLL) works. The phase-detector output is filtered to remove any audio and retain only the dc component, which is a measure of the drift in the local oscillator, and then applied as an "error signal" to the local oscillator to keep it locked in phase with the incoming FM signal.

The filter keeps the voltage-controlled oscillator (VCO) from following the audio, so we pick the audio off ahead of the filter, and we have an FM detector (Fig. 1).

For AM reception, we do things a little differently. We lock the VCO to the incoming signal, just as before, but we add things.

It works out to be something very much like SSB reception; many years ago QST pushed something they called "exalted carrier reception," which involved using two i-f strips, one extremely sharp to pick out the carrier from between the sidebands for high-gain amplification, and the other to accept the sidebands.

Our phase-locked AM reception is just about the same, except that we generate a new carrier in the VCO which is phase-locked to the incoming carrier, and use that to demodulate the sidebands. The demodulation is accomplished in a "multiplier circuit" which is more familiar to most of us as a "product detector" or "mixer." Another low-pass filter shaves off the original input-frequency signal, the VCO signal, and the *sum* frequency from the product detector's output, leaving us the *difference* frequency, which turns out to be the audio we wanted to recover.

Note that this arrangement (Fig. 2) cannot demodulate an SSB signal since there is no carrier present for the phase-lock loop to lock onto. In such a case, it would attempt to lock on the strongest sideband component present with notably less than satisfactory results.

The 1957 circuit, by Dr. John Costas W2CRR and G. K. Webb WØAHM/2 (both with General Electric in Syracuse, N. Y. at that time), overcame the problem by deriving the control voltage for phase-locking from the sidebands rather than from the carrier. Unfortunately, their circuit has not yet been implemented on an integrated-circuit chip so far as we have

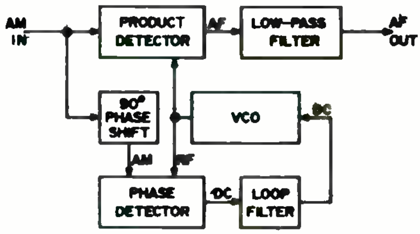


Fig. 2. AM detector using phase-locked loop is a bit more complex. Incoming AM is shifted 90 degrees in phase and VCO is then locked to it. VCO output is used as bfo input to product detector, to recover audio. Low-pass filter in audio signal path removes intermediate, VCO, and sum frequencies, leaving only the difference-frequency output of the product detector.

been able to learn and so we must wait a bit longer for that happy time.

Right now, though, we can build an FM receiver with excellent performance which will also be able to detect conventional AM, for a small fraction of the cost of conventional receiver circuits.

How Can We Use Phase-Locked Loops?

No new component, no matter how interesting it may be, is of much good to many of us unless we can put it to use. The logical question at this point, then, is "How can we use phase-locked loops?"

In the initial report describing the IC phase-locked loop, Dr. Alan B. Grebene listed nine electronic circuit functions for which he felt it was well suited:

1 – FM i-f strip and demodulator for commercial FM receivers.
2 – Commercial TV sound i-f and demodulator.
3 – Tuned AM detector.
4 – Self-contained SCA (storecast music) receiver.
5 – FM/multiplex telemetry receiver.
6 – Signal conditioner and limiter.
7 – Frequency-shift telegraph receiver.
8 – Frequency selective multiplier and divider.
9 – High-linearity FM detector for wide-deviation FM.

To his list, we can add several more directed specifically at amateur use:

10 – VHF FM mobile receiver for compact cars.

11 – VHF FM handheld receiver.

12 – Frequency synthesis for high-accuracy VHF FM transceivers (*Boelke, 73*, Feb. 1970).

Some of Dr. Grebene's suggested applications are outside our scope, such as the telementry receiver, signal conditioner, and TV sound demodulator. Several of the others telescope into a single application when signal frequencies are ignored – that of a versatile FM receiver. So let's see how to use the PLL (phase-locked loop) IC as an FM receiver, as an AM receiver, as a frequency multiplier, and as a FSK RTTY converter.

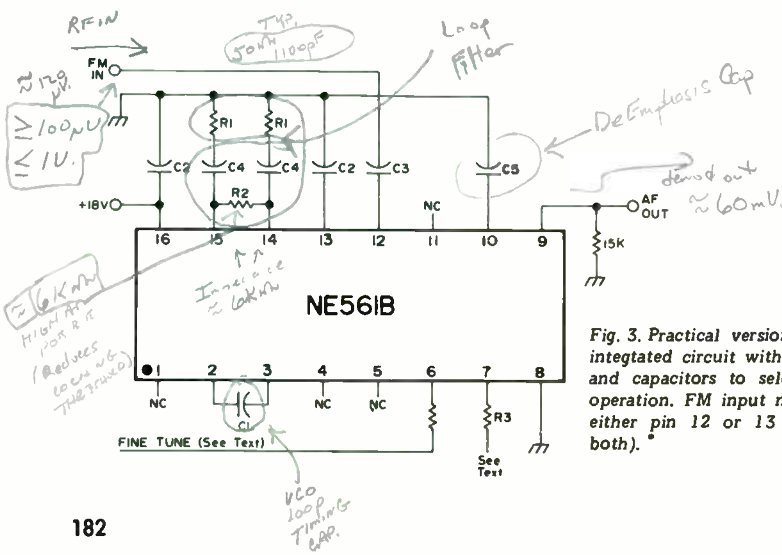


Fig. 3. Practical versions of PLL using integtated circuit with external resistors and capacitors to select frequency of operation. FM input may be applied to either pin 12 or 13 (or push-pull to both).

Let's emphasize that the circuits which follow are taken from the manufacturer's published applications notes and literature; we have not breadboarded any of them and so cannot guarantee results (but let us know if you have troubles with any of them).

FM Receiver

FM reception is simple with the IC PLL; that's the job it appears to have been invented to do in the first place. When it's running "locked" to a signal, the average dc level of the phase-detector output is directly proportional to the frequency of the input signal. As the input frequency shifts with modulation, it's this dc output that changes and causes the VCO to shift its frequency to remain locked on the input signal.

The only problem to face in building an FM receiver with the IC PLL is that of setting it up for the proper input signal frequency. The VCO center frequency is set by an external timing capacitor, and by varying the value of this capacitor we can work at any frequency from 100 Hz up to a guaranteed 15 MHz, with typical units operating to 30 MHz and operation as high as 60 MHz possible by a trick we'll pass on a little later.

The connections necessary, as well as the external components required, are shown in Fig. 3. Capacitor C1 is for timing; its value in picofarads for frequencies in the range from 100 Hz to 30 MHz is shown in the left-hand graph of Fig. 4. This capacitor provides a "coarse" adjustment of frequency which can be trimmed by variation of capacitor value. For "fine" frequency adjustment, current must be injected into pin 6 through a series resistor from the power supply; the right-hand graph of Fig. 4 shows the percentage of frequency change achieved for various values of current injection.

The other capacitors shown in Fig. 3 are bypass capacitors (C2), coupling capacitors (C3), and low-pass filter capacitors (C4). Their values are dependent to some degree upon the center frequency chosen. The deemphasis capacitor (C5) should be larger than 200 pF for commercial FM reception; its value is subject to experiment for communications purposes.

Resistors R1 together with capacitors C4 form the low-pass filter; it's fed by an internal impedance of 6 kΩ at pin 14 or 15. Typical values for R1 and C4 are 50Ω and 1100 pF.

The locking threshold of the circuit may be controlled by connecting resistor R2 across pins 14 and 15. Resistor R2 is normally left out, but at high input signal levels or high input signal frequency this reduction in threshold may be necessary to prevent instability. Approximately 6 kΩ is a typical starting point; the value of R2 should be as high as possible in any specific case, though.

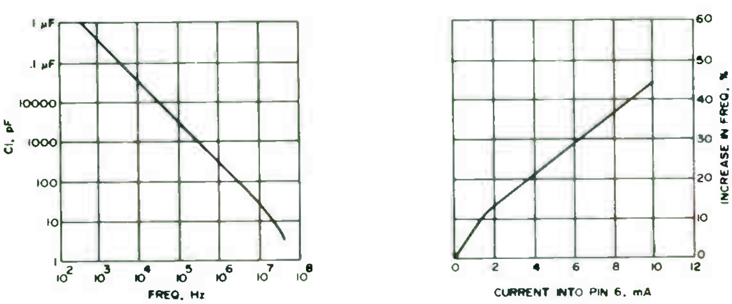


Fig. 4. These two graphs show how operating frequency of VCO in IC PLL is set. Graph at left shows approximate center frequency as value of C1 is changed through the range from 1 μF to 10 pF. This is coarse frequency setting, determining lowest operating frequency. Graph at right shows how frequency increases as current is fed into pin 6; approximately 45% increase in frequency can be attained by this means. This provides "fine tuning" control.

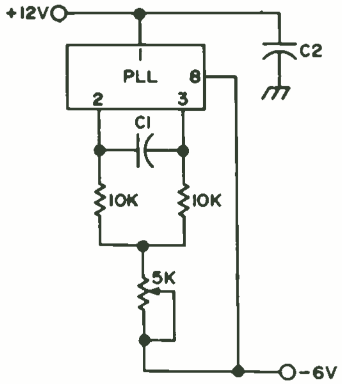


Fig. 5. The NE560/NE561 IC PLL is rated for operation only up to 15 MHz, although many units operate satisfactorily up to 30 MHz. Operating frequency for all units can be increased to approximately 60 MHz by connecting pins 1, 2, 3, and 8 as shown here. This modification can be applied to any of the accompanying circuits. The 5 kΩ pot serves as a fine-tuning control, replacing the current injection into pin 6. If current injection is to be used, the pot can be omitted from this circuit.

Tracking range may be controlled by current injection into pin 7 through resistor R3. If R3 is omitted, tracking range will be approximately 15% of center frequency. When R3 is set to a value which permits 0.65 mA of current to flow into the 600Ω impedance at pin 7, this figure is reduced to approximately 3%.

Other controls are possible, but should not be necessary when the PLL is used as an FM receiver. Input signal level should be at least 120 μV to either pin 12 or 13 (the unused input pin should be ac-grounded through a bypass capacitor). Output at pin 9 across a 15 kΩ external resistor (which must be in the circuit; this is an open emitter in the IC) should average 60 mV.

For VHF operation of the PLL, two 10 kΩ resistors and a 5 kΩ pot should be added externally as shown in Fig. 5. According to applications engineer Ralph Seymour, this modification extends the frequency range up to 60 MHz. The 5 kΩ pot provides convenient fine-tuning of center frequency, and may be omitted.

Because of the comparatively high signal level required, and the low input frequency (when compared to the 2 meter band, for instance) a VHF FM receiver for ham use of the PLL would require a converter ahead of the PLL circuit. A block diagram of such a hookup appears as Fig. 6. Note that the converter could be crystal-controlled to produce a 28 MHz output; by adjustment of current into pin 6 the PLL can easily be tuned over the full band from a 30 MHz center frequency. This provides, for less than $50, an FM receiver based on the time-honored 75A4 principle of crystal-controlled front end and tunable i-f.

AM Receiver

The AM reception capability of the PLL is something of an "extra." The PLL is locked to input signal carrier frequency and its output is used as the local oscillator to a built-in product detector. A 90-degree phase shift is required to obtain proper operation.

Figure 7 shows the hookup. Bypass and coupling capacitors, all shown as 0.1 μF, should have low impedance at operating frequency (this circuit is intended to cover the BC band rather than HF or VHF). Capacitor C1 is selected to obtain the proper center frequency, and C2 sets the bandwidth by filtering off high-frequency audio output.

The phase-shift network (R1–2, C3–4) provides 90 degrees of phase shift for the rf

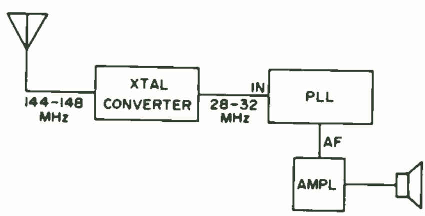


Fig. 6. For ham use as a VHF FM receiver, a crystal-controlled converter ahead of the PLL is necessary. This converter need only provide a moderate output signal level, however, and can be broadband with all receiver tuning being accomplished at the PLL circuit.

input signal; the sum of R1 and R2 should be less than 5 kΩ (2 kΩ each was the value used in the prototype) and C3 and C4 should have reactance equal to the values of R1 and R2. For BC operation, 82 pF was chosen.

In this hookup, the low-pass filter is not critical, since no audio is taken from the loop itself; so a simple 0.01 μF capacitor from pin 14 to pin 15 suffices.

Tuning may be done in either of two ways. The first is more straightforward but the second is elegant. The first way is to vary the value of timing capacitor C1. For BC operation, C1 should tune from a minimum of 220 pF (for 1600 kHz) to a maximum of 620 pF (for 500 kHz). The classic receiver alignment technique of padding capacitance is used at the high end of the range, but at the low end varying the current into pin 6 (fine-tuning control) takes the place of adjustment of inductance.

The second method of tuning the receiver uses a fixed value for C1. This value is whatever is required to make the VCO operate at 940 kHz (geometric mean frequency) when the current at pin 6 is zero. Pin 6 is then connected through a 1.2 kΩ resistor to the arm of a 5 kΩ pot (Fig. 8) across the 18V power supply. Varying the pot tunes the receiver across the BC band using only the fine-tuning feature.

This receiver requires an antenna and a good ground; it must get at least 100 μV from pin 9 to ground. Operation will be improved by using a broadband untuned rf amplifier, but care is necessary to assure that the PLL is not overdriven (maximum input signal if 1V rms). Maximum audio output is 2V p-p, and typical output is about 60 mV.

Frequency Multiplication

The PLL IC can be used as a frequency multiplier in several ways. The simplest is merely to set the center frequency to some multiple of the actual input signal frequency. However, as the higher (and thus weaker) harmonics are used for locking, the lock range decreases. If input frequency is fairly stable and rapid tracking is not required, this technique can be used to multiply by 2, 3, 4, or 5 times any input signal. Output of the VCO at pin 5 is a square wave.

Action of the PLL as a multiplier can be improved by converting the input signal to a square wave, which has much stronger harmonics than does a sine. When this is done, any output frequency up to 15 MHz can be produced from its tenth subharmonic (multiplication of up to 10 times).

Since the PLL output is already a square wave, rich in harmonics, this offers an opportunity for use as a frequency marker. Two PLLs in series could produce a 1 MHz standard from a 100 kHz crystal, and in turn a 10 MHz standard from the 1 MHz standard. One great advantage of the

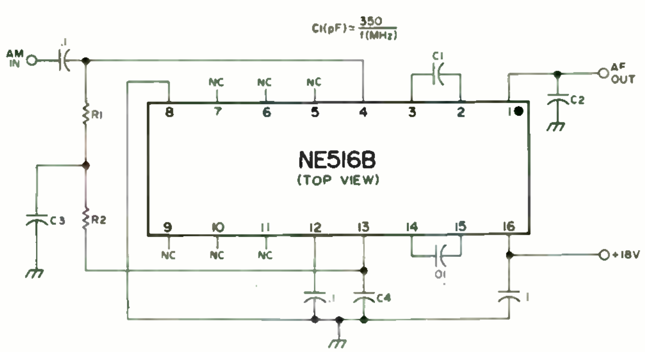


Fig. 7. This is a BC-band AM receiver using the NE561 IC PLL (the NE560 does not include AM detection capability).

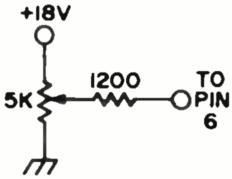


Fig. 8. This simple arrangement for connection to pin 6 of NE560/NE561 provides fine tuning over a 3:1 range. The 1.2 kΩ resistor limits maximum injection current. Timing capacitor should be set for lowest frequency with pot arm at ground, and 1.2 kΩ resistor then trimmed to set highest frequency desired, with pot arm at hot end.

PLL multiplier as compared to a multivibrator or a conventional tuned-amplifier multiplier is that the order of multiplication can be changed merely by changing center frequency (such as with switched timing capacitors). This would permit generation of a 3.5 MHz square wave as the seventh multiple of a 500 kHz standard, giving you strong markers at the bottom edge of every HF ham band.

The PLL multiplier will divide frequency just as easily as it multiplies. If center frequency is adjusted to be one-third that of the input signal, the circuit still locks. This action occurs only for *odd* submultiples, however; for even divisions (half, quarter, eighth) it doesn't work. You can divide 3, 5, 7 or 9 only.

Extensions of these features make possible the construction of simple frequency synthesizers; too many PLLs are required to make them practical at present prices, but when costs come down they may well be worth investigating. Collins Radio's book "Fundamentals of SSB" (out of print since 1962, unfortunately) gives the principles involved if you're interested.

The hookup for use of the PLL as a frequency multiplier or divider is shown in Fig. 9.

FSK Converter for RTTY

Using the PLL as an RTTY FSK converter is almost identical to its use as an FM receiver, because FSK (or AFSK) is merely a means of carrying binary or telegraph (mark–space) information by means of frequency modulation. The frequency shift involved is usually rather small – but so is that for FM, in comparison with the center frequency.

The FSK input, either at communications-receiver i-f for FSK or at audio frequency for AFSK, is applied to the FM input of the PLL. The loop filter capacitor is made smaller than usual to eliminate any possibility of "overshoot" in the output pulse, and a three-stage ladder filter removes the carrier component from the output (this is necessary only for AFSK). The center

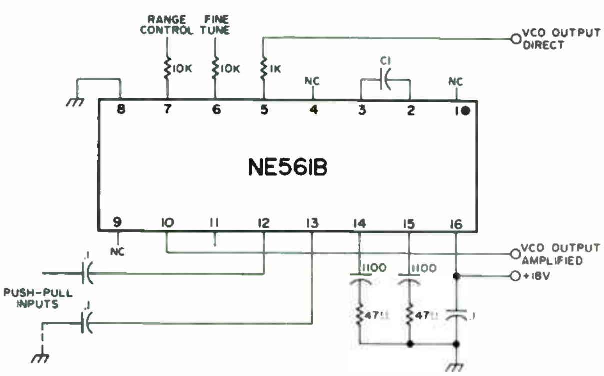


Fig. 9. PLL makes fine frequency multiplier or divider. For this application, audio output connections are ignored and the VCO output is used instead. If input is single-ended, one of the two push-pull input leads should be bypassed to ground as shown by dotted lines. Circuit will multiply up to 10 times, and divide input frequency by 3, 5, 7, or 9. C1 and fine-tuning adjustment must be set for operation near desired output frequency. When input is applied, VCO will lock to exact multiple or odd submultiple of input if it is within locking range and of adequate strength.

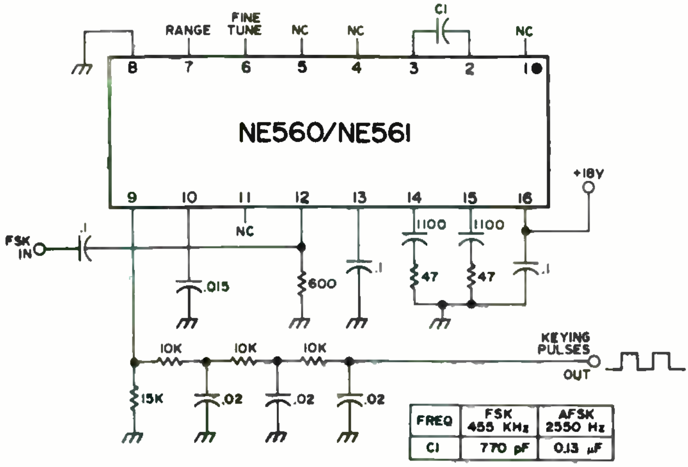


Fig. 10. RTTY converter circuit is taken from computer data set applications note; data set is same as AFSK converter, but gets input signal from telephone line and so is not subject to such high levels of interference as is RTTY. Input may be either at i-f or audio frequencies; table shows values of C1 for both cases. Output consists of pulses which may drive a keying circuit for selector magnets.

frequency is adjusted to produce about 12V at the output when the input frequency is at its lower figure. When input frequency rises to its other figure, output will rise a maximum of 4V as the VCO tracks the input-frequency change. This voltage change is coupled out through output coupling capacitor to drive any external circuits desired, such as a magnet driver.

The circuit is shown in Fig. 10. Figure 11 shows an alternate circuit using a different type of PLL chip together with an IC voltage comparator at the output to change the output levels to values which are compatible with normal digital IC chips. In this case, center frequency is adjusted to produce a slightly positive voltage at the output, when input is at the low frequency.

SCA Adapter

The PLL's capability to receive FM at almost any center frequency comes in handy if you like background music from

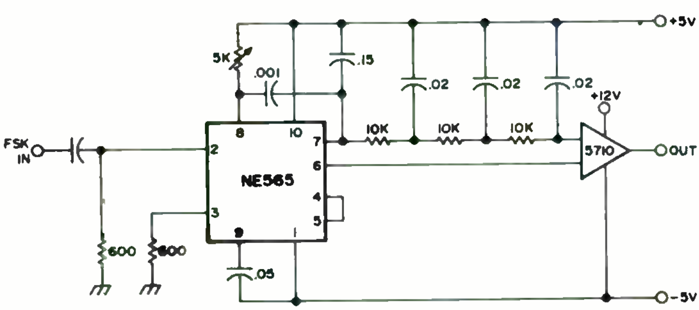


Fig. 11. Alternate RTTY circuit uses NE565 IC, which is so new that its price is not listed yet. Maximum frequency of 565 is 500 kHz. This circuit is designed to drive digital IC devices, and type 5710 voltage comparator is included to adjust output level to values suitable for digital ICs. Pot is for frequency adjustment.

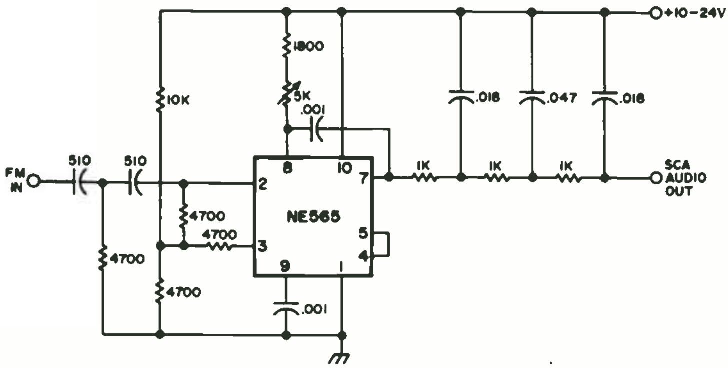


Fig. 12. Background music adapter for FM receivers uses type 565 IC PLL. 510 pF capacitors and 4.7 kΩ resistors at input form high-pass filter to keep audio from FM set from overloading PLL; ladder filter at output removes everything above about 10 kHz to keep from overloading audio amplifiers following. Pot allows frequency to be adjusted to 67 kHz to pick off SCA subcarrier.

the "storecast" services sold by many commercial FM stations. These storecast signals are transmitted as FM of a 67 kHz subcarrier which is itself a part of the normal FM signal. To receive them with the PLL, all you need do is set up the PLL for a center frequency of 67 kHz and tap off an output from your FM receiver ahead of the deemphasis filter (which bypasses the 67 kHz signal to ground) to feed the PLL. PLL output will then be the background music, which is free of all interruptions such as commercials or even station identification.

One precaution may be necessary. The signal from the FM receiver to the PLL input should go through a high-pass filter to prevent any possibility of overload by the much stronger audio of the normal broadcast program or any accompanying stereo information at 38 kHz. A typical hookup using the NE565 PLL is shown in Fig. 12.

The Source

About the only thing we haven't yet told you about the PLL IC chip is who makes it and where they can be obtained.

The manufacturer is Signetics Corporation, a subsidiary of Corning Glass Works and one of the leaders in the integrated-circuit industry for a number of years now. The PLL IC is only one of many chips in their line (their price list as of 4 May 1970 required 21 pages merely to list prices of current IC products).

Signetics is located at 811 E. Arques Avenue, Sunnyvale CA 94086, and we obtained our information from Ralph Semour, linear applications supervisor.

They make the PLL in several different models, with model numbers to match. The one shown in most of these diagrams is the NE561B, which is the most general in application. It includes the AM-detection capability, and quoted price in lots of 1-24 is $37.50 each. The NE560B is virtually identical but does not include the AM-detection features, and lists at $30 each. Models NE565 and SE565 (the N indicates *commercial* temperature range; the S means *military* range) are too new to be listed; they operate only up to 500 kHz and so presumably would be lower in cost. The 565 is used in the circuits of Figs. 11 and 12. Finally, the NE562 is intended for direct interface with digital logic systems and, like the 565 it is too new to have an established price yet.

Pin connections for the 560, 561, and 565 are shown in Fig. 13.

The N5710T comparator shown in Fig. 11 sells for $2.62 each in lots of 1–24.

All of these (with the possible exception of the 562 and 565, which were still *preliminary* designs at this writing) may be obtained on special order from the manufacturer. In addition, the major mail-order

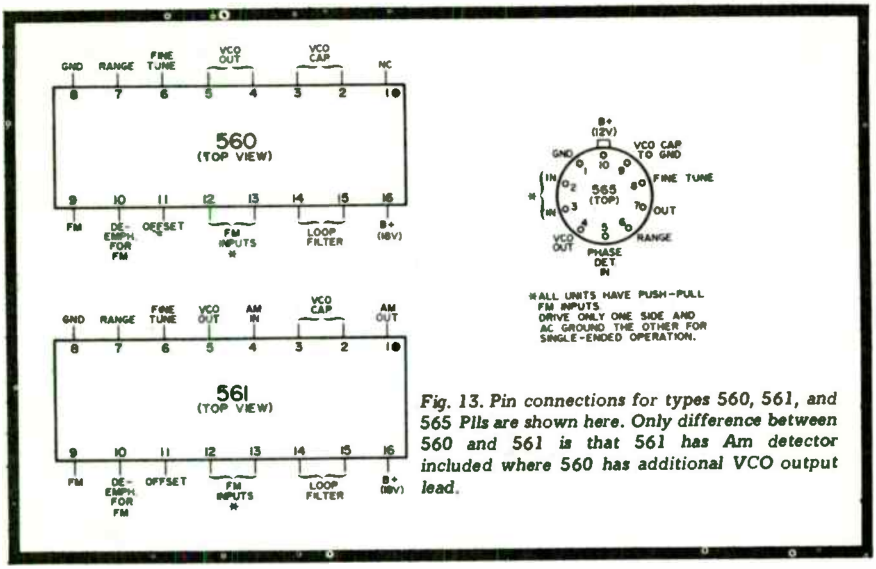


Fig. 13. Pin connections for types 560, 561, and 565 Plls are shown here. Only difference between 560 and 561 is that 561 has Am detector included where 560 has additional VCO output lead.

houses catering to the industrial electronics trade may be able to obtain them although no catalog available to us lists Signetics as one of the lines carried in stock (the firm deals primarily with the original-equipment-manufacturer market). Inquiries should be addressed to Mr. Seymour and should mention this article.

. . . K5JKX ■

Index

H

I

K

L

M

N

O

P

Q

R

S