SOLID-STATE PROJECTS
for the Experimenter

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Solid-State Projects for the Experimenter

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Preface

Today's experimenter—amateur radio operator, audiophile, and the hobbyist with a variety of interests—just can't seem to get his fill of projects. And with all the inexpensive solid-state components available, there seems to be no end to the unique devices one can build—and put to practical use.

Here are more than 60 projects, gleaned from the pages of 73 Magazine, of interest to anyone in electronics. The devices range from a simple transistor tester to an electronic counter, from a capacity meter to a ham TV receiver. The idea of this collection is not only to provide you with interesting and practical projects to build, but also to help you become more intimately acquainted with such modern components such as ICs, FETs, thyristors, varactors and zener diodes, etc. Have fun!

Wayne Green

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CHAPTER 1

Zener Diodes

One of the many solid-state devices now available to the radio amateur builder is the zener diode. Properly used, it serves as a reference voltage source capable of delivering considerable current. Unlike a battery, its life is indefinitely long, although it must be supplied with a continuous current for many of its applications. This article contains the basic information required to intelligently design and use zener diodes, and some selected sources of information from the rather sparse supply are listed in a bibliography at the end.

For instance

The number one application of zener diodes is probably dc voltage regulation for transistor circuits. Suppose you have just purchased a new 1N2974 zener for about $5.10. The catalog says 10 watts, 10 volts, 20% tolerance. Since you lack experience with zeners, you breadboard the intended circuit to pick up a few details on what zener regulators do. Perhaps the circuit is that of Fig. 1. Careful! The illustrated circuit is more suited to my purposes than yours! You turn it on and this might be what happens:

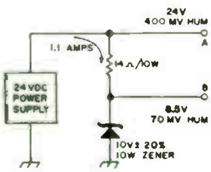

Fig. 1. A possible hit-or-miss zener regulator circuit. What's wrong with it?

A voltage measurement reveals regulation at 8.5 volts rather than the indicated 10 volts. You notice the voltage seems to be creeping up as you watch the meter. A signal tracer probe on point A finds lots of hum; but there is some at B too. Perhaps if some current is drawn from B the hum will decrease. A load test shows little effect on the hum except that under very heavy load the hum increases . . . some other noise too! Thinking about that, you smell smoke; it's the series resistor overheating. As you reach over to turn off the power supply, you notice the zener voltage seems to have dropped to zero. This reminds you that the zener has been operating without a heat sink. It's failed completely! Terrible. You can avoid the damage to wallet and peace of mind by using the following material. After reading, get out a pencil and some old envelopes or something, and design zener circuits. You'll soon catch the idea!

Zener facts

Zener diodes are supplied in a large variety of packages. The controlling factor is how much heat must be dissipated. A majority of the zeners available are supplied in a diode-like glass package for up to 250 milliwatts, a wire-mounted cylinder resembling a resistor or a silicon rectifier for the 1-watt size, a stud-mounting package for the 10-watt size, and a larger package resembling a power transistor for the 50-watt size.

A catalog search brought out hundreds of zeners rated from 250 milliwatts to 50 watts. Some engineering books mention zeners as small as 50 milliwatts (Cutler) to as large as 100 watts (Littauer). Operating voltages ranged from 3 volts to 200 or so. The least expensive were priced at 75 cents (General Electric Z4XL series). $20 seemed to be the upper limit, with a large variety in the $4-$7

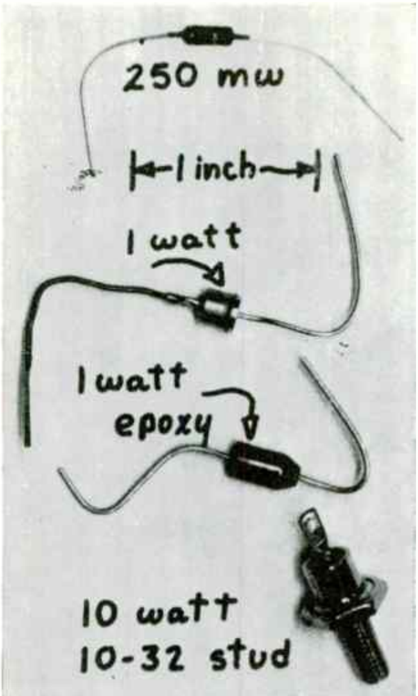

Fig. 2. Some typical zener diodes. They look just like ordinary diodes. Perhaps some of your diodes are zener diodes!

range. Price tends to increase with higher power rating and closer tolerance, but some good zeners are available in epoxy packages at low prices.

Some ordinary silicon diodes and transistors may be used as zeners. General Electric says some of their epoxy cased transistors can be used in this way. But most zeners are slightly special silicon diodes, designed to dissipate the fairly large amounts of heat produced in normal zener operation. Small zeners dissipate the heat along their leads or into the air; 10 watt and larger zeners are built like power transistors. The semiconductor material is brazed to a copper stud or surface which provides the route for dissipating excess heat into the required heat sink. Fig. 2 shows some typical zeners.

Only a small part of the package called a zener diode is actually the working element. This key piece is a semiconductor PN junction formed on a silicon wafer by a process involving some heat and considerable accuracy. Fig. 3 shows the interior of a 10 watt zener. In normal zener operation the PN junction is biased opposite to the direction of easy conduction, at a voltage great enough so it conducts anyway. Sounds rough, but works fine. The arrangement is called reverse bias, and the zener always appears in the schematic with its arrow pointing toward the positive supply line. Fig. 4 is a graph of current plotted against voltage for normal zener operation.

A small voltage invokes very little current. As the voltages approaches 10 volts, the current increases quite drastically toward some terrific value as the zener begins to act like a short circuit. If there is no current-limiting resistance in the circuit, the zener will promptly perish. This very rapid current upon voltage dependence gives the zener its useful voltage regulating property. The region of the curve in which the current first begins to rise is called the zener knee, and the normal operating region is called the zener plateau. The useful plateau is limited at one extreme by the current required to keep the zener action alive, and at the other by temperature increase sufficient to destroy the zener.

The zener regulating voltage depends on how thick the PN junction is. If the PN junction is very narrow, the zener will regulate at a low voltage; we leave these details to the manufacturer. But depending on the structure of the zener it may show an increase or a decrease of voltage as it gets warmer! Zeners under about 5 volts will show a decrease in voltage, over about 5 volts a rise, with increasing temperature. A happy choice of voltage and current will give a zero drift: 40 mA

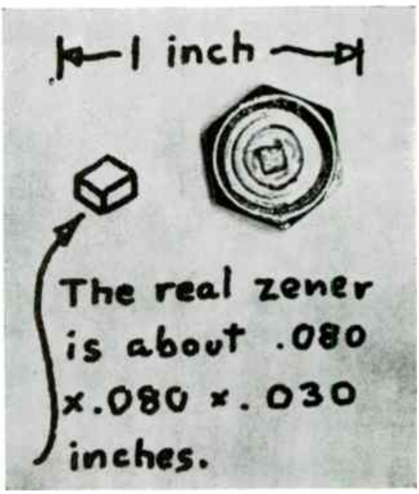

Fig. 3. Inside the case of a 10-watt zener. The actual PN junction is in the tiny square at the center of the circle.

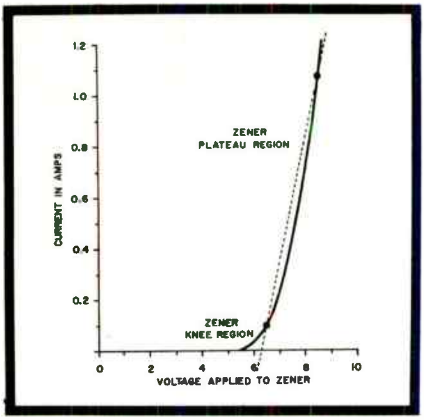

Fig. 4. How zener current depends on applied voltage for a 10 watt zener. The rise beyond the knee is so sharp the zener must be supplied from a current-limiting resistor or circuit.

at 4.8 volts, to 3 mA at 6 volts. Review the manufacturer's specs if real stability is required. Two or more zeners in series will show a smaller temperature voltage drift than a single equivalent higher-voltage zener.

Zener specifications

All zeners are supplied with a voltage rating, and a tolerance. Like resistors, the standard tolerances are 20%, 10%, and 5%. The nominal values are usually chosen in the same way as those for resistors, resulting in voltage ratings that should sound very familiar. The system is based on the twelfth root of ten for 10% zeners, so you will find for instance, 3.3, 3.9, 4.7, 5.6, etc., voltage ratings. Fig. 5 lists some zeners and their properties. If you are trying to regulate to a critical voltage, a germanium or a silicon diode may be placed in series with the zener (small increment) or in series with the load (small decrement).

The temperature drift problem can be minimized by keeping the zener cool. This conflicts with power handling ability but tends to guarantee long life despite experimental accidents. Like all semiconductor materials, if a zener PN junction gets too hot, the doping atoms begin to jump into new sites. This is very bad for the zener! Since the junction may withstand temperatures as high as 200 degrees Centigrade, zeners are not remarkably fragile. But they cannot withstand the kind of overload even small power supplies can produce.

Type	Price	Nom. Voltage Test Current	Watts	Tolerance	Dynamic Resistance in Ohms
1N4728A	$1.93	3.3@76 mA	1	5%	10
1N4733A	1.93	5.1@49 mA	1	5%	7
1N4735A	1.93	6.2@41 mA	1	5%	2
1N4739A	1.93	9.1@28 mA	1	5%	5
1N4747A	1.93	20@12.5 mA	1	5%	22
1N4752A	1.93	33@7.5 mA	1	5%	4.5
1N957B	2.95	6.8@18.5 mA	0.5	5%	4.5
1N3016B	3.70	6.8@37 mA	1	5%	3.5
1N2970B	7.30	6.8@370 mA	10	5%	1.2
1N2804B	10.65	6.8@1.85 A	50	5%	0.2
Z4XL6.2	0.75	6.2@20 mA	1	20%	9
Z4XL6.2B	0.84	6.2@20 mA	1	10%	9

Fig. 5. Types, prices, and characteristics of some typical zeners. Note variations in dynamic resistance.

Zener wattage, as in any resistor, equals voltage across the zener times current through the zener. Check manufacturer's specs if much power is to be handled or if operating near maximum ratings. For breadboard and quick-and-dirty construction the fingertip test will do: too hot for a five-second fingertip touch equals too hot.

The zener's ability to stabilize and filter a power supply output is indicated by its dynamic resistance. Low dynamic resistance is desirable. Suppose you wish to have power supply output stay within one-tenth volt of nominal in spite of 100 mA variations in current. By Ohm's law that works out to one ohm: this is the dynamic resistance required. High wattage zeners, near 6 volts, have better dynamic resistance than any others; high voltage zeners have very poor dynamic resistance but are not required for most semiconductor circuits.

The practical effects of dynamic resistance can be brought out by drawing equivalent circuits to show what's involved. An equivalent circuit is something used by engineers to simplify circuit problems. The dotted box suggests "we imagine the actual device acts like what's in there." From a hum viewpoint, the zener improves the situation as if the entire circuit were a voltage divider. The upper resistor is the required series resistor R_s and the lower resistor is the zener's dynamic resistance. This is usually listed in the catalog entry. If the dynamic resistance is one ohm, and the series resistor is fifteen ohms, the usual way of working out voltage-divider circuits tells us the hum will be reduced by a factor of 16. That doesn't remove it! Typical hum from a capacitor filter low-voltage supply is a half volt to three volts. This would result in

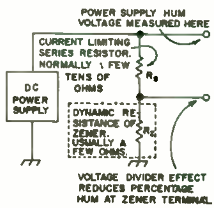

Fig. 6. Equivalent circuit for estimating how much a zener regulator will reduce the percentage and amplitude of power-supply hum.

anticipated hum figures 30 to 200 mV; still plenty of hum.

Fig. 7 shows the way to estimate how much the zener voltage will change under load. This figure is very different from the actual circuit so do not feel stupid if it's not clear! Try this with an actual zener when you've finished the article. Measure the zener voltage at a current near the knee. Measure it again at near maximum zener current. It will be higher. Now write down that higher voltage next to the 'inside battery.' If you draw current from this equivalent circuit, the voltage will drop because of losses across the inside resistor. This is just what the real zener circuit did: the small zener current corresponds to maximum load condition from the equivalent circuit. The voltage change divided by the current change gives the value of the series resistor. It will be the dynamic resistance again. This shows that knowing the dynamic resistance is very useful in reckoning the effects on regulated voltage of changes in load current.

Zener noise

The useful ability of zeners to control circuit hum and noise is somewhat compromised by their natural ability to generate signals of their own. The zener regulating process is like an electrical discharge, and can produce similar noise. Good zeners produce very little noise, but some may become quite loud in the zener knee region. If your new circuit seems to be troubled by erratic frying and hissing noises, and this is traced to the zener regulator; the two solutions are increasing zener current to keep it out of the knee region, or the addition of a capacitor to take up the noise. Try 0.1 μF to start.

Varactors are reverse-biased silicon diodes. Since zeners are also reverse-biased silicon diodes, do they have an associated capacitance? They certainly do; it may be as large at .01 μF. This capacitance is unimportant in normal operation, and probably has a beneficial effect in reducing zener impedance at high frequencies. Perhaps this capacitance can be put to other uses! Might be an easy way to

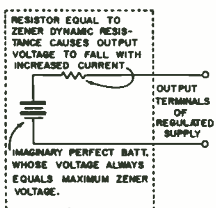

Fig. 7. Equivalent circuit for reckoning voltage drop with increased loading of a zener regulated supply.

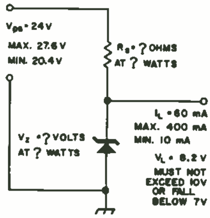

Fig. 8. Designing a reliable zener regulator. The question marks indicate values to be worked out.

double up to 40 from 80 meters. Another possible application would be oscillator tuning; perhaps a cheap zener would work better than a cheap silicon diode.

Zener regulator design

All zener circuit designs depend on the same basic facts of the zener's voltage, tolerance, wattage rating, dynamic impedance, and perhaps temperature drift. The worst case from the zener's viewpoint is DC power regulation, with a steady current supply from a voltage source capable of destroying the zener if the series resistor fails or is shorted. Zener regulators are very useful and deserve a close examination.

Only professional engineers should design zener circuits to operate at the limits of the zener's capabilities. The amateur, by leaving generous margin for error, can simplify the design problem to a point where only the simplest math and less than complete information on the zener's capabilities will be sufficient for a reliable design. Apart from possibly very serious misunderstandings on the amateur's part, the major troubles that might arise are the relatively generous tolerances on most inexpensive zeners, the wattage problem, and voltage change under load resulting from dynamic resistance.

The design process commences with finding some working figures. Refer to Fig. 8, a design sketch made just before carrying out the following procedure. Power supply voltage, zener voltage, and load current information are collected, along with their estimated maximum and minimum values. A class B audio amplifier of a few watts capability could account for the fairly large current variations shown in the diagram; this is a rather extreme case. But it could happen! If the load were a receiver small-signal circuit, an oscillator, or a transmitter VFO; the load would be practically constant. After collecting this information from figures, estimates, educated guesses, and by breadboarding, the circuit is designed on average values. When a series resistor and a zener are chosen, their anticipated properties are checked against possible extreme conditions of voltage and amperage.

Variations in load current are made up by opposite variations in zener current. Minimum zener current flows when maximum load current is taken from the regulator. This adds up to the first requirement for the zener: it must carry a current greater than the anticipated load swing. The regulation fails if the zener is starved into the knee region, and if the zener is overheated by excessive current, catastrophe is likely. Often a well-placed large capacitor in the load circuit will absorb the drastic swings. The zener required in Fig. 8 must carry more than the swing of 390 mA. If the minimum and maximum loads were both greater by 1 A, the swing would remain the same and so the zener minimum current requirements would be unchanged. However, with increasing loads flowing past the zener, a point may come at which turn-on or turn-off of the circuit results in transient overloading of the zener.

A safe minimum zener current is 10% of its maximum rating. A fresh zener accompanied by manufacturer's specs may be used at much lower levels, taking the specs as a reliable guide. For surplus zeners the 10% lower limit is recommended unless a test shows the particular zener will stand further starvation. Since the voltage requirements are already determined, the choice is made on the basis of wattage. Don't be afraid to use an oversize zener; it's safer and the larger zener will have a lower dynamic resistance. The absolute minimum wattage required equals the maximum possible voltage across the zener times the maximum expected current through it. There seem to be no 5 watt zeners; for most regulator purposes the choice is limited to 1 or 10 watts. Or 50 watts if you wallet can stand the drain. The zener in Fig. 8 should carry about 450 mA under minimum-load conditions; at 9 volts that works out to just under 4.5 watts. A 10-watt zener is indicated.

Now the series resistor R_s can be chosen. The voltage at its upper end is fixed by the power supply, and at its lower end by the zener. The series current is the maximum load current plus the minimum zener current finally decided on . . . in the case of Fig. 8 this was a total of 450 mA. This same current flows if the load drops to 10 mA, since the zener now takes 440 mA. Knowing the voltage across the resistor, 24 volts minus 8.2 volts, or 15.8 volts, it appears that a resistance of 35 ohms and about 7 watts is the minimum value.

This leaves no margin for error. A better choice is an Ohmite adjustable wire-wound resistor, 50 ohms, 25 watts. More than half the resistor will be in the circuit. If only half or 25 ohms were used, it would still be rated at 12.5 watts, so that with this choice the success of the design seems probable.

Now we return to the zener. Any electrical slop in the design can be taken up by the resistor, and we know that a 10-watt zener is required. Referring to the catalog, we find a 1N2973B, 9.1 volt 5% zener. We expect the circuit can withstand the possibly slightly high voltage. If not, we'll change the circuit. We choose a zener at the high end of the range because its voltage will drop under load: a dynamic resistance of 2 ohms means 2 volts drop per amp decrease in zener current, or in the case of Fig. 8 the voltage will swing over a range of 0.23 volts. Adding to this the .45 volts possible error due to zener tolerance, and adding that to the 9.1 nominal zener rating gives about 9.8 volts as the largest we should expect to see in the circuit. Subtracting the same figure from the 9.1 nominal figure gives a minimum of 8.4 volts in the case of an extremely low valued zener. These results are within the previously decided requirements.

There is still the question of changes in power supply voltage. What happens if line voltage changes drastically? This cannot be answered simply; some line voltages are more changeable than others! In Fig. 8 a 10% variation either way was just picked out of the air. This is probably large. Now if the zener voltage is fixed at 10 volts, which it will never quite reach in the actual circuit, and if the supply voltage drops to 20.4 volts; the series resistor being at 35 ohms with 10.4 volts across it now passes only 335 mA. Not enough, so we adjust the resistor to 23 ohms, and it now passes 450 mA. But now we must try the other extreme: the power supply voltage rises to 27.6 volts and we suppose the zener to be an 8.4 volt type. Then we find 19.2 volts across a 23 ohm resistor. That is about 840 mA. The zener won't overheat if it is properly mounted, but the resistor must dissipate 16 watts. Since less than half of it is carrying current, we must go to a smaller resistance at the same wattage in order to get enough dissipating surface. The choice is a 25 ohm 25 watt resistor, still adjustable. Or if you're a little apprehensive about going that close to the limit, a 50 ohm 50 watt resistor can be purchased at an 83 cent increase in price.

That completes the design. This process avoids the following kinds of grief: Zener failure due to overheating; zener voltage out of specs; circuit loads zener regulator into the knee region; regulator fails due to starvation or overheating at extreme line voltage values.

Correcting zener voltage

The combined effects of high prices and large tolerances are hard to beat! But a resourceful amateur need not lose a project just because his nearest zener isn't quite near enough. Zener voltage can be adjusted up or down by correct use of small germanium or silicon diodes. The price is a slight increase in dynamic resistance and in temperature drift. Some knowledge of the properties of forward-biased diodes is required.

A silicon or germanium junction diode, carrying a forward current of 10 to 100 mA, depending on its size, has properties very like those of a low-voltage zener. In fact, there are no zeners under about three volts, and diodes are used in just this way to fill the remaining gap down to near zero. Beyond the early stages of conduction, a few microamps or mils, the diode voltage changes very little with current. Its dynamic resistance is quite low. Diodes can be used in zener circuits as if they are little batteries, to achieve a slight increase or decrease in apparent zener voltage. The voltage measured across the zener itself is not affected.

The voltage at which the diode regulates depends on its material: germanium or silicon. A germanium diode well into conduction will show a stable voltage of around 0.3 volts; a silicon diode regulates above 0.7 volts. A transistor base-emitter or base-collector junction could be used in place of the real diode; it's a PN junction too and will show the same behavior.

To achieve a small increase, the diode is placed in series with the zener, as shown in **Fig. 9A**. The reverse-biased zener and the forward-biased diode point in opposite directions. The drawing is arranged so that positive current—a convention—flows down. **Fig. 9B** shows the diode in series with the load, so that its voltage subtracts from the zener voltage. They seem to be pointed in the same direction, but the current flows against the zener and with the diode. This is certainly confusing and will require some careful thinking. Try it; make your mistakes on a breadboard where they show clearly and are inexpensively remedied!

Amplified zeners

Being relatively high priced and having rather large tolerances, zeners may seem rather useless to many amateurs. But a small, inexpensive zener can be combined with a transistor, making a simple two-terminal circuit that will stand in very well indeed for a 50-watt or even larger zener. This particular transaction shows an unusual measure of profit: besides greatly reduced price and substantial easing of power limitations, dynamic resistance may be improved and becomes little affected by using diodes in series with the zener to build up its voltage. The effects of temperature upon voltage are increased but this will rarely be important. The current handling ability is multiplied by the transistor β, but the temperature drift is only that of the individual diodes in series.

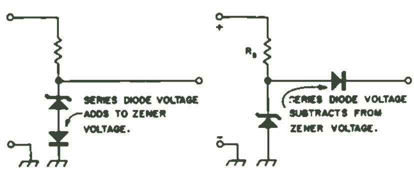

Fig. 9. The effective regulating voltage can be adjusted by correct application of ordinary germanium or silicon diodes.

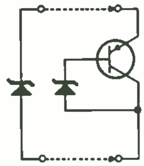

Fig. 10. Schematic of an amplified zener. It's all there!

For instance, a Texas Instruments 2N251A at $2.25 plus a General Electric Z4XL6.2 at 75¢ adds up to $3.00 for a shiny new, somewhat adjustable zener, rated about 50 watts depending on the beta of the transistor. This is comparable to the 1N2804B, priced at $10.65. That's what makes amplified zeners interesting!

The complete circuit is shown in **Fig. 10**. It does seem rather bare in comparison with most transistor circuits, but everything that's really required is there: one zener and one transistor. This two-terminal circuit closely resembles the emitter follower regulator, and if a resistor were added from the zener diode/transistor base connection up to a higher voltage to ensure liberal zener current, it would be an emitter follower regulator. But the resistor can be omitted if the amplified zener's knee region is avoided, and then the current divides between the transistor and the zener according to the beta of the transistor.

An amplified zener is shown in **Fig. 11**. This one costs an estimated 50¢ and gives very good test results. Its knee region seems to end at about 4.2 mA and the actual zener diode is not overheating at 1.6 amps regulator current. A current increase to about 1.6 amps boosts the voltage from 5.6 to 6.8 volts, for an average dynamic resistance of 0.67 ohms. Using the voltage-divider method and hum input-output measurements, its dynamic resistance at 1.6 amps is 0.59 ohms. The very best zeners are little better than that. There is one hidden pitfall: the transistor's leakage current increases with temperature. This extends the knee region to higher current values.

This is how the current division works out. There are only two terminals; the current must

Fig. 11. A real amplified zener, and its measured characteristics. This one is good for about 20 watts. Made of junk box parts, its estimated cost is 50¢.

go in one and out the other. It takes two routes in between. Suppose the base-collector zener is carrying m milliamperes. The transistor base-emitter junction supplies this current, and as a result an additional current, β times larger, flows from emitter to collector of the transistor. The total current I is the sum of these two, so that we write

$$I = m + \beta m = m\ (1 + \beta)$$

The one plus beta in the parentheses is not particularly different if we leave out the one, provided the beta is greater than ten or twenty. It usually is in a usable transistor; the difference between ten and eleven is 10%, small by electronic standards. For most purposes it's simpler yet true enough to say all of I goes through the transistor, the circuit regulates at the zener voltage plus the transistor BE voltage, and the zener heating current is I divided by β. The error is trivial.

For example, the Z4XL6.2 is rated at one watt, the 2N251A at 90 watts, and suppose a β measurement under approximate operating conditions gives a result of 50, well within specs. Remember that β is quite evanescent, depending upon collector current in addition to great variations between transistors of the same type! The maximum allowable zener current is 160 mA, since .160 amps times 6.2 volts equals the rated one watt. Then 50 times 160 mA gives a maximum of 8.4 amps current. The amplified zener regulates at 6.4 volts, since the live germanium transistor will show about 0.2 volts from base up to the emitter which is added to the zener voltage. If you really want to dissipate 50 watts, the transistor should have higher β or a 10-watt zener should be used; stay away from calculated limits!

The dynamic resistance of the amplified zener will be the inside real zener's dynamic resistance divided by the transistor β. This works out to one-fiftieth of 9 ohms: 0.18 ohms. This value is so low that the power transistor's characteristics may enter into the final result; the final value will still be well under an ohm. This result is not appreciably spoiled by adding series diodes to pad up the zener's apparent voltage.

As the current through the amplified zener is reduced toward zero, the real zener and the power transistor both weaken. The combined effects are rather uncertain, so that breadboarding with the actual components is a good, safe practice. Find the knee by measurement; remember that a capacitor across the zener will reduce its noise generating capabilities! A 6 μf electrolytic capacitor eliminated a rushing noise near the knee region in the amplified zener shown in Fig. 11.

Other zener applications

Zeners do not go very well in parallel. One will tend to hog the current. There is no need for parallel zeners anyway; an amplified zener will do a better job. But zeners can be connected in series to provide two or more regulated voltages. And in this case the ground can be between the zeners, rather than at one end of the power supply. If you ground both points, it won't work!

The design of switching circuits is considerably simplified if the usual collector voltage supply is supplemented by a lesser voltage of opposite polarity. The second supply is used to drain off unwanted leakage currents, turn diodes and transistors hard off, and for other applications. This relatively slight increase in the designer's armament eases many tough circuit problems. By using a pair of zeners in series, the desirable pair of voltages can often be obtained without going to the time and expense of a separate second power supply and all its problems of cost, space, weight and regulation. Fig. 12 shows how simple this arrangement is.

Without a load circuit, the same current flows through both zeners. If some current is side-tracked around either zener, the voltage

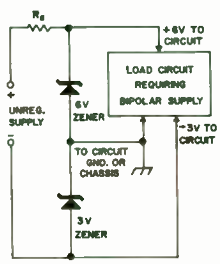

Fig. 12. A pair of zeners in series to provide both positive and negative voltages with respect to circuit ground, using a single power supply.

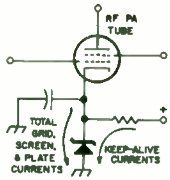

Fig. 13. Zener diode biasing for an RF power vacuum tube. Bypass capacitor recommended, appropriate for operating frequency.

across it remains constant. The regulation isn't disturbed at all if some current is taken out around both zeners, and this is the normal application. The usual considerations about starving and overfeeding zeners are applicable here, and the beginning designer should remember that the two supply circuits don't necessarily require the same currents at the same time.

The zener diode shows little promise as a limiter and no schematic for this application is included. Ordinary diodes are distinctly superior. Zeners require too high voltage: 3 volts or more. In normal circuits that's in the high-power range. The clipping should be carried out well before the signal gets this large. Also, the zener will clip at normal silicon diode levels as soon as its PN junction is forward biased. Additional diodes would be required to prevent this, unless unsymmetrical clipping were intended. Zener clipper circuits appear impractical.

The only remaining field in which vacuum tubes retain some superiority to solid-state amplifiers is large-signal rf power amplification. The zener diode can fill a very useful spot here. It can replace the cathode bias resistor, offering a bias voltage quite independent of tube current. Fig. 13 shows a zener in this application.

Because the zener acts like a battery, most of the high voltage is taken up by the vacuum

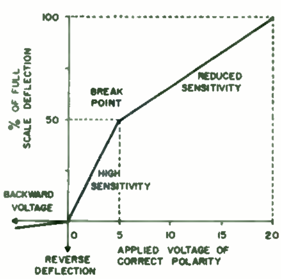

Fig. 14. How an improved meter might be calibrated to show small and large voltages with comparable accuracy. Reverse voltage does not bang the needle backward.

tube. The zener merely guarantees the bias. It never runs down or emits corrosive chemicals, and has a lower internal resistance than the batteries used for this application in the old gear described just after WW2. If the tube is to be biased to cutoff, the zener can be supplied with enough current to keep out of its knee region by means of a resistor up to the high voltage or over to the adjacent transistor circuit which should be providing the rf to the power amplifier. The zener should be bypassed for rf.

Zener meter

A zener diode can be wedded to a meter circuit with very useful results. To understand the utility of this match look at the usual linear meter scale. Suppose it reads to 20 volts. A one-volt reading will be way down at one end, and a lower range is required to make it readable. A small change huge percentage increase at the low end equals in scale space a small change tiny percentage increase at the high end. That's not a very equitable distribution! For example, it would be convenient to check transistor emitter-base and emitter-collector voltages without changing ranges.

The required benefit is achieved if the circuit can be made to show variable meter sensitivity. Fig. 14 shows a realizable result, detailed below. The first half of the meter scale is taken up with the zero-to-five volts range. The five-to-twenty volts range occupies the second half of the scale, without switching. The poor sensitivity to voltage applied in the wrong direction is a valuable by-product of scale tailoring with a zener diode.

At first glance this circuit appears to have been designed by a network expert. A closer look reveals that the values of the resistors may be deduced, one at a time, by thinking out the inside requirements of the circuit. R1 and R2 will fall first. If the meter is to read 5 volts at half scale with 5 volts applied, R1 plus R2 must come to 200kΩ since this will pass the required 25 microamps. The Lafayette meter's resistance of 1kΩ is insignificant in comparison to this value. Supposing at 5 volts the zener hasn't quite broken down, it must have just 3 volts across it. The voltage across R1 must be 2 volts, and at 25 microamps the resistance must be 80kΩ. That leaves 120kΩ for R2 since the pair must add up to 200kΩ. We already know the zener; the problem is two-thirds solved.

Now we proceed confidently to the determination of R3. At 20 volts applied the meter reads full scale, therefore is carrying 50 microamps. This current through R2 must bring the junction between R1, R2 and the zener to 6 volts. Now there must be 14 volts across R1 and that yields 175 microamps through it. The

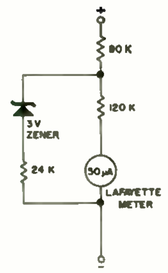

Fig. 15. A circuit that will produce the characteristics shown.

meter gets 50; 125 microamps pass through the zener and its series resistor. We have it! From the junction through the zener we lost three volts, by design; the remaining three volts at 125 microamps fixes R3 at 24kΩ.

A breadboard check shows that the circuit behaves about as shown in the graph. This graph preceded the design, and the actual circuit is influenced by the characteristics of the zener in its knee region. Because the zener comes into conduction gradually as applied voltage is increased, rather than abruptly, the actual scale change from steep to flatter occurs along a rounded curve. A new calibration scale must be constructed empirically. That is, each point must be located by applying the indicated voltage and marking or listing the resulting meter deflection. This good idea needs further development; it requires enough current to disturb many transistor circuits.

The constant-current generator circuit closely resembles the amplified zener. Only a resistor has been added. But the constant current generator guarantees a certain fixed current, rather than the amplified zener's reliable voltage. Its operation depends on the resistor; the zener provides a reference voltage and the transistor, acting as an emitter follower, holds that voltage across the resistor. The resulting current, determined by Ohm's law, is independent of voltages applied to the outside circuit terminals if the transistor is biased into its operating range.

A working circuit is shown in Fig. 16. Remember that the power dissipated by the transistor is determined by its collector current and voltage, not by the values at the rest of the circuit. As in the amplified zener, if the transistor β is large enough the zener current may be ignored. The computation proceeds in this way: the 6 volt zener fixes the voltage across the resistor at 5.8 volts, because 0.2 volts is lost across the base-emitter junction of the germanium transistor. If a silicon transistor were used, the resistor would see 5.3 volts. Since a current of 100 mA is to be guaranteed. the resistor must therefore be 58 ohms. A fixed current of 10 mA would require a 580 ohm resistor since the voltage across it is held constant. But that might not work so well since the zener could be starved for current; perhaps the zener could be biased elsewhere and its voltage carried over to the transistor base.

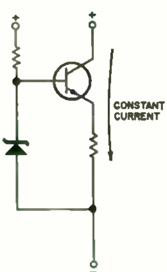

Fig. 16. A zener diode combined with a transistor to make a constant-current regulating circuit.

This is an excellent circuit for eliminating hum. The hum current cannot pass the constant current circuit: no hum! But before this circuit can be put to work in a usable power supply, it must be provided with an appropriate load. The fixed current will generate a large voltage across a large resistor, a smaller voltage across a smaller resistor, and a zero voltage without blowing up anything across a short. This fail-safe feature can be retained while correcting the terrible regulation problem by adding a zener regulator. The constant current is just right for biasing zeners; it is inserted in place of the usual series resistor, and a really good supply results. You should know how to do that by now!

Finally, the last circuit is a realizable substitute for a continuously variable zener. It looks very much like a Darlington pair used as an emitter follower. A current of 10 mA or so from a zener regulator puts a fixed voltage at one end of a pot, decreasing to zero at the ground end. This voltage is stable if very little current is drawn. But the current required by the Darlington pair to regulate at a certain voltage will be the through or zener-like current divided by the β of the first transistor, the result divided again by the β of the second transistor. A milliamp will determine one to ten amps! The illustrated circuit will regulate from about one to 15 volts. The capacitor is required to take out hum coming around through the zener reference voltage source. A supply using the two circuits above shows regulation as good as simple feedback-regulated supplies, combined hum and noise of about 0.6 millivolts, and adjustability over a wide range.

Surplus zeners

Zener diodes are available at prices well under par from several sources. The routes by which these zeners enter the surplus and ham markets are not at all apparent, but it seems

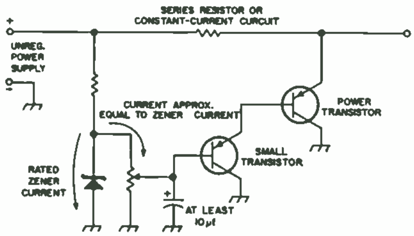

Fig. 17. A very close relative of the amplified zener. Acts like a variable zener.

that in many cases these zeners are rejects having no place in any electronics gear, amateur or otherwise. It also appears that some suppliers—note plural!—do not test their zeners as well and carefully as advertising statements seem to indicate.

Assorted zeners from one supplier were tested for zener voltage and dynamic resistance. Most tended to regulate in the general 20% region, but a few were drastically off. Many of these zeners, 10 watt stud mounting types priced under a dollar each, had fairly high dynamic resistance. Perhaps that is why they were available! A second collection, about 30 assorted zeners adding up to the attractive price of $10 plus shipping, appeared considerably less economical after careful checking and tests. Some were mounted backwards in their cases, many showed poor regulation, a few were phenomenally noisy, others did not zener at all, and one had a broken lead. The more expensive varieties did not seem to average any better than the cheapest ones. There is a moral here. If you are going to use surplus zeners, check regulating voltage, dynamic resistance, and noise characteristics of each zener before you put it in that nice new circuit. Don't take it on faith; the chances that it is not as indicated may be as bad as one in two.

This experience suggests that the most effective way to buy is to purchase new stock zeners, or else test before buying. It may help to initiate a general practice of testing zeners promptly upon receipt, and returning bad ones to the supplier. Be certain the test is correct! Or perhaps you have found a good source of tested surplus zeners; if so, make the most of it and tell your friends. A zener is a zener, and it's the device, not the label, that is required in the circuit.

Testing surplus zeners

A batch of surplus zeners can be tested most effectively if the operation is performed in several steps. The first pass eliminates the obvious duds, the second sorts out the remaining zeners into broad voltage ranges. A third, perhaps, determines if a particular zener can be used in a specific application.

Several instruments are required for complete testing. Also a few resistors and clip leads, a place to work, some scratch paper, and marking paint. A high-sensitivity multimeter or a dc VTVM serves for voltage measurement. Another multimeter or a milliammeter provides for current measurement. An optional ac VTVM is useful for checking dynamic resistance by the hum voltage divider method. A signal tracer will serve very well for detecting the slight hiss a few zeners show in the knee region, or the raucous racket at higher current levels indicating the zener should be discarded. Finally, a magnifying glass assists in detecting mechanical faults on the surface of the package.

The surplus market is low man on the totem pole. It's quite safe to expect a specific zener has something wrong with it, which brought it to the supplier and then you[2]. The testing operation is a sort of detective game played to find out if the fault will or will not interfere with its use in a piece of ham gear. This game can be played most productively if the goals are known. First, does it show semiconductor properties at all? Second, what do they seem to be? Finally, does a closer inspection show they are really there, and that obvious faults are absent?

Modern technology and the manufacturers have conspired to make this game more difficult than it might be. A given zener may be a double-anode device, usable in either direction. Or it may be a zener and a diode, practically the same thing but rather different in intent. And there is a chance it is an amplified zener: a zener and a transistor in one package. The amplified zeners seen so far have been high-power devices, but this may change any time.

A first inspection serves to eliminate broken zeners, ones with bad leads, cracked cases, and other faults. An obviously abused condition is certainly grounds for rejection. At this time the wattage can be estimated by comparison with known zeners and catalog descriptions. A few zeners are shown in **Fig. 2.** Low-wattage accurate zeners may be placed in large cases for better temperature control; high-wattage zeners are indicated by the provision of some means for mounting to a heat sink.

Then the power supply is set up with its negative output terminal to ground. A resistor is placed in series with its positive terminal, chosen to limit the current to near 10 milliamperes. A 40 volt supply would require about 4000 ohms, anything over a half watt would do. The zener goes between the output end of the resistor and ground. Regardless of its condition it cannot receive more than 10 mA; this is a safe arrangement.

The first test is to measure the lowest voltage across the zener at this current, trying both directions. If there is a polarity mark or band, the least voltage should be seen when the band or "cathode" end is toward negative ground. If the voltage is under about 0.6 volts, the device is not a zener and further testing is not required. If it is in this range, and if doubling the current by halving the series resistance from the supply produces only a small increase in voltage, the device is showing proper characteristics for a forward-biased silicon PN junction. If this cannot be achieved, it may be a faulty device, or it may be one of the more complex varieties, mentioned but otherwise carefully avoided in this article.

A breakdown test is now appropriate. The cathode end is turned toward the positive supply, and a measurement of voltage gives the approximate zener regulating voltage. If the resulting voltage is the power supply voltage, then no current is flowing and the device may be a rectifier whose inverse voltage, or a zener whose breakdown voltage, exceeds that available. A current doubling should, again, have very little effect on the stabilized voltage. This test indicates that the device shows zener characteristics.

Knowing the approximate zener voltage and wattage, the supply circuit can now be revised to bias the zener to anticipated normal operating conditions. At this time the dynamic resistance can be estimated by the hum reduction method or by the voltage change over current change method. Typical values for test current and dynamic resistance are available from most catalogs.

If the extra lead is not objectionable, the signal tracer can be left attached to the zener during these tests. With practice, good zeners can be sorted from bad ones almost by ear alone, on the basis of hum and noise. But if this has not been done, a final check for noise should be carried out. Raucous, splattering noise indicates immediate disposal of the zener. A fine-textured hiss at low current levels is permissible, unless it shows a tendency to increase with time or current. Larger zeners should be firmly rapped with an insulating rod to check for loose internal connections.

Zeners that have passed all tests might be marked with fast-drying modeling paint, in resistor color code, as to their values. The paint will also serve to indicate that they have passed a fairly comprehensive test.

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*Of course, many zeners reach the surplus market as manufacturers' over-run, production ends, etc. These zeners, which are available from many suppliers, are generally good, new diodes. *Ed.*

CHAPTER 2

Integrated Circuits

If you are going to keep abreast of the most modern solid state circuitry, it's time to start thinking about using linear integrated circuits. Motorola and RCA have recently made available, off distributor shelves, monolithic integrated circuits suitable for operation up to about 100 MHz. And, the best news is that these latest solid state innovations are well within the pocketbook of the average ham. The introductory single piece price of these high frequency devices is in the $4-5 dollar range. (RCA has just announced a 40% price reduction so the price will be less than listed in the catalog.) RCA is also offering an FM *if* amplifier limiter integrated circuit containing ten transistors and a voltage regulator for only $2.00 and an integrated circuit amplifier-discriminator with twelve transistors and regulator for $2.65.

However, the two circuits that look the most promising for ham use from the price and versatility viewpoint, are the Motorola MC 1550 and the RCA CA 3005. Both of these devices are broadly classified as rf-*if* amplifiers, and either one is useful at frequencies from dc to beyond 100 MHz. They can be used with an external tuned circuit, transformer, or resistive load in applications such as:

A—Mixers
B—Wide and narrow band amplifiers (r-f, i-f and video)
C—Oscillators
D—product detectors
E—low-power modulators,

and probably many more with the application of a little ham ingenuity.

I don't wish to leave the impression that the devices mentioned are the only ones available. Similar devices are made by other manufacturers, for example, Westinghouse and Philco, to mention a couple, and also there are a host of other linear integrated circuits broadly classified as dc amplifiers, audio amplifiers and video amplifiers. Fairchild has just announced the availability of the μA 703 rf amplifier with specification and price in the same range with the Motorola and RCA devices discussed in this article. Moreover, as I will get around to later, some of the low-cost digital integrated circuits can be extremely useful for rf purposes.

Circuit Operation

Now, let's get down to details, first as to what is inside these integrated circuits, how they function, and then, some typical hook-ups.

The MC 1550 and the CA 3005 are quite similar in construction and operation, in that both use a balanced differential amplifier. A simplified schematic that can be used to get familiar with the operation of both integrated circuits are comprised of three devices is shown in Fig. 1. Essentially, these integrated circuits are comprised of three transistors. The current to the emitter-coupled differential transistor pair is supplied from a constant current sink transistor.

The voltage V_2 and resistor R_5 establish the current I_{D1} in diode D1. Since D1 and Q1 are built on a tiny monolithic silicon chip their base-emitter voltage characteristics will be quite similar. Therefore, the emitter current of Q1 will, for all practical purposes, be equal to the diode current. This current established in Q1 will be shared in some manner by Q2 and Q3 depending

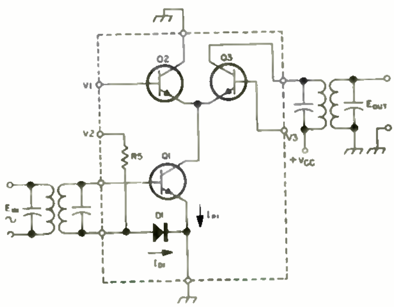
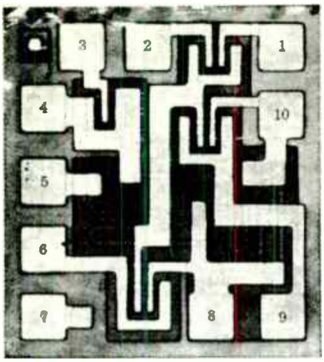


Fig. 1. Simplified schematic diagram of integrated circuit rf-*if* amplifier.



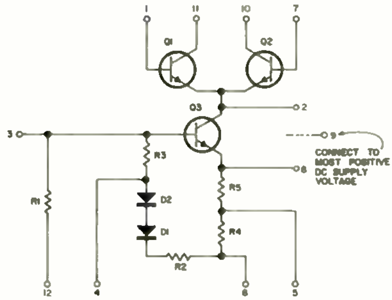


Fig. 3. Schematic diagram of the RCA CA 3005 integrated circuit rf amplifier.

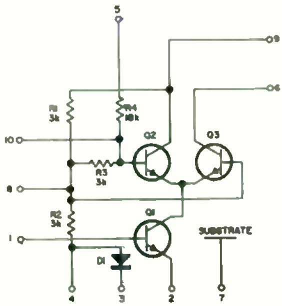


Fig. 2. Schematic diagram and photograph of the Motorola MC 1550. Terminal numbers refer to leads on TO-5 type package.

upon the voltages at V_1 and V_2. If V_1 is at least 114 mV greater than V_2, Q_3 will not conduct and all the current will flow through Q2 and Q1. Under this condition, the gain of the entire module is at a minimum. However, if V_1 is less than V_2 by 114 mV or more, all the current will flow through Q3 providing maximum circuit gain. This characteristic should give a hint as to one of the several possible applications of a voltage applied to the base of Q_2, i.e., it's a very good point to apply AGC voltage.

Now, let's consider signal operating conditions, the incoming signal is applied to the base of Q1 and the output signal is taken from the collector of Q3. Thus, QI and Q3 are functioning as a common-emitter common-base pair which is better known as a cascade configuration. This configuration has a very distinct advantage because it considerably reduces the internal feedback as compared to a single transistor. The fact that the internal feedback is extremely low means that these circuits (the MC 1550 and CA 3005) are very stable, and you won't have to concern yourself with neutralizing.

Another performance advantage is the AGC capability of these integrated circuit devices as compared to a single transistor. The application of an AGC voltage to Q2 has negligible effect on the operation of QI, hence, the input characteristics of QI remain constant. Thus, there is no detuning of the tuned input circuit with changes in the AGC voltage.

Both the RCA and Motorola circuits can be operated as a differential amplifier with minor external modifications in the wiring. Some of the options of using the cascade configuration or the differential configuration will be covered later.

The CA 3005

Now let's take a look at some of the circuits RCA proposes for their versatile CA 3005 integrated circuit.

The CA 3005 can be operated at various levels of supply voltage from 3 to 9 volts and from single or dual dc power sources. Fig. 5 shows the various methods of connecting supply voltages for both the differential and cascade amplifier configuration and for single and dual supplies. Fewer

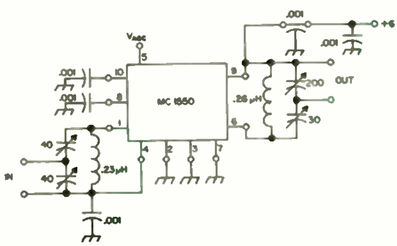
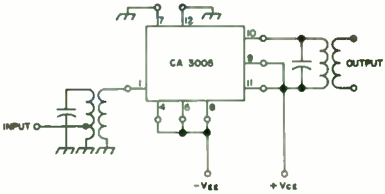
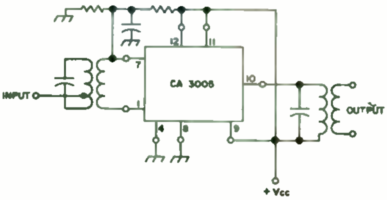


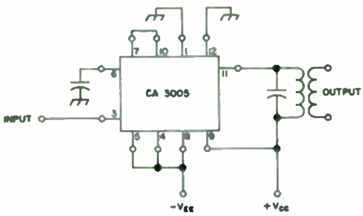
Fig. 4. Typical 50-60 MHz tuned amplifier. Gain is 30 dB with 0 volts AGC and bandwidth is 5 MHz.



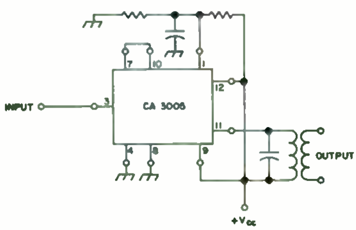
(A) DIFFERENTIAL-AMPLIFIER CONFIGURATION OPERATED FROM A DUAL SUPPLY



(B) DIFFERENTIAL-AMPLIFIER CONFIGURATION OPERATED FROM A SINGLE SUPPLY

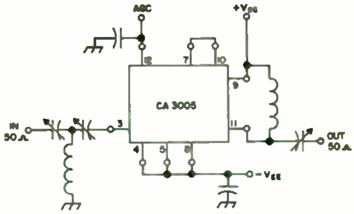


(C) CASCODE-AMPLIFIER CONFIGURATION OPERATED FROM A DUAL SUPPLY



(D) CASCODE-AMPLIFIER CONFIGURATION OPERATED FROM A SINGLE SUPPLY

Fig. 5. Supply connections for the CA 3005 integrated circuit amplifier. A. Differential amplifier configuration operated from a dual supply. B. Differential amplifier configuration operated from a single supply. C. Cascade amplifier configuration operated from a dual supply. D. Cascade amplifier configuration operated from a single supply.



DC SUPPLIES	POWER GAIN (dB)	
V_{EE} & V_{CC}	30 MHz	100 MHz
±6V	36.0	20.0
±4.5V	33.0	18.5
±3V	22.0	15.0

Fig. 6. Typical power gain performance of a cascade configuration at various supply voltages.

external components are needed with the dual supply. To clarify what is meant by a dual supply, take the case of operation from 9 volts. Two nine volts batteries are needed, one for the positive supply Vcc, the other for the negative supply Vee. The other terminal of both batteries is grounded. Notice that when only one supply is used, an external voltage divider and by-pass capacitor is needed for the CA 3005. The MC 1550 has this voltage divider built in the circuit.

The circuits in Fig. 11 illustrate what you can expect from the CA 3005 operating as an *rf* or *if* amplifier at 30 MHz and 100 MHz from several supply voltages. Of course, several of these integrated amplifiers can be cascaded to provide additional gain. However, the CA 3004 which has emitter resistors that provide increased signal handling capabilities is recommended in place of the CA 3005 when several stages of *if* amplification are needed.

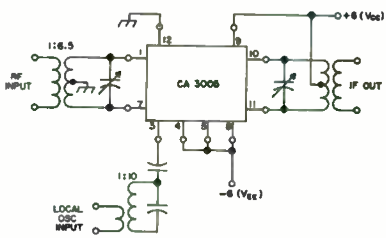


Fig. 7. Typical circuit diagram using the CA 3005 as a mixer.

Mixer applications

The CA 3005 integrated circuit can also be used as a mixer converter, low power modulator and as a product detector. A typical example of a mixer application is shown in Fig. 7. The local oscillator signal is applied to the base of Q3 and the rf signal is applied either single-ended or double ended to the bases of transistors Q1 and Q2. A mixer-oscillator combination, which could be considered as a complete front-end on a single chip, is shown in Fig. 8.

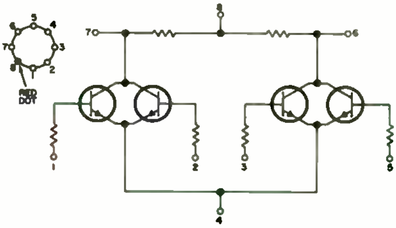


Fig. 9. Schematic diagram of the Fairchild μL 914 dual gate logic integrated circuit.

Using digital circuits in rf applications

The Fairchild μL 914, which is a dual gate logic circuit, is of particular interest for *rf-if* circuitry. This epoxy device costs only 80 cents and prices are still going down. For this meager sum, you receive the equivalent of four transistor and six resistors. However, for the *rf-if* applications, you can only use two of the transistors. The complete schematic for the μL 914 is shown in Fig. 9. The 914 can be used at frequencies up to about 20 MHz. At 10 MHz the gain is about 30 dB falling off at frequencies above 10 MHz. Note in the typical circuit shown in Fig. 10 that AGC can be applied giving excellent gain control. Another point to note, also, is that the 914 can handle inputs of about 150 mV or less. Greater signal voltages will cause limiting, hence, the 914 also makes an excellent FM-*if* limiter.

With this broad introduction to the Motorola MC 1550, the RCA CA 3005 and the Fairchild μL 914, integrated circuits and the typical application examples, I am sure that many of you will soon be plugging them into sockets. And, before long, there will be construction articles for converters, receivers and even QRP transmitters.

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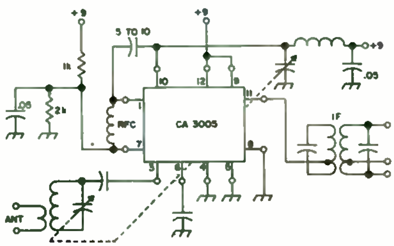

Fig. 8. CA 3005 can function as a complete front end. Part of the circuit acts as a mixer and the other parts as a local oscillator.

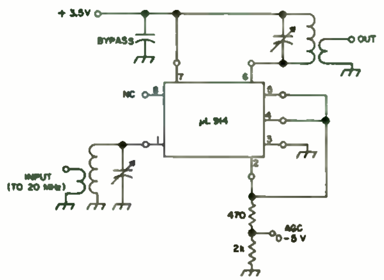

Fig. 10. Fairchild μL 914 shown in a typical rf-if circuit. Input signals up to 20 MHz and levels up to 150 mV can be handled by this device.

CHAPTER 3

Silicon Transistors Used as Zeners or Varicaps

Need a low cost zener diode? Or how about a low cost voltage variable capacitor (varicap)? How many times have you had the desire to regulate that mobile converter, but didn't want (or couldn't) spend the 2 or 3 dollars additional for a zener diode? Or how about that tunable converter you wanted to remote tune by voltage, but, oh, the price of the varicaps? Well, stop! Remember that silicon transistor with the broken lead, or the one with the open collector, or was it the emitter lead that was open? Anyway, if you threw it away, you could have thrown away that low cost zener or varicap you needed.

First let's see what you could have done for a zener. By reverse-biasing the emitter-base junction of a silicon transistor, you have a very handy zener diode. Some silicon transistors even exhibit better zener diode characteristics than some of the diodes sold specifically as zeners.

Of course, the first obvious test is to determine what the zener voltage is for your specific transistor. Generally, most (there are always exceptions, of course) silicon NPN transistors will exhibit a zener action somewhere between 6 and 11 volts. (Pretty ideal for mobile regulators). Fig. 1 shows the hook-up for determining the zener voltage. A value of 470 ohms for R_s is sufficient to limit the current to a safe value while determining the zener voltage. Connect a variable power supply as shown and connect a VTVM from emitter to base. Notice that the collector is not connected and is not needed in this application. Now, slowly increase the input voltage while monitoring the voltage output on the VTVM. At a specific voltage input, the voltage out will stop increasing. Any further increase in the input voltage beyond this point will now cause only a very slight increase in output voltage. The voltage as read on the VTVM is your zener diode voltage. The next question asked is, "OK, but what range of current can I regulate?" A rule of thumb here is; divide the voltage obtained as the zener voltage into the free air dissipation rating of the transistor. For example, if the transistor zenered at 10 volts and is a 300 mW device, the 10 volts into 300 mW gives 30 mA. This would be a safe operating limit. However, tests indicate that this isn't necessarily the maximum limit, but it is unlikely that you would regulate a circuit drawing more.

OK, now you know what the zener voltage is and have an idea as to the amount of current you can regulate. Let's apply this to a more specific example. Suppose you want to regulate that mobile converter's oscillator stage. Let's say your transistor zeners at 10 volts. With 10 volts on your oscillator stage, it draws 3.75 mA. Since it's mobile, you will vary between the 12 volts from the battery to approximately 14.7 volts at maximum generator output. Fig. 2 shows the hook-up. The only requirement is to determine the value of R_s so that the transistor (oops—I mean zener) will regulate properly with a variable input. To determine R_s, the following formula is used:

$$R_s = \frac{V_L - V_Z}{I_L + 0.1\ I_L}$$

where: V_L = lowest voltage input
V_Z = zener voltage
I_L = load current

Therefore, in the example, $V_L = 12$ volts, $V_Z = 10$ volts, and $I_L = 3.75$ mA. Using these values and solving for R_s gives a value of 485 ohms for the series resistor. A 470 ohm resistor would do quite nicely. As a check, let's determine what the maximum current will be through the zener. This will occur at

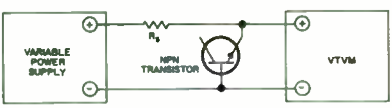

Fig. 1. Checking the zener voltage of diode junctions in NPN transistors. R_s can be about 470 ohms for these low voltage diodes. It limits current flow to keep from damaging the junction in the transistor.

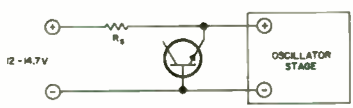

Fig. 2. Using a transistor as a zener diode to stabilize a transistor oscillator stage. Selection of R_s is discussed in the text.

the 14.7 volt input. At this level, R_s would have to drop 4.7 volts. This represents a total current through the 470 ohm resistor of 10 mA. The oscillator stage draws 3.75 mA. Therefore, only 6.25 mA is flowing through the zener, which is well within the dissipation rating of the device. In like respect, at the low voltage input, R_s drops only 2 volts which represents a total current of 4.25 mA. The zener draws only .5 mA in this case.

If your stage, which you desire to regulate, has a variable current requirement as well as a variable voltage input, use the following formula to determine the series resistor, R_s:

$$R_s = \frac{V_L - V_Z}{I_{Lmax} + 0.1\ I_{Lmax}}$$

where: I_{Lmax} = maximum load current

In like respect, if the input voltage is constant and the load current variable, use the same formula just given, with input voltage used in place of V_L.

These formulas are based on the premise that for conservative designs, the empirical factor of 10% of the maximum load current should be used for minimum zener current. In other words, the zener then is capable of regulating from this minimum current up to the value of maximum zener current as governed by the dissipation rating of the transistor. If your change in input is small, a figure of 20% may be used to better advantage for a little better regulating action.

Table 1 shows typical zener voltages meas sured on various transistors. All but the 2N709 are available for under $1.00 as compared to zener diode prices ranging from several dollars and up. Notice that some (the exceptions) like the 2N94 have a very high zener which limits the usable current range. (Possibly low current B+ regulators?)

Pay particular attention to the Fairchild 2N3567. This device sells in the neighborhood of 60 cents and exhibits extremely good zener action. It also exhibits another characteristic to be covered next.

Now let's forget that emitter lead and use instead the collector lead in conjunction with the base lead. Fig. 3A shows a typical varicap tuned tank circuit and 3B shows a NPN transistor connected for use as a varicap. C_1 in the figure isolates the varicap from DC. For this application, device dissipation is of little concern as only leakage current is flowing and will be quite insignificant. As with any series connection of capacitors, some consideration must be given to C_1. If this capacitor is quite large, compared to the varicap, then the tuning range of the tank circuit will be in direct proportion to the maximum change of the varicap's capacitance with applied voltage. If it is smaller than the varicap, then the change in frequency with the change in the varicap's capacitance with voltage will be small.

There are two means by which you can determine whether or not your silicon transistor will be suitable as a varicap. Some will have only a minor change in capacitance with voltage changes and others will have a greater change. The first method and by far the simplest is to refer to the data sheet for the transistor in question. If you're lucky, this capacitance change with voltage will be graphically plotted. The graph to look for is the output capacitance versus reverse bias voltage. This is listed on the data sheets as C_{ob} ($I_E = 0$).

The second method is to actually connect the transistor in question to a tank circuit such as that shown in Fig. 3B. Use a large by-pass such as a .001 for C_1 and a 100 K resistor for R_1. Place a suitable coil of known inductance across the transistor and capacitor as shown. Apply a DC voltage of 0.5 to 1 volt and then using a grid dip, find the resonant frequency of the tank circuit. Increase the supply voltage until further increases have little effect on the resonant frequency. Compute the value of the transistor's capacitance at the low voltage level and at the high voltage level. This gives you the range of capacity versus the voltage change required to produce this capacitance for the transistor in question.

Depending upon the application you have in mind, it's advisable to actually plot the

Transistor	Typical Zener Voltage
SE2001	6.1
2N706	6.2
2N3642	6.4
2N709	7.5
SE4002	9.8
SE6001	10.2
2N3567	10.2
2N94	50.0
2N233A	55.0
2N212	50.0

TABLE 1

capacity versus the voltage. As with a true varicap, maximum capacitance occurs at a low voltage and minimum capacitance at the higher voltages. The rate of capacitance change is greater at the lower voltage changes and vice versa. For tuning applications, you would be concerned primarily with capacitance change obtainable for the range of voltage you have available. However, for an ap-

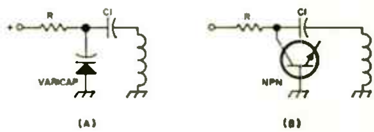

Fig. 3. Use of a varicap or transistor as a voltage variable capacitor in a tank circuit.

plication such as producing FM, it is advisable to choose a section of the curve where linearity is achieved and bias the varicap to this value of voltage through a suitable divider. Then by applying an audio voltage to the varicap, a linear swing plus and minus may be achieved. Be sure that your bias point is sufficiently high so that the level of applied audio voltage doesn't overcome the bias on the varicap causing it to conduct.

Remember the Fairchild 2N3567 mentioned before for use as a zener? Well, here it is again. This little device exhibits excellent varicap characteristics. The capacitance and the capacitance change is ideal for a wide range of applications such as tuning, AFC, FM, etc. **Fig. 4** shows the typical C_{ob} of this device. As an example of the range possible with this particular device, consider a control voltage from 0.5 volts to 10 volts. This represents a capacitance of approximately 28 pF and 13 pF respectively. This is a capactance ratio of 2.15 to 1 which represents a possible frequency ratio of 1.46 to 1. In other words, you could tune from 40 MHz to 58.4 MHz or by proper choice of the coupling capacitor you could easily cover 50 to 54 MHz.

Incidentally, this 2N3567 can also be used, of all things, like a transistor. It is designed primarily for amplifier and switching applications. It exhibits a 40 volt collector to emitter voltage and 300 mW dissipation. Purchasing three of these devices at approximately $1.80 would enable you to build a fairly low cost zener regulated, voltage tuned, oscillator or RF stage.

As mentioned previously, some transistors only change several pF with voltage change, but don't overlook the possibility for FM transmitters where you only require 25 kHz or less swing where a small change in capacitance would be sufficient.

The applications given in this article are all for NPN silicon transistors since the most common silicons are NPN. There are a number of inexpensive PNP silicons now available and they can be used if you reverse the voltage shown.

So next time you start to pitch that NPN transistor with the open, or broken lead, stop and consider the other possible applications for it as a zener or varicap.

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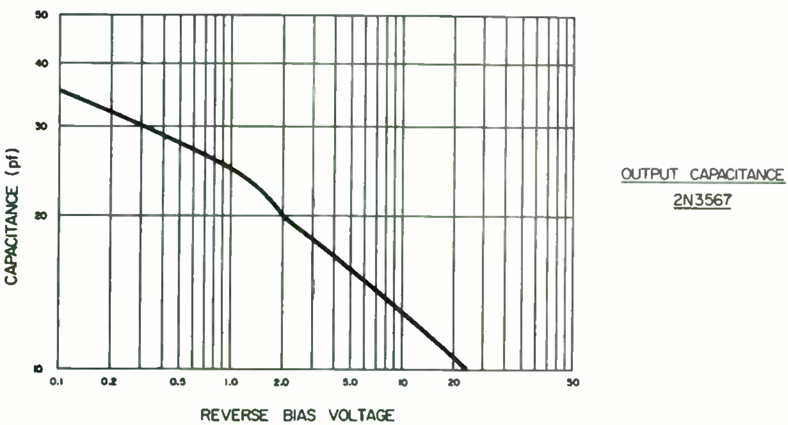


Fig. 4. Capacitance versus reverse bias voltage for the base collector junction of the Fairchild 2N3567 transistor.

CHAPTER 4

Basic Transistor Circuit Design

Foreword

Over the past few years, emphasis has shifted more and more toward the application of solid-state devices in amateur equipment. Most new circuits in the ham magazines are transistorized, and many valuable tube collections and heavy power supplies gather dust while their owners scrounge transistors, diodes and small batteries. Such is progress.

For the serious builder the time usually comes when no existing designs quite satisfy what is required and he begins to think about a design of his own. This is the point where many worthwhile projects come to an abrupt halt; this is where many hams decide that transistors are just too complicated and that all solid-state designers are wizards. The procedures I propose here are satisfactory for all but the most rigorous design requirements and may be used in the design of professional as well as amateur equipment.

Information required for design

The characteristic curves of the device (transistor or diode) should be at hand before any proper design can be attempted. While some information may be available from transistor manuals or transistor testers, only manufacturer's curves describe how the device will work under any given conditions. Manufacturers will usually supply curves upon request.

For practical thinking, an intimate knowledge of semiconductor physics serves little purpose. In terms of design, transistors are as simple as tubes. A vacuum tube is a valve, the current flowing from cathode to anode being controlled by varying the relative amplitude of the control-grid to cathode voltage. Simple. A transistor is a valve, the current flowing from emitter to collector being controlled by how much current flows into the base. Field effect transistors should be considered solid-state triodes for our simple approach—just as simple and just as adequate for design thinking.

The operation of NPN and PNP transistors is identical—only the polarities of the voltages applied to the transistor (bias) are different. Transistors are biased with the following rules in mind (see Fig. 1):

Moving the base level closer to the B minus level (by decreasing R_B) causes more base current to flow, and therefore more collector current to flow. Note that with a PNP transistor the base must be negative with respect to the emitter, and the collector must be *more* negative with respect to the emitter than the base.

The valve principle may be clearly illustrated by examining the collector curves for a typical transistor in Fig. 2. These curves could describe the operation of *either* a PNP or NPN transistor.

A constant collector-emitter voltage (V_{CE}), is chosen by drawing a vertical line through any desired V_{CE} on the scale (7 volts). Whenever this vertical line intersects with a base current curve, a horizontal line is drawn from that point to the collector-current scale.

Point X3 is the point where, with a V_{CE} of 7 volts, a base current flow of 0.3 mA causes a collector current of 10 mA. Increasing the base current to 0.4 mA moves our operating point to X1, and causes the collector current to increase to 20 mA. Note that these curves describe the transistor operation for a given set of conditions.

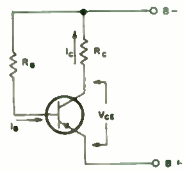


Fig. 1. Simple biasing circuit for a PNP transistor. I_C is the collector current, I_B is the base current, and V_{CE} is the voltage between the collector and emitter. Since I_B is much smaller than I_C, the base-biasing resistor R_B is much larger than R_C, the collector resistor. This circuit may be used with NPN transistors by simply changing the polarity of the supply voltages.

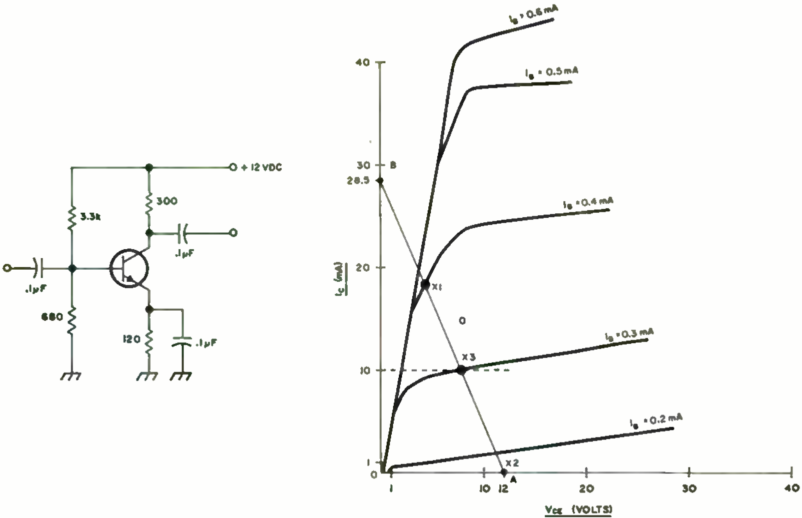


Fig. 2. A PNP transistor amplifier and its dc (static) load line. Point X3 is the quiescent point determined by the emitter-collector voltage of 7 volts. Saturation occurs at X1, cutoff at X2.

Design procedures

Power output

For a safe design, use a transistor which is rated at ten times the required power output (at room temperature of 25°C). This seems extreme, but transistors are derated quite sharply (dissipation-wise) as temperature rises.

Frequency

The upper frequency limit of a transistor is usually specified as f_{hfe} or f_{α}. If the frequency rating given is "f_{hfb}", divide the f_{hfb} figure by the h_{fe} rating for the transistor to obtain f_{hfe}. This is the point where the stage gain will be down by 3 dB (half-power point). For reliable design use a transistor whose minus 3 dB frequency is ten times the maximum frequency of operation.

Supply voltage

Most amateurs feel more at home with a positive supply voltage so let us use NPN designs. The available supply will determine the transistor. The transistor rating to consider here is BV_{CEO}, the collector-emitter breakdown voltage. If the amplifier has a resonant circuit in the collector circuit, the BV_{CEO} rating should be four times the supply voltage (V_{CC})—indeed, if the stage is being modulated, the BV_{CEO} *must* be at least 4 V_{CC} or breakdown of the collector-emitter path may occur. These large safety factors may seem extreme, but are desirable for trouble-free designs.

Our design examples will use the *2N7388*, a fictitious silicon 100 MHZ NPN transistor, rated at 0.5 watt at 25°C; BV_{CEO} = 50 volts and I_{CMAX} = 1 Amp.

Setting the operating point:

The first step in the design—once the circuit has been selected—is to set the dc operating point. This point of operation affects the gain of the stage, and determines the power drawn from the supply under no-signal conditions. The usual requirement is to obtain the most gain. In portable gear, gain may have to be sacrificed to keep power drain low, or higher-gain transistors may have to be obtained.

The choice of operating point can be made by examining the plot of h_{FE} (forward current gain) against I_C (collector current). The h_{FE}/I_C curve for the 2N7388 might look like Fig. 3. The procedure we will follow in

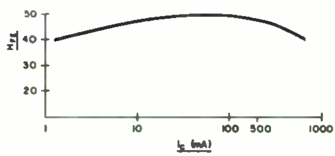


Fig. 3. Plot of the forward current gain, H_{FE}, as a function of the collector current. The maximum H_{FE} of 50 coincides with I_C = 100 mA, but moving up or down from this point does not alter it appreciably. To maintain linearity in amplifiers, the quiescent point is chosen on a reasonably flat portion of this curve.

choosing our operating point is as follows:

1. Examine the peak of the curve (maximum h_{FE} point) and see what I_C it requires.
2. If this current drain seems extreme, move to the left of this peak until a compromise between gain and current drain is obtained.
3. Try to work over a reasonably flat portion of the curve so that changes in I_C around the chosen operating point do not affect the stage gain very much.

Choosing our operating point from the curves of Fig. 3, we see that a maximum h_{FE} of 50 occurs at $I_C = 100$ mA. If we choose $I_C = 10$ mA, h_{FE} is still about 45. Moving I_C either way from the 10 mA point does not change h_{FE} appreciably. We will therefore set our dc operating point with no-signal input (quiescent or "Q" point) at $I_C = 10$ mA and $h_{FE} = 45$.

Once the operating point is set, we can quickly establish the rest of the dc conditions in our circuit by examining the plot of I_C versus V_{CE} (Fig. 4).

While looking at these curves, it would be a good idea to examine the circuit we ultimately hope to design, and how Messrs. Ohm and Murphy can combine to confuse our thinking.

The simple, basic amplifier circuit shown in Fig. 5 will work. The "practical" circuit shown is the same basic amplifier, but R_E has been added and R_B is now the parallel combination of R_1 and R_2. The ac operation of the two circuits is identical, but the practical circuit is stable and reliable with changes in temperature. Let's look at the dc operation of this amplifier, with its various currents and voltage drops. To simplify things we shall assume that I_E (emitter current) is equal to I_C (collector current). We can do this without introducing significant error.

The supply voltage itself will depend on the value of load resistance required. The supply voltage must be high enough so that the I_CR_L voltage drop does not approach the value of supply voltage at the quiescent operation point. This would cause the transistor to "run out of V_{CE}"—with resultant distortion. Try to arrange the I_CR_L voltage drop so that it is not much greater than one-quarter the supply voltage at the operating point.

The currents and voltage drops are shown in Fig. 6. The base-emitter voltage (V_{BE}) of a conducting silicon transistor is approximately 0.7 volts, and we will assume that it is so. Ohm's law tells us that I_CR_E (drop across emitter resistor) plus I_CR_L (drop across load resistance) plus V_{CE} (collector-emitter voltage) must add up to our supply voltage. While the collector curves will only show the placement of the Q point with respect to V_{CE}, we must consider the various voltage drops when choosing the V_{CE} with respect to the supply voltage. For example, with a 6 volt supply, if we choose a V_{CE} of 5.5 volts, only 0.5 volts will appear across the

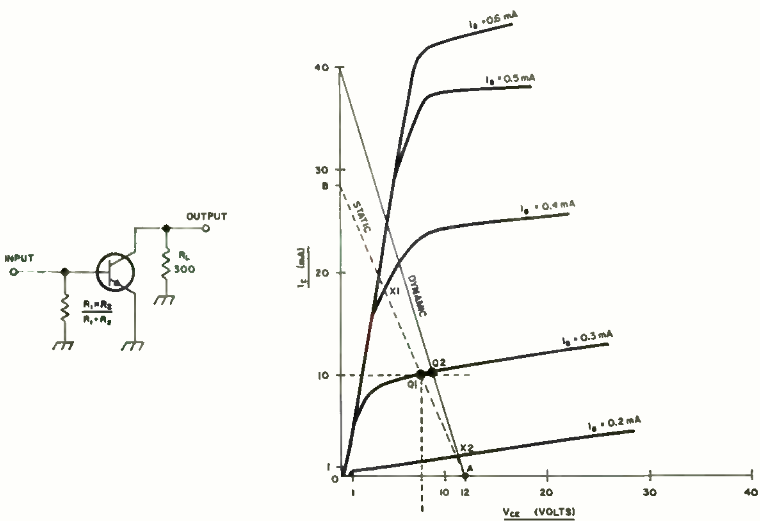


Fig. 4. Comparison of the static and dynamic load lines of a transistor amplifier. Note that going from dc to ac shifts the position of the load line. The circuit of the amplifier is shown on the left. The base-bias resistor is shown as a parallel combination of the two resistors that are actually used because this is the resistance 'seen' by the signal.

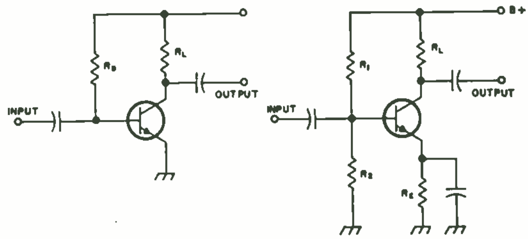


Fig. 5. The simple transistor circuit shown on the left will work, but the more practical circuit on the right is more stable and reliable with changes in temperature.

load resistance and R_E; with a 3 k load (a practical value), the I_C would have to be less than 200 μa. Keeping these various pitfalls in mind, let's look again at the collector curves (Fig. 4).

We want to place our Q point on the 10 mA I_C level (shown dotted horizontally). Let's assume a 12 volt power supply and a 300 Ω load. What value of V_{CE} do we choose to fix our Q point?

A general rule to follow is to have 1/10 the supply voltage across R_E; with a 12 volt supply this leaves 10.8 volts. With 10 mA I_C, the drop across R_L is 3 volts (10 mA x 300 Ω). Our V_{CE} then is 12 V − (3 V + 1.2 V.) = 7.8 volts (shown dotted vertically). The point of operation of our transistor will be the intersection of the 10 mA I_C level and the 7.8 volt V_{CE} position. Marking this intersection as Q_1, it is seen that under these conditions our base current will be 0.3 mA. The power drawn by the transistor at the Q point (with no signal) is 7.8 V x 10 mA = .078 watts, or 78 mW. Since the 2N7388 transistor is rated at 0.5 watts, this is an adequate margin of safety.

As a check on the base current (I_B) we can use the approximation $I_B = I_C/h_{FE}$. In our case, this works out to be 0.22 mA (10 mA/45). Considering the fictitious nature of our curves, this puts us in the right ballpark. For reference at the operating Q point:

I_C = 10 mA
V_{CE} = 7.8 volts
P_{DISS} = V_{CE} I_C = 0.78 watts (no signal)
I_B = 0.3 mA

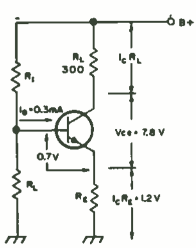


Fig. 6. The element currents and voltage drops of the transistor amplifier.

DC circuit values

Having defined the Q point, we can easily specify values for R_L, R_E, R_1 and R_2. For clarity, let's show the known conditions on the dc circuit (Fig. 6). We specified R_L as 300 Ω. If R_L cannot be specified, use the largest value that the transistor and available supply will allow. This results in minimum loading by the following circuitry.

For stable operation with temperature changes, it is desirable to have 1/10 of V_{CC} across R_E. By Ohm's Law, R_E = 1.2 V/10 mA = 120 ohms.

We can calculate the values of R_1 and R_2 by following a few simple rules.

A. To prevent changes in the transistor from affecting the circuit, the current through the series string of R_1 and R_2 should be at least ten times the desired base current. The current through R_1 is greater than that through R_2 by the amount of the desired base current. By Ohm's Law: $R_1 + R_2$ = 12 volts/3 mA = 4000 Ω.

B. The voltage (with respect to ground) at the base is approximately the sum of the R_E voltage drop plus the voltage across the base-emitter junction (0.7 volts).

V_B = (1.2V + 0.7V) = 1.9 volts.

C. V_B appears across R_2, which by Ohm's Law is 1.9 volts/3 mA = 634 Ω.

D. R_1 = 4000 − 634 Ω = 3366 Ω.

E. Checking R_1 by Ohm's Law, R_1 = (12 − 1.9 volts)/3 mA= 3360 Ω. This is within 1% of our previous calculation.

The nearest standard resistor values will be quite satisfactory (680 and 3300 ohms respectively). Discrepancies between calculated currents and operating currents are caused by reading curves inaccurately, slide-rule errors and variations in transistor characteristics.

AC circuit values

The capacitors used for input, output and by-passing should have a low reactance at the operating frequency. Values of 0.1 μF to 0.01 μF are usually suitable.

Once we have the amplifier designed, it is handy to know what it will and won't do.

Back to the collector curves we go again—this time to examine the "load lines" of the amplifier (Fig. 2 and 4).

To draw the load line, the two extreme operating conditions of the amplifier must be considered.

1. No collector current at all. With no voltage drops across R_E or R_L, V_{CE} is equal to the supply voltage (point A on the curves).
2. So much collector current that the entire supply voltage appears as voltage drops across R_E and R_L, and V_{CE} is zero. This collector current is calculated by Ohm's Law:

$$I_C = \frac{V_{CC}}{R_L + R_E} = \frac{12 \text{ volts}}{420\ \Omega} = 28.5 \text{ mA}$$

This point is at B on the curves.

Once points A, B and Q are located on the curves, a straight line is drawn joining the three points. This is called a load line; since it was derived from the dc operating conditions, it is the dc or "static" load line.

The static load line describes the dc operation of the amplifier. For example, if we drive the amplifier with more than 0.4 mA base current, the transistor operates above point X1 on the load line, and goes into saturation. If we put in less than 0.2 mA base current, the transistor goes into a "cut-off" state. Operation must be between X1 and X2. Note that if this is the case, equal changes of I_B either side of 0.3 mA produce equal changes in collector current. This is the requirement for linear, distortion-free operation.

The ac or "dynamic" load line can be drawn just as easily. For ac, the bypassed R_E does not exist; the maximum collector current (ac) is now V_{CC}/R_L or 40 mA.

Note that the Q point has shifted slightly to the right along the $I_B = 0.3$ mA line, from Q1 to Q2. This slight shift in operating point can, for practical purposes, be ignored.

We now have our load lines. So what? We can obtain quite a bit of information from

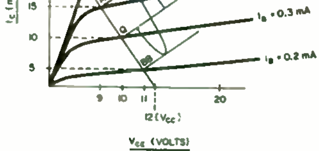


Fig. 7. Expanded view of the operating area with imposition of the base-input signal. Past the limits set by 'AA' and 'BB', the change in base current ceases to be linear and distortion will result if the signal is driven beyond these points.

them. We are interested in the ac operation of our amplifier, so let's use the dynamic load line. An expanded view of the operating area is shown in **Fig. 7**.

First, let's check the power dissipation. From the load line we see the ac operating point sits at $I_C = 10$ mA, $V_{CE} - 10$ volts. This means a Q point dissipation of 100 mW, well within the rating of the transistor.

Next, let's mark the operating limits on our load line. These limits must be so set that equal changes in base current around the Q point (up and down the load line) produce equal changes in collector current. Past the limits shown as AA and BB in **Fig. 7** this condition is not met and distortion will occur. **Fig. 7** shows a sine-wave input centered about the Q point, producing the following results:

Collector current swing of 10 mA (5 mA to 15 mA)

Base current swing of 0.2 mA (0.2 mA to 0.4 mA)

Current gain is approximately 50 (10 mA/0.2 mA)

Output voltage of 2 volts peak-to-peak (9 volts to 11 volts)

Maximum power developed in the load under these conditions is $(0.707\ I_C \text{ peak})^2 \times R_L$. Substituting the figures from our curves:

$P_{out} = (3.525 \text{ mA})^2 \times 300\ \Omega = 3.73$ mW.

Fig. 4 shows that in the ac circuit the bias network equivalent resistance is in parallel with the input to the transistor. The transistor input impedance is approximately equal to 26 h_{FE}/I_C (mA). The 4000 ohm equivalent resistance of the base bias network does not alter this impedance appreciably.

The *input impedance* of the circuit is about 117 ohms and the *driving power* required is approximately equal to the product of the square of the input (base) current and the input impedance

$P_{in} = (I_B)^2 \times Z_{in} = (0.3\text{mA})^2 \times 117\Omega = 10\mu$W

The *power gain* is P_{out}/P_{in} or $3.73/.01 = 373$

Note that these results only apply if a 300 ohm ac load is presented to the transistor. If this circuit drives other circuits, a new ac load line must be drawn, and new results calculated.

With our two load lines as an example, it might be a good idea to discuss the effects of various load impedances presented to the amplifier. Fig. 2 shows the amplifier with no external circuitry attached: we saw that the dc load was 420 Ω ($R_E + R_L$) and the ac load was 300 Ω (R_L). What happens to the amplifer when it us used to drive another circuit?

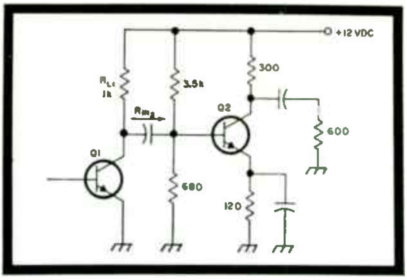


Fig. 8. To construct the ac load line of an amplifier, both the input and output circuits must be included. For example, to draw the dynamic load line of amplifier Q2, the effect of the driving transistor, Q1, and the 600 ohm load must be considered as illustrated in the text.

To consider the ac loads on an amplifier, we must consider all coupling and bypass capacitors as short circuits and direct short circuits across all dc power supplies. To illustrate this, examine **Fig. 8.**

Fig. 8 shows the amplifier driving a 600 ohm load, and being driven by another transistor, Q1.

The load on Q1 is $(R_{L1} R_{in2})/(R_{L1} + R_{in2}) = (1000 + 117)/1117 = 100$ ohms. A 100 ohm ac load line would be required for Q1. The ac load on Q2 is (300 x 600)/900 or 200 ohms.

The ac load line must now be re-drawn, using 200 ohms as the ac load. The 300 ohm ac load has been shunted, and that load line no longer applies. The dc load line of course still holds.

Let's consider an alternative–Fig. 9 shows the amplifier coupled to the load through a transformer. Ignoring the dc resistance of the transformer winding, the dc load line remains as before. The ac load has changed drastically.

Remembering that transformers can transform voltage, current *and* impedance,

$$\text{Primary impedance} = \left(\frac{\text{Pri turns}}{\text{Sec turns}}\right)^2 \text{x Secondary impedance} = \left(\frac{10}{5}\right)^2 \text{x 600 ohms}$$

$= 2400$ ohms

Thus the 600 ohm load is presented to the transistor as a 2400 ohm load. In addition, the 300 ohm load resistor is in series with this, making a total ac load of 2700 ohms. By bypassing the top of the primary winding with an 0.1 µF capacitor, we can eliminate R_2 from the ac load line.

The point which this load impedance discussion should make is that the ac load impedance presented to the transistor must be decided by the "most likely to succeed" ac load line. The load impedance this load line represents is the *total* ac load, and must include all non-bypassed bias circuits, transformer-coupled loads, and what have you. The dc circuits in the collector, and any collector transformer turns ratios must be adjusted so that the transistor sees this load impedance. In other words, the load line is first established, and *then* the output circuit is designed to present the proper load impedance.

With the information presented here it should be possible for the average ham to design simple transistor circuits at low frequencies. A treatment of high-frequency design should be covered as a separate topic, but the basic ideas and biasing methods would certainly apply. ■

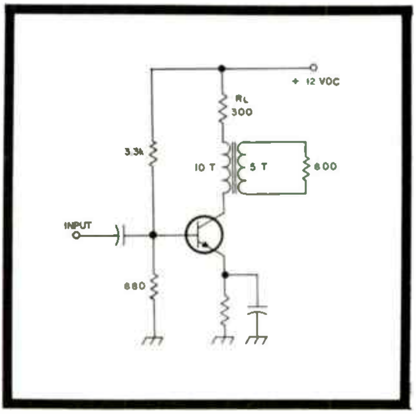


Fig. 9. Transistor amplifier with a transformer-coupled load. When constructing the dynamic load line for this amplifier, the impedance transformation ratio of the transformer must be included in the calculations.

CHAPTER 5

Logic ICs for Amateur Use

The title is not meant to imply that amateur operators are *illogical* in any way, but rather to identify this article as one in which integrated circuit (IC) logic elements are used in ways for which they weren't specifically designed. That is, it will attempt to show some of the many ways that amateurs can use digital IC's in circuits that are non-computer oriented.

Historically, there are two reasons why the digital IC (micro-logic) became readily available at low cost before the linear IC. One of these reasons was the rapid growth of the digital computer industry; increasing both individual computer size and the number of computers in production. Size, cost, and reliability requirements of the new digital computers offered a rich prize to the semiconductor industry if it could come up with an IC to suit computers. The second reason digital IC's came first is the fact that logic circuits are easier to make than linear circuits. Logic circuits generally require only that their transistors be in one state or another (for instance, "on" or "off") and this requirement is relatively easily met by mass production units.

Out of all the research that was poured into the realization of the digital IC for the computer industry, several "logic families" of IC's emerged. These logic families have all made successes in computer use to one extent or another and no one family has yet obtained a clearcut advantage over the others on all counts. The present major logic families are: Resistor Transistor Logic (RTL), Diode Transistor Logic (DTL) Transistor-Transistor Logic (TTL), and Emitter Coupled Logic (ECL).

RTL Integrated Circuits have become the least expensive, most available IC's on the market. In small quantities (1-99) the price of a simple J-K flip-flop has dropped to $1.35 and that of a dual two-input gate to $.80. A number of semi-conductor manufacturers, Motorola, Fairchild, Sperry, Texas Instruments, and others all make the RTL line; and at least between *some* units, voltages are compatible. There are two mainly-used packages, the TO-5 can with 8 or 10 pins and the "Dual-Inline Package" (DIP) with 14 pins.

The basic building block of the RTL family is the gate shown in Fig. 1. This gate can be expanded into two, three, and four-input types as shown in Fig. 2. In the gates shown in Fig. 1 and 2, a + 1 volt input to *any* input will saturate a transistor and pull the output *down* from the +3.5V supply level to saturation.

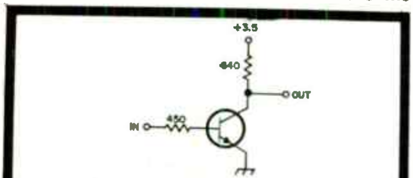


Fig. 1. The basic transistor gate.

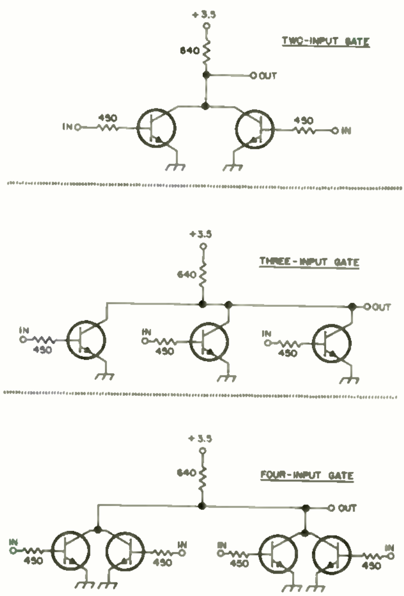


Fig. 2. Various gate arrangements—two input, three input and four input.

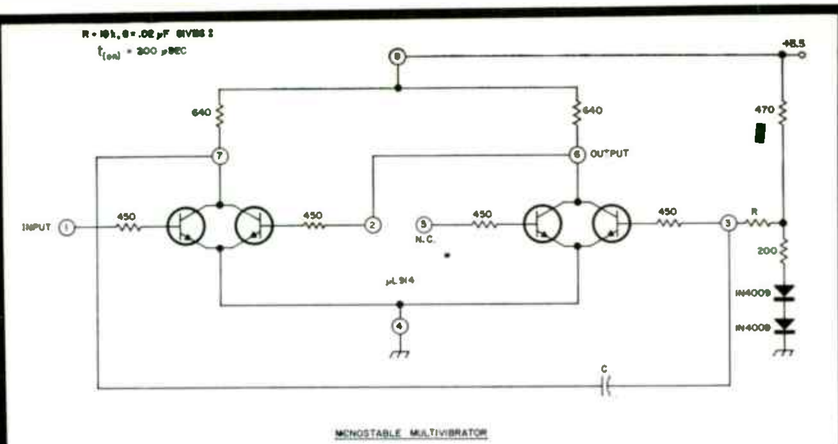


Fig. 4. A monostable multivibrator using the µL914.

One of the least expensive and most versatile RTL gates that is in general use is the Fairchild µL914. This is shown in Fig. 3, It can be used (in addition to its normal use as a gate) as a monostable, bistable, or astable multivibrator. The bistable multivibrator connection of this chip can be purchased as the µL902, a type "RS" flip-flop. The connection of the µL914 in various types of multivibrator circuits is shown in Fig. 4, 5 and 6.

Although the µL914 is easily used as a type "RS" flip-flop, it is simpler to use the µL923 type J-K flip-flop for most purposes. The real advantage of using the J-K is the simplicity one attains in dividing by different numbers. Even fairly large prime numbers may be divided using J-K's, and no critical feedback capacitors are required. Since the J-K has many ports, there a number of ways to divide most numbers. Some examples of dividers using J-K's are shown in Fig. 7.

There are some points of care which must be observed when using these RTL IC's. The individual J-K will draw between 20 and 25 mA of current at 3.6 volts, so use good "fat" supply leads. This care in buss (and ground) lines is essential, because the RTL family has the lowest noise immunity (for spikes on the supply line) of any of the logic families. Also, when in doubt, it never

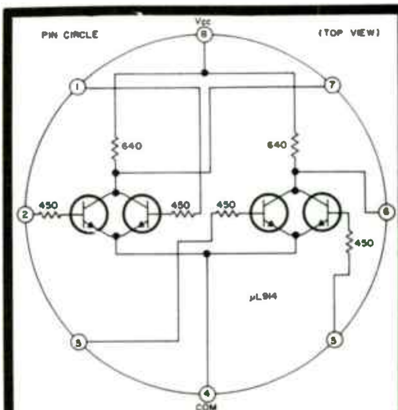


Fig. 3. Basing diagram and circuit of the Fairchild µL914 RTL gate.

HEP Number	MC-Number	Description	Amateur Use (other than logic)
553	303	Half-adder	
554	304	Bias-driver	Regulator
556	306	Three-input gate	Schmitt trigger, free-running multivibrator, amplifier
558	308	J-K flip-flop	Divider, one-shot multivibrator

Table 2. Comparison of the Motorola HEP line totheir MC300 IC elements.

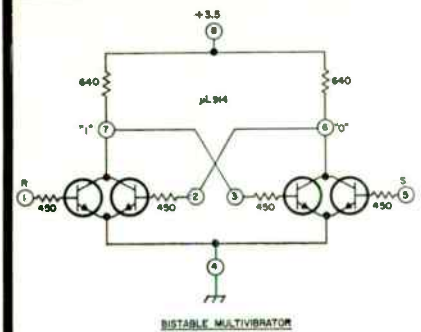


Fig. 5. A Fairchild μL914 in a bistable multivibrator circuit.

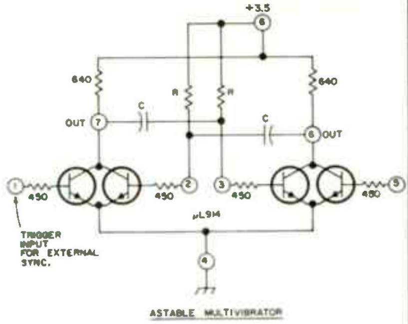


Fig. 6. An astable multivibrator using the μL914.

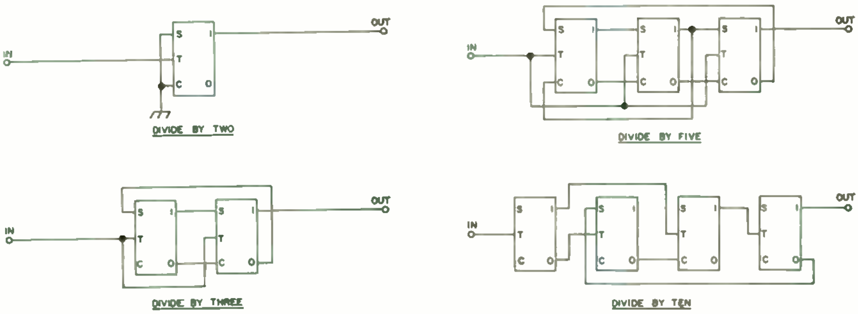


Fig. 7. J-K flip-flop divider circuits; divide by two, three, five and ten.

hurts to put a 330 μF-6V tantalytic capacitor right across the IC supply terminals.

The RTL J-K flip-flops require a fast rise-time waveform to trigger them properly. Try to keep your rise-time to less than 1 μsec if possible. For instance, the ÷ 5 circuit of Fig. 7 was unreliable when the input rise-time approached 3 μsec.

The convention in the Motorola and Fairchild RTL family is to add load factor numbers adjacent to the pin numbers of the IC diagrams. For instance, a μL914 gate input is *three* units of loading and a μL914 gate output will drive *16* units of loading. This load factor scheduling is completely consistent within the Fairchild μL900 series, even though some members of this family are lower power units than others. The Motorola

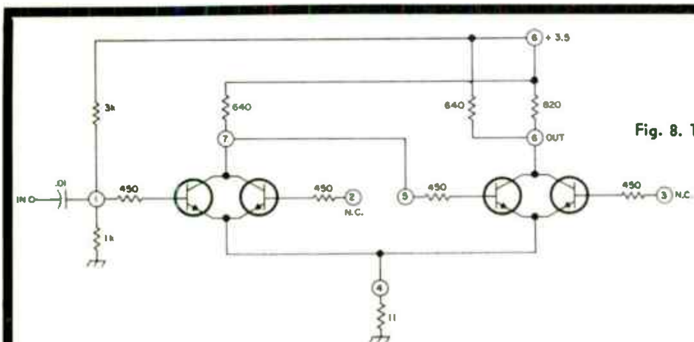
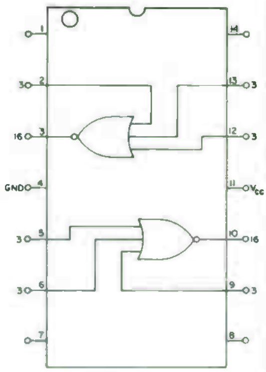
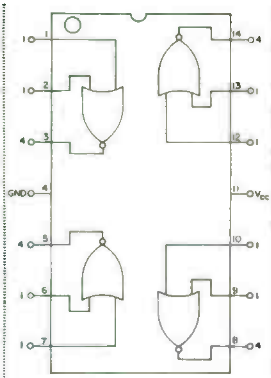


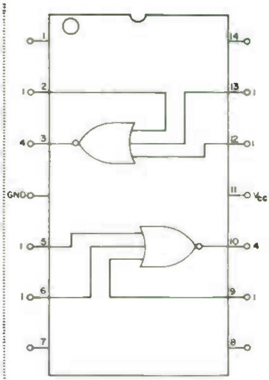
Fig. 8. The Schmitt trigger.



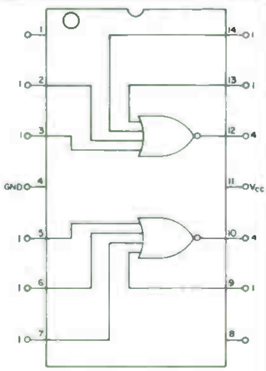
MC715P DUAL 3-INPUT GATE



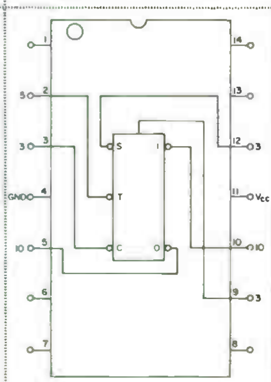
MC717P QUAD 2-INPUT GATE



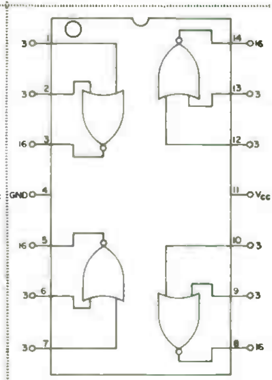
MC718P DUAL 3-INPUT GATE



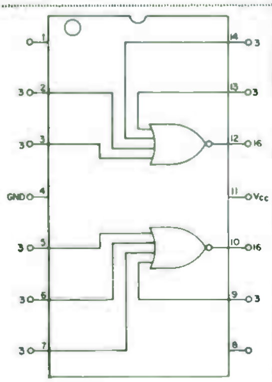
MC719P DUAL 4-INPUT GATE



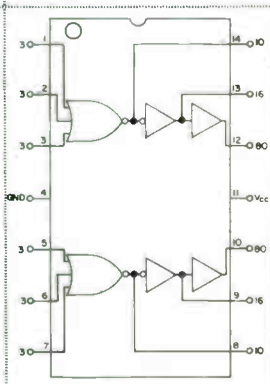
MC723P J-K FLIP-FLOP



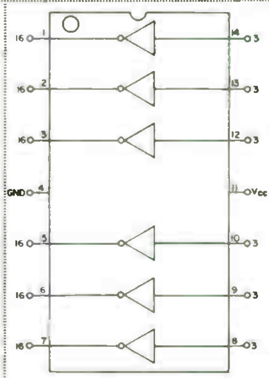
MC724P QUAD 2-INPUT GATE



MC725P DUAL 4-INPUT GATE



MC788P DUAL BUFFER



MC789P HEX INVERTER

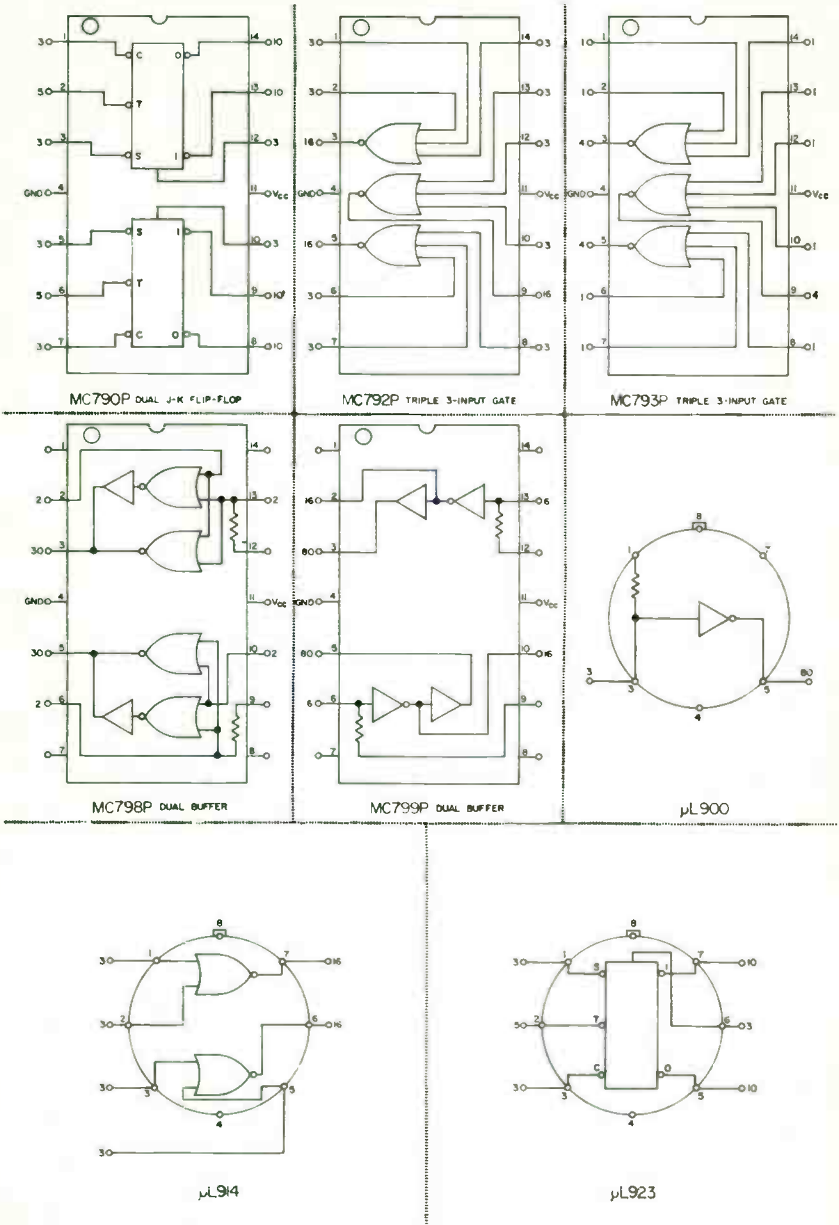


Table 1. Basing diagrams, circuit logic and load factors for popular IC packages. The MC-numbered units are manufactured by Motorola; μL-units by Fairchild.

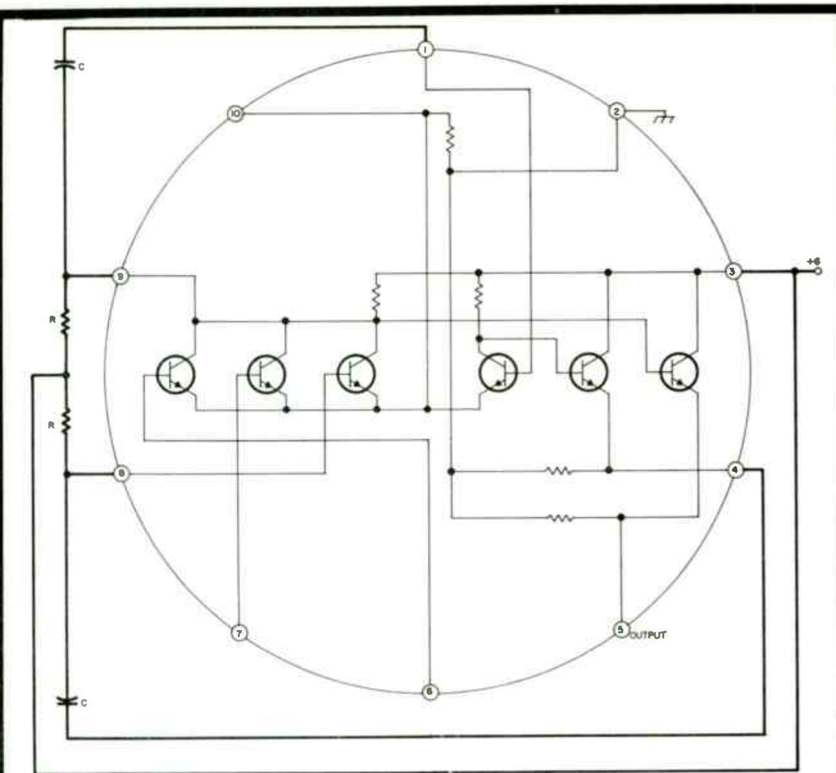


Fig. 9. The HEP 556 connected as a free-running multivibrator. For an output frequency of 1200 Hz, R = 100k and C = 0.01 uF. For greater output level at the expense of waveform distortion, pins number 2 and 10 may be shorted together.

MC-700P series uses the same supply voltages and logic voltage levels; and the load factor designations are also compatible with the Fairchild μL900 family.

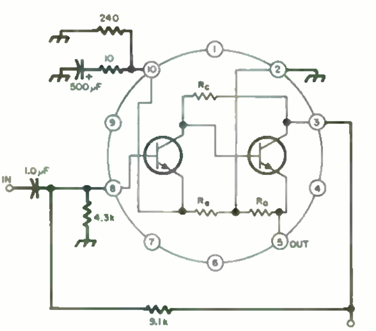


Fig. 10. Using the HEP556 as a low input impedance, low-level amplifier. The voltage gain of this circuit is approximately 20.

The only types of RTL IC's that will appeal to those with low budget projects will be the less-expensive units that are packaged in epoxy or plastic. Three of the Fairchild μL900 family are available in epoxy: the μL900, μL914, and μL923. Also the entire Motorola MC700P family is plastic and available at about the same price level *per logic function* as the Fairchild μL900 series.

The Motorola MC700P line includes only one single function unit, the MC723P—a J-K

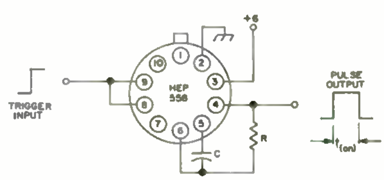


Fig. 11. The HEP558 connected as a one-shot multivibrator. The main consideration when using this circuit is to keep the value of R less than 160 kilohms; the $t_{(on)}$ time will be approximately equal to 1.4 RC.

flip-flop. The rest of this DIP family consists of multiple function units. The MC790P is an outstanding one; it is a dual J-K flip-flop at $2.00. This brings the price per J-K down to $1.00, the lowest in the industry to this author's knowledge.

Table 1 shows the various economy-plastic RTL units available from the two lines discussed, with their loading diagrams and prices. It is important to keep in mind that the pin-numbering of *these* IC's is *as viewed from the top of the package;* this is the reverse of the way transistor basing diagrams are usually shown. There are several articles available on these RTL units that can be helpful: references 3, 4, and 5.

Without sounding biased in favor of RTL, the author feels that this family represents the best one to "cut one's teeth" on. The reasons for this feeling are simple: (1) RTL is inexpensive, so first experience comes cheap, and (2) since there is a resistor in nearly every lead to the internal transistors of the IC chip, your mistakes are not likely to destroy the units.

ECL (Emitter-Coupled Logic) is another family that should be of interest to the amateur. Motorola has recently made four types of MECL (Motorola Emitter Coupled Logic) available in their HEP line. This HEP line has the distinct advantage of being available nearly anywhere in the U.S. and also through mail order firms such as Allied Electronics. Table II shows the types of HEP IC's that are available, and the similar industrial ver-

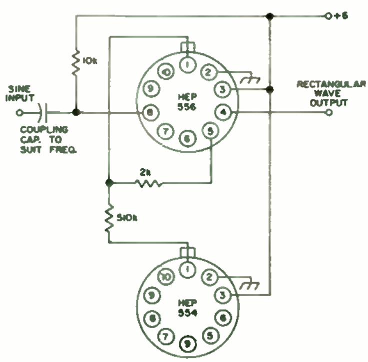


Fig. 12. Using the HEP554 and HEP556 as a Schmitt trigger. The 510k resistor should be 510 ohms.

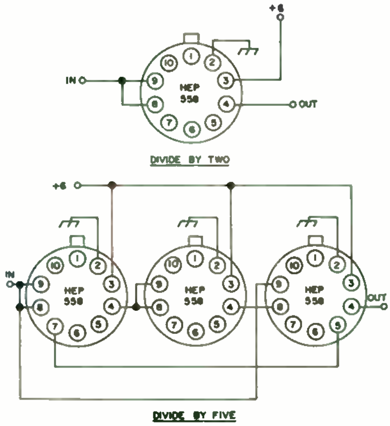


Fig. 13. Connecting the HEP558 IC as a divider—divide by two and divide by five.

sions of each. The HEP versions of the standard Motorola MC300 series are not obliged to have the same specifications as their industrial versions, but the cross-referencing is still helpful. By consulting the Motorola Application Notes for the MC303, MC304, MC306, and MC308, a large body of helpful information can be gleaned on uses for the HEP553, HEP554, HEP556, and HEP558.

The MECL family is unlike most other logic families in that a logic level change does not cause any component transistors to saturate. This means that the MECL family can operate much faster than others, since no saturated transistors have to be pulled out of saturation during switching. The industrial MC300 units can be used up to 30 MHz switching rates, so, we can expect to find *some* of the HEP units that will approach this rate, too. This inherently faster operation is reflected in small propagation time through counting elements. This allows us to build serial dividers of *large* prime numbers, like 17, at fairly high frequencies.

Fig. 9 through 13 show several uses for the HEP integrated circuits. The applications to which these circuits are put, will be left to the readers' needs and ingenuity. Some of the units which can be built using HEP integrated circuits are described in detail in a booklet by Motorola.

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CHAPTER 6

Crystal Oscillators

Most standard crystals, surplus or otherwise available inexpensively, are for vacuum tube circuits. If used in transistor oscillators, these crystals usually oscillate on slightly different frequencies than marked, since most transistor circuits operate at low impedance levels unlike tube circuits.

Transistor crystal oscillator circuits fall into two categories: those which operate near the series resonance of the crystal (f_s) and those which operate near the parallel resonance of the crystal (f_∞). These operation points are illustrated in **Fig. 1.**

Several oscillators of the series resonance type are shown in **Fig. 2.** Note that in each circuit, if one were to remove the crystal, and substitute a blocking capacitor, the circuit would continue to oscillate at about the same frequency. That is, the crystal in the oscillating circuit looks like a small resistance.

The circuits that utilize the parallel resonance of the crystal all operate slightly on the inductive side of f_∞. Thus, these circuits are all, after they have been broken down to basics, Colpitts oscillators. Two examples of this type of oscillator are shown in **Fig. 3**; the only difference between them is that one is grounded-base and the other grounded-emitter.

If one picks the C_{be} and C_{ce} capacitances such that they yield 32 pf in series (or supplements them external to the transistor to give 32 pf) then the crystal will oscillate on its marked frequency in most cases. Of course, the voltage division ratio of C_{be} and C_{ce} must also be adjusted to give a feedback factor commensurate with the β of the transistor.

There are a host of other transistor crystal oscillators, which work to one degree or another. These, for the most part, are subject to adjustment and the individual transistor's characteristics, costing the designer the very stability he used crystals to obtain. Possibly the most satisfactory circuit for use with "32 pf" crystals is the very tube circuit for which it was designed—but replacing the tube with a field effect transistor. In this way, the impedance levels are of the same order and the circuit is readily dealt with by technicians experienced in tube circuitry.

For a short period of time, when FET's (Field Effect Transistors) were first introduced, their prices were quite high. However, although there are still some in the $50/each category, a number of reasonably priced FET's are available for experimenter use. Siliconix now has their U146 and U147 at about $3.00; these P-channel Silicon units are of the 2N2606 family.[1]

FET's can be thought of as near equivalents of vacuum tubes. The drain corresponding the plate, the gate to the grid, and the source to the cathode. This correspondence is a much better one than the collector-plate, base-grid, and emitter-cathode set for bipolar transistors and tubes. As in a tube, the current through the FET is controlled by the *voltage* between the control electrode and source of current carriers. **Fig. 4** shows the representations of FET's along with a tube symbol to demonstrate the likeness.

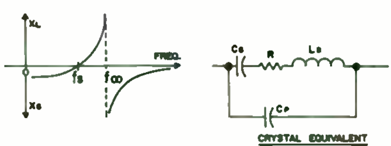


Fig. 1. Series (f_s) and parallel (f_∞) resonant points for crystals.

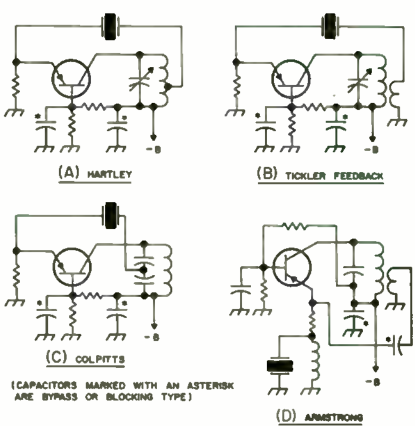


Fig. 2. Typical series resonant transistor crystal oscillators.

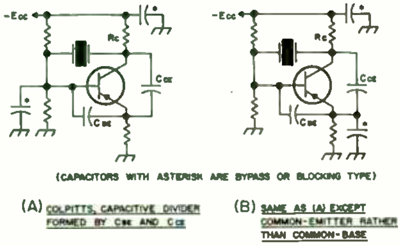


Fig. 3. Parallel resonant crystal oscillators.

The N-channel is most like the vacuum tube because one applies positive voltage to its drain and negative bias voltage to its gate for amplifier operation. The P-channel types are the other way around with negative voltage on the drain and positive bias. (The MOS types—which stands for Metal Oxide Silicon—are still rather expensive for amateur use.)

To illustrate how one can go directly from a well-established tube circuit to an equivalent FET circuit, let's steal the 100 kc crystal calibrator from the Collins 75S1. The original is shown in **Fig. 5**, and its triode equivalent in **Fig. 6** (we must draw the *triode* equivalent since we are going to replace the tube with a triode device). The only difference between these two circuits is that the output coupling of the original (electron coupling) has been left out. Let us now, again, redraw the circuit substituting an FET and lower the "plate" voltage a bit to suit the E_{dss} of the transistor, as in **Fig. 7**.

A parallel resonant circuit for the desired oscillator frequency is placed across the "plate" load to allow sufficient dc to flow and to make a low impedance tap available for output.

The 8-50 pf trimmer capacitor can be adjusted a bit for each crystal to bring the oscillation frequency right to that which is stamped on the crystal can. In most cases, the output frequency will fall within the crystal tolerance with no adjustment necessary.

Since we had to dispense with the electron-coupling method of deriving the oscillator's output, it would be well to add some output isolation to this FET equivalent. This can be done by adding an amplifier stage, using a conventional silicon transistor, as in **Fig. 8**.

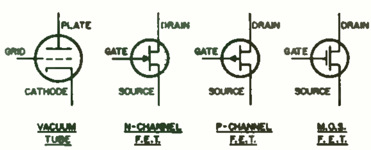


Fig. 4. Symbols for FET's and MOS FET's.

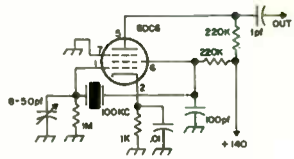


Fig. 5. Collins 100 kc calibrator.

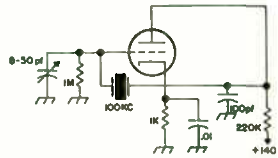


Fig. 6. Triode equivalent of Collins 100 kc calibrator shown in Fig. 5.

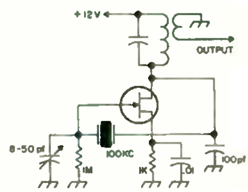


Fig. 7. FET equivalent of Fig. 6.

In the above example, the FET used was a Texas Instrument TIX-882, a germanium FET which was available for about $3.00 several years ago. This type (N channel) was used to

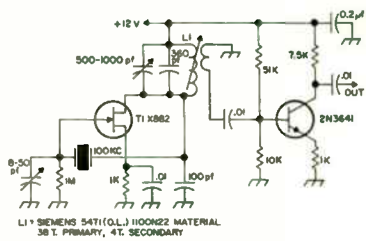


Fig. 8. Isolation amplifier added to Fig. 7.

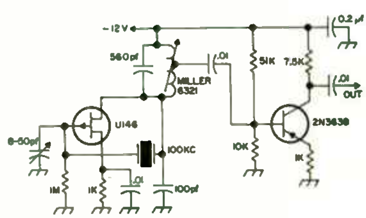


Fig. 9. P-channel FET oscillator and isolation amplifier.

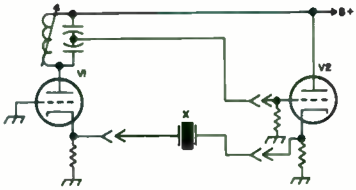


Fig. 10. A Butler oscillator as analysed by sections.

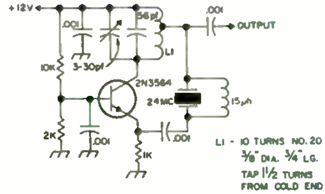


Fig. 11. In this circuit for crystals designed for series mode operation, an inductance is added in parallel to the crystal to resonate the holder capacitance and prevent spurious oscillation.

allow the tube-circuit-conversion example to use a positive supply, for illustration purposes only. The TIX-882 is no longer available, but a silicon N channel FET could be substituted.

Since it is the P channel devices that are now available inexpensively (U146 and U147), **Fig.** 9 shows the circuit (again redrawn) for one of these.

The main intent in the foregoing section was to demonstrate how one can use FT 171, FT 241, FT 243, and similar surplus crystals with FET oscillators so that they oscillate as marked. However, in the more recently available surplus, there are some crystals which *are designed for* series-mode operation. Military types in the HC10/U (coaxial) case are for series operation: CR9 and CR 24. Also, many of the crystals in the HC6/U (hermetically-sealed metal cases with ½" spaced, 0.050" pins) are for series operation: CR19, 23, 25, 26, 28, 30, 32, 35, 45, 51, 52, 53, 54, 65, and 75.

Some of these crystal units were designed for tube type oscillators of the "Butler" type, but will work well in the type of oscillators shown in Fig. 2. The Butler oscillator achieves the low impedance driving source and low impedance input by use of a cathode-follower and a grounded grid stage respectively. The Butler oscillator is shown in **Fig. 10,** divided into its component sections. As can be seen, V_1 is a Colpitts Oscillator whose capacitive-tap does not return directly to the cathode but drives V_2, a cathode follower. V_2 drives the cathode of V_1 through the low series-resonance resistance of X.

This method of utilizing the series-resonance frequency of a crystal with a tube circuit is the "hard way," although an equivalent FET Butler oscillator could be built. It is far easier to use a single bipolar transistor oscillator like one of those in **Fig. 2.** An actual circuit is shown in **Fig.** 11, for a 24 mc crystal oscillator using a CR 24 coaxial crystal.

. . . **W6GXN**

RECEIVERS AND CONVERTERS

CHAPTER 7

Building Blocks

Here is an idea you might like to try. A small receiver, which is suitable for use with most of the myriad of semiconductor converters, can be built for $20 and the aid of a good junk box. If a receiver is not what you need, there are still some circuits and ideas to suit your fancy.

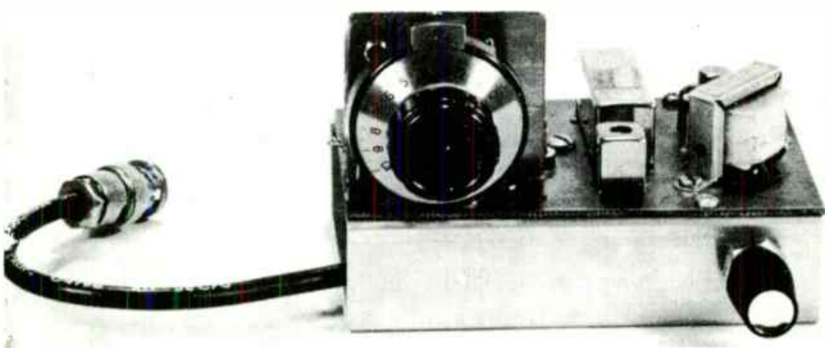
The use of discrete electronic components is becoming obsolete in cases where a standard circuit is used. Instead, we use assemblies, or integrated circuits, which are usually superior in design to the original approach. As you will see with this receiver, the use of building blocks is also a means of economizing.

Circuit description

By using an RCA CA3020, which costs $2.80, I was able to build a quality 500 mW audio system at a price competitive with foreign-made audio assemblies. The *if* amplifier and detector is a Miller 8903. This assembly consists of two units, an 8901 and an 8902, which will provide 55 dB gain at 455 kHz. The cost—$5.75. My only misgiving on this unit is its frequency. That 910 kHz image is a problem when the receiver is used with VHF converters.

After some deliberation, I chose to build the mixer stage with a junction field effect transistor (JFET). The epoxy 2N3819 by Texas Instruments fills the bill for less than $1.00.

Tuning of this stage, and tracking with the oscillator, provides the only real challenge of the system. The values shown tune from 12 MHz to 18 MHz, but other ranges are easily obtained by changing the tanks. I was able to maintain good tracking with two adjacent rotor plates removed from the antenna section of the tuning capacitor. It was also necessary to bend the outside rotor plates slightly, for good mid-band tracking. Since the transistors are fairly well isolated from the tanks, tracking checks were possible with a grid-dip oscillator with no power applied to the receiver. Slight adjustment of the an-



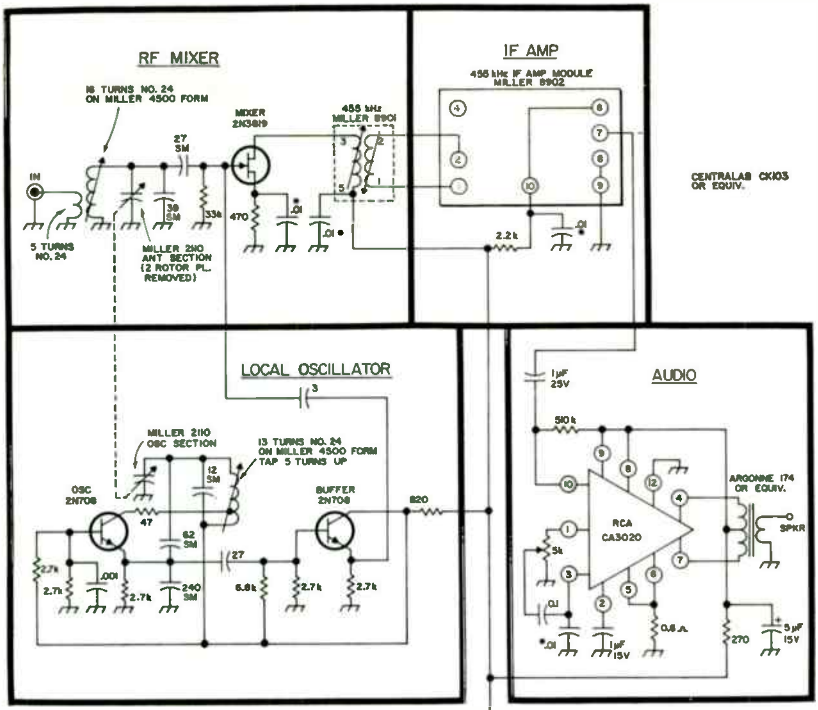


Fig. 1. Circuit diagram of the building-block receiver.

tenna tank is necessary when power is applied.

If a local oscillator is loaded by the mixer when strong stations are tuned, it will be pulled from its normal frequency. A buffer stage, in the form of an emitter-follower, will overcome that problem. George Daughters, WB6AIC, used a rather elaborate two transistor buffer in his HBR-TR receiver (*QST*, April 1967), but this seemed unnecessary in the unit described here. I did copy George's oscillator with good results, however. Although the 2N708 was used twice in this circuit, there is really no preference for that type. RCA 40237's, or various plastic types, would be less expensive.

The entire unit was mounted on a 2⅞ x 4⅛ inch piece of PC board, which by luck, almost fits a Bud CB1626 chassis. A good PC board designer might lay out the printed wiring and eliminate the 12-pin socket used for the integrated circuit. Eventually, the receiver will be part of a more classic enclosure. The Argonne 36 mm vernier dial was photographed to show the intended mounting.

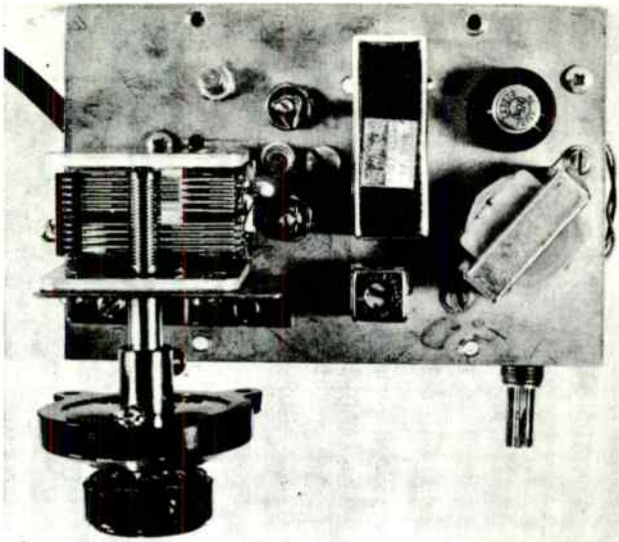
Performance

While this may not be the ultimate in modern day receivers, its performance is good. I have used it in conjunction with a 2-meter converter similar to that of K6HMO (*73*, October 1966). Selectivity is excellent, but the sensitivity might be improved. The JFET does not easily overload; however, it appears to have a minimum threshhold which acts like a squelch.

An earlier attempt used a bipolar mixer which had better sensitivity, plus more noise and overload problems. Despite the problems, some experimenters may find the bipolar type to be preferable.

In all, this circuit satisfied my immediate need for a small, inexpensive, superheterodyne receiver, and it opens new doors for experimentation. Some ideas, which seem practicable using these building blocks as stepping stones, are suggested below:

1. A dual conversion system—by adding a 4.5 MHz *if* and crystal oscillator. The RCA CA3022 integrated circuit might be used for the amplifier.



Top view of the building-block receiver showing the location of the components.

2. An rf stage—because of the noise and selectivity characteristics described, a broadband stage might be suitable.
3. A product detector
4. A higher power audio system—by substituting an Argonne AR163 for the present output transformer, the CA3020 will drive a power transistor such as the RCA 40250 or the 2N3054.

It is my intention to work on some of these ideas for a new mobile receiver, but I would not be disappointed if you beat me to completion.

. . . WB2EGZ

CHAPTER 8

The 2Q-A22 Transistor Communications Receiver

A strong desire to duplicate the popular Drake 2-B receiver in transistor form, prompted the building of the receiver shown. It is the result of over two years of experimental design, building, rebuilding, testing and listening. The block diagram in **Fig. 1** closely resembles that of the Drake, and for that reason I have named it the 2Q.

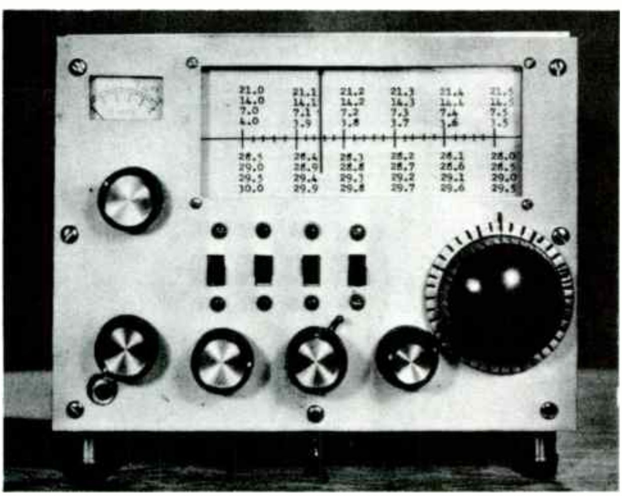
The completed receiver is a triple conversion superheterodyne, covering all amateur bands 10 through 80 meters. It has excellent sensitivity, selectivity and stability. Cross modulation has been reduced to a minimum by the use of FET transistors in both the rf and first mixer stages. Such features as bandpass tuning, FET detector, S-meter, agc and a 100 kHz crystal calibrator are included.

The circuit shown in **Fig. 2** is actually the result of two that were built. The first design, following the usual transistor circuit theory, matching impedances, etc., resulted in a receiver that lacked the necessary sensitivity and selectivity. Cross modulation was also a problem because bipolar transistors were used in the front end. The second design is the result of a concentrated effort toward obtaining maximum selectivity by the use of small capacity coupling where possible, high Q tuned circuits, and tapping collectors down on the coils to preserve their Q. Cross modulation was reduced to a minimum by using FET transistors in both the rf and first mixer stages and by using a separate rf gain control.

The circuit

Much has been written on transistor circuitry during the past few years so I will be as brief as possible and describe only those points which I think important or unusual.

Capacitive coupling is used throughout the front end (preselector). It uses high-Q toroid coils and slug-tuned coils. The simple switching provides the necessary selectivity and ease of adjustment desirable when compact construction is used. Ami-Tron toroids were not used for the 15- and 10-meter bands due to the lack of space for the necessary trimmer capacitors, but their use is definitely recommended for all bands. The selectivity and stuffing ratio gained by their



Front view of the 2Q transistorized receiver. Bottom, left to right, are the phone jack, bandswitch, *if* gain, band-pass tuner, selectivity switch combined with rf and volume control, and main tuning. Switches, left to right, are agc S-meter switch, dial light, 100 kHz calibrator and bfo.

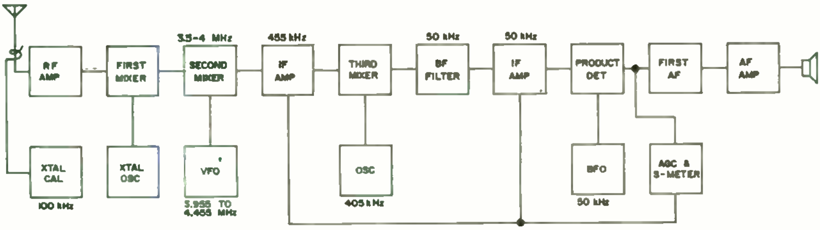


Fig. 1. Block diagram of the 2Q, a completely transistorized communications receiver of modern design using FET's in the front end.

use is very necessary. The tuning capacitor, a two-gang trf unit, was reduced to 200 pF per section. Space for the rf choke was solved by placing it in the crystal oscillator compartment.

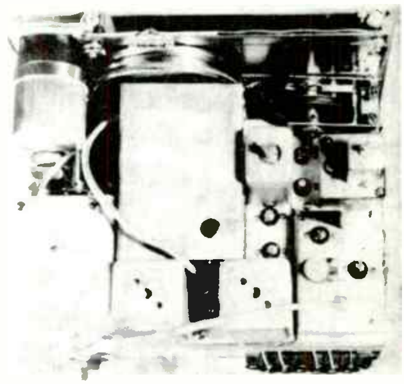
The FET mixer, using source injection, is capacitively coupled to the second mixer. The circuit, possibly of my own design, was preferred to a gate injection circuit. The only FET's available were N-channel 2N3823's, but possibly some of the cheaper ones will work as well.* I intend to try the Motorola MPF 105 FET when I can locate a distributor who stocks them. Alignment of the front end is simply a matter of adjusting turns, spacing, and trimmer capacitors, until the amateur bands are staggered across the preselector dial.

The 3.5 MHz-4.0 MHz variable *if*, mixer and oscillator section, consists of a high-C Colpitts oscillator, and a base injected mixer, with an output at 455 kHz. Only the highest quality components should be used here, since it is a major frequency determining circuit.

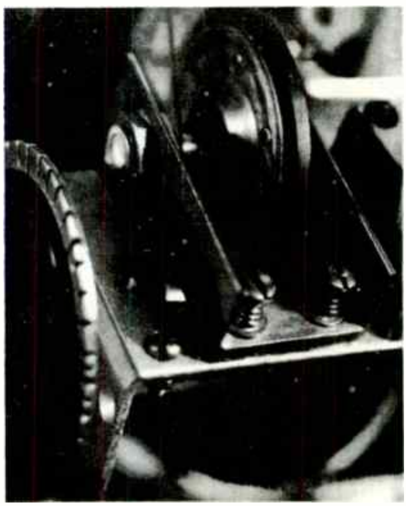
Only one stage of amplification was found necessary for the 455 kHz *if* section. The mixer is capacitively coupled with base injection at 405 kHz from a high-C Colpitts oscillator giving a 50 kHz output. Here again the oscillator is a major frequency determining circuit and care should be used in its construction. The 455 kHz *if* coils can be any high-Q center tapped units, preferably using toroids or cup cores. This is a good spot for a mechanical filter; something I intend to try in the near future.

The band-pass tuner was constructed using coils wound on 1" diameter powdered iron toroids from an old telephone company audio filter. The ones I used were blue and numbered A9301572. The tuning is done with a three-gang trf type broadcast tuning capacitor, with a stop added to limit its travel to about 20 degrees, starting from maximum. The switch uses a hollow ¼ inch shaft, with

* 2N3819 FET's seem to work as well as the more expensive 2N3823's. With the 2N3819, the only circuit changes were in the rf amplifier—the source was grounded and B+ changed to 14 V. Motorola MPF-103's have been tried too. At 90¢ each they seem to work as well and their specs are almost identical.



Top view of the 2Q receiver, showing the layout of the various parts.



Close up view of the dial tuning mechanism.

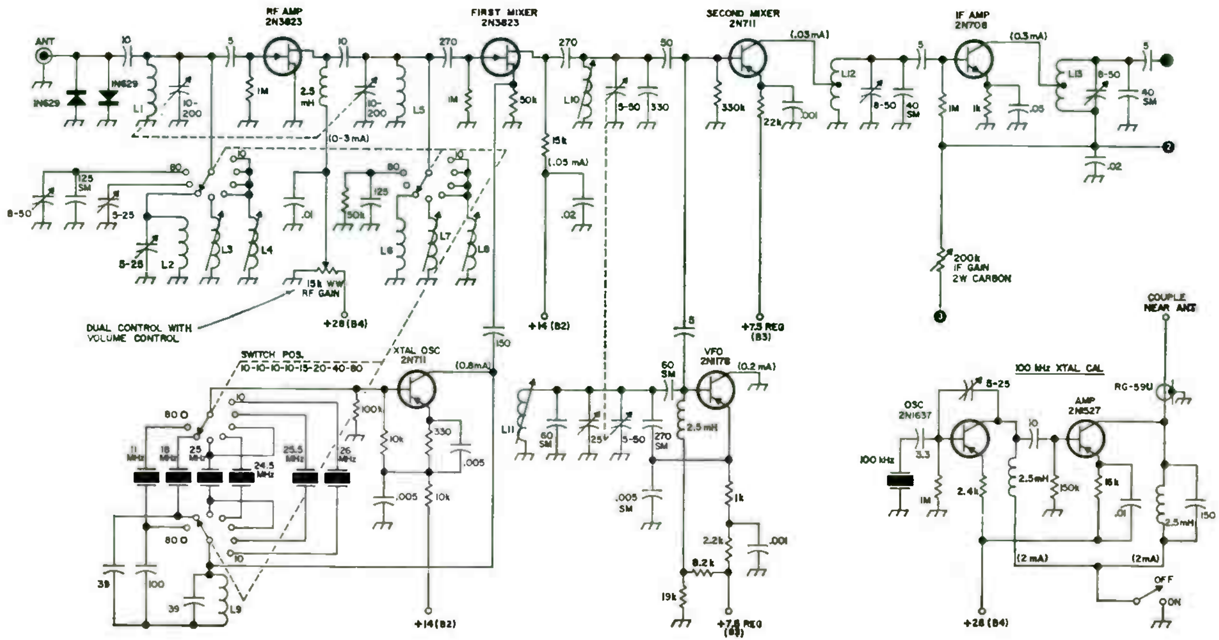
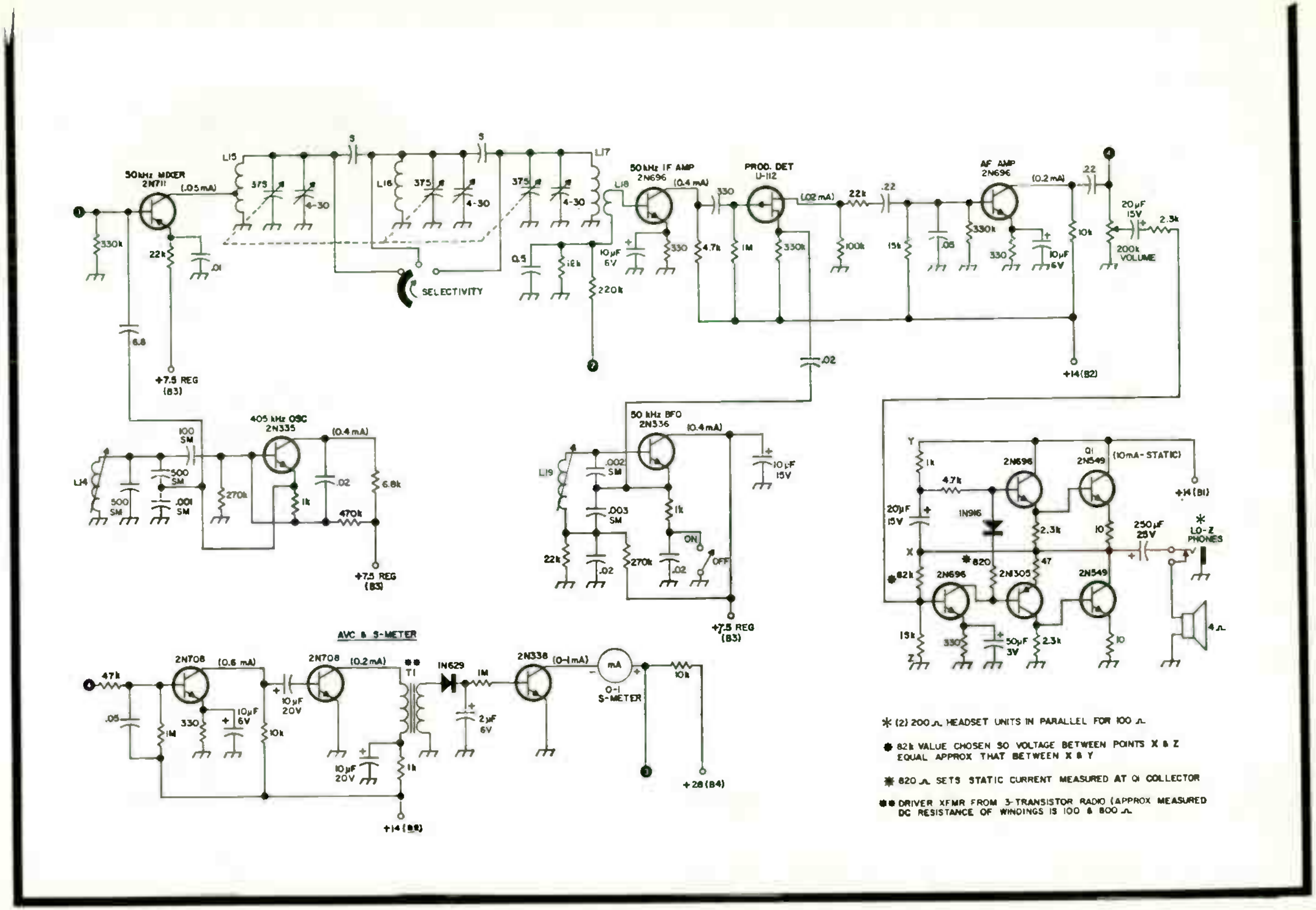


Fig. 2. Schematic diagram of the 2Q communications receiver. The currents shown in parenthesis are the collector currents for each stage. The rf chokes used were 3-pi types on ⅛-inch iron cores taken from a surplus computer board although miniature 2.5 mH units should work ok. Coil values are given in Table 1. Later experimentation by W5ETT indicates that the bias network used with the 2N708 455 kHz *if* amplifier was not too tolerant to different transistors. He recommends removing the 1M base-bias resistor and replacing it with a 330k resistor *and* a 27k resistor from base to ground.

*T-68-2 and T-50-6 toroid cores may be purchased from Ami-tron Associates, 12033 Otsego Street, North Hollywood, California. Price 50¢ each plus postage.



50kHz MIXER
2N711
(.05mA)
L15
L16
L17
L18
375
4-30
3
SELECTIVITY
330k
22k
.01
6.8
+7.5 REG (B3)
0.5
18k
220k
10µF 6V
50kHz IF AMP
2N696
(0.4 mA)
330
4.7k
1M
PROD. DET
U-112
(.02 mA)
330k
22k
.22
100k
15k
.05
AF AMP
2N696
(0.2 mA)
330k
10µF 6V
10k
20µF 15V
200k VOLUME
2.3k
+14 (B2)
.02
405 kHz OSC
2N335
(0.4 mA)
100 SM
500 SM
.001 SM
L14
270k
1k
.02
6.8k
470k
+7.5 REG (B3)
50 kHz BFO
2N336
(0.4 mA)
L19
.002 SM
.003 SM
1k
ON
OFF
10µF 15V
22k
.02
270k
+7.5 REG (B3)
Y
1k
4.7k
2N696
Q1
2N549
(10mA-STATIC)
+14 (B1)
20µF 15V
1N916
2.3k
10
250µF 25V
LO-Z PHONES
X
820
2N1305
47
82k
2N696
2N549
15k
330
50µF 3V
2.3k
10
Z
4Ω
AVC & S-METER
47k
.05
2N708
(0.6 mA)
1M
330
10µF 6V
10k
10µF 20V
2N708
(0.2 mA)
T1
10µF 20V
1k
+14 (B2)
1N629
1M
2µF 6V
2N338
(0-1mA)
mA
0-1 S-METER
10k
+28 (B4)
(2) 200Ω HEADSET UNITS IN PARALLEL FOR 100Ω
82k VALUE CHOSEN SO VOLTAGE BETWEEN POINTS X & Z EQUAL APPROX THAT BETWEEN X & Y
820Ω SETS STATIC CURRENT MEASURED AT Q1 COLLECTOR
DRIVER XFMR FROM 3-TRANSISTOR RADIO (APPROX MEASURED DC RESISTANCE OF WINDINGS IS 100 & 800Ω

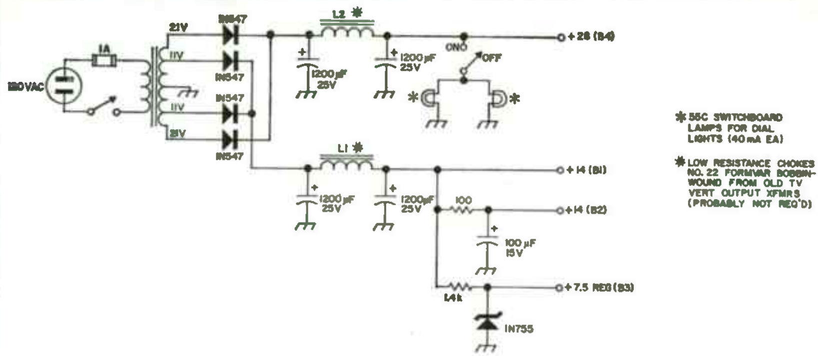


Fig. 3. AC power supply for the 2Q communications receiver.

the TC shaft being operated through it.

The FET detector using a P-channel U112 or 2N2497 has plenty of bfo injection and works very well on SSB.

Good S-meter action and a certain amount of gain control is provided by the circuit shown by simply reducing the amount of voltage applied to the *if* transistors.

I had some trouble getting the 100 kHz crystal calibrator aligned with WWV, so it was necessary to devise the circuit shown. WWV may be received on the receiver, during daylight hours here, by putting it on 7 MHz and tuning the preselector to minimum capacity.

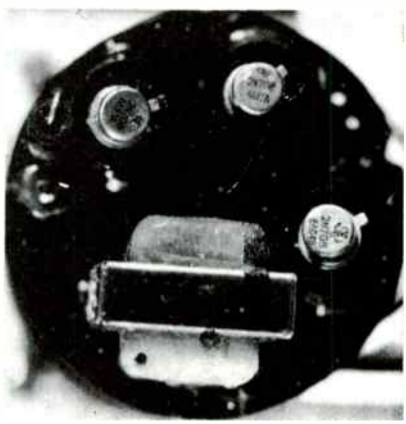
Table 1. Coils for the 2Q receiver.

L1	32 turns #22 Formvar on T-68-2 toroid core.
L2	20 turns #22 Formvar on T-50-6 toroid core.
L3, L4	20 turns #22 Formvar, 1/4" diameter with last 6 turns spaced to take a 1/2" long powdered-iron core.
L5	Same as L1 except 33 turns.
L6	Same as L2 except 23 turns.
L7	Same as L3 except 23 turns.
L8	Same as L3 except 23 turns.
L9	30 turns #24 enameled, 1/4" diameter.
L10	22 turns #24 enameled on 1/4" slug-tuned form.
L11	15 turns #24 enameled on 1/2" form, spaced diameter of wire. 3/8" powdered-iron slug.
L12, L13	110 turns, 6-strand Litz wire, tapped at 55 turns. Pi wound on 1/2" diameter ferrite cupped core 1 1/2" long. Three cups stacked to obtain necessary length after grinding out center of middle core.
L14	120 turns, 6-strand Litz wire, pi wound on 1/4" slug-tuned form.
L15, L16, L17	330 turns using 3 strands #29 enameled wire wound on powdered-iron toroid 1" diameter, #A930157-2. Toroid cores from old telephone equipment will work. L15 tapped 50 turns from ground end.
L18	10 turns #22 Formvar wound over L17.
L19	800 turns, 6-strand Litz wire, layer wound 1" long on 1/4" slug-tuned form.

From this point, the rest of the receiver is simply audio, six transistors in all, with a transformerless audio circuit taken mostly from a GE transistor manual. The power supply, one left over from another project, is no doubt overfiltered. Any well-filtered dc source of 14 V and 28 V will do. The receiver draws 20-125 mA depending on volume. The dial lamps use an additional 40 mA each. The receiver will work well on only 12-14 V, but the S-meter and AGC will be out of the picture.

Choice of transistors

The transistors used are by no means the only ones which will work in the receiver. My choice was made largely from tests with the ones which were available in my transistor junk box. Either PNP or NPN will work in most circuits, NPN being preferred in



Back view of the S-meter. One of the 2N708's (Q15) was replaced with a 2N338 after this photo was taken for better agc and S-meter action.

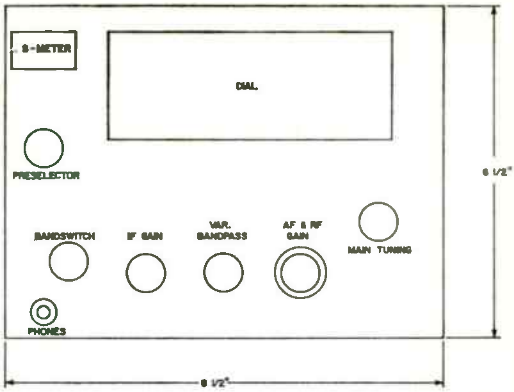


Fig. 4 Front panel layout used by W5ETT in the original model of the 2Q receiver.

most cases for *if* and oscillator traisistors. Oscillator types should be those that have no internal connections to the case. The use of sockets for all transistors is highly recommended.

Construction

The receiver cabinet measures 8½" long x 6½" high x 6½" deep. The receiver is divided into a number of sub-assemblies mounted on a main chassis, made of 14-gauge aluminum. The sub-chassis are of 21-gauge aluminum.

Only the 50 kHz *if* amplifier and audio stages were built on the main chassis. The S-meter and agc circuitry were mounted on the back of the S-meter. Fig. 4 is a rough layout of the front panel.

The slide rule dial has a tuning rate of 45:1 or 45 turns of the tuning knob to cover 500 kHz. This gives at least 25 revolutions on the 40 and 20 meter bands. The mechanism consists of a weighted knob on a ¼" shaft driving a 2" rubber tired wheel (Jenson #J1490-01) on a ¼" shaft driving a dial cord to a 3½" dial drum on the tuning capacitor. The dial scale was made on white paper (pasted to a piece of stiff cardboard) using a black ball point pen and a typewriter. The dial drum was made from a reinforced, nickle-plated lid from a peanut butter jar.

Conclusion

No wild claims shall be made for this receiver except to say it is the best homebrew receiver I have ever owned. Only 5 feet of wire strung up in the shack has been found necessary for good reception. Many hours were spent just listening and hearing signals that I could never hear with my old 14-tube homebrew receiver. I would like to thank Jim Miles W5KWJ for his comments and encouraging me to write this article.

. . . W5ETT

CHAPTER 9

Transistorized CW Filter & Monitor

Just about every transceiver on the market still lacks adequate built-in provisions for CW work—that is, a CW keying monitor and sharper IF or audio selectivity for CW. The author previously built* a tube-type accessory unit that could be used with almost any transceiver to provide these functions. However, being an "outboard" unit it was not suited for use in mobile or portable applications. What was needed was a minaturized version which could be tucked away inside a transceiver case.

Circuit

After some experimentation, the author came up with the circuit shown in Fig. 1. Transistors QI, Q2 and Q3 comprise a so-called "active" audio filter which allows a good deal of selectivity to be obtained by only RC circuitry. Essentially QI, and the network in its base circuit make up a low-pass audio filter while Q3 with its network comprise a high-pass filter. The combination of the two circuits produces the selectivity characteristics shown in Fig. 2. The filter can be used between any two low-level, moderate to high impedance points in a receiver audio circuit.

The dashed line in Fig. 2 indicates the ideal if selectivity which a 2 kHz mechanical filter might provide. As can be seen the audio selectivity provided by the three stage "active" filter is a considerable improvement, at moderate attenuation levels, for CW work. If the graph were expanded to cover higher attenuations than 25 dB, it would be seen that the skirts of the active filter flare out beyond those of the mechanical filter. So, while the active filter will by no means provide the same steep skirted selectivity of a narrow crystal lattice or mechanical filter, it does provide enough selectivity, in a very simple form, for effective CW work.

The CW monitor is rf actuated and uses a unijunction transistor in a relaxation type oscillator circuit. The 25 k potentiometer is used as a tone control. A 10 k fixed resistor could be substituted for further simplicity. Output for headphones can be taken from either points A or B. If the outputs from the filter and monitor are parallel to be used with a pair of headphones, some experimentation will be necessary with the coupling condensers at points A or B to find a value which

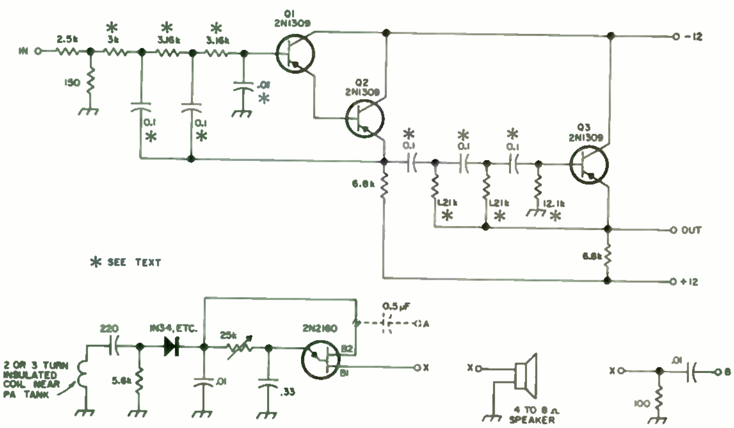


Fig. 1. Filter and monitor circuit which may be used with an SSB transceiver for excellent CW hamming. The −12 and +12 voltages may be obtained from any well-filtered point in the transceiver.

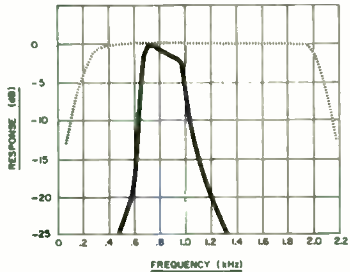


Fig. 2. Approximate frequency response of the filter shown in Fig. 1.

gives sufficient output level without loading the filter unduly (depending on headphone impedance).

The tone from the monitor, like any relaxation oscillator, is hardly very easy on the ears but satisfactory for the occasional CW user. The CW monitor shown in Fig. 3 is suggested if a smoother note is desired. The 12 volts necessary to power the circuit could be obtained from a RF pickup coil and rectifier, as with the unijunction type monitor, or from some point in the transmitter which provides 9 to 12 volts under key-down condition (across the cathode resistor of a rf stage in a grid-block keyed transmitter, for instance.)

Construction

Construction is simple and inexpensive. No adjustments, other than the tone control, are necessary.

How compactly the unit can be constructed depends solely upon the builder's ability to compact components on a perforated circuit board. Except, of course, for the rf pickup coil which must be placed by the PA tank coil, the unit will fit on a 2" x 2" perforated board.

All the transistors used are of the $1 variety. The only components that are critical are the resistors and condensers starred in Fig. 1. If the selectivity characteristics of the filter are to be attained, the resistors must be

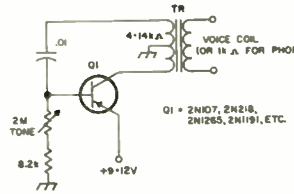


Fig. 3. Another tone oscillator which may be substituted for the one shown in Fig. 1.

of the 1% tolerance variety. The condensers must also be matched as closely as possible using, for example, a capacitance bridge or meter. If "off-the-shelf" 10% tolerance resistors and capacitors are used, performance will likely prove disappointing.

A great many, if not almost all, of the components necessary can be obtained by buying several of the computer boards available at three or four per dollar from various supply houses.

Usage

The unit can be wired into a transceiver so that it can be switched in and out of audio chain in the receiver. A still simpler, "no-holes" approach for those who only use headphones on CW and who have a medium or high impedance headphone jack on their

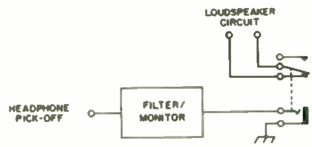


Fig. 4. A phone jack with additional switching circuit may be substituted for the regular phone jack.

transceiver is to replace the headphones jack with a multiple circuit unit, such as shown in Fig. 4. Plugging the headphones in the jack disables the loudspeaker and the filter and monitor are in the circuit ready for CW operation.

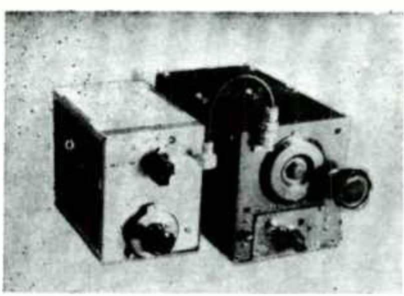
. . . W1DCG

CHAPTER 10

Intermediate Converter

A few weeks back I picked up the BC-band version of a command set at one of the local auctions. Now this is one of the nicer (and rarer) of the ARC-5 or 274-N series. It tunes 0.52 to 1.52 something or others (as this is about 23 years old, I guess they are megacycles) with divisions every ten kHz and about thirty turns of the knob to go the range. The intermediate frequency is 239 kHz and it has the same sort of variable-coupling cans that are used in the BC453, so sharpness is easily obtained.

With such admirable selectivity and band-spread, it seemed a shame that it couldn't be used to run one or more of my UHF converters into, so as to have a permanent setup for scanning 431.95 to 432.4, for instance, while working or trying to work another band. While it is hard to get any image rejection with a second *if* of two or three hundred kilohertz, there should be no trouble in doing it with the receiver tuning 1 to 1.5 MHz. This converter was made for that purpose. My six, two and 432 converters have a nominal intermediate frequency of 14 MHz, while that for 220 tunes 16 MHz *down* for 220 *up*. By using crystals providing beating frequencies of 13 and 17 MHz the ARC-5 tunes forward on all bands, which simplifies the mental arithmetic a



W1OOP's Converter Converter and modified BC command set.

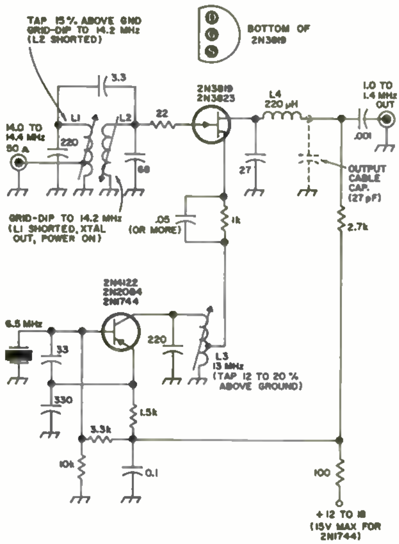


Fig. 1. Simple version of the Converter Converter by W1OOP. This is designed for covering 14.0-14.4 MHz, with 1.0-1.4 MHz output.

slew. (A slew is an archaic unit between three and ten dB.)

The oscillator uses surplus crystals at half frequency because I and Meshna had them. The mixer has an FET, since it made the whole business simpler. The 2N3819 (roughly the same as the TI-S34) can be thought of as a super-6CW4, or maybe a super-6CB6 with poor screening. It is quiet, it oscillates nicely at three hundred megohertz with only the leads to a twenty meter tank circuit, and the gain, while less than overwhelming, is adequate. According to the manufacturers specifications, the currents and voltages could be almost anything, but that is no problem for the man who is making only *one* gadget, because the values can be tailored to the particular FET. Suppose

Here's an intermediate converter for use between a UHF converter and a receiver tuning the broadcast band. It uses an inexpensive field effect transistor as a mixer for simplicity and excellent resistance to overloading.

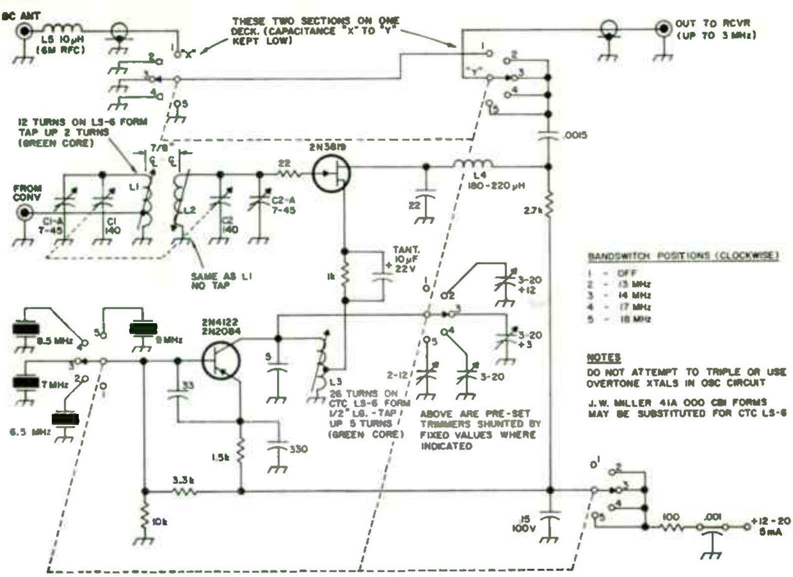


Fig. 2. More complete version of the Converter Converter, with band switching. Shield the quartz crystals if you live within 10 miles of a TV station.

we set up with a six to twelve volt battery on the drain (plate) and gate and source grounded to the negative terminal. We measure a current in the drain lead—it might be anything from 2 to 20 mA. For mixer operation, the FET should be turned all the way on only a bit of the time, and if we are trying to operate as a square-law mixer the gate voltage should swing from cutoff up to zero bias with oscillator drive, giving an average current about 40 percent of the zero-bias value, with drive, and a quiescent current 25 percent of the zero bias value. We can find the proper operating point by measuring the bias voltage required for one fourth the zero bias current, (no signal) and then setting things up so the source (cathode) resistor has that much voltage across it* when the oscillator is driving the mixer. My 2N3819 had 4.5 mA at zero bias. It also had UHF oscillations in the circuit, so I put that 22-ohm resistor

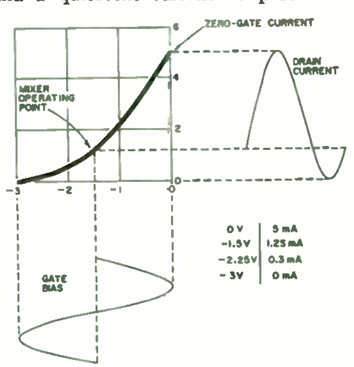


Fig. 3. Operation of the FET mixer.

*Measure through a good rf choke to avoid changing the amount of injection.

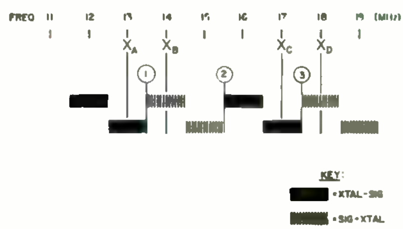


Fig. 4. Coverage of the Converter Converter. If the receiver tunes 0.50-1.50 MHz, there is no gap in coverage. If the receiver tunes 0.52-1.52 MHz, there is a 40 kHz gap at (1) and (3) and overlap at (2). Of course, if X_B and X_D are 40 kHz low, there is no gap. Coverage is 11.5-19.5 MHz. The coverage W1OOP wanted was 12-12.4, 14-14.4 and 16-15.5 MHz (the last backwards, for a 220 MHz converter with 236 MHz local oscillator.)

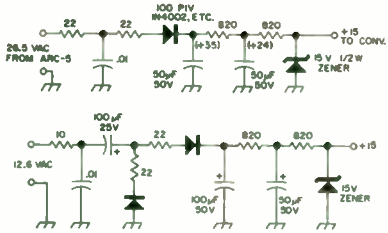


Fig. 5. Two suggested power supplies. The top one operates from 26.5 V ac from the command set and is the one W1OOP uses. The bottom one operates from 12.6 V ac and is hypothetical. In both, the 22-ohm resistors help eliminate hum modulation and hash as well as furnishing protection for the diodes.

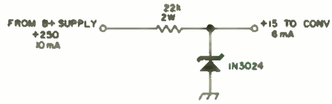
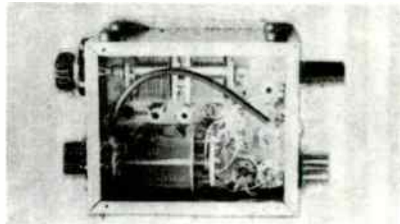


Fig. 6. You can get power for the converter from your receiver B+ supply.



in the gate lead . . . anything ten to fifty ohms would probably do.

With the second preselector tank shorted (I jammed a solder lug into C2) the current with oscillator going was 1.6 mA, but with C2 unshorted the current could be changed from 1 to 2.2 mA by tuning the preselector around. When working, the preselector tuned 1 MHz above the oscillator gave me 1.7 mA, while tuning to the low side dropped the current to about 1.4 mA. The big variations were when the preselector was tuned only a few hundred kilohertz either side of oscillator frequency. I judged that bias and injection were about right.

I used a 10-μF tantalum bypass on the source resistor, on the theory that it would reduce crossmodulation if any were going to take place, but any value over a 0.1μF should work. I had a lot of the small tantalums on hand.

Tuneup: The oscillator section should be got going first, as it is used in setting up the mixer bias as above. The preselector circuits then should be made to track over the range desired. If a dual 140-pF capacitor is used, the tuning range can be about two to one in frequency; for more range, use a larger capacitor. Many of the commercial ham-band-only receivers have a similar scheme. (They cannot track the preselector with the *if* tuning because sometimes they

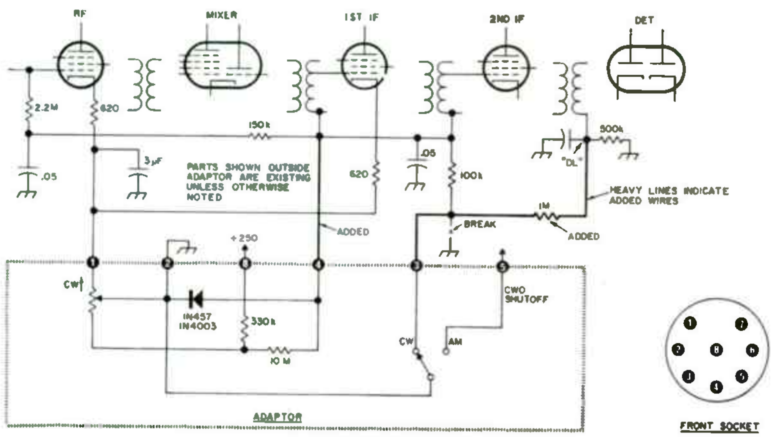


Fig. 7. Adding AGC and gain controls to the SCR-274N. The potentiometer is an Ohmite CB-2531, IRC-CTS Q14-120, Mallory U-28, RV4NAYSD253F or E. It has 25 kΩ total with about 2500 ohms to CW terminal at 50% rotation. The voltage on CCW terminal is low at low gain settings and about +15 volts with the control more than half on. This means no ACG developed until —2 volts on DL. Putting the switch in the CW position puts things back to the original 274-N status. You have to use some ACG delay or there's not enough audio. Replace the antenna connector with a BNC or phono jack for converter use.

are going in different directions!) No *rf* stage is used because of the gain provided by the UHF converter ahead, but the two coils are loosely coupled by being side by side (0.875 in or 2.215 cm center to center) with windings in the same direction and the top end of each coil hot. (With this polarity, a little capacitance coupling will *add* to the inductive coupling.) The two circuits have to tune the right range (or a little bit more, but no less) and should do it together, but they don't have to track with a dial or an oscillator.

I started with the slugs all the way out and went to minimum capacitance on the gang capacitor. Then using the trimmers, I put both coils at about 20 mHz. A piece of drill rod in the coil *not* being grid-dipped keeps things simple. Dip one, shift drill rod, dip other. Then to low end (11.5 mHz) and maximum capacitance put slugs where they need to be. Then, using signals from an antenna or from the dipper, you can peak the trimmers on a signal at the high end and peak the iron slugs on a low end signal once or twice and the job is done. The preselector action should be very obvious. Write down a few dial readings so you can hit them again in a hurry.

One word of caution: although I put the crystals outside the box (in sockets) I found that there was pickup from my rather local TV stations, curable by putting a shield over them. Therefore, I suggest putting the crystals inside the box. The input is a two-stage bandpass filter, the output is a low-pass filter (cuts off a bit over 3 mHz) and there is not any excuse for hash from TV signals, nor for TVI from the oscillator, if things are laid out correctly, and the power leads filtered.

Because the power drain is only about 5 mA, the juice could be stolen from the B-supply in many cases. (Be sure the regulator diode used is dependable!) The power supplies shown will work on twelve or twenty-four-volt filament windings. Without readjustment, it should be ok to use with any receiver tuning either the broadcast band or up to 3 mHz, for instance the 1.5 to 3 mHz Arc-5.

Later tests show that the $1 Motorola MPF105 FET works very well in this circuit.

. . . W1OOP

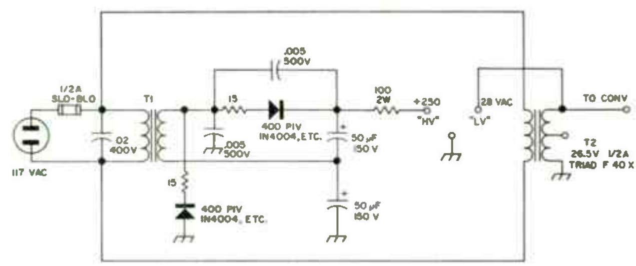


Fig. 8. Simple conversion power supply. T1 is a Lafayette isolation transformer XF or Triad N-51X (20 VA isolation transformer). Use the Triad backwards to keep B+ down. The 0.005 µF capacitors and 15 ohm (not critical) resistors reduce hash in the BC band.

CHAPTER 11

FET Converter for 10, 15 & 20 Meters

It is the author's goal to build a completely transistorized 10—160 Meter station. Lack of a low priced rf power transistor has postponed construction of the transmitter and, until recently, construction of the receiver was not attempted since it was felt that a vacuum tube front end was superior to any transistorized front end the author was capable of building. The availability of reasonably priced field effect transistors has changed the latter situation. With the high input impedance and almost perfect square-law transfer characteristic of the FET, the input circuit loading and susceptibility to cross modulation of a conventional transistor front end are easily avoided. Accordingly, the first step in building a receiver, designing and building a 10-20 meter crystal-controlled converter, was undertaken.

The converter schematic is shown in Fig. I. Motorola 2N4224 FET's are used as the rf amplifier and mixer, while a bipolar 2N1180 is used as a transistor oscillator. A pair of Motorola MPF105's could probably be substituted for the 2N4224's, at a third of the cost, but this has not yet been tried.

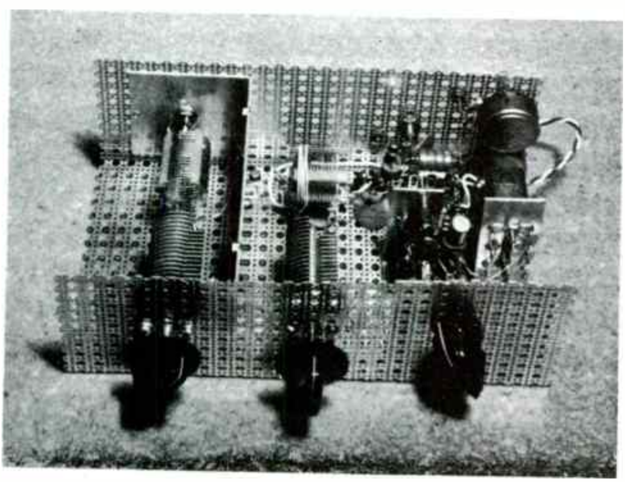
The rf amplifier is designed to provide only enough gain to override any noise generated in the converter, so as to minimize susceptibility to cross modulation. This small amount of amplification, in conjunction with the sharp tuning characteristics of L_1-C_1 and L_2-C_2, yields a front end that is every bit the equal of its vacuum tube counterpart.

Mixer injection is accomplished by means of a "gimmick" capacitor connected to the mixer gate. Source injection was found to be satisfactory, but was not used because of the bandswitching problem it introduced: another switch section and a long lead to the mixer source were required. Source injection would be preferable if local oscillator radiation proved to be a problem, since it places two FET's between the oscillator and antenna rather than one, as is the case when gate injection is used. (The reverse transfer capacitance of the 2N4224 is 2 pF compared to a grid-plate capacitance of 0.02 pF for the 6BZ6 pentode, a common rf amplifier.) A suitable cource injection circuit is shown in Fig 2 for those who may prefer it.

The oscillator circuit was borrowed from another article[1] and is conventional in design.

Adjustment

After the converter is completed, connect it to an antenna and an 80 Meter receiver.



Top view of the converter. Note that the input circuit is shielded. Photo by Chuck Marshall.

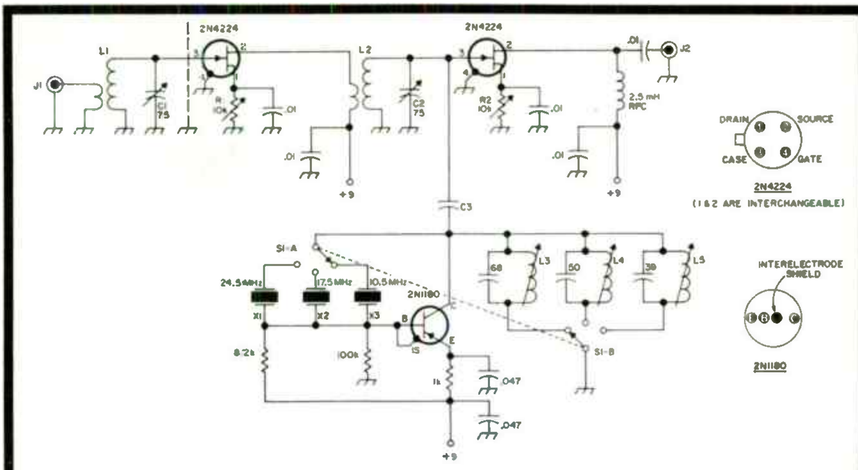


Fig. 1. Schematic diagram of the FET converter for 20, 15 and 10 meters.

DO NOT CONNECT THE POWER SOURCE UNTIL ALL TRANSISTORS ARE IN THEIR SOCKETS. Set the bias pots, R_1 and R_2 to mid-range and place S_1 in the 20 meter position. Then adjust L_3, C_1, and C_2 for a peak in 20 meter signal strength. As the receiver is tuned across the band, C_1 and C_2 will have to be re-peaked every 25 kHz. or so. Place the converter in operation on 15 and 10 meters in a similar fashion. Adjust the bias pots for best converter operation, after which they may be replaced with fixed resistors or wired permanently into the circuit.

Parts List

C_3—Two 1-inch lengths of insulated hookup wire twisted together.

L_1—Primary, 12 turn No. 20, 16 t.p.i., ¾-inch diam. (B&W 3011), Secondary, 3 turns No. 20, 16 t.p.i., ¾-inch diam. spaced 1 turn from primary.

L_2—Primary, as L_1, Secondary, 3¼ turns insulated hookup wire wound on cold end of primary.

L_3—35 turns No. 30 enam. wire, close-wound on ¼-inch diam. iron-slug form (Miller 20A000RBI useable).

L_4—25 turns No. 30 enam. wire, close-wound on same type form as L_3.

L_5 15 turns No. 30 enam. wire, close-wound on same type form as L_3.

Results

The results to date have been pleasing. Although the converter has not seen much use on 10 meters, it appears to work well on that band. On 15 and 20 meters it easily holds its own with the 6BZ6-6U8A converter in the ARRL Handbook[2] as regards cross modulation. The overall gain of the FET converter is slightly less than its vacuum tube counterpart, but turning up the receiver rf gain remedies that. Any signal that can be copied with the vacuum tube converter can be copied with the FET converter.

. . . K6DQB

References

1. Harris, "Transistor High-Frequency Converters," **QST**.
2. "A Crystal Controlled Converter for 10, 15, and 10 Meters," The Radio Amateur's Handbook.

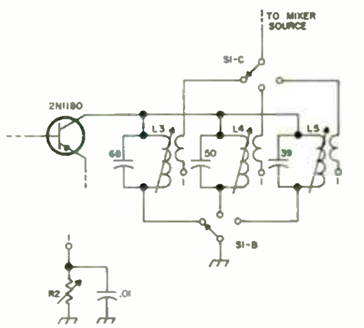


Fig. 2. Alternate circuit employing source injection. Coils L3, L4 and L5 are identical to those in Fig. 1, except that each has a three turn secondary.

CHAPTER 12

FET Converter for 40 & 160 Meters

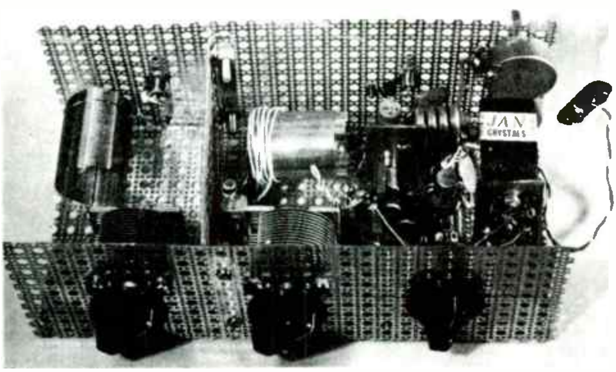
In a previous article , the author described a 10-20 meter FET converter which was to become part of the front-end of a 10-160 meter transistorized receiver. The results were sufficiently encouraging to warrant building a modified version, covering 40 and 160 meters. Eventually, the two converters will be combined with an 80 meter FET tuner, which is still in the breadboard stage, for coverage of the six high-frequency amateur bands.

Design

The converter schematic is shown in Fig. 1. A Motorola 2N4224 is used as the rf amplifier, while an MPF105 is used as the mixer. The local oscillator is a 2N1180. An attempt to use an MPF105 as the rf amplifier resulted in an unstable stage. Perhaps more attention to lead length and dress would have tamed the stage, but this was not investigated. A pair of 1N100 diodes is connected across the converter input to prevent excessive rf voltages from being applied to the 2N4224 while transmitting.

While there is some question as to the need for an rf amplifier at frequencies below about 10 MHz, one was included, principally to minimize local oscillator radiation by way of the antenna. However, if the converter is to be used with an *if* other than 80 meters (as described later), an rf amplifier may be helpful in reducing image response, depending on the *if* chosen.

The difference between the 10-20 and 40-160 meter converters lies in the rf amplifier, mixer, and local oscillator tuned circuit constants and the local oscillator injection frequencies. The values shown in Fig. 1 are appropriate to tune the rf amplifier between 1.8 and 7.3 MHz and to provide the injection frequencies required to heterodyne 40 and 160 meters to 80 meters. With the crystals specified, 160 meters is tuned between 3.7 and 3.5 MHz, and 40 meters is tuned between 3.7 and 3.9 MHz on the 80 meter *if*. Note that if 7.0 MHz is hetrodyned to 3.5 MHz, and the 40 meter band is to be tuned "frontwards", an injection frequency of 3.5 MHz is required, and the local oscillator will interfere with reception on the low end of the band.



Top view of the FET converter for 40 and 160 meters.

Photo by Joe Cohen

If you have been looking for a way to listen in on 160 meters, here's a converter for 160 and 40 which feeds into an 80-meter tuner. The 40-meter coverage is ideal for the novice whose equipment limits him to 80 meters.

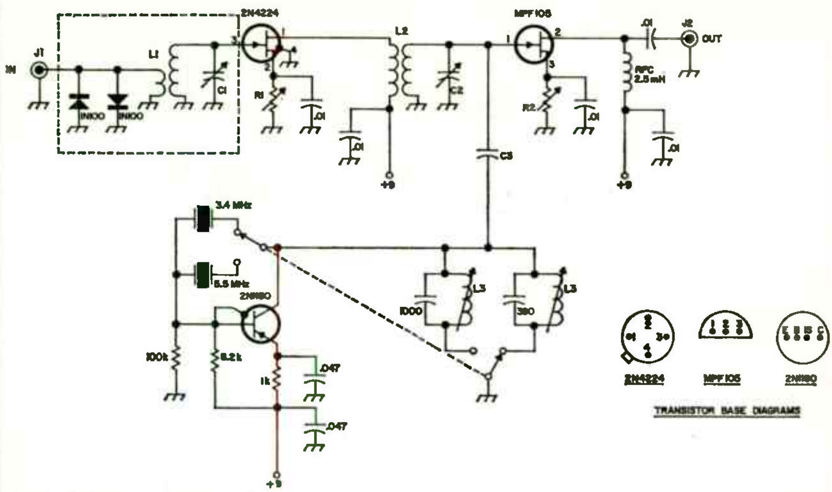


Fig. 1. Schematic diagram of the 40 and 160 meter converter. Field effect transistors are used in the rf amplifier and mixer stages, a bipolar device in the crystal-controlled oscillator.

Table I Local oscillator coil and crystal requirements for an *if* of 220 to 550 kHz

Frequency Range	Oscillator Coil	Tuning Oscillator Padder Capacitor	Injection Frequency
7.0 - 7.3 MHz	15 turns	330 pF	6.8 MHz
7.3 - 7.0 MHz*	15 turns	330 pF	7.5 MHz
3.5 - 3.8 MHz	36 turns	390 pF	3.3 MHz
3.8 - 4.1 MHz	36 turns	390 pF	3.6 MHz
2.0 - 1.8 MHz	26 turns	1000 pF	2.2 MHz

* This arrangement is recommended for the CW operator since it provides a higher *if* at the low end of the 40 meter band for better image rejection.

Use with *if*'s other than 80 meters

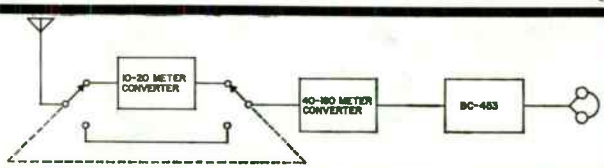
For several weeks the 10-20 and 40-160 meter converters were used with a BC-453 as the station receiver at K6DQB. The hookup used is shown in Fig. 2. In this case, an extra crystal(s) is used and 80 meters, as well as 40 and 160 meters is heterodyned to the *if* tuning range of the BC-453 — 220 to 550 kHz. The 80-meter range of the low-frequency converter is used as the *if* for the high-frequency converter for 10-20 meter reception. A table of local oscillator coil and crystal requirements is shown in Table 1. With the injection frequencies shown, coverage of 10 meters is incomplete. Two additional crystals, 3.9 MHz in the low-frequency converter, and 25.5 MHz in the high-frequency converter, would provide the necessary heterodyning frequency combinations required for full coverage. Since the author has not tried this, no values are specified in the table.

Results

As with the 10-20 meter converter, the results with this converter have been pleasing. In particular, susceptibility to cross-modulation appears very slight.

. . . K6DQB

Fig. 2. The two converter —BC-453 receiving arrangement.



CHAPTER 13

Single Transistor Converter

The following converter is your answer to a sensitive and inexpensive receiver for your car. When used with the usual car broadcast radio it will provide excellent monitoring of the two, six, and ten meter bands as well as police and fire frequencies on the low and high bands. Total cost is about three dollars with a well stocked parts box, or a kit can be obtained by writing Webber Labs, 40 Morris Street, West Lynn, Mass. If you choose to build your own you will need a crystal and a special high frequency transistor available as surplus from the above address for a total cost of $4.00.

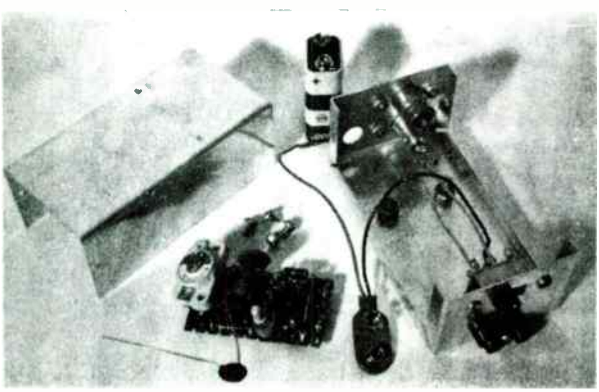
Two different versions are shown, tuning 28-54 mc and 100-200 mc respectively, the only difference being in Coil L1. If desired, plug-in coils may be used but due to the low cost of the parts it is easier to build two converters.

The unit is housed in a small 2"W x 4"L x 1⅝"D Minibox. Parts are mounted by their leads on a 1" x 2" piece of perf board and proper connections made on the underside of the board with the extra lengths of component lead. Because of the high frequencies involved wiring should be kept short and as much point to point as possible. Placement of parts is straight-forward and should cause no trouble if the pictorial is followed. Be sure to leave room at each corner of the board for the mounting holes.

When winding coils L1 and L2 use the specifications shown in the parts list. Because of the small size of the coils it may be easier to solder one end of the wire to a lead on the coil form and then turn the form between your fingers, feeding the wire as you turn. After every few turns or so, apply a small bit of candle wax to the completed part and heat the coil slightly to melt the wax in. This will prevent the wire from slipping off the form. When making the tap on L1 in the 28-54 mc model, be sure to continue the winding in the same direction as before.

Next, drill or punch holes in each end of the minibox and mount the antenna jacks, switch and crystal socket as shown. Drill holes in each corner of the perf board and mount it in the bottom of the box with small grommets or spacers to prevent the leads from shorting against the chassis. Finally, make the connections from the board to the mounted components on the box as shown in the pictorial.

In the authors' model a nine volt battery was used for power and taped to the outside of the OTC. Because of the small amount of power required by the circuit, the battery will last most of its shelf life. Power may be tapped from the car battery and R3 changed but this will result in more noise through the engine's electrical system.



A one transistor converter for 28-54 or 108-176 mc

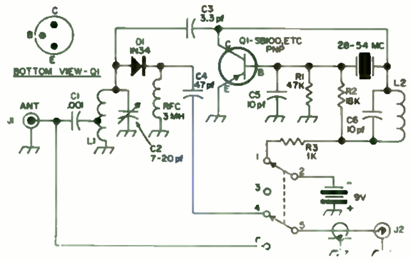


Fig. 1. L1 for 100-200 mc is 4 T #16 wire, ¼" diameter, tap at 1 T from ground. For 30-50 mc L1 is 36 T #30 on ¼" dia form ½" long, tap 11 T from ground. L2 is 36 T #30 on ¼" diameter form ½" long.

With a crystal plugged into the socket the converter is ready to go. To determine crystal frequency, use the following formulas:

For the 28-54 mc converter

$F_1 - 1.2 = f_c$

For the 100-200 mc converter

$\frac{1}{3}(F_1 - 1.2) = f_c$

F_1 is the frequency you wish to tune in megacycles

f_c is the oscillator frequency in megacycles.

The answer to the above formulas is the frequency of the crystal in megacycles. When ordering from Webber labs it is not necessary to figure the frequency but be sure to specify the frequency of the station you wish to monitor.

Plus in a jumper lead made of about twelve inches of coaxial cable with a male Motorola plug on each end into J2. Plug the other end into the radio and the antenna into J1. With S1 set in the off position turn the radio on and tune the dial. The usual commercial stations will be heard. If not, check the wiring of S1 and the antenna circuits. If all is okay, push S1 to the on position and tune the radio to 1200. A slight hiss will be heard and if the crystal frequency is correct you should hear a signal when the station for which you are listening transmits. It may be necessary to adjust C2 slightly to obtain maximum strength. By tuning the dial a wide range of frequencies may be heard. However, for stations which are far apart in frequency, different crystals will have to be used. The farther away from the crystal frequency the lower the sensitivity.

The converter may be mounted permanently under the dash or placed in the glove compartment of the car. It's about the easiest way to get good reception of these interesting frequencies for mobile use.

. . . WA4SAM, W1DVG

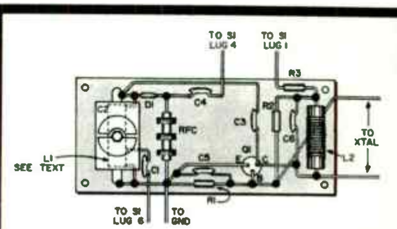


Fig. 2. Layout. The converter is built on a 1" x 2" piece of perforated board.

CHAPTER 14

IC-IF Strip

Price reductions brought about by increased production and improved manufacturing techniques have brought the use of integrated circuits within the financial grasp of every serious amateur. Most of the available units have been for digital circuits, but a new group of linear devices has been rapidly expanding. These circuits include almost all the more versatile circuits—audio amplifiers, video amplifiers, and most recently, rf amplifiers. Since these devices are often as low in price as discrete transistors, and performance is substantially better, it is advantageous to consider using them in home projects.

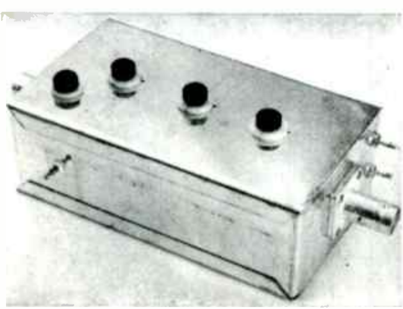
To see how easily the devices may be used and to provide a useful circuit application, a 30 MHz *if* strip was built using the Fairchild rf *if* amplifier, the μA03E. This device is a six-lead, epoxy, monolithic (single-chip) integrated circuit that has five transistors and two resistors connected in an emitter-coupled configuration. The amplifier can be used with transformers as interstage coupling elements; has all the biasing internally; and does not saturate for large overloads. Neutralization of the stages is not necessary because of the very low internal feedback of the circuit. Such an amplifier could be used with existing microwave converters or radar equipment.

Fig. 1 shows the schematic of the emitter-coupled amplifier and its associated bias network. The biasing of the amplifier may be understood by assuming that all parts within the circuit are well matched and the transistor current gains are high enough that the base currents can be neglected. It is also assumed that the transformer windings, particularly in the input circuit, have negligible dc resistance.

From Fig. 1 it can be seen that the current in the output transistor Q_4 will be approximately equal to one-fourth of the supply current. Since the emitter current of Q_5 is invariable and equal to one-half the supply current, any change in collector current of Q_4 will be reflected in the collector current of Q_3. In other words, the output current will be switched from zero to one-half the supply current. The output current, I_o, is shown as a function of input voltage, V, in Fig. 2.

Quantitatively, the collector current of the biasing transistor, Q_2, is substantially independent of the transistor characteristics and is simply a function of the supply voltage and a single resistor.

All of the transistors are assumed to be identical, hence the collector current of the



85 dB gain at 30 MHz with a 7.5 dB noise figure.

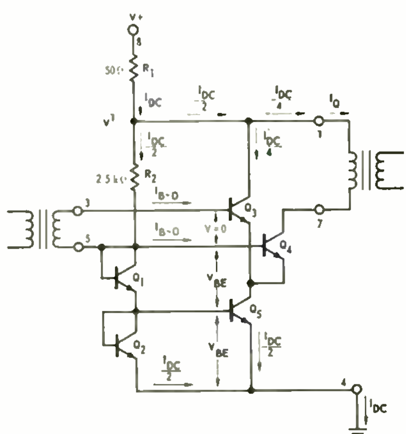


Fig. 1. Schematic diagram of the Fairchild μA703E integrated circuit, including bias currents.

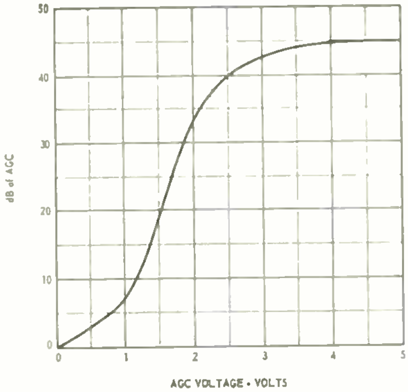
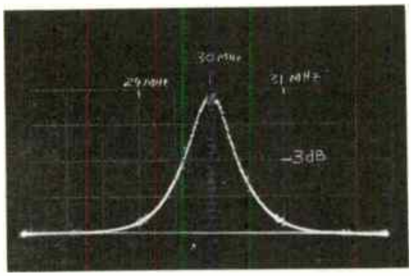


Fig. 2. AGC characteristics of the 30 MHz *if* amplifier.

current source, Q_5, is equal to that of Q_2 because their bases are fed from a common voltage point[1]. The collector current of Q_5 splits evenly between Q_3 and Q_4 with zero input signal. When the amplifier is driven, this current is alternately switched between Q_3 and Q_4. To prevent saturation of Q_4 when the amplifier is driven with a large signal, the load resistance must be low enough that current limiting occurs before the output voltage drops to $2V_{BE}$. For a transformer coupled output, the load resistance (R_L) must be less than or equal to $2R_2$.

This type of amplifier is referred to as a "differential pair" and is somewhat analogous to a seesaw—if one end of the seesaw is moved, the other end moves a corresponding distance but in the opposite direction. The biasing transistor, Q_5, acts like the fulcrum of the seesaw. (see Fig. 3).

A complete analysis of the μA703 is available as an application note entitled "Designing with the μA703 Monolithic RF *IF* Amplifier" from the applications group, Fairchild Semiconductor, Mountain View, California 94040.



Frequency response of the 30 MHz *if* strip. The —3 dB points in the response curve result in a bandwidth of 800 kHz. With a positive 12 volt supply, this amplifier provides 85 dB gain at 30 MHz with a noise figure of 7.5 dB.

Fig. 4 shows the schematic diagram of a 30 MHz *if* strip using the μA703E. This

1. "Some Circuit Design Techniques for Linear Integrated Circuits", Fairchild Technical Paper #33.

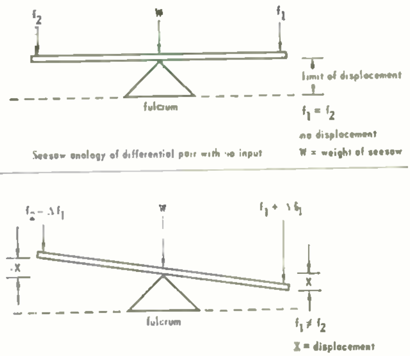
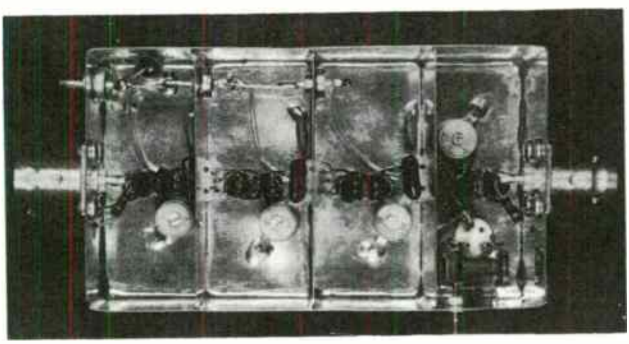


Fig. 3. Simple see-saw analogy of the differential pair with signal applied.



Below panel view of the integrated circuit 30 MHz *if* amplifier. Note the extensive shielding between stages and position of the toroidal interstage transformers.

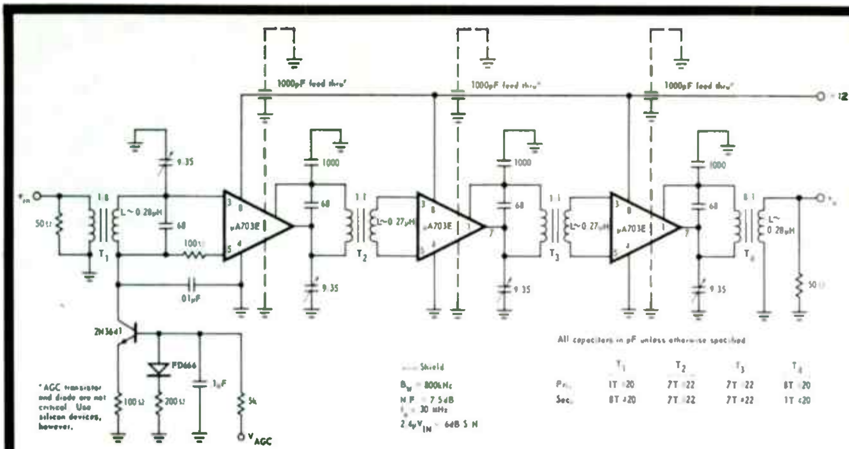


Fig. 4. Schematic diagram of the 30 MHz *if* amplifier. The interstage transformers are bifilar wound on cores similar to the T-37-10 cores available from Ami-Tron Associates.*

amplifier has a gain of approximately 85 dB, a noise figure of 7.5 dB and a bandwidth of 800 kHz. The input signal required for a 6 dB signal to noise ratio is 2.4 μV. Interstage transformers are bifilar wound on Micro-metals T44-10 cores, and have Q's of the order of 125. Placement of parts should follow the photo or be very close in order to realize the full gain of the amplifier without running into oscillation. In general, use care, and be careful, the results will be worth the effort.

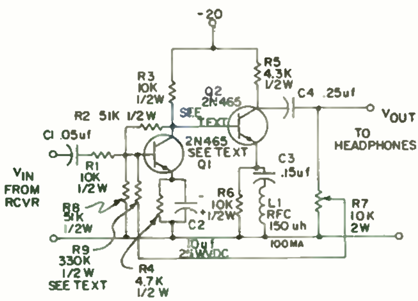
The author wishes to acknowledge the effort of Dave Capella, whose patience and ability were required to build and test the amplifier. ■

*Ami-Tron Associates, 12033 Otsego Street, North Hollywood 91607.

CHAPTER 15

Audio Bandpass Filter

Some electronics companies are now advertising that they use digital computers to design their filter networks. For the ham who needs a very selective audio bandpass filter, but does not possess an IBM 650 and can't afford ten bucks for a high quality audio inductor, the following equipment is described. It is a two-stage transistor audio amplifier with a series resonant LC circuit in the forward loop, and uses positive feedback to increase the Q of the cheap and dirty tank circuit to as high a degree as might be desired. Its input is plugged into the headphone jack of the shack receiver, and the op's phones plug into the amplifier output. At a bandwidth of 80 cycles it has a peak gain of 20 db, and for slow CW signals (and there are some on the novice bands) the bandwidth can be decreased to the limit of intelligibility. Just for fun, the author increased the feedback to just short of instability, and measured a 3db bindwidth of 3.5 cycles at a center frequency of 1070 cycles. Made the signal sound like someone "strummin' on the ol' banjo", but it just shows what the little thing can do.



The schematic shows the design to be fairly conventional, as far as the two amplifier stages are concerned. The power supply is a 22½ volt battery, current drain being only 2½ mils. The first stage is rather stiffly biased, and uses negative voltage feedback from the collector to the base. This serves to improve gain stability, and also to reduce the output impedance of the amplifier, thus improving the power transfer to the succeeding stage, which near resonance has a low impedance input. The emitter resistor of the first stage is heavily bypassed, which tends to make this stage have a low input impedance. Coupling from the input is through a capacitor and series 10K resistor, which simulates pretty well a head phone load on the communications receiver.

The base of the second stage is tied directly to the collector of the first, thus simplifying bias and coupling. Note that the emitter resistor of this stage is paralleled by a series capacitor and inductor (150 millihenry rf choke). At frequencies far removed from resonance of the LC combination, the emitter impedance of the second stage is essentially the emitter resistor, but near resonance the impedance of the series LC begins to drop. Thus, near resonance, the low series impedance of the tank predominates and the flow of base current is increased, assisted by the low output impedance of the first stage.

It will be noted that two 180 degree phase reversals take place in the amplifier, one in each stage. Thus, the output voltage, taken from the collector of the second stage, is in phase with the input voltage, and the feedback from the pot through the fixed resistor is regenerative. In essence, this means that as more signal gets through the amplifier, more is fed back, and since the gain of the amplifier increases near resonance, due to the series LC circuit, the output voltage rises sharply near resonance. This is what transforms a low-Q circuit into a highly selective audio amplifier.

Construction is straightforward. Parts layout is not in the least critical, and considerale latitude is allowed in the transistors. In fact, almost any low power audio transistor can be used. The only component which requires care is the fixed feedback resistor. Values should be tried until one is found which just produces oscillation when the pot is fully advanced to the maximum feedback position. Thus, when you're ready to operate, plug the amplifier into the phone jack of the receiver, the headphones into the amplifier output jack, and advance the regeneration control until a whistle is heard. Back off a bit, and you're in business. . . . Hansen

CHAPTER 16

Six-Meter FET Converter

Field effect transistors (FET) seem to be the answer to converter design for the 50 MHz amateur band. The cross-modulation problems common with ordinary transistors and even with tubes are no longer a real headache when using these new transistors. Ordinary transistors are subject to overload and cross-modulation with more than about 20 millivolts input which means that local stations can ride in on weaker signals anywhere in the amateur band. An rf stage ahead of the mixer even with tubes (less overload characteristics) will usually amplify a local station 100 KHz or so away from the desired signal enough to cross-modulate it in the mixer stage. FET types of transistors as mixers have extremely good characteristics for reducing cross-modulation and will even permit the use of an rf amplifier in most locations. Ordinary transistors and even some tube mixer types will often overload enough with an rf stage circuit to make them useless in some locations.

The FET units have been expensive for use in the vhf region and often have exhibited poor noise figure values. The writer recently obtained some new FET plastic-cased transistors for approximately one dollar apiece from a Texas Instrument distributor. These were TIM12 units which have very low NF and good gain values at 50 MHz. A circuit of a good 50-MHz converter is shown in Fig. 1 and illustrated in the photographs. The converter was built on a scrap piece of copper-plated board 1½ x 6 inches for mounting into a 6 inch wide aluminum chassis. The noise figure measurement between 50 and 52 MHz was from 1.5 to 2.5 dB. This is very low and means that in nearly all locations, antenna noise pick-up will completely override the receiver noise.

The cross-modulation capability was checked by connecting two signal generators to the input jack. One signal generator was connected to the converter input thru a 10 dB pad and, with no modulation, was set to give an S5 or S6 signal reading in the *if* receiver when the whole system was tuned to this signal frequency. Then another tone modulated signal generator was turned on at about 1 MHz off frequency (connected directly to the converter) and its output attenuator adjusted until some over-riding tone modulation was heard on the cw signal generator. It took more tone signal than could be obtained thru the attenuator which was supposed to have 100,000 microvolts maximum output.

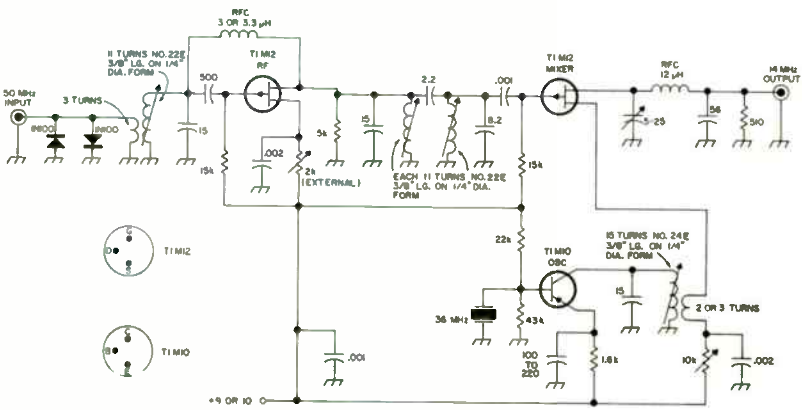
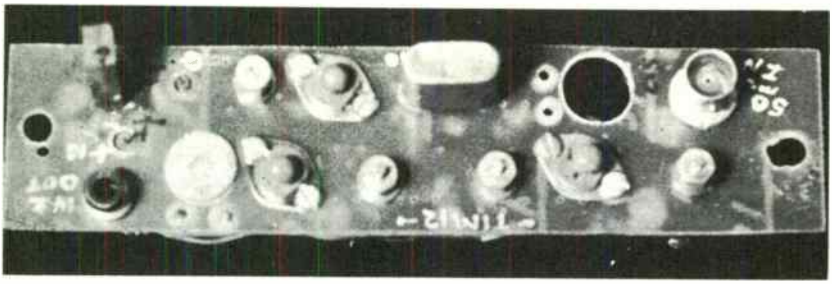


Fig. 1. 50 MHz converter using field effect transistor rf amplifier and mixer. The FET's cost about $1 each. This converter has a noise figure of around 2 dB and great resistance to cross-modulation.



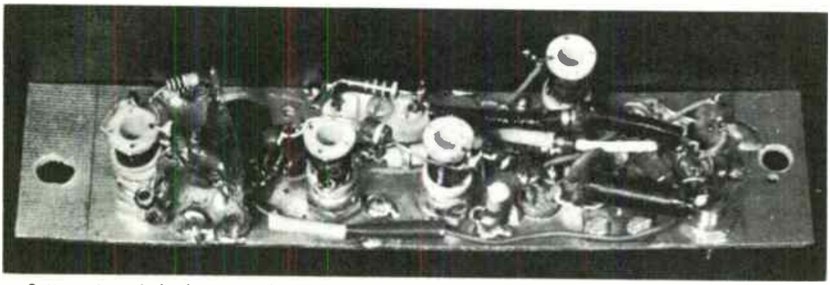
Top view of W6AJF's six meter FET converter. The extra hole by the BNC input jack was used for neutralizing trimmer that proved unnecessary.

The "one volt" output jack produced appreciable cross modulation. It was estimated that it took about ¼ volt to produce objectionable cross-modulation. It was necessary to have a large resistor pad between the converter and the *if* receiver, and to have the two test signals separated far enough apart so the cross modulation problems in the *if* receiver were negligible. It is surprising how poor some homemade and some commercial radio receivers are for cross modulation in the 14 MHz region. It would seem that FET transistors should be used in all 14 to 18 MHz and 5 or 2 MHz if and mixer stages right up to the main sharp mechanical or crystal filter in the *if* section. A 20 dB pad on the *if* receiver input helped to reduce these effects while trying to check the converter only. The added pad or attenuator was only a stopgap cure since the real cure is to use a better designed *if* receiver.

Surprisingly, the IN100 back-to-back diodes in the receiver input were not troublesome in these cross-modulation tests. These diodes are standard on all my converter inputs in order to provide some transistor protection from moderately high powered transmitters at this station. The 1N100 diodes have a low capacitance, reasonably high back resistance and quite low forward resistance and are low cost types. Connected across the coax input jack, the diode loss is very low and it does provide some added protection against destructive surge voltages from the antenna system or switching relay.

The converter rf stage required some neutralization by means of a 3 or 3.3 microhenry rf choke connected between the input and output tuned circuits. This resonates roughly at 50 MHz with the gate to drain capacitance of a TIM12 which is typically about 3 pF. Even with this amount of inductive neutralization it was necessary to load the tuned input circuit down to quite a bit less than 1000 ohms by means of the antenna link of three turns. The FET has high input and output impedance and a 5000 ohm resistor across the output tuned circuit was also needed. A variable source resistor of 2000 ohms was mounted external to the converter to permit easy rf gain adjustment.

The FET mixer stage in this unit has gate signal input and source oscillator injection. A small Trimpot, 0 to 10,000 ohms, provides bias for the mixer stage. This pot and the oscillator pick-up link of 2 to 3 turns were adjusted to provide minimum



Bottom view of the low noise, low cross-modulation FET converter. The copper shield is across the rf amplifier socket. The solenoid rf choke at the other end is part of the pi network output circuit.

cross-modulation effects. Actually a 2-kΩ or 3-kΩ fixed resistor would be quite satisfactory for this type of transistor and oscillator injection voltage. The latter is greater than with ordinary transistor mixers, but should be a little less than that which gives maximum mixer gain. At the maximum gain value, the cross-modulation effects are worse. The mixer output circuit is a pi coupling network tuned to about 15 MHz. The dc path resistor across the output jack can be made much lower in value if a wider *if* frequency response is needed. The value will be somewhere between 50 and 500 ohms for most *if* receivers. If the latter actually looks like 50 to 70 ohms, the dc shunt resistor can be of a higher value.

The 36 MHz crystal oscillator uses a TIM10 or any other VHF transistor which will produce strong 36 MHz output with one or two milliamperes of collector current. The emitter bypass capacitor produces regeneration and its value will usually range between 100 pF and 220 or even more for most types of PNP transistors. The FET TIM12 units are P-channel which is similar to PNP transistors for battery supply polarity. Some FET units are N-channel which require the same supply voltage polarity as NPN transistors. The TIM12 has an odd base arrangement of leads (see Fig.1) as compared to ordinary transistors. This can cause some confusion in wiring up the transistor sockets and requires a little care in checking over the circuit wiring before fixing up the converter.

As a final comment, this converter showed a 25 to 30 dB improvement in cross-modulation as compared to several other 50 MHz converters using ordinary vhf transistors of several types. It also had a better NF than the other converters. The spurious signal responses were less due to the FET mixer.

. . . W6AJF

CHAPTER 17

Simple 432-MHz Converter

The new Texas Instrument TIXMO5 transistors were rated for operation at 200 mc but are surprisingly good at 432 mc. These units compare very favorably with transistors costing many times as much, and at about 50 cents apiece, a few extra can be bought in order to get some very choice ones for the front end of a 432 mc converter. The writer found that about one out of every three were red hot for 432 mc operation and the other two out of three were still better than other $3.00 types generally used at 432 mc. The only problem in their use is mechanical breakage of these plastic cased units. This can be minimized by using the new transistor sockets made for type TO-18 cased transistors since the three leads do not have to be spread out as when using the larger (TO-5 type) sockets. The writer managed to break a few transistors in the first converter built here so a new one, shown in the photographs, was built with the new smaller sockets. No more breakage was encountered but it is a little upsetting to pay as much for a socket as for a transistor. Direct soldering of the transistors into the circuit might be an alternative but makes it hard to select the lower NF transistor for the first rf stage. This arrangement was used finally in the first converter in place of the large sockets though one transistor was damaged in the process of soldering. The small but expensive sockets are really the best solution.

432 mc antenna filter

In Fig. 1, a dual circuit antenna filter is shown which was built into an aluminum chassis box 12 x 2¼ x 2½ inches in size. This unit works reasonably well for receiving and for transmitting at low power perhaps up to 50 watts output. With a pair of 4CX250R tubes in the final amplifier, the aluminum box gets hot at the low impedance ends and circuits go out of resonance. The circuits would stay in resonance for a minute or so with 400 to 500 watts of rf power then the box gets warm to the touch and losses go up as the circuits go out of resonance and dielectric losses also increase. Then end result was retirement of this antenna filter to the receiver front end only since its losses, when cold, are approximately one db. The signal loss then is not too objectionable for normal 432 mc signal reception. The next project at W6AJF will probably be a heavy duty dual coaxial circuit unit built of copper since the writer likes to use high power at 432 mc occasionally. The filter in the transmitter output is highly desirable to nearly eliminate lower and higher order frequencies from getting into the big antenna array. More TVI problems have been encountered here on 432 mc than on 220 or 144 mc band operation, so a good antenna filter is needed. Solid state stereo phonographs and FM band receivers increase the "TVI" problem for many amateur VHF operators.

432 mc converter

At 432 mc, common base rf stages are usually much easier to get into proper operation than with neutralized common emitter systems. The converter shown in Fig. 2 and in the photographs uses two common base rf stages with forward gain control to set the gain just below oscillation or high regeneration operation. The advantages of forward gain control have been discussed in previous sections on transistor converters. The mixer stage seemed

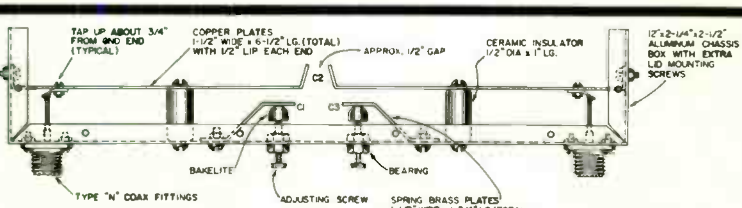
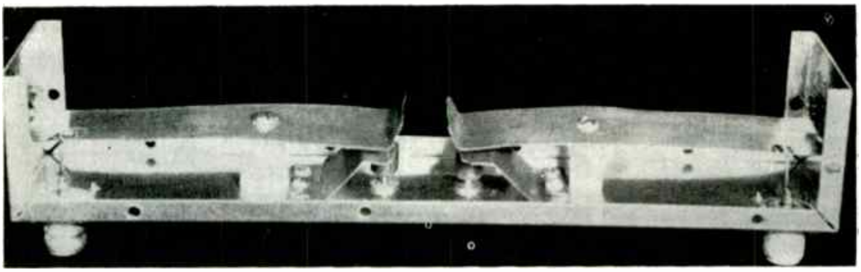


Fig. 1. 432 mc antenna coupler-filter suitable for receiving and low power transmitting use.



432 mc antenna coupler-filter.

to function best with signal and oscillator injection into the base circuit. Emitter injection was tried but resulted in mixer oscillation due to the added inductance in the emitter lead at 432 mc. The mixer output circuit in Fig. 2 is a simple tuned circuit since only about .5 mc bandwidth was needed near 14 mc, the if output. If wider frequency coverage is desired, the pi network used in the 144 mc converter, previously illustrated, will cover 4 or 5 mc but with somewhat less mixer gain.

The oscillator uses a 46.444 or a 139.333 overtone crystal with the collector circuit tuned at or near 139.3 mc. The emitter bypass condenser, a small 5 to 25 pf adjustable condenser, permits either type of crystal to be used. It is set for best oscillation in either case. The coupling condenser to the "fast" diode tripler was made adjustable in order to achieve optimum load on the oscillator and maximum output from the diode tripler stage. The 1N914 is a fairly low capacitance high speed computer diode (silicon type) which is better than a 1N82A as a frequency doubler or tripler to 418 mc. It is also suitable for use as back to back protective diodes in the front rf stage though the 4 pf shunt capacitance per diode means some coil turn juggling to keep the input tuned to 432 mc. Type 1N100 diodes were used in the unit shown here since they seemed to have less shunt capacitance and antenna relay leakage from the transmitter wasn't too severe at W6AJF. The coax relay was a hard to find type more suitable for operation in the UHF region than the standard Dow relays available for lower frequency operation at W6AJF. Type N fittings are more efficient at 432 mc and the Dow relays at this station all had the other type of fittings which are suitable for the VHF bands but not as low a SWR rating at UHF.

The grounded base input rf stage has a tendency to oscillate with a change from one antenna to another or to a signal generator. A small variable condenser was shunted across the coax input jack and just enough capacitance added to stabilize this stage. The input coil still has to be adjusted for best NF with a noise generator. The second rf stage will likewise oscillate if the emitter coupling coil is too far away from the 432 tuned copper strap circuit. Too loose coupling will also make the preceding tuned collector circuit working Q value too high and tend to make the converter tune too sharply and be too regenerative. The rf gain control will not function properly if

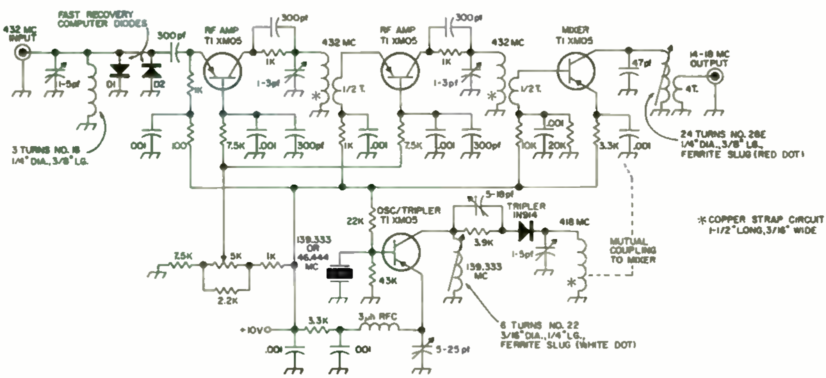
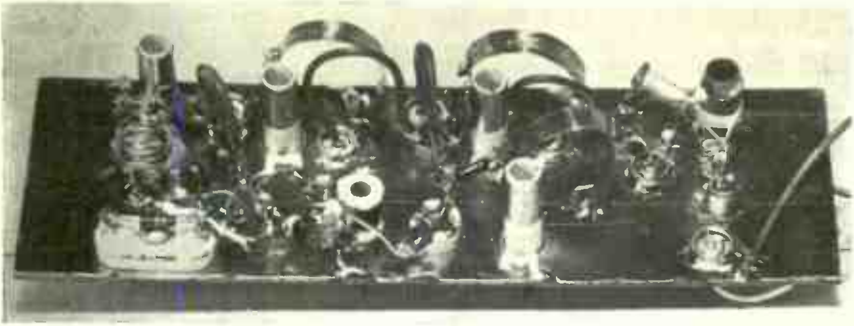
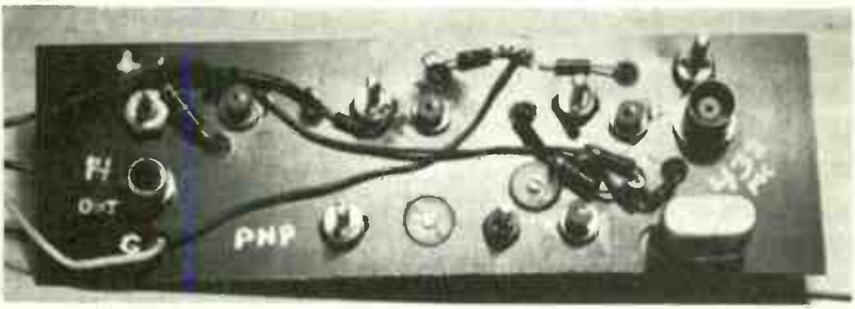


Fig. 2. 432 mc low noise converter using 50¢ TIXMO5 transistors.



Bottom of the 432 mc converter.

the two rf stages are not loaded correctly and are excessively regenerative. Any 432 mc converter takes a little time and care in adjustment for best weak signal reception. When you get a 432 mc converter into such good operation that occasional auto ignition noise is very noticeable it is probably red hot for 432 operation. It takes a 432 paramp to do better on 432 weak signals. The measured NF of this converter was 4 db which is better than any other transistor converter (except an expensive 2N2857 unit) tested here. It was definitely superior to grounded grid nuvistor and 416B converters tested here on the same noise generator. It would seem that solid state devices are really here to stay. The 4dbNF measured here after a long period of adjustments, is a relative figure since some other noise generator might read a 2 or 3 or 5 db figure. Noise measurements above 200 mc can be very individualistic but are still useful in getting the best NF possible from a converter in the UHF region.



Top of the low noise 432 mc converter.

CHAPTER 18

Two-Meter FET Converter

In the last year or two several bipolar transistors (conventional junction transistors) have been introduced which provide excellent VHF characteristics. A low noise VHF converter using these devices will be lower in cost, provide better performance, and use simpler circuits than a vacuum tube converter. Bipolar transistors costing only a dollar or two will produce noise figures as low as dB at 144 MHz. These devices share one serious shortcoming, however. They are very susceptible to cross modulation.

Cross modulation

One feature that all bipolar transistors and diodes have in common is a transfer function in which the current flow in a forward biased pn junction, e.g. the base-emitter junction of a transistor, is proportional to $e^{qv/kt}$, where q is the charge on an electron, k is Boltzmann's constant, v the applied voltage, and t the Absolute temperature. $e^{qv/kt}$ can be expanded into a series as

$$e^{qv/kt} = 1 + qv/kt + \frac{(qv/kt)^2}{2!} + \frac{(qv/kt)^3}{3!} + \dots$$

The even order terms produce harmonics, even order combination frequencies, and dc terms. These products are usually not troublesome. It is the odd order terms which cause cross modulation.

At room temperature, 290°A, kt/q is equal to 26 millivolts. For voltages v greater than 26 mV the exponent qv/kt has a value greater than unity, it can be seen that the third order term increases rapidly when qv/kt is greater than unity. If v is composed of two voltages $E_a \sin\omega_1 t + E_b \sin\omega_2 t$ a little algebra will show that there will be a component at ω_1 with amplitude proportional to E_b and a component at ω_2 with amplitude proportional to E_a. This causes cross modulation.

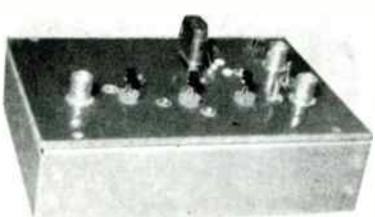
The action of the fifth and higher order odd terms is similar to that of the third order term.

Therefore, we can conclude that with bipolar semiconductors applied voltages around 26 mV. will result in serious cross modulation. It is important to note cross modulation can occur not only in the first stage of a receiver, but in any stage where two or more large amplitude signals are present.

The designer of a receiver front end is faced with a serious problem. A highly selective filter which would prevent all but one signal from reaching a transistor or diode will be a lossy device. In order to have a noise figure as low as possible it is desirable to precede this filter with enough gain to make its effect on noise figure negligible. However, this gain may result in the whole 2 meter band being amplified at once by 30 dB or more, and large voltages may easily result when strong signals are present. Some readers may have had trouble with FM and TV stations showing up in their transistor converters or receivers. It is usually insufficient front end selectivity and resultant cross modulation which are responsible.

The FET at VHF

The field effect transistor has an almost perfect square-law transfer function, i.e. odd order terms are almost nonexistent. The result is greatly improved cross modulation performance. In addition, recently introduced FETs are capable of very low noise performance in the VHF and UHF range. Noise figures as good as any vacuum tube or bipolar transistor are readily obtainable.



This converter uses inexpensive field effect transistors as mixer rf amplifiers. It has a noise figure of less than 2.5 dB and very low cross modulation.

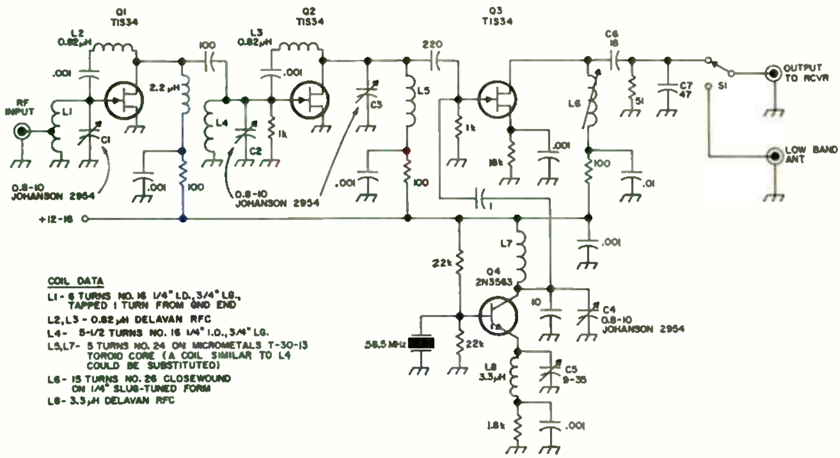


Fig. 1. Schematic of K6HMO's low-noise, low-cross modulation two meter converter using inexpensive field effect transistors (FET's). This converter can give a noise figure of under 2.5 dB, gain of 27 dB and very good rejection of cross modulation products.

The device selected for this converter, the TIS34, is made by Texas Instruments and is encased in an economy plastic package. The price is $2.80. The TIS34 is very similar to the 2N3823—it appears to be the same chip—which has been available for some time for about $12. The data sheet for the 2N3823 is much more complete than the one for the TIS34 and it was used in the design of this converter.

Construction

The FET converter shown in Fig. 1 was constructed on a piece of printed circuit material cut 4x6 inches. The completed converter was mounted in an upside down 4x6x1½ inch aluminum.

The circuitry is very conventional and makes use of single-tuned transformers for impedance matching and to provide the required selectivity.

The noise figure test for the TIS34 specifies a source impedance of 1000 ohms. This is obtained with the tapped input inductor. The drain load impedance is simply the parallel combination of the real part of the FET input impedance and R1. The imaginary part of the input impedance is tuned out with L4 and C2. L2 is a standard rf choke used for neutralization by resonating with the drain-gate capacitance at 144 MHz. The second stage is essentially identical to the first. The rf amplifiers are simple, stable, and use very few components.

The design of the mixer was almost entirely empirical because of the lack of large signal data. The best compromise between noise figure and gain resulted when the FET was biased to a few hundred microamperes with the source resistor, (remember cathode bias) and then driven on with the local oscillator power. Better performance was obtained with the available local oscillator power when the local oscillator was introduced at the gate rather than at the source. The mixer output circuit is resonant at 28 MHz. The 51-ohm resistor terminates the receiver used with the converter and with L6, C6, and C7 determines the drain load impedance of Q3.

Briefly, L8 and C5 are resonant between the fundamental and the third overtone frequency. In this case L7 and C4 resonate at twice the third overtone frequency at 117 MHz. This results in an if frequency of 27 to 31 MHz. Other if frequencies could be used merely by selecting another crystal frequency and scaling the mixer output circuit to the desired frequency. For example, for 14 to 18 MHz use a 65 MHz crystal and double L6, C6, and C7.

The switch at the *if* output is used to switch the communications receiver between the converter and a low frequency antenna.

Performance

The completed converter has a noise figure of just under 2.5 dB measured on a Hewlett Packard Noise Figure Meter. The gain from 144 to 28 MHz was measured and found to be 27 dB. Cross modulation measurements were made in order to compare performance with a conventional bipolar transistor converter. The setup shown in Fig. 2 was used

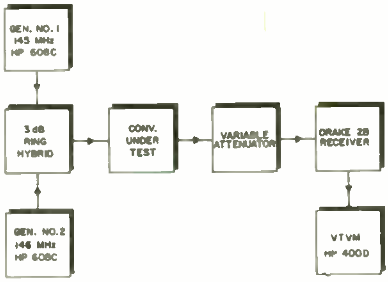


Fig. 2. Test setup used for cross modulation measurements.

for these measurements. The attenuator was included in order to assure that cross modulation observed was not produced in the communications receiver.

The converter-receiver combination was tuned to generator #1 at 145 MHz. Generator #1 was modulated 30% with a 1000-Hz tone, and the output level to the converter was 5 microvolts. The modulation on generator #1 was then turned off. Generator #2 was tuned to 146 MHz and modulated 30% with a 1000-Hz tone. The output level of generator #2 was then increased until the signal from generator #1 appeared to be modulated 1% (30 dB down on the VTVM). In ordinary use 1% represents a just detectable case of cross modulation.

This test was performed on both the converter described here and another using bipolar transistors. The results are shown below along with other comparative measurements:

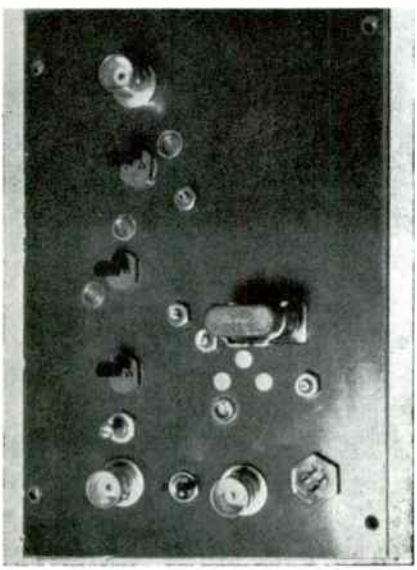
	Bipolar	FET
Noise Figure	3.1 dB	2.4 dB
Gain	24 dB	27 dB
Rf input for 1% cross mod.	0.5 mV	27 mV

As can be seen the FET lives up to the claims made for it. 27 mV at the input means about 1 volt at the gate of the mixer where the cross modulation should be occurring. This improved performance means that fellow down the street would have to increase his power 2916 times in order to produce the same amount of interference.

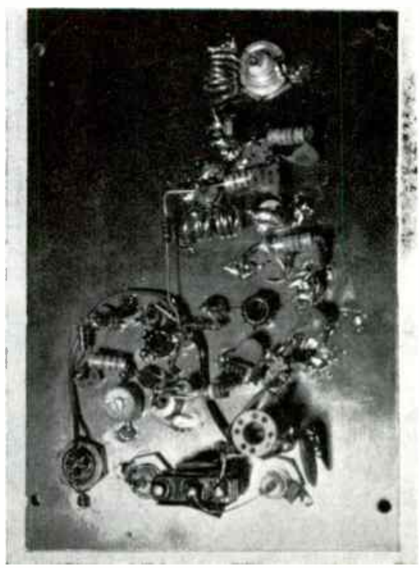
On the air tests at the author's shack have confirmed the improvement in cross modulation resistance. With the bipolar converter, the modulation of a local repeater could be heard on every other signal on the band. With the FET converter, no detectable cross modulation has been observed from the repeater or any other local station in several months of operation.

Miscellaneous

Power for the converter here is supplied by a transistor transmitter used with it. There is plenty of room, however, to build a power supply into the converter. A suitable power supply is shown in Fig. 3.



Top view of the FET two meter converter.



Bottom view of the FET two meter converter.

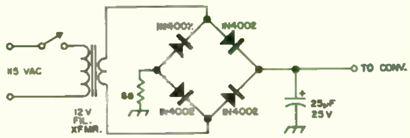


Fig. 3. Power supply suitable for use with the FET converter.

The trimmer capacitors used here are quite expensive. They sell for more than the FETs. Almost any other good quality type would be a suitable substitute. Small ceramic trimmers could be used at a considerable saving.

For those who have an especially acute cross modulation problem, and who are willing to experiment, even better cross modulation performance than that obtained with this converter could be obtained with the common gate configuration. The common gate configuration is similar to grounded grid in a vacuum tube. The noise figure will be about 0.5 dB higher and the gain a little lower, but because of the lower impedance levels the voltages will be lower and remember it is the voltage that leads to cross modulation.

. . . K6HMO

CHAPTER 19

A Converter for 1296 MHz

The converter is intended for ham communications on the 1296 mc band and could easily be adopted for moon bounce operation. An if frequency of 30 mc permits the use of any communications receiver capable of tuning this range.

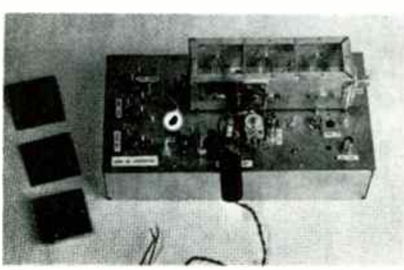
Six transistors are used in functions previously performed by tubes. The use of transistors reduces the amount of heat generated both by the *if* stages and harmonic generators, thereby increasing frequency stability. A diode is used as a frequency tripler in the last stage of the frequency multiplier section. A low noise mixer diode is used in conjunction with a low noise *if* preamplifier (2.7 db) to provide extremely low noise operation. This converter has a measured overall noise figure of 8.5 db with a power gain of better than 50 db. The image response was at least 10 db down from the signal frequency.

Circuit Description

Fig. 1 shows the schematic diagram of the entire converter.

Philco's 2N1742 MADT transistors are utilized in the oscillator and the three frequency doubler circuits. All are operating in the common-base configuration. A 1N147 type mixer diode is used as the diode frequency tripler. The 2N1744 MADT type which is specified for oscillator service was considered for the multiplier stages but was found to have lower power gains than the 2N1742 which due to its higher gains provided more drive to the diode tripler stage.

The 1266 mc heterodyning signal for the 1N263 diode mixer is generated by a 52.75 mc crystal controlled oscillator, three Class "C" frequency doublers and a diode frequency tripler.



Top View of 1296 mc Converter.

In the oscillator output circuit, capacitor C_1 and coil L_1 are tuned to 52.75 mc. The 1.5 to 7.0 mmfd trimmer in the emitter circuit is adjusted for maximum output consistent with stable oscillator operation. With certain crystals it may be necessary to add about 1.0 mmfd of capacitance between the collector and emitter terminals. Winding L_2 couples the output from the oscillator to the input of the first doubler. Capacitor C_3 and coil L_3 tune the output of this stage to 105.5 mc. A tap on coil L_3 couples the 105.5 mc output to the emitter of the second doubler. Capacitor C_4 and coil L_4 tune the output of this stage to 211 mc.

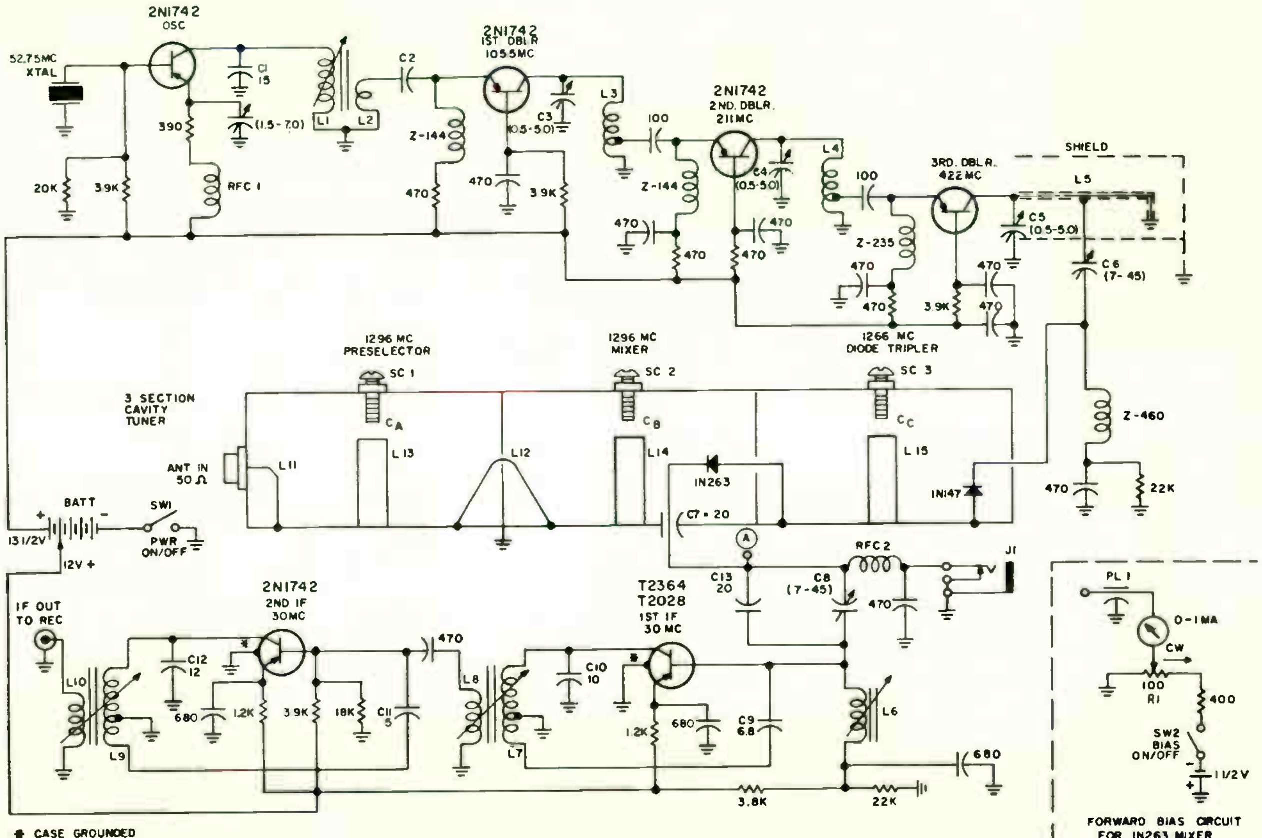
A low impedance tap on coil L_4 couples the 211 mc output to the input of the third doubler. A strip line tank circuit consisting of capacitor C_5 and coil L_5 tune the output to 422 mc. A copper type shield serves as the enclosure for this tank circuit.

Driving of the 1N147 multiplier tripler diode to 1266 mc is through the variable matching capacitor C_6 tapped onto L_5. Inductor L_{15} with the capacitance C_e formed by the tuning screw SC_3 is tuned to 1266 mc.

Inductors L_{13} and L_{15} and capacitors C_A and C_B formed by the tuning screws SC_1 and SC_2 respectively, are preselectors tuned to 1296 mc. Loops L_{11} and L_{12} couple the signal from the antenna through the preselectors to the 1N263 diode mixer.

The 30 mc output of the mixer is amplified in two common-emitter *if* stages. Coupling to the T2364 first *if* amplifier is through a network consisting of coil L_6 and capacitors C_7, C_8 and C_{13}. C_8 and L_6 are adjusted for the lowest noise figure. The *if* system noise figure was measured to be 2.7 db. A T2028 can be used in this stage with a slight deterioration in the noise figure. Capacitor C_{10} and coil L_7 are tuned to 30 mc. Inductive neutralization is provided by the tapped output coil L_7 and capacitor C_9.

Secondary winding L_8 provides the coupling to the 2N1742 second *if* amplifier. Capacitor C_{12} and coil L_9 are tuned to 30 mc. Inductive neutralization is obtained by the tapped output coil L_9 and capacitor C_{11}. The converter



* CASE GROUNDED
.ALL CAP VALUES IN mmf.

FIG. 1

output is coupled to the input of the receiver through secondary winding L_{10}.

Application of forward bias to the 1N263 diode mixer gives about a 3 db improvement in noise figure. Jack, JK_1 and plug, PL_1 provide the means of introducing the forward bias in the mixer circuit. R_1 and the 1266 mc diode tripler tank circuit are adjusted for best noise figure. The 1N263 mixer current is in the range of 300 to 400 μa with forward bias and about 200 μa with the bias removed. The mixer can be operated without forward bias if the loss in noise figure can be tolerated. After aligning the converter for maximum gain the input preselectors may need to be trimmed for the minimum noise figure.

Chart 1 indicates the approximate current drawn by the individual stages. The current that flows in the multiplier stages is dependent upon driving power and tuning.

Osc.	1st Dblr.	2nd Dblr.	3rd Dblr.
5.0 ma	3.0 ma	5.0 ma	2.0 ma

1st IF	2nd IF	Bleeder Current	Total Current
0.6 ma	1.7 ma	1.2 ma	18.5 ma

Alignment Procedure

The *if* system should be first tuned to 30 mc. This is done by connecting the output of the *if* section to the antenna terminals of the 30 mc receiver and applying bias. Modulated output from the signal generator is inserted at point A. The mixer diode is removed from the circuit during this part of alignment. Adjust the coils of L_6, L_7 and L_9 for maximum output on the receiver.

The multiplier section can best be adjusted by use of a grid dip oscillator. With the GDO operating as a wave meter, it should be coupled to the oscillator tank coil and core of

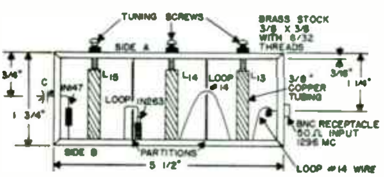


FIG 2

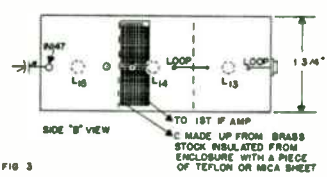
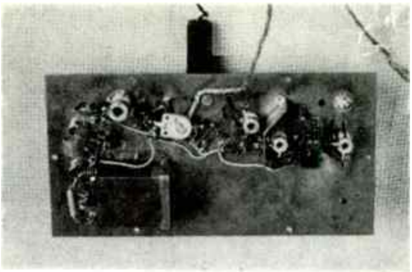


FIG 3



Bottom View—Note the copper enclosure for L5 in upper right hand corner.

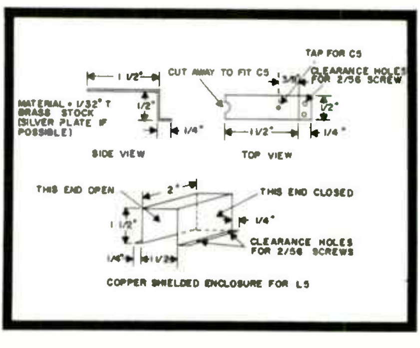
L_1 adjusted for maximum output. The variable trimmer (1.5-7.0 mmfd) in the emitter circuit should also be set for maximum output. With the wave meter set to 105.5 mc and coupled to L_3, adjust capacitor C_3 for maximum indication.

The wave meter is next set to 211 mc coupled to L_4 and C_4 adjusted for maximum indication on the wave meter. Finally, with the wave meter coupled to L_5 at a frequency of 422 mc, adjust C_5 for maximum indication on the wave meter.

By connecting a 0-5 ma meter across the 22K resistor in the dc return for the 1N147 diode tripler, all of the multiplier stages and capacitor C_6 can be trimmed by tuning for a maximum indication on the meter. Remove the meter after this alignment.

The 22K resistor in the 1N147 bias circuit develops back bias or self bias for the diode. The multiplying action is more efficient when the diode is biased in this manner.

With the antenna connected to the input terminal on the cavity tuner, adjust tuning screw SC_3 and C_6 for maximum mixer current as measured at Jack JK_1. With a 1296 mc signal applied to the input connector, adjust tuning screws SC_1 and SC_2 for maximum output. If a noise generator is available, apply to antenna input and adjust all of the tuning



screws and potentiometer R_1 for minimum noise figure. Only slight adjustment of the tuning screws is necessary to optimize the noise figure.

Chart II indicates the approximate power output delivered by the oscillator and the succeeding multiplier stages.

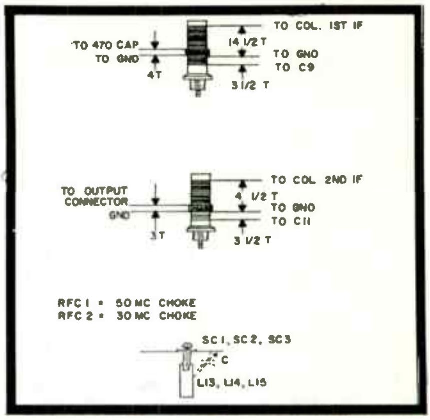
Osc.	1st Dblr.	2nd Dblr.	3rd Dblr.
10 mw	14 mw	9 mw	6 mw

Construction Data

The entire converter was constructed on a 9½" by 5" brass plate 1/16" thick. Figs. 2 and 3 show the details of the cavity tuner. Silver plating of the cavity tuner is advisable but not essential. . . . W3HIX

Coil Data

L1 7 turns #28 enam. coper wire space wound to occupy 5/16" on ⅜" dia. from slug tuned with powdered iron core.
L2—4 turns #28 enam. copper wire interwound with low potential end of L1.
L3—5¾ turns #18 tinned copper wire ½" I.D. air wound to occupy ½" output tap ⅓ turn from ground end.
L4—5¼ turns #18 tinned copper wire ¼" I.D. air wound to occupy 7/16".
L6—13 turns #30 enam. copper wire spaced to occupy ½" winding area on ¼" form with powdered iron core (VHF Grade).
L7—18 turns #30 enam. copper wire close wound on ¼" form with powdered iron core, ground tap 14½ turns from collecter end.
L8—4 turns #30 enam. over low potential end of L7.
L9—18 turns #30 enam. closewound on ¼" form with powdered iron core (VHF Grade) ground tap 14½ turns from collecter end.
L10—3 turns #28 enam. wound over low potential end of L9.
L13, L14, L15—1 7/16" long ⅜" dia. copper tubing tuning screws SC1, SC2, SC3 enter the hollow copper tubing to form tuning capacitors as noted below.



CHAPTER 20

FET Pre - Amp

In recent months, numerous articles have described VHF rf amplifiers using encapsulated FET's. Unfortunately, they are usually unavailable at the local electronics store. This article is about an FET that's not quite so cheap, but which is available off the shelf at any Union Carbide transistor distributor. The only drawback, if there is one, is the cost of the 2N4416. They were $7.10 apiece in February, but since have come down to about $5.00 each in small quantities.

The manufacturing data sheet for the 2N4416 says that it has a gain of 18 dB with a noise figure of 2 dB (maximum) at 100 MHz, tapering off to a gain of 10 dB and a NF of 4 dB (maximum) at 400 MHz. There is also a set of curves that indicates that the average 2N4416 is about 1 dB better than the specification.

Almost as soon as the development of the transistor became known to the general public around 1950, amateurs started working on ways and means of pulling the tubes out of existing equipment and replacing them with these new and more efficient devices. The main difficulty was that the conventional bipolar transistor is a low impedance device and does not readily fit into circuits with inductances wound to accommodate high impedance circuits such as the grid of a pentode. The advent of the FET has changed all this.

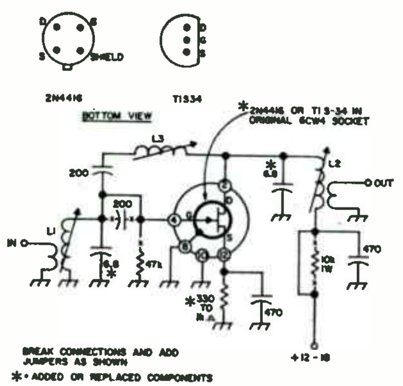


Fig. 1. Installation of the 2N4416 or TIS-34 FET in a two-meter nuvistor converter. The 330 ohm source resistor shown here is a typical value only—the correct value is determined experimentally for the proper amount of drain-to-source current and best noise figure as explained in the text.

144 MHz preamplifier

Recently two 2N4416's were obtained and plugged into an existing International Crystal 144 MHz preamplifier in place of the original nuvistor. Three simple wiring changes were actually needed and one more was added as a slight improvement.

Fig. 1 shows the International Crystal circuit with the added jumper wires. The N-channel FET can be plugged right into the nuvistor socket if care is taken to get the right leads on the transistor into the right slots in the socket. The drain goes to the nuvistor plate connection (pin 2), the gate goes to the nuvistor grid connection (pin 4), and (here is the tricky one) the source goes to the nuvistor *filament* connection (pin 12). The shield goes to any ground connection on the nuvistor socket.

The reason for connecting the source to the filament lead was to take advantage of the 470 pF by-pass capacitor already installed at the nuvistor socket. Now all that remains is to ground the source (old filament lead) through a suitable bias resistor. Note the use of the word "suitable". The 2N4416 data sheet says to adjust the bias to obtain a source current of 5 mA—of the 2N4416 transistors that I tried, one required 330 ohms bias and the other required 1000 ohms bias with the same 12 volt drain supply.

The manufacturer actually shows a variable voltage in series with the gate in their experimental 400 MHz circuit (Fig. 2). Once the bias is set for the correct source current (5 mA), there appears to be no significant difference in the performance of the transistors that I have tested to date.

Referring again to Fig. 1, you will note that the following changes were made to the original circuit: short out the grid bypass capacitor, short out the 10 kilohm plate (drain) dropping resistor and add a source bias resistor (correct value determined by trial) between the old filament input terminal and ground. Clip one end of the 47 kilohm grid bias resistor loose since this will now appear directly across the old grid input coil L2 and serves no useful purpose except to lower the Q of L1. Finally, connect a source of B plus (12 to 20 volts) to the shorted-out 10 kilohm resistor with the negative to ground.

If you don't own an International Crystal preamplifier, you can build up the same circuit from scratch leaving out unnecessary parts and using a standard transistor socket. I suggest that a shield be placed across the middle of the transistor/nuvistor socket to separate the input from the output to prevent instability.

It is interesting to note that the only remaining difference between the Union Carbide test circuit (Fig. 2) and the modified nuvistor preamplifier (Fig. 1) is the relative position of the neutralizing coil L3 and the 200 pF blocking capacitor.

For some reason, all the published circuits of FET VHF preamplifiers that I have seen show the coil on the gate side of the capacitor. The nuvistor style preamplifer shows the coil on the plate (or drain) side of the capacitor as in Fig. 1. I left the neutralizing coil on the drain side of the capacitor to facilitate conversion. Neutralization was relatively easy but there is some interaction between L2 and L3. Every time you adjust one it is necessary to touch up the other. This might be improved by moving the neutralizing coil to the other side of the blocking capacitor.

I also tried a couple of TIS-34 FET's in my converted nuvistor preamplifier and found that the gain was about the same but the noise was about one dB higher.

Any N-channel VHF FET will work in the circuit shown in Fig. 1 with only slight differences in gain and noise figure. If you use a different make of transistor, do a little checking before you make your purchase and compare manufacturer's data sheets.

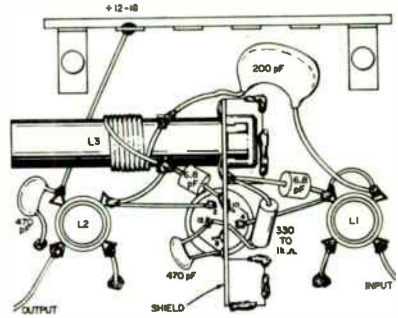


Fig. 3. Layout of the two-meter FET preamplifier.

432 MHz preamplifier

Since I had such good luck with the 144 MHz preamp, I decided to try a similar approach on 432. The result is both state of the art in performance and about the ultimate in simplicity at one and the same time. Because I couldn't believe what I thought I was hearing, I took the thing all the way up to Sonoma and we plugged it in ahead of Frank Jones', W6AJF, latest 432 MHz transistor converter. The result was a .9 to 1.2 dB improvement, depending on which test set-up we used. To quote the maestro himself, "that's pretty good." Practically speaking, the noise figure can't be much worse than 2.5 dB.

This circuit was taken directly from the test circuit shown on Union Carbide's data sheets. The only unusual thing about the circuit is the method of impedance matching used in both the input and output circuits. This consists of a pair of 2 pF capacitors, and you can't get much simpler than that. What's more—it works.

Because I had to have a place to start, I used the inductor dimensions from Frank Jones' transistor converter. As it turned out, I had to lop 1/4 inch off each one. Following the Union Carbide directions, I adjusted the

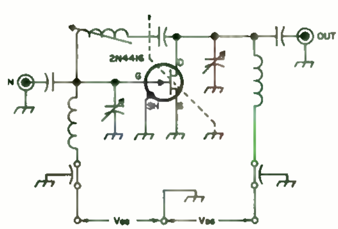


Fig. 2. 2N4416 400 MHz rf amplifier circuit recommended by Union Carbide.

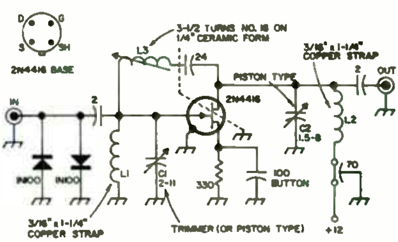


Fig. 4. 432 MHz preamplifier using the 2N4416 FET. This amplifier provides a noise figure of approximately 2.5 dB at 432 MHz with 12 dB gain.

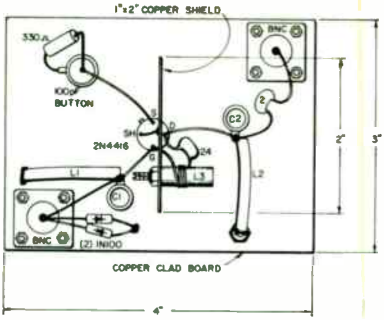


Fig. 5. Layout of the 2N4416 432 MHz preamplifier built by W6OSA. The neutralizing inductor L3 should be shown on the gate side of the shield.

drain-source current for 5 mA. This required a 330 ohm source resistor for one 2N4416 and 1k for the other one of the two I had on hand. The noise figure of the device requiring only 330 ohms bias is slightly better than the second. I don't know whether this is a universal rule, but I would be interested to hear if others experience the same effect.

Neutralizing was surprisingly easy. I started with a three turn coil and had to add another half-turn. I can now make the preamp oscillate by de-tuning the neutralizing coil, L3, to either side with the slug.

The gain of the preamp is 12 dB, plus or minus one. This is probably not enough if you are planning to run it directly into a diode mixer, but when added to the front end of an existing converter, either tube or transistor, the results are sensational. My own line up consists of a 2N4416, followed by two 6CW4's, into a 2900 hot-carrier diode mixer. Not only is the 2N4416 preamp highly stable and nonregenerative, it had a most beneficial effect on that first 6CW4.

For those who have not worked with FET's before, here are a couple of hints. The drain to source junction is almost a dead short with zero bias between source and gate. You can confirm this with an ohmmeter *not* on the X1 scale, please). This is why they are normally operated with reverse bias for class A operation. The required bias is most easily obtained by putting a resistance in series with the source. This is the same idea as cathode bias in vacuum tube circuits. The amount of bias can be adjusted for best noise figure. However, the rule of thumb is to set it up for half the rated I_{DSS} (saturated drain-source-current). In the case of the 2N4416, this works out to be about 5 mA I_{DS}.

When using an FET as a mixer, the I_{DS} should be very low with no local oscillator injection—less than 1 mA. The best place for injection is the source, but this does require more local oscillator power. You will, in fact, probably need an extra transistor in your L. O. chain to get the necessary power compared to what you can get away with if you put the injection on the gate. However, bipolar transistors are pretty inexpensive these days.

Getting back to our 432 MHz preamp, the construction is simple even if you are a mechanical "drop-out" like me. I used a socket for the FET because I wanted to be able to plug in different ones. Apparently there is no need for those "zero-length" leads that we used to read about in connection with bipolar transistors on 432 MHz. The output tuning is quite sharp, the input is broad. This is why I used the high quality piston capacitor on the output and the ordinary trimmer on the input. You can use pistons in both places if you desire. I used the 70 pF feedthrough bypass capacitor because it was the first one I could find around the shack. Any value up to 500 pF should work equally well.

If you are running any kind of power on 432 MHz, like a 4X150, Frank Jones recommends using two 1N100 diodes back to back from the input connection to ground. It is pretty cheap insurance.

The ground end of the gate inductor, L1, is soldered directly to the copper circuit board base of the preamp. The ground end of the drain inductor is soldered to the top of the 70 pF feedthrough capacitor.

Any value of drain voltage from 12 to 18 volts can be used—I use a 12 V battery eliminator. I suspect that if you go much under 12 V you will start to degrade the performance. Everything else pretty much speaks for itself.

Postscript

I consulted some technical publications and learned that the rule of thumb for class A amplifiers is to use $I_{DSS}/2$ (drain-source-current-saturated divided by 2), and whatever amount of bias it takes to arrive at that condition. I next conducted a little experiment with the four TIS-34's that I had on hand to see what the variations between individual FET's might be. The results of that survey are rather amazing too (see Table 1). The funny thing is that all four samples work equally well when placed in the circuit and adjusted for the proper current. The above rule of thumb does not apply to mixer applications, incidentally. In that case, the I_{DS} should be much lower, about .35 mA.

		Source Resistance					
	TIS-34	0	100	200	300	500	1000
	#1	8.5	5.8	4.9	4.0	3.8	2.2
I_{DS}	#2	11.0	8.0	6.4	5.5	4.2	2.4
in	#3	5.9	4.4	3.7	3.4	2.3	1.2
mA	#4	16.0	11.5	8.1	6.5	4.9	3.1

I tried running one TIS-34 at zero bias and found that the I_{DSS} started out at about 8.5 mA when first turned on and drifted down to 6.5 mA after about 15 minutes. (Note that FET's draw less current as they heat up—sort of reverse thermal runaway.) In any event, I don't recommend the zero bias mode of operation. I tried grounding the source as recommended by K6HMO* and pinned the needle on my meter. My advice is to experiment until you get the best results, but start out with at least 500 ohms for the TIS-34 and 1k for the 2N4416.

. . . W6OSA

* "A low-cost FET two meter converter," page 74.

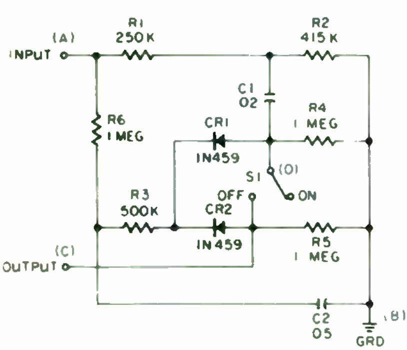
CHAPTER 21

Transistorized Noise Clipper

Most modern communications receivers are equipped with factory-installed noise-limiting devices; but either impulse-noise problems are becoming more severe or the "stock" noise limiters are becoming less effective. When an acute noise problem arises—as when first encountering the notoriously noisy mobile conditions of six-meter operation—and the built-in automatic noise limiter seems inadequate in its clipping depth, most hams spend some time on research. In the quest for easy and practical methods of impulse-type interference rejection, a mobile operator may be surprised to see the volumes that have been written on the suppression of noise at the source.

Unfortunately, source-type or "active" noise control can be a formidable task. The prospect of thorough bypassing, grounding, and shielding is often too involved to consider. But simply because the "passive" noise rejection of the built-in noise limiter doesn't always prove satisfactory, it is by no means proof positive that one need take "active" measures. The majority of commercially made communications receivers employ half-wave noise clipper circuits. That is, the positive audio half-cycle is clipped while the negative half-cycle remains unchanged. This allows a substantially undistorted audio output with a reasonable degree of impulse noise chopping. While this is generally acceptable, the more severe noise problems demand full-wave control.

While most hams are aware of the shortcomings of the conventional ANL circuits, few want to alter the appearance or circuitry of their "store-bought" rigs through modification or the addition of a home-brew accessory. An active amateur may have a transmitter under the dash panel of his automobile, plus a receiver or converter, and maybe even a power supply. The inclusion of another circuit would mean another black box—which is generally taboo with the xyl's.



So the limiter described here might just be the perfect answer to those "incurable" mobile interference problems. It effectively provides full-wave noise limiting and can be built into a unit small enough to mount in the actual receiver (under the chassis). And there is no need to change the receiver's physical configuration or electronic circuitry. Inclusion of the solid-state clipper into a receiver involves only soldering at three (or possibly four) points, and the disconnection of one wire (where a noise limiter already exists). This is no more effort than the soldering of a transistor into a circuit.

Theory of Operation

The limiter as described here was designed by Dick Hughes (W6CCD) of Pomona. He started with a standard circuit which employed a vacuum tube type dual diode, then substituted semiconductors in place of the tube. The inherent incompatibility of the original components with semiconductors necessitated other changes. The result is shown in the diagram. In this circuit, highly effective clipping with very low distortion is accomplished by clipping only on the positive signal pulses. The negative pulses are reduced in amplitude to a value approximating the clipped positive pulse. This gives a relatively noise-free output sine wave whose characteristics are essentially the same as those of a clean signal from the detector.

Actual clipping level is determined by capacitor C2. The higher the value, the greater will be the amount of signal clipped from the positive audio pulse, and the greater will be the amplitude of the negative cycle. The voltage difference between the unclipped negative half-cycle at C1 and the unclipped reduced-amplitude negative half-cycle at C2 determines the level, or amplitude, of the signal at the output.

Experimentation has shown a value of .05 microfarad to be optimum for conditions of serious noise in receivers with a bandwidth of 2 or 3 kc. In broader receivers, the noise could be clipped closer with little increase in signal distortion. In highly selective receivers, however, with a bandwidth of 1 kc or less, the value of C2 might be raised slightly to assure distortion-free performance while still providing adequate impulse-type noise limiting functions.

The circuit is adaptable to any vacuum tube circuit and to hybrid circuits employing vacuum tubes in the audio preamplification stages. The filter is installed in series with the audio line between the detector and the grid of the first audio amplifier.

To summarize: the distortion is average, while the audio dampening characteristics are a considerable improvement over a conventional half-wave noise clipper. The effective clipping action under conditions of severe noise is dramatically distinct.

Construction

The following parts are required for the noise limiter:

R4–1 meg	C2–.05
R5–1 meg	D1–1N459
R6–1 meg	D2–1N459
C1–.02	S1–SPST

The resistors may be ¼ watt, if obtainable (a necessity where extreme miniaturization is required). The switch is not required, and may be omitted in installations where the circuit will be on at all times. The diodes may be replaced by either 1N457 or 1N458 as long as both diodes are the same. The only difference in these diodes is the peak inverse voltage rating, which will not be exceeded in any case. There is nothing critical in the selection of diodes for the noise eliminator. Anv conventional diode with an extremely high back resistance and relatively low forward resistance will suffice. The diodes must be able to conduct under normal conditions without decreasing amplitude and to offer a nearly infinite resistance to pulses in the nonconducting state. All values shown are plus or minus ten per cent. The circuit point marked A is the audio input. B is floating ground or ground. C is the audio output. If there is already a noise limiter in the circuit, and it is not desirable to remove it or utilize the switch in the new circuit, simply disconnect the existing limiter from the audio input and prevent it from coming into contact with other wires. This will disable the existing noise limiter. Tie new limiter point A to the same contact point. Tie audio output C to audio-output point of existing limiter (the existing limiter may remain tied in at this point). If an on/off switch is to be used, the switch will connect between C and D. These two leads should be shielded to prevent pickup of noise and hum. When C and D are shorted (switch closed) the noise limiter is off.

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TRANSMITTERS

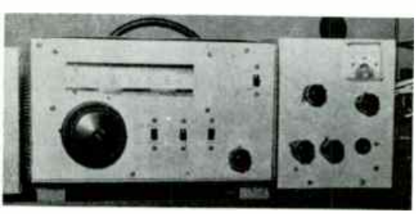
CHAPTER 22

Transistorized SSB Xmtr for 20 & 80 Meters

The Transistor has now been developed to such a stage that relative high power can be delivered on the HF bands. It is therefore time to start using these new transistors in transmitter construction instead of tubes. The transmitter, to be described in this article, covers the eighty and twenty meters phone bands, and has an output of approximately one quarter watt. However this power level could easily be raised to 10 watts or more by using newer transistors.

Before going into details with the transmitter, it might be useful to recall that a transistor is not a tube, and consequently is not supposed to act as such. The greatest difference lies in the fact that transistors are power driven devices, whereas tubes are voltage driven devices; this, however is about the only difference one needs to remember when building and adjusting transistorized equipment.

Regarding the description of the transistorized 80 and 20 meter SSB transmitter, transistors offer many advantages over tubes when used in lower power SSB exciters. Heating is no longer a problem and the resulting troubles with frequency drift is eliminated. The power drain is much lower, which makes the rig much handier for portable use. The long-time stability of transistors seems to be very good, so a transistorized rig should not call for realignment as frequently as would be necessary in tube rigs. These are only a few of the reasons the author built a transistorized rig.



Shows the transmitter front view. The knob lower right corner is the band switch. The unit to the right is the linear amplifier.

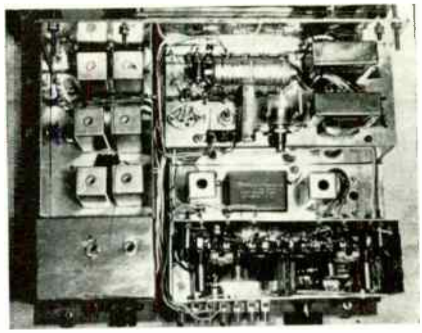
The block-schematic in Fig. 1 shows the different parts of the transmitter. Carrier is generated on 9 mc, and is fed to a balanced modulator, to which the audio from a 3 stage audio amplifier is also feed. From the modulator the signal—a DSB signal—goes to a 9 mc McCoy crystal filter, where one of the sidebands is cut away. From the filter the signal (which is now a SSB signal) goes to an amplifier and further on to a mixer stage. Into this stage is also injected a signal from a 5 mc oscillator. At the output of the mixer either the sum or the difference signal might be selected (3.8 or 14 mc) and fed to a stage of amplification before going into the final amplifier with an output of approximately 250-300 mw. Although this is not much power, it is sufficient for local contacts, and more than enough to drive the author's linear amplifier well beyond the 300 watts power limit in this country.

The Audio Amplifier

The audio amplifier consists of three transistors. The microphone to be used with this transmitter is a high-Z xtal mike, and since the input impedance of a common emitter stage is rather low, some sort of matching must be connected between the mike and the amplifier if there is to be no great loss in gain and quality. An input transformer might be used if it is already available, but a far better method would be to use a common collector stage as an impedance match. Such a stage has a very high input impedance, and an output impedance which will match quite well to the following common emitter stage. Between this stage and the output stage there is connected an audio gain control. In the collector circuit of the output transistor there is a driver transformer for a push-pull audio amplifier; only half of the output winding is used to feed the balanced modulator. All coupling- and by-pass capacitors in the audio amplifier have been made quite large, so that a good audio quality (that is, a reasonable amount of bass) can be obtained. The audio amplifier is not a critical part of the transmitter, so it is not necessary to bother with temperature stabilization. The ordinary dc-feedback method shown in schematic will work very well.

The Carrier Oscillator

The carrier is generated on 9 mc., and there is a separate oscillator for each sideband. There are several reasons for the use of two oscillators. The oscillator section had to be mounted in the rear part of the chassis, and it was therefore quite difficult to place a switch in such a manner that it could reach from the rear to the front panel and still leave enough room for other parts of the rig. Another, and probably better reason, is that it is quite difficult to get the crystals to oscillate on the proper frequencies because of the rather high input capacitance between base and emitter. It is therefore necessary not to add any more capacitance to the circuit than needed. The proper sideband is selected by switching the collector voltage to the oscillating transistor. Output is taken from the collector and fed to a broadband transformer wound on a ferrite toroid. The windings are bifilar, and there is no significant resonance in the transformer. Output is fed to the balanced modulator. The crystals are oscillating in a circuit similar to the tube circuit recommended by the McCoy factory, and it should be quite easy to get the oscillators on the proper frequencies by tuning the trimmers. Since these trimmers also control the feedback in the oscillators, it might very well happen that the oscillators have unequal outputs. This can be corrected by adjusting the resistors in the collector supply leads.

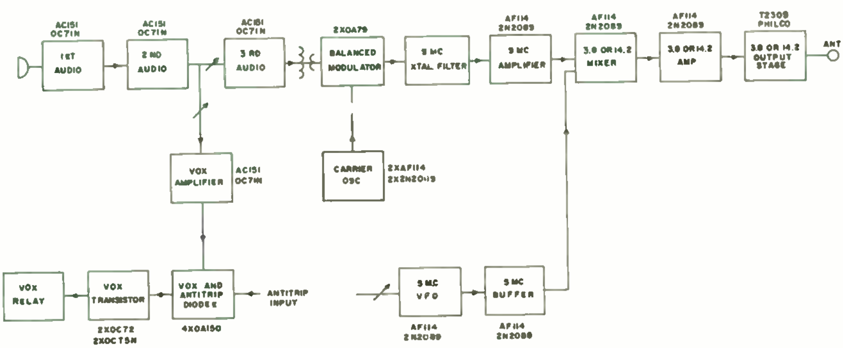


Shows the transmitter from above, to the left are the rf amplifiers, with the mixer near the front panel. To the right are first the power-supply mounted above the oscillator section. Instead of using a 48 volt ct transformer, two 24 volt transformers were connected in series. Next comes the filter section, and below the carrier and audio sections.

The Filter Section

The balanced modulator is of a very conventional design, using two shuntfed diodes. The coil is bifilar wound to assure a good balance. The carrier-nulling is done with a 1 kiloohm carbon potentiometer and a small variable capacitance. The transformer L_2 is a modified 10.7 mc miniature transformer. One of the sections is removed and a bifilar coil L_1 is wound close to the cold-end of L_2. The top of L_2 is coupled to the filter through a 5-30 mmfd capacitor. The filter is shunted by a resistor on 560 ohms to give a proper termination. The coupling network is an L-network, and it is very easy to adjust. The output terminal of the filter is also shunted

TRANSMITTER BLOCK DIAGRAM



by a resistor on 560 Ohms, and a tuned circuit with resonance around 9 mc, this resonance is not at all critical, because the coil is shunted with the 560 Ohms resistor. The two shunt resistors are probably not necessary at all, but they help make the tuning very easy. The filter is followed by an amplifier of conventional design, the collector circuit L_4 and L_5 is a standard 10.7 mc transistor *if* transformer. Most types on the marker will easily tune down on 9 mc, but if not they may be padded with an outside capacitor on 10-20 mmfd. There is no impedance match between L_4 and the transistor. It is not needed since the stage delivers gain, but does not add to the selectivity. Signal output to the rf-mixer is taken from the base link L_5 and fed through a piece of coaxial cable.

The Variable Oscillator

The variable 5 mc oscillator is probably the most critical part of the entire rig. If the frequency stability is poor, the exciter is of little value. Luckily, it is rather simple to make the VFO as stable as SSB operation calls for. One big advantage is of course that the transistor does not generate any heat. This almost eliminates all problems with warm-up drift. However, transistors have one fault which might make it difficult to build an oscillator with good long-time stability. This is the rising junction temperature, which is caused by the collector leakage current. The first step towards a stable transistor oscillator must therefore be to make a good DC stabilization. DC stabilization is also necessary because a changing collector current will cause a change in the transistor's load impedance and load capacitance. All capacitors must be of the best quality; as in a normal tube oscillator, silver mica capacitors and air trimmers must be used. Needless to say, all components must be rigidly mounted. The oscillator which tunes 5.1-5.35 mc, is followed by a buffer stage. With the coverage listed the exciter will tune 3.9-3.65 mc and 14.1-14.35 mc. The buffer stage is a must if good stability is to be obtained. This stage is a normal common emitter, and no trace of frequency shift was noted under modulation. The voltage for the oscillator and the buffer stage should be stabilized by a zener diode or powered from a separate battery, to eliminate pulling when the rig is used as a mobile exciter.

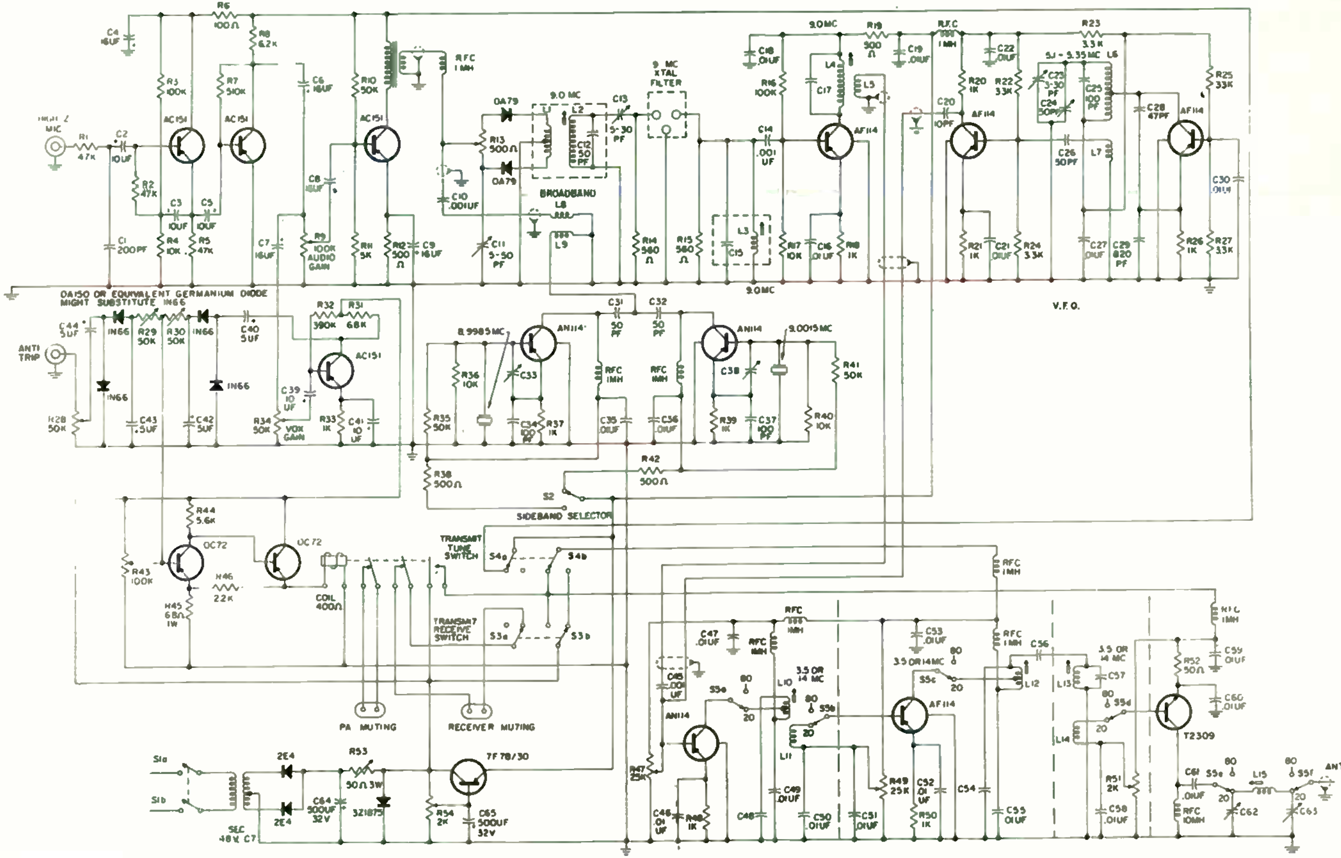
The RF Amplifiers

The rf mixer is a normal additive mixer stage. No attempt has been made to try a balanced mixer, since the mixer is followed by so many tuned circuits that there have been no troubles with any type of spurious radiation from the rig. The base circuit of the mixer should be well shielded, so that there is no possibility of rf pick-up from the linear amplifiers. Remember that the power level at this point is very low and not much rf is needed to make a feed-back chain. The proper base bias is selected with a small "set and forget" potentiometer. In the collector circuit there is a tuned circuit for 20 and 80 meters. The collector is tapped down on the coil L_{10} to give a good match, so that ample gain and selectivity can be obtained.

The mixer is followed by a common emitter rf amplifier stage. The rf signal is fed to the amplifier via a link L_{11} wound on the cold-end of L_{10}. One side of the link is bypassed to ground, both in the mixer and in the amplifier section. This is necessary to prevent self-oscillation and pick-up. The collector is tapped in on the collector coil L_{12} so as much gain as possible can be obtained. The coupling to the output stage is done through a band-pass filter. This band-pass filter is necessary to assure a good unwanted signal suppression. In contrast to voltage-driven tubes, transistors may cause loading because they are power-driven. The selectivity is therefore much lower in a transistorized circuit than in a tube circuit. Consequently more tuned circuits are needed in a transistor rig to obtain a certain degree of selectivity than would be necessary in a tube circuit. Drive for the output transistor is taken from a small link L_{14} on the cold-end of the last coil L_{13} in the band-pass filter, and fed to the base of the transistor. This stage is also connected as a common emitter stage. The output circuit is a pi-network since it was the easiest to use. Impedance matching from the collector to the tuned circuit and further on to the antenna must be correct. If not, the stage will deliver a very distorted signal to the antenna. The author has had little success in calculating the proper values for the pi-network, so the values used were obtained by trial and error. In all but the last stage, the value of the base voltage divider or potentiometer is relatively uncritical, although lower resistance gives better stabili-

Notes on diagram

C33 and C38 both 4-30. Adjust C33 and C38 to correct carrier frequency and R38 and R42 for the same output on both sidebands.
Adjust R53 so the no-load zener current is approx. 125 ma.
Adjust R54 so the emitter voltage under load is 9 volts.
* Indicates silver-mica capacitors.
L2, L3 and L4 are standard 10.7 mc transformers.
C12, C15 and C17 are selected so the circuit will resonate on 9 mc.
L1 2x5 turns bifilar wound on cold-end of L2, L5 3 turns on cold end of L4.
Unmarked capacitors are mmfd. N means x1000.



HIGH Z MIC
ANTI TRIP
OA150 OR EQUIVALENT GERMANIUM DIODE MIGHT SUBSTITUTE IN66
VOX GAIN
AUDIO GAIN
BROADBAND
9 MC XTAL FILTER
9.0 MC
8.9985MC
9.0015MC
V.F.O.
SIDEBAND SELECTOR
TRANSMIT TUNE SWITCH
TRANSMIT RECEIVE SWITCH
PA MUTING
RECEIVER MUTING
COIL 400Ω
SEC 48V. CT
ANT.

ty. But in the final stage the potentiometer should not have greater than 2 kilohm resistance, or it would be difficult to get the stage to work as a linear amplifier. This is simply because the linear draws base current, and when used as a power amplifier, the change in base current is quite high. If the resistance in the circuit now is high, this will cause a great swing in base bias, resulting in distortion. By keeping the resistance low, a greater change in current can be allowed, without affecting the base bias. Remember that transistors are power driven, and will therefore draw base current when operating in class A, B and C.

As noted in the schematic, there is no tuning controls but the VFO. This is because it is possible to tune the rf amplifiers in such a manner that uniform gain is obtained in a band-pass of around 100 kc on 80 meters and around 400 kc on 20 meters, measured on a dummy load. Outside of the band-pass the gain drops off rapidly. This rig is tuned with a center frequency on 3750 kcs and 14175 kcs, and can cover 3.7 to 3.8 mcs and 14.1 to 14.35 mcs. The main SSB band in Europe is 3.7 to 3.8 mcs, but the transmitter can easily be tuned in the American phone band. However, only the range 3.8 to 3.9 mcs may be covered, without changing the values of the oscillator components. The output transistor, a Philco T 2309, delivers around 250 to 300 mw's output.

The Transistors

Little has been said about the transistors used in the transmitter; the reason is simply that the transistors the author used, might be difficult to obtain in the USA since they are made in Europe by Siemens. Furthermore it is quite difficult to find usable equivalent transistors, as may be seen from the variety of types in the Philco, Texas and PSI catalogues. Some data on the type of transistors used might help in selecting proper substitutes (See table I).

All germanium PNP types. OA 79 and OA 150 are rf germanium diodes with high back resistance. Almost any two matched diodes will do in the balanced modulator.

The VOX

It is not necessary to say much about the VOX circuit, since it has already been well covered in an earlier issue of this magazine. However, it might be necessary to mention the fact that the supply voltage to the VOX and audio stages must be quite stable in order to avoid troubles with self triggering. The first power supply used in this transmitter had a voltage drop of about 0.5 volt when the transmitter was activated. This voltage drop looked like a "one shot pulse" to the audio amplifier, and was amplified enough to be able to trigger the VOX stage. The pulse frequency was around 1 c/s., so this kind of trouble might be prevented by using a small coupling capacitor between the audio amplifier and the VOX amplifier, or by stabilizing the supply voltage. In fact, since the voltage for the oscillators must be stabilized, it is no extra trouble to stabilize the entire power supply.

Mechanical Layout

Although it was decided to etch all circuitry, it was soon found that this type of mounting was excellent for making a neat and clean-looking construction, but not very useful when components had to be removed or changed. Instead, the rig was made by placing the different stages in small sub-chassis. This proved to be very useful when testing the different stages. The entire exciter measures approximately 6 by 9 by 11 inches. Although it is not a miniaturized rig, it could easily be made much smaller without affecting anything but the dial. A good and stable exciter is not worth much if you don't have a dial which can tell you your exact frequency, so the dial will determine how small the rig might be. The variable oscillator and the rf amplifiers should be carefully shielded to prevent pick-up and self-oscillation. All coils are housed in cans. The photos show how the different stages have been placed.

The Alignment

The alignment of the exciter begins with the power supply. First, adjust R_{53} so the no-load zener current is approximately 125 mA. Next, the transmitter is connected to the power supply. Be sure that the potentiometers R_{47} and R_{49} are set with the arm to the ground side, and R_{51} set with the arm to the negative side. Switch S_3 and S_4 to transmit, and adjust R_{54} so the voltage on the low-voltage line is around 10 volts. Later,

Table 1

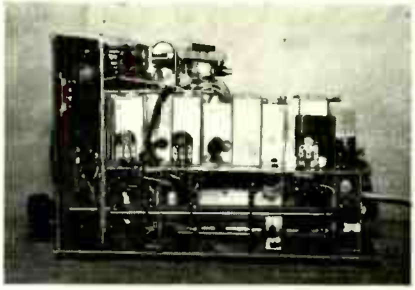
	Suggested Replacement
AC 151 audio trons . col diss. 30 mw h_{fe} 60 — V_{cbo} 18 V. —	OC71N — Amperex
AF 114 VHF trans . col diss. 50 mw h_{fe} 15 0— V_{cbo} 20 V.	2N2671 — Amperex
OC 72 audio trans . col diss. 100 mw h_{fe} 75 — V_{cbo} 32 V. —	OC75N — Amperex
TF 78/30 audio trons . col diss. 3 W h_{fe} 50 — W_{cbo} 30 V. —	2N178 — Motorola or 2N176

when all amplifiers have been adjusted correctly, the control is readjusted so that the voltage is 9 volts. Now disconnect all stages but the audioamplifiers. All 3 stages should draw 3-4 ma at 9 volts. Connect a high-Z mike to the input and a pair of phones over the output terminals of the audio transformer with the balanced modulator disconnected. Speak into the mike and check for a good audio quality. If an audio generator and a VTVM is available then check the response. Be sure to have sufficient bass; the drop-off should be around 200 to 300 c/s., but not higher. The output with the gain control full clockwise should be at least 0.5 volt across a 500 ohm load.

The next stage to be checked is the VOX. First, apply collector voltage to the stages; next, adjust R_{43} so the relay just goes in the receiving position (great voltage drop across the relay coil). You will find the adjustment of the VOX balance (R_{43}) quite critical; set the potentiometers as near to the switch position as possible. Now advance the VOX gain (R_{31}) while speaking into the mike. If nothing is wrong the VOX relay will start to trigger. The delay might be adjusted slightly with the R_{30} and R_{43} potentiometers. Adjust the controls until a desired VOX operation is obtained. Under normal VOX operation the switch S_3 should be in the receiving position.

Now connect the two carrier oscillators in the circuit and connect a VTVM from the hot-end of L_9 to ground; check to make sure that both oscillators are oscillating. Place both trimmers C_{33} and C_{38} at full mesh. If a frequency meter or an accurate receiver on 9 mc is available, the correct positions of the carrier oscillators might be fixed with C_{33} and C_{38}. If these facilities are not available, the carrier oscillator will be fixed later when the rig is on the air in a ham band. Also adjust R_{38} and R_{42} so the two oscillators are giving the same output. Note! (this should not be done until the carriers are placed in correct relation to the filter).

If a high sensitive VTVM is not available, which is not likely to be the case in an amateur station, the next step will be to test and adjust the variable oscillator. Connect the oscillator and the buffer to the 9 volt line. If the oscillator is oscillating properly there should be around 2-4 volts of rf on the collector of the buffer transistor. Make sure that the oscillator is covering the desired range with a grid-dip meter or a receiver. As an example tune the oscillator so that the output frequency of the exciter will be 3.8 mc. The oscillator will then be on 5.2 mc. Switch the bandswitch to the eighty-meter position. Resonate with a grid-dipper or a transistor-dipper L_3 and L_4 to 9 mc, and adjust L_{10}-L_{12} and L_{13} to 3.8 mc. Set the collector current



Looks into the rf amplifier section, note the heat-sink on the output transistor made of a piece of tinned iron bent around the transistor.

of the rf mixer to approximately 4 ma by R_{47}. Connect the receiver through a small capacitor to the hot-end of L_{11} and tune to 3.8 mc. Start to speak into the mike, and tune the receiver back and forth until you can hear the transmitter. Then adjust L_2 and C_{13} and readjust L_3 and L_4 for maximum signal and best quality. Adjust R_{13} and C_{11} for carrier null, and try to move C_{11} to the other side of the balance control if no result is obtained in carrier nulling. Adjust C_{33} for the proper sideband position, switch to the other sideband, and repeat the treatment with C_{38}. Readjust the carrier balance controls for maximum suppression on both sidebands. The setting of L_2 and C_{13} might affect the carrier balance, so adjust these once more.

Having come this far, you are pretty well on the way. Disconnect the receiver from L_{11}, and place a VTVM on the collector of the mixer. Peak the core to maximum output; also adjust R_{47} to maximum output (the adjustment of R_{47} might be more critical on twenty than on eighty meters). The rf voltage on the mixer collector should be 0.5 to 1.5 volts. Now connect the remaining amplifiers to the 18 volt line and a 50 ohm dummy load across the antenna terminal. Place a milliameter in series with the emitter of the T 2309. Place the VTVM from the hot-end of L_{11} and adjust L_{12}-L_{13} and R_{49} for maximum deflection on the meter (approximately 2-3 volts). Move the VTVM across the dummy load and connect also the receiver through a small capacitor to this dummy load. Listen to the receiver while slowly advancing the bias control R_{51}. At a certain setting the signal will change from highly distorted to a clear signal. Note the idler current (approximately 20 ma), and adjust $C_{62\text{-}63}$ and L_{15} for maximum output. The position of R_{51} will be found to be quite critical. You have now almost completed half of the tuning procedure; the only remaining thing to be done is to retune the rf amplifiers so it covers the desired range with a uniform output.

Repeat the entire procedure outlined above on 20 meters. The bias potentiometers should also be readjusted since they might be a little more critical on this band than on 80 meters. There should be no trouble in getting the exciter to cover the entire twenty meter band. If the shielding is good, there should be no trouble with parasitic or self oscillation.

Although the exciter has been designed for fixed use, it could very easily be modified for mobile use. Changing the 18 volt line to 12 volt should not affect the operation, but will of course give a somewhat lower output. To avoid frequency changing and drift in operation while driving, the oscillators should be fed from a dry battery or a pair of cascade coupled zener diodes.

The transmitter has now been in use for about one year and has shown very good operation facilities. The low power drain enables the rig to be tuned on at all times, so the frequency stability can be even better. The author got the rather bad idea of placing the power supply on top of the oscillator section. And since the transformer generates some heat, this heating could affect the frequency stability. By leaving the power on at all times, the temperature will be kept rather constant, thus eliminating the warm-up drift. However, the power drain is below 5 watts, so you don't have to bother about the electric bill.

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COIL DATA

L_1 4 + 4 turns bifilar wound on cold end of L_2.
L_2 one section of a standard 10, 7 mc if transformer.
L_3 same as L_2
L_4 same as L_2, L_5 base winding.
L_6 16 turns Airdux 816, collector taped 8 turns from the hot end.
L_7 3 turns on the cold end of L_6, spaced 2 turns.
L_8,L_9 broad-band transformer wound on a ferrite toroid L_8 12 turns, L_9 3 turns.
L_{10} 80 meters: 20 turns, taped 5 from hot end.
20 meters: 14 turns, taped 5 from hot end.
L_{11} 80 meters: 4 turns.
20 meters: 3,5 turns.
L_{12} as L_{10} on both bands.
L_{13} as L_{10} but no taping.
L_{14} 80 meters: 4 turns.
20 meters: 2,5 turns.
L_{15} 80 meters: 25 turns.
20 meters: 10,5 turns.

All coils are wound on a 8 mm form = 5/16", close-wound with number 20 wire. The cores are all high permeability ferrite cores.

CAPACITANCE VALUES

C_{16}	80	meters	350	mmfd	20	meters	90	mmfd
C_{34}	80	meters	350	mmfd	20	meters	90	mmfd
C_{36}	80	meters	50	mmfd	20	meters	15	mmfd
C_{37}	80	meters	350	mmfd	20	meters	90	mmfd
C_{62}	80	meters	1000	mmfd	20	meters	250	mmfd
C_{63}	80	meters	3000	mmfd	20	meters	750	mmfd

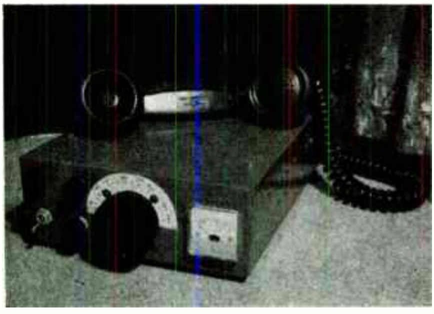
CHAPTER 23

Six-Meter Portable Station

The equipment which is to be described was built by K1IWI and is the revised and improved version of W1OOP's rig. The original idea was to make a better receiver which could be used in conjunction with the Springfield CD Walky-talkies, for net operation at sports-car hillclimbs, at spots not suitable for parking a car. When the receiver was finally done, there was so much room that we put in a transmitter also. With the right hill under you, it can be a lot of fun to use, and the battery life is long enough for a full day of operating.

The set consists of a crystal-controlled converter, a tunable *if* receiver covering 7 to 11 mc/s which is also handy for listening to CHU, a crystal-controlled transmitter taking 25 mc crystals, and a class-A modulator with clipping. There are 17 to 20 transistors, depending on whether you incorporate the optional final, BFO and squelch. The drain on receive is under 20 ma at 12 volts, and on transmit about 250 ma. With eight MN1500 penlite cells ($2.64 worth) you are set for a long day of hamming, or about 30 hours' net operation with a loudspeaker. Two F4BP lantern batteries ($1.58 total) will go about three times as long. The ten-cent penlight cells will poo out after a half dozen transmissions, so don't bother with them. K1IWI goes mountain-topping with a pair of hot-shots and claims they last forever.

The receiver, so far as we can tell, is good enough. There is some cross-modulation from strong local stations, but many "tube" receiving setups are as bad. The sensitivity is good, about 5 db noise figure including the losses in the protective circuit. The selectivity of the prototype was at one time adjusted to be 3 kc wide at 6 db down and 20 kc at 60 db, about the same as the early HRO's, but the coupling was later modified to make the receiver a little wider for net operation, by increasing the coupling capacitor between the pair of *if* coils to the value shown. It's still more selective than the older Gonsets.



Receiver drift is small, and what there is, is not caused by the transistors. Regulated voltage is used on the tunable (second) oscillator and its associated mixer to get away from battery voltage shift as the cells recover from a transmission. Transmitter drift is much less than found in tube transmitters using the same crystals, although the circuit was set up for maximum output.

The transmitter makes as much noise as can be expected from flea power. With a two-element portable beam Bob talked to five states from the top of Mt. Monadnock (the one you walk up) while W2NSD/1 was not using the ether. It is even fun to work W4's and W9's when you are about a watt—and hard, too. The point is that there are many spots where you can go with a lightweight portable and not find a bus-mounted kilowatt there ahead of you.

The antenna tends to be heavier than the transceiver. Hi-Par makes a portable beam (we don't have one) which should be fine. For walk-around use, like at hamfests, a center-loaded piano wire whip does the job. In any case, there is a problem setting up the transmitter, and you should have a sensitive tuned field strength meter with some provision for monitoring with headphones. Currents don't change much with tuning, but modulation quality is critically dependent on the tuning of the final. There is sidetone provided to the earpiece of the WE handset, but it only tells you that the modulator is working. Rf power transistors for use at 50 mc are a bit of a problem. Since both the driver and final have modulation applied, the required breakdown voltage is a little less than four times the battery voltage for each stage—say 48 volts for BV_{CER}. If BV_{CBO} is what the manufacturer specifies, it should be over 60, in our experience, i.e., type 2N1506 performed ok but 2N1505 and 2N2297 appeared to break down on modulation peaks, causing flat-topping. (A Tektronix 545A with suitable plug-in will allow you to view the rf envelope across a 50-ohm load, if you feed the audio in as an external trigger.) The high-frequency performance of types used for 27-mc CB rigs is seldom good enough, but it might be worth while to try a few dozen to find a good one. In general the ratings on rf transistors don't include AM, but a type rated at 1 watt out at 28 volts at 70 mc should be good for a quarter watt AM carrier output at 14 volts at 50 mc if you have enough drive (typically at least 110 mw.). When better transistors are made, you can expect that we'll try to use them at even higher frequencies (where as usual they won't work very well!)

The reason both final and penultimate stage are modulated is that with low gains and what is roughly class-B operation, modulating one stage doesn't seem to work. Also, neither the final nor the driver should be loaded for carrier conditions, but (like a Gonset linear) they should be tuned and loaded for best "upward" modulation. Since the transistor gain is much higher at, say, ten mc than at 50, there is a good chance of having parasitics at several mcs, though not much likelihood of trouble above the output frequency. A trick peculiar to transistors comes when the collector (in these NPN's) swings in a negative direction as the base goes positive to the point that current flows from base to collector, and for an instant in the rf cycle the output and input are connected. In one instance, this was found to react back to cause the oscillator stage to stop, giving an extremely overmodulated-looking output (the modulation envelope pinched off completely) and an extraordinary amount of buckshot. FM comes much easier.

The transmitter schematic is shown in Fig. 1. Useful carrier power output is about 1.5 watt,* with fairly good audio quality. Someone compared the modulation to a Zeus, but we heard that his receiver is working better lately since he got it fixed. To transmit, plus 12.6 is applied to the complete transmitter by the send-receive switch or relay. A type 2N706 or similar is used in the 25-mc oscillator. The circuit tunes like a triode tube oscillator, that is, the tank should be a little bit on the high side of crystal frequency. With a few exceptions, CB oscillators will work fine in this position, but not all makes of 2N706 work well. Fairchild and PSI seem to be better for rf work than others.

The second stage doubles from 25 to 50 mc. The oscillator tuning may be checked by observing rectified base current across the 820 ohm resistor. A relatively high-c collector tank is used at 50 mc. Note that some transistors of a type may have higher than rated breakdown voltage, but don't bet on which ones.

The driver is pretty ordinary. Neutralization is not used; large-signal amplifiers cannot be neutralized exactly, because the base-to-collector capacitance is a function of the voltage between the two elements, i.e., it's a varactor. The modulation can be applied in either the positive or negative lead, since there is no heater or filament insulation problem, and we chose to put it in the negative lead in order to make it possible to use a center tapped choke in conjunction with a pnp power transistor. The rf chokes and ferrite beads specified are what we used; a small solenoid wound on a 47-ohm resistor would probably do as well as the beads, if the latter are not obtainable.

* Bob's Heathkit "Sixer" read 1.0 watt on the same wattmeter.

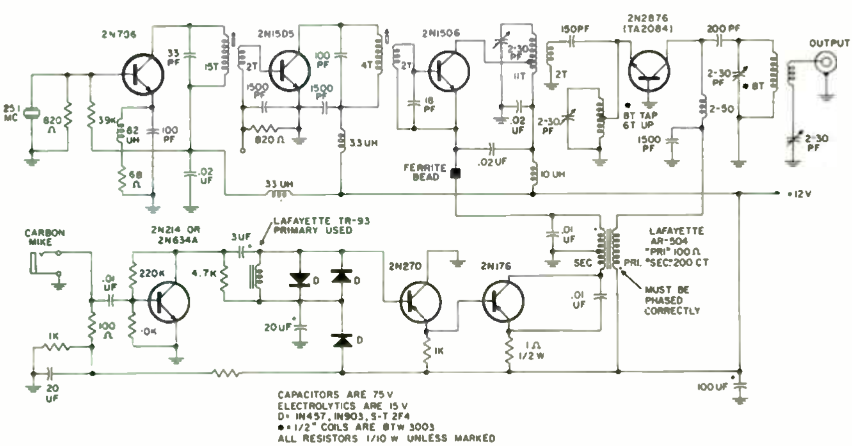


FIG. I

The modulator is designed for a carbon mike. To use a crystal mike (ceramic, if you will be in the hot sun) would take two more transistors. The first stage is NPN, since we are using negative ground, and its collector current flowing through a silicon diode provides bias for the other transistors. Two silicon diodes across a miniature choke (a transistor radio output transformer) clip the speech waveforms at about 1.3v peak-to-peak. An emitter follower using just about any PNP alloy transistor drives a power transistor operating class-A. If you cut the carrier when you have nothing to say, a class-A modulator amplifies square waves as efficiently as class-B. Any auto radio output transistor will work, although the 2N1172 (TO-37 style) miniature unit would be much smaller and about the same price. The unbypassed emitter resistor in that stage is adjusted to such a value that the current through the two halves of the choke is about the same. The center-tapped choke will be about half the size of the choke that would be needed for Heising modulation.

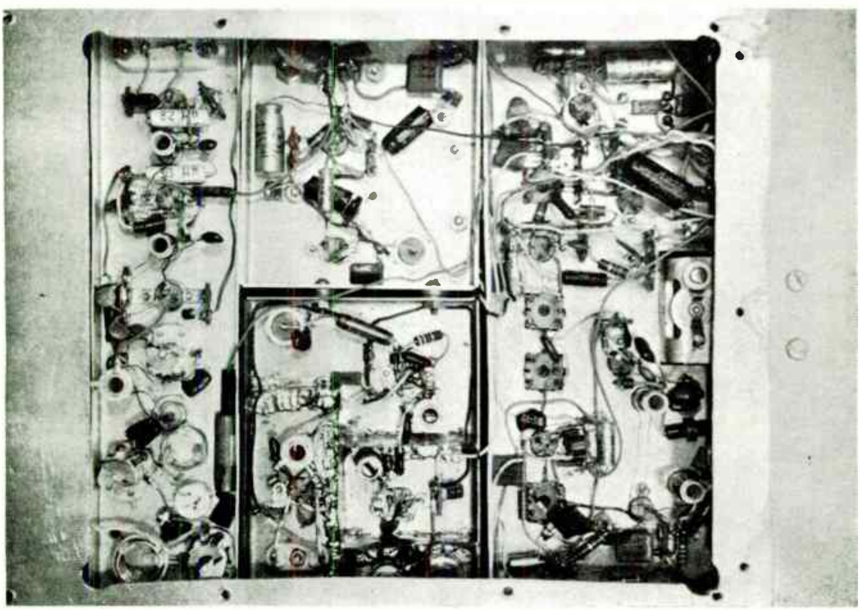
Construction—General

Bob built his transceiver in a commercial 8 × 12 × 3 aluminum chassis. The smallest face becomes the front panel. Four chassis were built up to fit the available space, with the receiver full width next the panel, then two half width units for the modulator and the converter, and finally the rf section in a chassis about an inch and a half by eight. Batteries went in the rear, but the photos show the fourth stage on the transmitter, with external batteries being used.

As the photos show, the open side of the chassis was fitted with a cover plate, while the other face was mostly hacked away to allow access to the other side of the subassemblies without removing them.

The main mechanical problem is to get adequate tuning precision and mechanical stability in the receiver; it should be fitted so as to make sure the dial and tuning capacitor line up right, which is best done by first mounting the dial, then connecting the T C, and then spotting where the tuning capacitor mounts on the chassis. Send-receive switching is done with a wafer switch. Only two poles are really needed, as the audio amplifiers are separate.

Heat and heat-sinks. The silicon transistors in the rf section will work satisfactorily without any heat radiator, although they get hot. Thermalloy #2211 heat sinks (in Lafayette's catalog) help quite a bit, and it wouldn't hurt to put the same type on the audio output transistors, if TO-5 size are used. The modulator power transistor is mounted on the aluminum chassis with the mica insulator supplied (at least it is with Delco types) for electrical in-



Transmitter | Modulator Converter | Receiver sulation. Check for and remove all burrs in the area covered by the mica. Run your tongue over it to check for smoothness.

Components. The individual units were made up on .040 aluminum chassis to simplify grounding and shielding problems, except the converter, which was made of 0.031 brass, so that shields could be soldered in place. Sockets were used for the transistors, a great convenience when there is a desire to try various types. We used saddle mount sockets (Elco 05-3301) with four pins which would accept TO-5 transistors directly, as well as TO-40, 2N43, 2N78, 2N270 and TO-11. Other types such as TO-7, which has a shield lead, TO-44 which has leads close together, TO-1, TO-18 and such can be plugged in by arranging the leads properly. The sockets were mounted with 2-56 brass hardware. Solder lugs for no. 2 were made by using the small end from standard size solder lugs. Model railroad hobby shops have the small bolts and nuts.

Resistors. We used several styles. Our advice is to use ohmite (Allen-Bradley) QUARTER watt types. The half watt use too much space, the tenth watt type are impossible to wire without tweezers and a jeweller's loupe —and the color code is hard to see—but the quarter watt size is just fine. Lafayette had some Japanese desposited-carbon resistors which were very fragile, and the small Globar

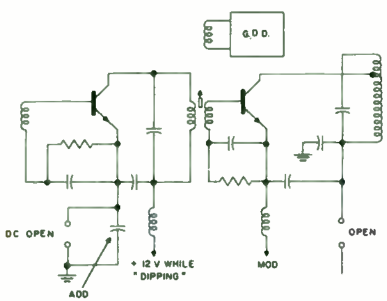


FIG.2

To tune up a transistor interstage with a grid dipper, you apply collector voltage to the previous stage, make sure the base return (coil or whatever) is in place, and disconnect the emitter lead. The driven stage should have collector voltage disconnected. Now you have 12 volts or so between collector and base of the previous stage, so the C-B capacitance is approximately what it will be when things are going, and the next stage has zero bias. Couple loosely to get a sharp dip; if the GDO signal is more than a few hundredths of a volt current will be drawn and the capacitance will be changed. A transmitter interstage set up to dip is shown here.

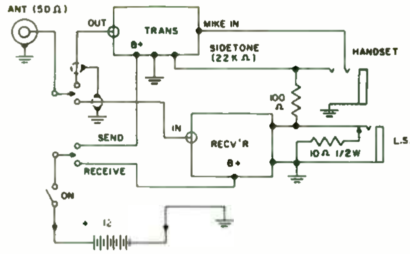


FIG. 3

resistors came apart when the soldering iron got them too hot. An Ungar pencil iron on a Variac worked well, but the smallest GE iron with a resistor in series (cut out by a foot switch) was even more convenient.

Capacitors. The electrolytics were C-D type NLW (easy on the leads, some broke off first bend) or assorted Japanese. The coupling and AGC electrolytics might be Mallory TAM or similar, though the aluminum type is satisfactory. The rf and *if* bypasses were mostly 75-working-volt ceramic types from Lafayette, with some 10-w.v. ceramics used as base-return-to-emitter bypasses where there normally is no more than two volts. You've never heard a noisy component until you put 12 volts on a 10-volt ceramic capacitor. Like frying bacon.

Battery mounting. If internal penlite-sized batteries are to be used, they may be mounted in commercially available clips which hold a group of two to four cells, bolted to a wall of the case. We have had trouble with these after rough handling (the transceiver is just the right size to drop!) from the batteries getting askew and not making good contact. It is suggested that the batteries should be accessible for voltage check under load or physical inspection *without a screwdriver.* A piece of insulating material held in place over the clips would also prevent cells coming loose. With the F4BP lantern batteries there should be no problem, since it is easy to keep the binding posts tight. Certain types of closed-circuit type plugs or the equivalent could be used to hook into an external power source. The MN1500s appear to take several partial "recharges," so maybe it would help to plug into the car battery with the internal cells still connected: a blocking diode (1N91) is suggested to avoid trouble from reversed polarities or low external voltage.

Converter. (Fig. 5.) The 50 to 7 mc converter was originally adapted from a QST article. Like most such, there were enough changes that there is no justice in involving the original author. The main thing is to remember that there is no substitute for a good transistor in the first stage. The rf amplifier is used grounded-base, for reasons now only a dim memory. A silicon computer diode (not just any type, check the numbers) is used as a limiter at the input in hopes that strong 50-mc signals will not immediately melt the first transistor. Transmitter leakage is not bad, but what if the transceiver with whip attached is carried by a mobile just when he goes on transmit? A good check on the diode is that it should make no difference on weak sigs, on or off.

The double-tuned coupling helps keep Radio Moscow off six meters, but if some oscillator juice should get back to the rf input, there is still a chance of interference. We used 0.010 copper (sheet? foil?) for shields. The oscillator circuit is a legacy from the QST article. The circuit shown in the transmitter, adapted for PNP, should work as well. The main thing is that the oscillator should start reliably and not move around with voltage changes. Required drive to the mixer is a couple of milliwatts. Mixer injection is somewhat fussy. Because of the fancy bias network, mixer current will only change a few per cent with the proper amount of injection. If in doubt, use less, so as to reduce the birdies and the spurious responses.

Roughly speaking, the 50-mc input impedance is the same grounded-emitter or grounded-base, about 30 ohms at the usual currents. The rf stage might as well be a premium type (still under three dollars) and the rest can be any type recommended for TV or FM use. See the section on the transmitter for tips on grid-dipping. The mixer plate lead feeds a

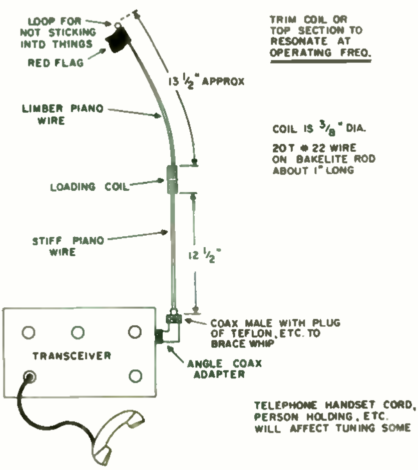


FIG. 4

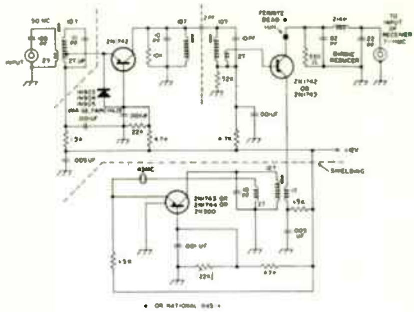


FIG. 5

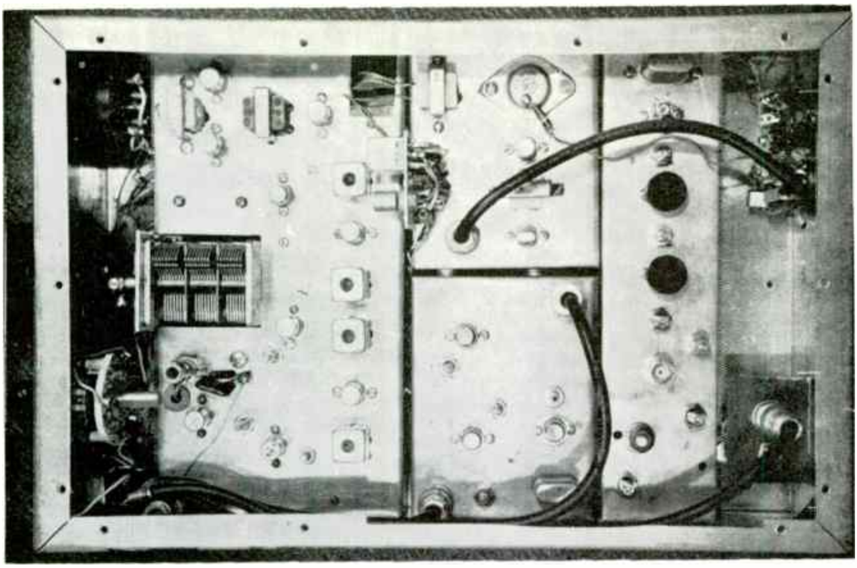
All resistors 1/4 watt, all capacitors mmfd, all coils on 1/4" ceramic tuned forms.

pi-section low pass filter which is intended to reduce the quantity of 43 mc signal delivered to the 7-11 receiver. If this is very strong, it will combine with harmonics of the second oscillator in the second mixer to give birdies (in pairs) each time the 2LO harmonic gets 455 kc away from the 1LO frequency. The pi-section interstages are also intended to reduce the transmission of 43 mc. Tapped-coil coupling, used in the prototype, was quite bad in this respect, as the taps seemed to resonate in the 40-50 mc region.

The converter and receiver amplifier stages are unneutralized. Per-stage gains are fairly low by design, and the MADT, PADT and "drift" transistors used have very low collector-to-base capacitance. They are cheap enough that there's no point in substituting.

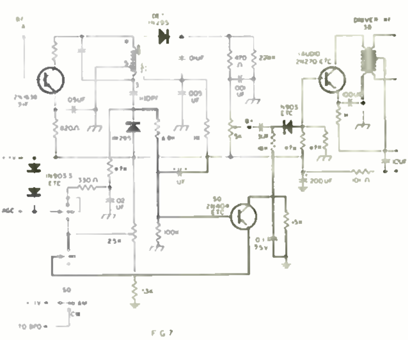
The 7-11 mc "tunable *if*" is a complete receiver in itself. (Fig. 6.) Rf gain is fairly low to reduce overload in the broad-tuning stages, but there should be no difficulty in getting all that's needed in the 455 kc amplifier. There is a tuned rf stage, a mixer and separate oscillator, two stages of *if* at 455 kc, a diode second detector, a separate agc rectifier, and two stages of audio. Optionally, there is squelch and a BFO, with provision for manual gain control. Audio output power is enough for mobile use. Most of the parts can be found in the Lafayette catalog, or salvaged from defunct transistor radios. The tuning capacitor used was 365 mmf per section and had nearly semicircular plates, so that with the series capacitors shown, the low end of six is spread a lot, but the high end is still covered. Try and get a sturdy one, and mount it so that straining the case won't twist the capacitor frame.

If you have never built and tracked a superhet before, this is one heck of a time to start. The main idea is that two gangs of the variable capacitor tune circuits from 7 to 11 mc/s, while the third, in this case, tunes an

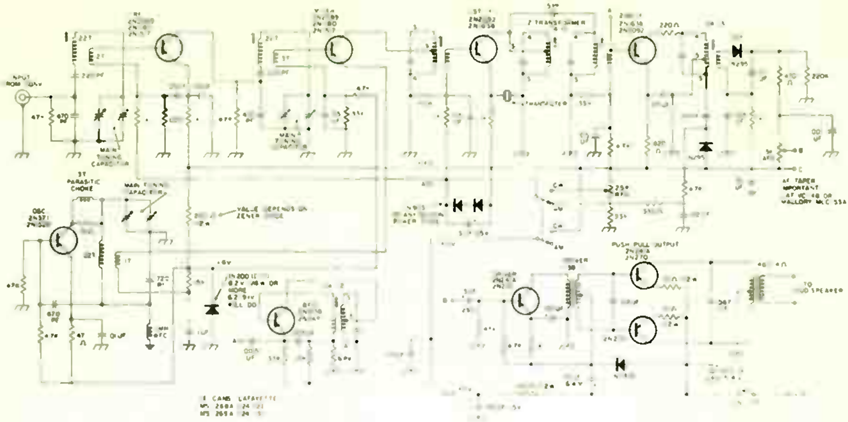


oscillator from 6545 to 10545 kc, 455 below the signal frequency. The trimmers built into the tuning condenser (if there are none, wire some 3-30's in) are used to make things come out on the high end, and the slugs in the CTC coils are used to put the coils on the right frequency at the low end of the range. Since the slugs move things by the same percentage at each end and the trimmers have about twice the effect at the high end as at the low end, trimming at each end alternately several times is likely to converge to the proper condition, where the circuits track all the way. It is very convenient to have some sort of signal generator or test oscillator on hand when doing this, but a grid dipper can be used as a makeshift substitute. The 220, 470 and 720 pf capacitors which are in series with the coils and tuning capacitors should be silver mica type, and within about 5 per cent of the right value. Do *not* use "BC" or "GP" type ceramics—in fact any ceramic capacitor that seems small enough to use is probably not good enough to bother with. Arco CM15's or DM15's come small and with excellent electrical characteristics.

In this receiver, AGC is applied only to the 7 mc rf stage and the first *if* amplifier. AGC on the 50-mc rf stage might help, but was not included. It is not desirable to cut the gain or collector current of the second *if*, as all the power it can deliver is sometimes needed. In order to get good AGC action with transistors, the last *if* amplifier must deliver enough power to the AGC rectifier to allow it to develop sufficient voltage to buck out the forward bias normally applied to the base circuits of the various rf and *if* amplifiers. For a 50 K ohm bias divider the required current is 12.6/50,000 or about 250 microamps. 250 ua through 5,000 ohms (4,700 plus 330) is 1.25 volts. The power needed is thus about 1/3 milliwatt under carrier conditions or 1.25 mw peak. If the last *if* stage is running 1 ma of collector current at about 11 volts, there is a maximum of 5 mw rf power output available class A into a 10K load. If one fourth of the theoretical maxi-



C1 may be changed to 150 or 200 mmfd if strong locals still overload or block on AM.
Changes shown in heavy lines.
* RF gain in CW pos.—sq. in sq. pos.



mum is needed to develop the DC for AGC use (assuming a perfectly efficient rectifier, which is not likely at one volt out) things are pretty thin, and we cannot afford to cut the collector current of the second *if* stage at all. If the taps on the last transformer are wrong, no signal, no matter how strong, will develop cutoff bias, and the receiver will wipe the modulation off strong local signals. If there is some doubt, feed the BFO into the last *if* input full strength and see if the emitter current of the first *if* (as measured by the voltage drop across the 1K resistor) can be forced to zero and beyond. The "beyond" is to allow for modulation peaks.

The control voltage in the absence of a signal is clamped by a couple of silicon diodes to about 1.1 volts forward bias. When the AGC starts working, the diodes unclamp, but the *if* voltage at the AGC rectifier has to be about a half volt before this takes place, that is, the AGC is "delayed." Any good silicon junction diode—power, top hat, alloy, diffused, or planar—will work here as the AGC clamp. The types suggested are small. The detector and AGC rectifiers are 1N295, similar to 1N34 but small and tested as a dectector.

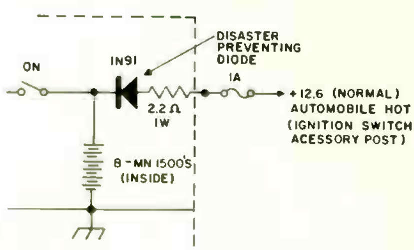
When a squelch is used, it is desirable to set tihngs so the squelch opens before the AGC starts to work. When the BFO is turned on, the AGC must be disabled, and to do this, we shift the line to a manual gain pot. The clamps still function to limit the maximum-gain forward bias to a safe value.

The oscillator circuit is a modified Colpitts. It is not as easy to get going as a vacuum tube oscillator because there is not any amplitude-controlling mechanism in a transistor oscillator equal to the grid-leak-and-condenser we are used to in tubes. Because the oscillator is around 7 mc, and a good oscillator transistor should be suitable for much higher frequency (so that the transit time, which varies with voltage and current and temperature, will be small compared with ninety degrees at operating frequency) there is danger of parasitics at higher frequencies, so a parasitic choke was put in.

As in the first mixer, the amount of drive is critical, and a bit of fiddling with the one-turn pickup coil (move it along the form to vary injection) will be needed. The oscillator runs on the low side with the values shown.

The *if* amplifier as drawn has effectively five tuned circuits. The Clevite TF-01A resonant emitter bypass has about as much rejection of off-frequency signals as another transformer. If you can't get one, use an 0.1 mf capacitor in place of it. The *if* transformers are small ferrite-core jobs, apparently U.S. made, which have an unloaded "Q" of about 140 at 455 kc. As the drift transistors have higher output impedance than the alloy types which were common when these were designed, we have our choice of more gain (by tapping the collectors up) or more selectivity (by using the former collector tap).

We went for selectivity, except in the last transformer, which is heavily loaded by the diodes. Most 455 kc transistor *if* cans found



in small six-transistor radios can be used in place of those specified; the connections are fairly standard. If more selectivity is needed, two coils coupled by a capacitor could also be used between the mixer and first *if*, with only a small reduction in overall gain. The coupling capacitor between the paired coils is run to the tap, so that a larger, easier to control capacitance can be used.

To align the *if* amplifier, a signal may be "stolen" from another receiver, by wrapping a wire around the last *if* plate lead in your HQ 129 or whatever, and tacking it to the base lead of the mixer in the transistor receiver. Tune the HQ 129 to a nice strong broadcast station and start twiddling screws. The single-tuned coils are just peaked for maximum output measured with a Simpson, etc. across the audio gain control (about one volt at most) and the double-tuned pair is adjusted by clipping about 100 mmf across the primary terminals of one can and trimming up the other, then moving the 100 mmf to the primary terminals of the other can and peaking the first.

The audio amplifier is conventional. It differs from most small transistor radio audio amps in that it runs on 12 volts and puts out half a watt rather than 9 volts and a quarter watt. The idling current of the class-B amplifier should be about 2 ma. Less gives scratchy quality, more uses too much electricity. The diode gives a no-signal bias which varies with temperature in the right way to compensate for the thermal characteristics of the output transistors. If a 1N2326 is not available any cheap alloy PNP transistor may be used, connecting to the base as cathode and hooking collector to emitter for anode. A small adjustment in the idling current can be made by changing the value of the emitter resistor in the first audio stage.

The BFO circuit shown will work with almost any alloy transistor, but the loading on the signal circuits will be a little less with the drift unit specified. Any interstage *if* transformer will work. The BFO is tuned to band center. No pitch control is provided.

If squelch is desired, it is inserted as shown in Fig. 7. The manual rf gain pot is used to set the squelch level, when that function is in use.

In the first model, one transistor was switched between BFO and squelch. The BFO was too near the front end of the receiver, and BFO harmonics were all over the dial. Segregating the BFO fixed this: in the photos the BFO assembly is tacked on above chassis. In the receiver shown in the photos, a transistor is in the squelch socket, but the socket is not wired.

A list of possible transistor types for the receiver is given in Table 2. The recommended types cost from fifty cents each for some audio types to about a dollar each for the 7-11 rf and two and a half for the hot six-meter rf stage. You may find something satisfactory in your pickle jars full of slightly surplus semiconductors, but it's not too likely. The types used in Japanese AM (not AM-FM) radios will not be suitable. Required collector breakdown voltage is at least 20, and 30 is better.

. . . W1OOP

Table of Transmitter Transistor Types

A. Silicon NPN RF Power transistors
Oscillator: Some 2N697, 706, 707, 708 some 2N718, 753, 759, 760, 913, 914, 915, 916, 957, 2N834, 2N1338, 1505, 1506, 2297.
Doubler: 2N707, 708, 915, 1505, 1491, 2297. If modulated, may have voltage breakdown problems. 2N1506, 2N1492, 2N1493, 2N1342, 2N2218, 2N3118 should be O.K. modulated.
250-miliwatt final: 2N1506, 2N2876, 2N2631, PT531, TA2084, 2N1978, 2N3118. Nothing more than five years old.
1.5-watt final: 2N2876, TA2084, PT657.
NPN AF amplifier: 2N35, 78, 167, 169, 214, 388, 445, 634, 635 etc.
PNP AF driver: 2N43, 188, 241, 270, 396, 404, 407 and many more.
Modulator: 2N1172, 2826, 2827, 2N301, 176, 276, 342, 553, 554.

Table of Receiver Transistor Types

Converter-RF 2N1742, 2N2494, 2495, PADT-28, 2N502A, RCA 2N2873, Philco T 1694, 2N1177 (last choice)
Mixer 2N1743, 2089, 1177, 1179, 2N1745, 2N1517
Osc. 2N1744, 1743, 1178, 2084, 1517, 1745, 1868, 2N501
7-11 RF 2N2089, 1180, 1517, 2084, 2N384, 370, 2N1726, 1747
mixer 2N2089, 1180, 1517, 2084, 2N372, 2N274, 2N247
oscillator 2N371, 1526, others will work, may need changes in feedback.
IFs, 2N1638, 2N2092, any listed above, for RF or mixer
BFO almost any computer or drift transistor, 2N1631, 1637, 247, 274. Squelch-anything.
AF driver & af output 2N270, 241A, 188A 525, 2N43A 2N1413, 2N1924, 2N1192
Zener diode = 1N200, or any up to 1½ watts, 6.8 to 8.2 volts nominal.

Note on transmitter transistors: The types used are Silicon NPN. If suitable transistors are not available on a beg, borrow, or buy surplus basis, it may be advisable to consider using the Amperex 2N2786 PNP germanium unit, announced some time after these transmitters were built. (Amperex Electronic Corp., 230 Duffy Ave., Hicksville, N. Y. has report #S113 on how to use it.) The main disadvantage is that the 2786s we have tested had roughly 30 volt collector breakdown, so that AM at 12 supply volts is not practical. The proper solution is to use series modulation from a 12v supply, or NFM, or a collector supply dropping resistor, adjustable so as to set up for maximum collector voltage that the transistor will stand and still modulate properly. The Amperex report suggests that 2N2207's can be used in the driver stages. The driver transistor costs under two dollars, the 2N2786 under five. Carrier output level for AM would be about 140Mw. For NFM, about half a watt out could be obtained.

CHAPTER 24

Midget Six-Meter Transceiver

The new Philco MADT transistors, though inexpensive, have brought about a minor revolution in VHF equipment and permit, for the first time, really miniature equipment to be constructed. Two of the transceivers shown here were made to illustrate this application. The unit uses five transistors in a superregenerative receiver and crystal controlled transmitter and modulator

While not designed for the DX'ing crowd on six meters, this little gadget has received over an 80 mile path using an inside dipole. The more usual range is about a half mile between two identical units using built-in whips. Considering the simplicity of the rig it is difficult to imagine why any amateur who occasionally travels wouldn't pack one of these little gems in his suitcase so he could get in touch with local hamdom.

Circuit

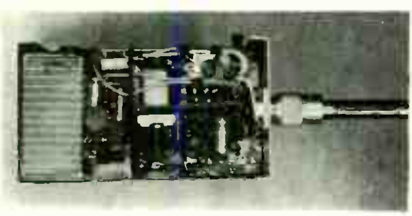
Q1 is a single transistor superregenerative receiver! Perhaps just a word should be put in here in support of this type of receiver. Heath uses this in their Sixers for the circuit is not only extremely simple, but very sensitive. It takes quite a superhet to do better on sensitivity. Selectivity suffers, and you can have some real problems if a very strong signal comes on near your frequency.

The detector is reflexed in that detected audio is fed back to the base of Q1 through C2 and is amplified. The operating point of Q1 is established by R1, R2, R3 and R4. R4 controls the regeneration. Tuning is achieved by varying C5. C7, C8 and R5 form a low pass filter to prevent the quench signal from overloading the audio section.

The transmitter uses two transistors, Q2 a fifth overtone oscillator and Q3 the final class C amplifier running a mighty 50 mw input. The output is on the order of 25 mw, down somewhere in the microbe-power division.

The audio/modulator uses two transistors, RC coupled, and a permanent magnet speaker which doubles as a dynamic mike. The receiver output transformer primary is used as a choke for Heising modulating the final on transmit, giving a healthy 90% modulation. Neutralization of the final might permit slightly higher modulation. The D1-C21 circuit prevents high voltages from the Heising choke from damaging Q5.

A four-pole-double-throw switch transfers all the circuits. Power is supplied by the usual 9 volt battery. In this unit a 5¼" x 3" x 2⅛" aluminum minibox was used and the circuit was mounted on a piece of double-sided copper-clad printed circuit board. The double-sided board acts as a fine shield between the audio and rf circuits. This board is excellent construction material for it is easily cut, drilled and soldered to. Shields can be soldered to it, making an extremely rigid assembly.

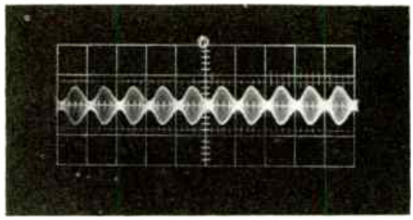


The whip antenna was mounted using a coax connector (photo) so that other antennas could easily be connected for better DX. The whip can easily be permanently mounted on a plexiglass or micarta mounting plate.

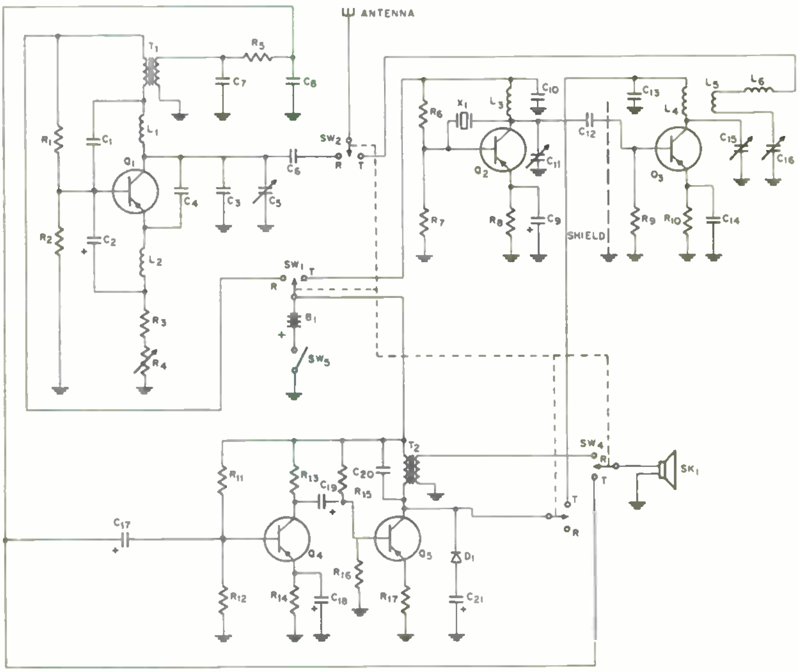
Tune-Up Procedure

The tune up procedure is straight forward. With S5 on, receiver regeneration control R_4 is adjusted for a strong rushing sound in the speaker. Final trimming of R_4 is best achieved while listening to an incoming signal. Capacitor C_3 may be adjusted slightly for bandsetting, and tuning capacitor C_5 should cover 2 megacycles of the six meter band. Quieting of the rushing sound should occur with an incoming carrier.

Transmitter tune up is also straight forward. The oscillator and final tank circuits can be initially set with a grid dipper. With S5 on, depressing the push to talk switch should cause immediate operation of the transmitter. Transmitter operation can be checked with a grid dipper, a field strength meter, or another



Actual photo of the transmitter output modulated with a whistle and observed with a 100 mc scope. Note relatively clean envelope.



Parts List

(½ watt carbon)
R1—39,000
R2—12,000
R3—2200
R4—5000 pot
R5—3900
R6—10,000
R7—1000
R8—100
R9—1200
R10—22
R11—220,000
R12—10,000
R13—3900
R14—100
R15—47,000
R16—4700
R17—22
C1—100 mmfd
C2—2 mfd 10v elect.
C3—10 mmfd
C4—6.8 mmfd
C5—15 mmfd MAPC var. with 4 plates removed
C6—4.7 mmfd
C7—.01 mfd
C8—.05 mfd
C9—.002 mfd 50v
C10—.01 mfd 50v
C11—4-30 mmfd trimmer
C12—15 mmfd
C13—.01 mfd 50v
C14—.01 mfd 50v
C15—4-30 mmfd trimmer
C16—280 mmfd trimmer
C17—30 mfd 10v
C18—30 mfd 10v
C19 30 mfd 10v
C20—.01 mfd 50v
C21—30 mfd 25v
Q1, Q2—2N1499A
Q3—2N1749
Q4, Q5—2N2374
D1—1N34A or equiv.
SK1—2½" 3.2 ohm speaker
X1—6 meter 5th overtone HC6U type
ANT—52" telescopic antenna
L1—10 turns ⅛" i.d. #16 enam
L2 6.8 uh rfc
L3—8 turns ⅛" i.d. #16 enam
L4—8 turns ⅛" i.d. #16 enam
L5—2½ turns #20 hook up wire over L4
L6—9 turns #20 plastic hook up wire ¼" i.d.
T1—Calrad CR60 20K-1K or equiv.
T2—Calrad CR40 1.2K—3.2 ohms or equiv.
B1—9v Battery, Everready #246 or equiv.
SW—4PDT push-to-talk switch, Lafayette SW92

Table I

Receiver: dc input current	15	ma
Osc: dc input current	4.5	ma
dc input power	40	mw
Final: dc input current	6	ma
dc input power	50	mw

receiver. If the oscillator fails to start when the push-to-talk switch is depressed, adjust C_{11} until it does so. Final transmitter adjustment is best achieved with the antenna fully extended and the case closed and held steady on a table with one hand. With a field strength meter nearby, adjust C_{11}, C_{15} and C_{16} for maximum output. Check to make sure the oscillator starts easily by depressing the push-to-talk switch several times. It should start each time. If it does not, back off on C_{11} slightly. Check for modulation by listening with another receiver. The signal should be crisp and clear. It is not necessary to hold the "mike" close or shout. Normal talking four to six inches from the "mike" should permit full modulation.

These little transceivers are a lot of fun, whether you pull them out at the local ham club meeting, talk all around conventions, meet hams in towns you are visiting, or even hook it up to your big beam and astound everyone. Reports of 5-9-plus have been consistently received over 20-mile paths with a four element beam. A little mountaintopping with this and a portable beam is a lot of fun and something you'll never forget.

. . . K3NHI

CHAPTER 25

160 Meter – 6 Watt Transmitter

If you would like to be one of the first hams in your area to go on the air with a home-brew transistor transmitter, this rig should appeal to you. It operates on the increasingly popular 160 meter band where TVI is no problem, and the average builder will have no troubles with feedback or neutralization. This rig is economical (a full set of RF transistors and audio transformers costs less than seventeen dollars), easy to build, and features conservative cool running design.

Other advantages of the rig are zero standby current, instant warmup, portability, 30% or more overall efficiency, no need for a high voltage supply, and the added bonus of having a 5 watt portable public address system available.

The signal is only 1½ to 2½ S units down compared to a 50 watt tube type rig on the other fellow's receiver. Good ground-wave coverage is obtained here in the city on phone, and more than once I could sense an upraised eyebrow when I reported the station power as two watts. On CW, Canada and nearby states have been worked with ease.

Circuit description

Four basic units comprise the transmitter. One is the chassis/panel assembly containing switching and metering, and serving as a mount for the three sub-assemblies. One of the subassemblies is the complete audio section, and the other two are RF sections. One of the RF sections is the final RF transistor and its heat sink; the other consists of a Pierce crystal oscillator, the grounded base buffer/driver, and RF final tank circuit components.

The audio section has two amplifiers, a driver, and a push-pull class B final, used either as the modulator, or, in the public address mode, as an audio output stage.

The RF sections all have slug-tuned, fixed frequency coils, since operation is at the one crystal frequency. Over 95% modulation is obtained by modulating the buffer/driver along with the final, rather than the usual method of modulating the final stage only, with its resultant 80 to 85% usual modulation limit. Link coupling is used throughout.

PTT (push to talk) is accomplished by SW2 serving as an on/off, antenna transfer, and receiver muting switch. Mode of operation is determined by SW3 for CW, Phone, Spot, or Public Address operation. Meter versatility is provided by SW1, to measure DC input voltage (0-20v), DC input current (0-1 amp), or RF rms output voltage (0-20v).

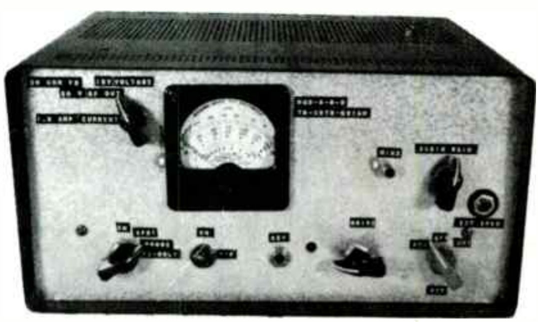
The power supply can be any well filtered DC source of 11 to 14 volts at 1 ampere. The author uses surplus four amp/hr Ni-Cad batteries, eleven in series, to provide about 13 volts for up to ten hours of operation before requiring recharging.

The silicon diodes connected back to back across the meter make it almost burnout proof, without appreciably affecting its accuracy.

Construction

Complete the chassis/panel, switching and metering assembly first. It then facilitates construction of the three sub-assemblies as each is built and tested.

The complete audio section is built on a 4" by 8" flat tin plate, and mounted flat on



the top of the chassis on the far right, as viewed from the front. Q4 and Q5 are mounted on the same large heat sink, but insulated from it by mica insulators. Q1, Q2 and Q3 do not need heat sinks. All audio components are mounted on this plate except the mike input jack, gain control R8, and the external speaker jack.

The RF final transistor, Q8, is mounted directly on its heat sink without any insulation, and the heat sink itself is insulated from the chassis. This section is mounted vertically in the center on top of the chassis, to the rear. Q8's base, emitter, and collector connections are brought over to the other RF section containing all other RF components.

Next assemble the other RF section on a 4" by 8" plate, and mount vertically along the far left top side of the chassis. Use in line construction, with the oscillator at the front, buffer/driver in the middle, and final tank circuit and decoupling components to the rear. The key jack and RF drive control mount on the front panel. Q6 and Q7 have fin type, press-on heat sinks.

Mounted on the rear of the chassis are the antenna input, output to receiver, output to receiver muting (grounding type), and power supply input jacks.

The photo indicates an on/off switch under the meter, and a center off position on the PTT switch. These were eliminated, as the schematic shows, because with the instant warmup inherent in transistor equipment, the PTT switch receive position becomes the off position. Also, the photo indicates different full scale meter values, due to the scale markings already on the meter used by the author.

Use sockets for all the transistors, as it helps greatly in testing, substitution, and enables short lead surplus transistors to be used.

Testing and alignment

The meter current shunt R7 is approximately 10 feet of #28 plastic covered copper wire, jumble wound, and adjusted in length until the meter reads full scale with one ampere current flowing through the meter circuit. Next adjust the voltage multiplier R6, until the meter indicates full scale with 20 volts DC applied. Adjusting R5, the RF rms voltage multiplier resistor is a little more complicated. Disconnect the final tank coil link from CR1,

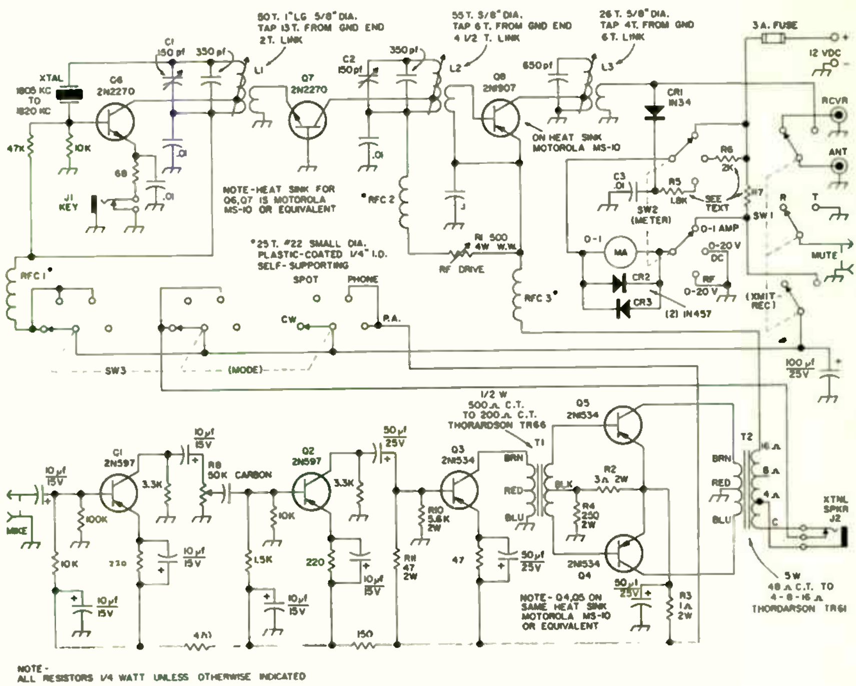
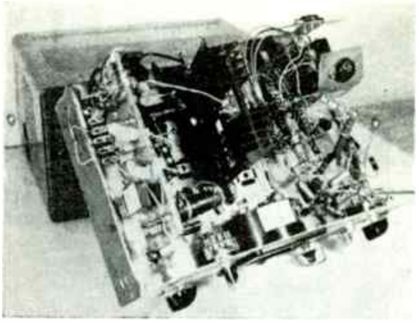


Fig. 1. 6 solid watts on 160 meters.



Neat interior construction is not all-important on 160 meters.

shunt C3 temporarily with a 1 μf paper condensor, and apply 6.3 volts from an AC filament source to the input side of CR1. Adjust R5 to give a 6.3 volt reading on the meter. This method gives reasonably accurate RF voltage readings for measuring the transmitter's output.

Test the audio section with a speaker connected to the secondary of the modulation transformer T2. The insulated speaker jack, J2, shoud be carefully checked to be sure it isn't grounded to the chassis; it would then be a direct short for the DC supply. If audio distortion is encountered, adjust biasing resistors R4 and R10 for minimum distortion, with a total idling current of about 75 ma for the entire audio section.

Substitutions may be made to keep down the cost of building, and in fact are recommended for all parts but the three RF transistors and two audio transformers.

Next, get the RF oscillator working. The buffer and RF final will draw no current until driven by the oscillator, as they are biased beyond cutoff without signal input. All three RF coils may be grid-dipped to approximate operating frequency, while wired in, transistors in sockets, and power off. The oscillator coil must be capable of being tuned slightly below crystal frequency, so as to provide the proper inductive reactance necessary for oscillation.

After the oscillator and buffer/driver are operating properly, connect the RF final transistor to its tank circuit and peak all circuits while operating into a dummy load (a 50 ohm 2 watt carbon resistor). Keep the RF output below 4 watts (14 RF rms volts on a 50 ohm line) when on CW, and at 2 watts (10 volts) on phone. If you go above these limits, you risk exceeding the peak collector current ratings of the RF final and early transistor failure, plus non-linearity on phone. At two watts out, the RF stages draw a total of 500 ma. On phone, this rises to 750 ma on audio peaks.

Always check and double check polarities, base, emitter and collector connections, and never operate the audio or RF stages without their proper load.

One requirement of this rig is that it must look into the proper antenna load, which should be a resonant antenna of 50 ohms impedance. The author, whose city lot can't accommodate a full size 160 meter antenna, uses his 40 meter dipole on 160 meters by letting the braid of the coax feeder float, and feeding the center conductor only, from an antenna coupler, which is adjusted with the aid of an SWR meter to the proper 50 ohm impedance. Very good results have been obtained from this antenna configuration, which is probably operating as a shortened vertical with top capacity loading!

The author hopes you will enjoy building and operating this rig as much as he has.

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CHAPTER 26

432-MHz Exciter

When I was first contemplating the construction of an exciter for my ATV station, I had hoped to have the entire unit solid state.

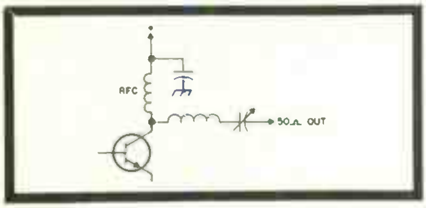


Fig. 1. Alternote output circuit to provide 50 ohm output for the 2N2950.

However, at that time, the prices of the transistors were quite high. I then settled for two tubes and a varactor.[1] Since then prices have come down, as much as 50% on one transistor in particular. It was decided that the plunge had to be made even if for no better reason than my own personal satisfaction.

Although the ultimate goal of the circuits shown is 432 mc energy, these can be broken down to give output power at the following frequencies:

48-50 mc---- 3 watts
48-50 mc----20 watts
144 mc----13 watts
432 mc---- 8 watts

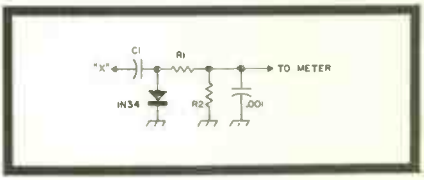


Fig. 2. Typicol metering circuit. One needed for each stoge.

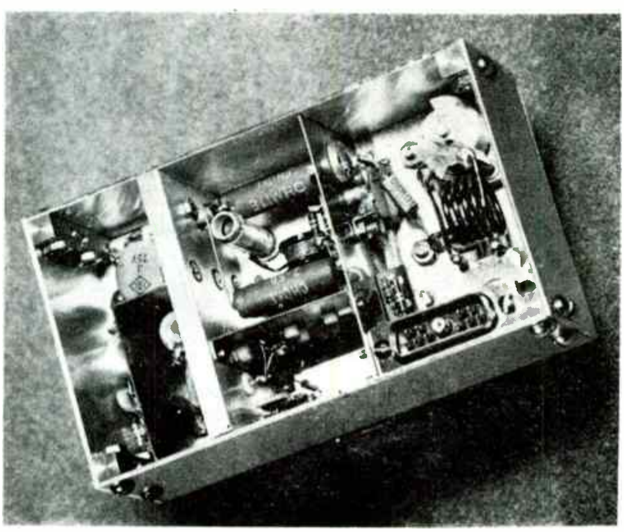
From the above it can be seen that by using the transistorized exciter alone, up to 20 watts can be obtained on six meters. By following this with the first varactor tripler, we then have a 2 meter exciter and of course the second varacter stage gives us the 432 exciter. If only 3 watts is desired on six, then the first three transistors are used. Refer to *Fig. 1* for an alternate output circuit using Q3 to drive a 50 ohm load. With this arrangement I have obtained up to 4.5 watts output with the 2N2950 but this is driving it rather hard. The 3 watt figure given is a safer and more conservative amount.

One thing I wish to bring up at this point is that all of the above figures are under CW conditions and at 24 volts. If amplitude modulation of the transistorized stages is desired, then the collector voltage must be kept down to around 12 volts (on the modulated transistors) with a resultant decrease in output.

It may seem strange to some but my first thoughts on construction was of heat sinks. Since compact construction was in order, a large dissipator was out of the question, yet a big dissipating area was needed for Q4. This paradox was solved with an IERC TO3-250-200. This sink provides ample dissipation while presenting a small area on the chassis base. This is necessary in order to maintain short lead construction under the chassis. Transistor Q3 is mounted on a piece of 1½" x 2⅝" x ¼" aluminum which doubles in duty as an interstage shield. Transistors Q2 and Q1 use smaller IERC dissipators in order to keep them down to proper temperature limits.

One of the most difficult things to get used to when using transistors is the low impedances. Once this fact is accepted, no trouble should be encountered. All circuitry is straight-forward without any fancy frills. In fact most of the basic circuitry was obtained from the data sheets on the Motorola transistors and varactors I used. The inputs of Q2, Q3, and Q4, as well as the output of Q4, are metered by the rectified RF method (see *Fig. 2*). A diode fed by a small capacitor samples the RF and feeds a voltage divider to provide sufficient DC to give an indication on a microameter. No values are given for these sampling circuits because these will change, depending upon what collector voltages are used, etc. Some may prefer to measure the collector currents but this is up to the builder.

The transistor exciter (Fig. 3) is built on a piece of 4⅞" x 2⅞" 16 gauge aluminum. The corner posts are made of drilled and tapped ¼" square brass rods. Four pieces of aluminum fasten to these rods to form a complete chassis. The two varactor tripler stages are built in a 1⅝" x 2" x 10" minibox; however, the entire circuitry can be built on a single larger chassis. The two chassis construction I used was to suit my particular requirements in my ATV project. The only things to be sure of in construction is to have short leads and proper shielding between stages. Power and RF connections on both chassis are brought out on miniature connectors which have their mates on a master chassis in my project. The coax connectors on top of the units shown here are just for testing. However, the builder can choose whatever method serves him best.



Bottom view of the 48 mc driver. Note Q3 mounted on ¼ inch heat sink/shield with a shield between the base and collector.

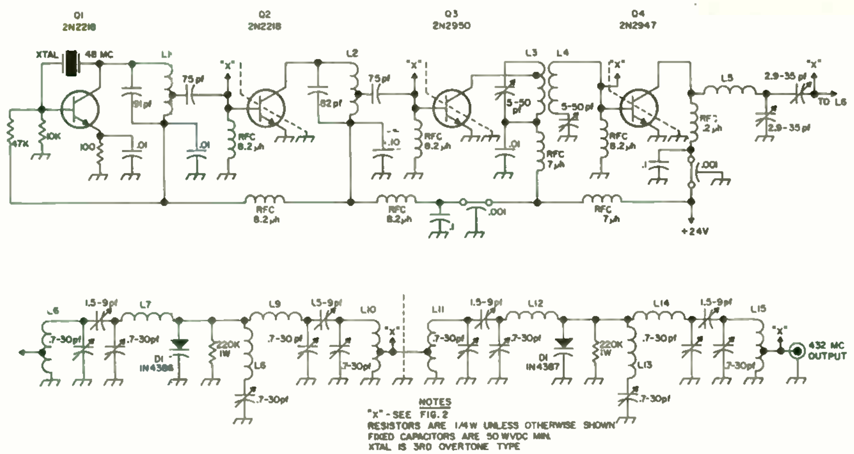
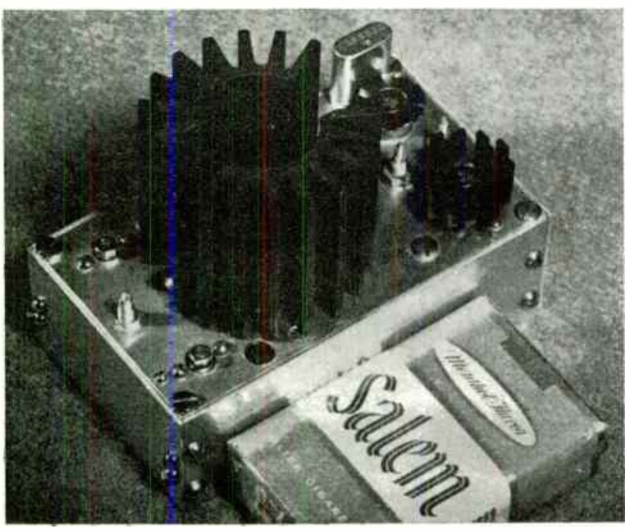


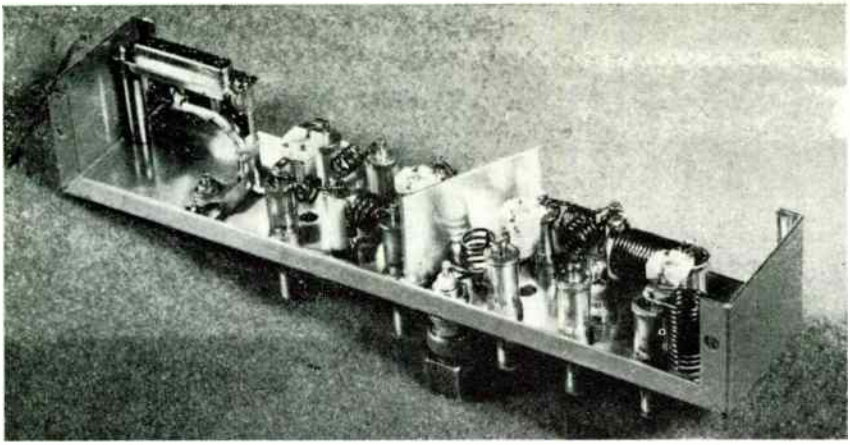
Fig. 3. Solid-state 432 mc exciter. The top section is the 6 meter driver good for 20 watts output. The bottom is the dual tripler using two varactors to get to 432 mc. Heat sinks are: Q1, IERC TXBF-032-025B; Q2, IERC LP5A1B; Q3, see text; Q4, IERC, TO 3-250-200.

When tuning up for the first time, it is recommended this be done stage by stage and disconnecting power to the preceding stages. All three amplifier stages are run Class C so when drive is removed there is no current flow; however, a stage that has drive but has its output out of resonance can be damaged. Tune up can be done much more safely with a lower voltage and a regulated supply with high and low voltage output is recommended.[2] If the exciter is tuned up at low voltage it will have to be re-peaked when full voltage is applied. This is because the junction capacity varies with voltage but all tuning will be quite close because of the high C circuits used in the outputs.

The two varactor triplers are of straightforward design which have been described many times before and we need not go into it again. The same applies to tune up but it is recommended that the first stage be tuned by itself before the second stage is connected. I have a



Top view of the 48 mc driver. This unit will produce 16-20 watts depending on the individual transistors used.



Bottom view of the varactor multiplier stages of the solid state 432 mc transmitter. On the right is the multiplier from 48 mc to 144 mc. Left of the shield is the multiplier from 144 mc to 432 mc.

jumper made of two right angle BNC connectors for tune up purposes. This is normally left connected but can be removed and a directional wattmeter inserted in the line for checking efficiency, etc.

Of course 432 mc is not the high frequency limit to the use of varactors. Quite the reverse is true in that varactors perform well and are most practical as the frequency goes up. The exciter described here could be followed by another varactor tripler (such as the Motorola MV-1808 as an example) and give about 4.5 watts output at 1296. Although I haven't tried this scheme yet.

1. "A Hybrid 432 mc Exciter."
2. "A Regulated Solid—State Supply."

Coil Table

L1 and L2	4 T #18¼" dia. Tap at 1¼ T from cold end. ¼" long, slug, tuned. X
L3	4½ T #18 7/16" dia. Tap at ¾ T from collector end. ¼" long.
L4	4½ T #18 7/16" dia. ¼" long.
L5	4½ T #14 9/16" dia. ½" long.
L6	12 T #16 ⅜" dia. Tap at 3 T from cold end. 1" long.
L7	13 T #16 ½" dia. 13/16" long.
L8	4 T #16 5/16" dia. 7/16" long.
L9	4 T #16 ⅜" dia. ½" long.
L10	4 T #16 ⅜" dia. Tap at 1 T from cold end. ½" long.
L11	4 T #16 ⅜" dia. Tap at 1 T from cold end. ½" long.
L12	4 T #16 ⅜" dia. ½" long.
L13	3½ T #16 3/16" dia. ⅝" long.
L14	2 T #16 ¼" dia. ⅜" long.
L15	3 T #16 3/16" dia. Tap at 1 T from cold end. 13/16" long.

CHAPTER 27

Varactor Tripler to 1296 MHz

With quite a few varactor triplers being used here in the Los Angeles area on 432 mc, an easy way to get on 1296 with a clean stable signal is to add another varactor tripler.[1] The tripler described here takes 5 watts

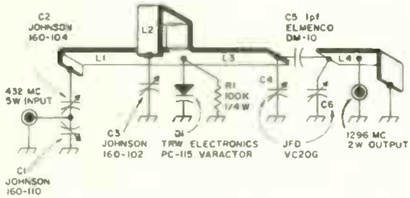
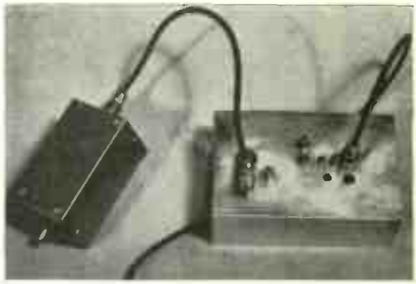


Fig. 1. Schemotic of voroctor tripler to 1296 mc.

input at 432 mc of FM or CW and triples it to 1296 mc with an efficiency of 40% or 2 watts output. AM can be tripled but the maximum input power cannot exceed 3 watts. The output can be used to drive a 2C39 type amplifier for higher power. The total cost using all new parts is less than $25.00.

1. Varactor Tripler to 432 mc's.

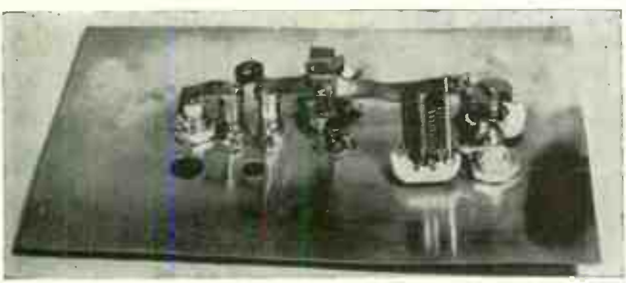
L1, C1, C2 are tuned to 432 mc and match the 50 ohm input to the varactor impedance. L2, C3 is series tuned to 864 mc as an idler



Tripler to 1296 driven by tripler to 432 (left).

so that the 2nd harmonic energy is circulated only through it and the diode to mix with the fundamental. L3, C4, and L4, C6 are two tuned circuits at 1296 mc to select only the 3rd harmonic energy, clean it up, and match to the load. Cavities can be used for less circuit loss and greater parasitic suppression but the added cost and effort does not seem to justify the slight improvement. The resistor R1 provides a DC return and self bias for the varactor so no power supply is necessary.

The cost of varactors jumps up sharply for input frequencies above 400 mc and powers above 5 watts. The PC-115 is $12.00 and for powers up to 16 watts input a TRW PV-002 for



Loyout of voroctor tripler. 432 mc input is on right with 1296 mc output on left. Voroctor con be seen in center in front of idler components. Note two extra holes.

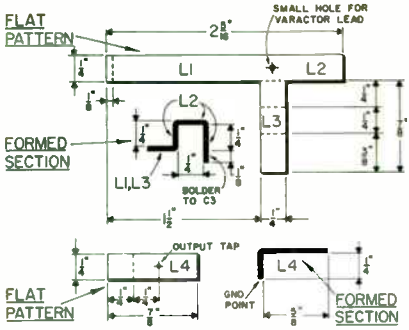


Fig. 2. Patterns of inductors.

$16.00 can be used. I recommend the PV-002 if you use the Hybrid 432 Exciter which is a great rig. They are available through most mail order parts houses. Newark Electronics at 223 W. Madison St., Chicago, handles the varactor. *Care must be exercised* if other varactors are used since dissipation, efficient frequency range, and impedance are very discrete.

Cut out the inductors as indicated in Fig. 2 from a piece of sheet copper.

Construction

Layout the brass board as shown. The board itself takes the place of the bottom plate on a 4 × 6 × 2 box chassis.

Refer to the pictures for parts placement and position. The varactor holder is made from a Cambion PLS6 coil form. If a PV-002 is used the form is not necessary since it is a stud mount. Break off the ceramic and take off the spring clip.

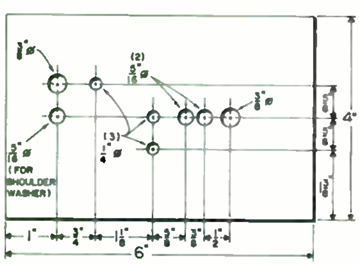


Fig. 3. Layout of tripler.

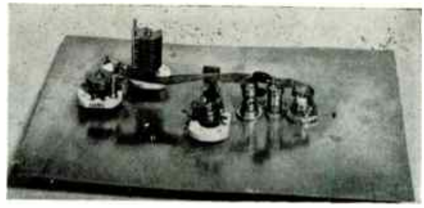


Photo of varactor tripler.

Mount the varactor holder on the board. Gently place the varactor in the holder with the brown band down. You may have to spread the holder fingers a little. Do not force the varactor in as the glass case is fragile. This holder provides a simple but effective ground and heat conductor for the varactor. Assembly from here on is straight forward. Make sure that the rotor of C2 is not grounded from improper seating of the fiber shoulder washer. The cold end of L4 is soldered to the BNC chassis connector body. Solder the hot lead of the varactor with no more than a 47 watt iron using care not to over heat as you would for any semiconductor. The JFD Variable Capacitors are fragile but they are the only ones that are good at this frequency. Take care when soldering around the top of the capacitor and L3 and L4.

Tune Up

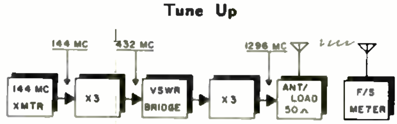


Fig. 4. Tune-up procedure.

The best way to tune up any varactor multiplier is to tune for the lowest VSWR between it and the generator. When properly tuned and matched, maximum power output and minimum VSWR coincide. It is assumed however, that most hams will not have a VSWR bridge capable of operating up to 432 mc so a field strength meter and the old method of "tune for maximum smoke" can be used. The L and C combinations have been chosen such that the wrong harmonic does not fall into any of their ranges. Each adjustment affects the other, so touch up each capacitor several times.

Now you're ready to get on the air on 1296 mc. Remember not to exceed the power rating of the varactor. Good DX'ing.

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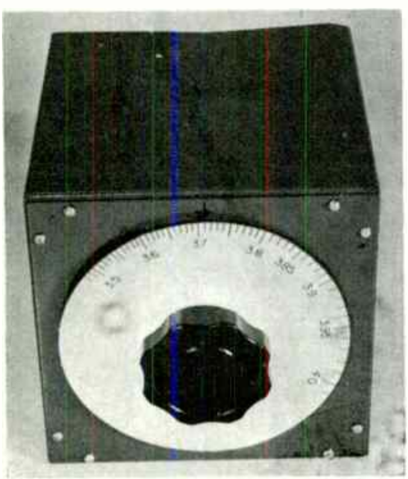
CHAPTER 28

FET VFO for 80 Meters

One of the big advantages of the field effect transistor oscillator is that it does not appreciably load down the tuned circuit. If you take a very close look at the circuit Q versus frequency stability curve you'll find that highly stable oscillators are coincident with high Q tanks. With the relatively low impedances encountered with the bipolar transistor[1], circuit loading is a severe problem which seriously affects frequency stability. In addition, the element capacitance of the bipolar device is very complex, varying with voltage, temperature *and* current. It is difficult to predict exactly what a given device will do because the rate of change of capacitance is a function of the bias level, and varies from device to device. On the other hand, the capacitance of the FET is almost completely unaffected by the source current, and the factors that influence FET capacitances always increase with temperature. Since the properly designed FET oscillator always has a positive temperature coefficient it is relatively easy to compensate.



The 80 meter FET VFO. The dial is made from a National AVD-250 planetary drive; the scale is printed on a paper disc which is glued to a 4 inch aluminum disc. The drive mechanism is on the rear of the panel; the aluminum disc is attached by means of screws and 1/4 inch spacers.

The complex capacitance of the bipolar device has *both* negative and positive temperature coefficients, and for wide excursions in temperature, bipolar transistor oscillators are very difficult to compensate. Never the less, transistors inherently generate very little heat, and in amateur applications this difficulty is often not apparent. I can remember spending three weeks of continuous labor trying to compensate a 3 MHz oscillator that was to be used on one of our space vehicles; the transistors finally had to be changed before a stable unit was obtained.

An example of the frequency variance with temperature of two oscillators, one with an FET and one with a 2N918 bipolar transistor is shown in Fig. 1. Note that whereas the frequency of the FET circuit decreases in a somewhat linear manner, the frequency of the bipolar oscillator first increases and then decreases. The negative temperature coefficient of capacitance dominates when the frequency increases and the positive coefficient as the frequency decreases. This type of curve obviously *can not* be compensated with a temperature sensitive capacitor in the tank circuit. It is also apparent from this graph that the bipolar circuit has a larger drift in frequency for a given change in temperature over most of the range. In many cases the ham oscillator is operated within the temperature range at the crest of the frequency curve where these undesired effects go unnoticed.

In the VFO described in this article, FET's are used in the oscillator and buffer stages and a bipolar transistor is used in the power output stage. The oscillator itself is a conventional series-tuned Colpitts or Clapp cir-

1. The bipolar transistor refers to conventional junction transistors made up of P-N junctions at the emitter, base and collector. In these transistors the current through the junction consists of both electrons and holes (absence of electrons). Because there are two types of current carriers (electrons and holes), these devices are referred to as "bipolar". On the other hand, the field effect transistor is a "unipolar" device because the current carriers are either electrons or holes, depending on whether it is an N- or P-channel device.

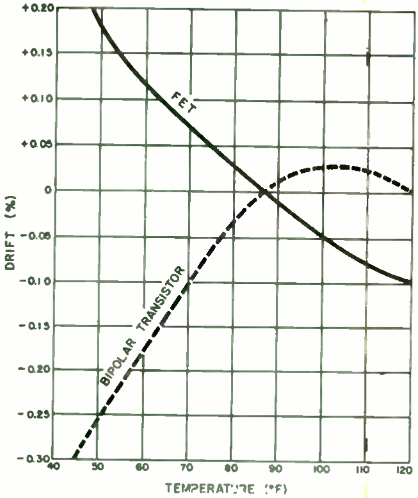


Fig. 1. Typical drift characteristics of FET and bipolar oscillators. This graph is based on a starting point of 30° C (86° F) and shows that the bipolar transistor has both negative and positive temperature characteristics while the FET capacitance has a positive temperature coefficient (causing the frequency to go down).

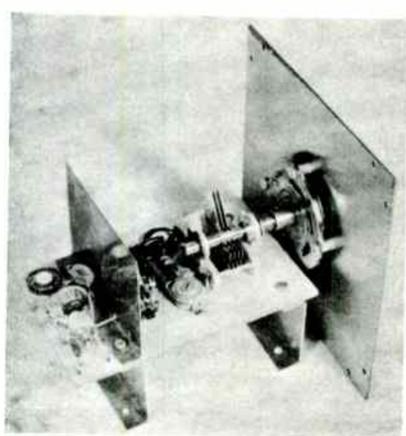
cuit with large silver mica capacitors providing the necessary feedback. The large value of these capacitors tends to swamp out any changes of capacitance within the device itself. The frequency drift of this unit was so small (probably because it was operated at room temperature near no vacuum tube heat generators) that no temperature compensation was required. In most cases no compensation should be necessary, but if thermal drift is a problem, a negative temperature coefficient capacitor may be added in parallel with the 50 pF tuning capacitor as shown by the dotted lines in **Fig. 2**. In the event this compensating capacitor is required, the 75 pF capacitor should be reduced by a like amount.

Since a compact unit was desired, toroidal coils were used to maintain high Q without massive air wound inductors. Since there is virtually no field around a toroidal inductor, they may be placed near other objects in the circuit without affecting their Q. Sometimes this low amount of external field can be a problem because it's difficult to couple a grid dipper to the circuit to find out its resonant frequency. In addition, toroids are quite susceptible to 60 hertz hum pickup, and this is accentuated by the high impedance FET, so care must be taken to shield the circuit properly. In the author's case the hum was completely eliminated when the circuitry was mounted in a metal box.

To reduce loading on the oscillator stage, an FET source follower is used as an untuned buffer stage. The low value of coupling capacitance, 5.6 pF, serves to further reduce the loading on the oscillator. To eliminate any possiblie pulling or resonance effects, no tuned circuit is used in the buffer stage. The stage shown in the schematic exhibits extremely high input impedance and is a very effective isolator; when the output stage is keyed, there is no perceptible change in the rf voltage across the oscillator tank and the frequency remains rock stable.

The stability of the supply voltage to the osicllator is maintained by the 400 milliwatt, 9 volt zener diode (HEP 101) across the drain to ground. With the circuit constants shown, the voltage on the drain of the MPF 105 field effect transistor remains within 1% of the Zener voltage as the supply voltage varies from 13 to 25 volts.

The output stage uses a low cost silicon transistor in a conventional class C rf power amplifier circuit. At 4 MHz this stage has about 25 dB gain and provides 2.0 volts RMS output; this is more than sufficient to drive most transmitters. If the second harmonic at 7 MHz is desired, it is recommended that another HEP 50 stage be added as a doubler. In the original model of this VFO the final stage was operated straight through on 80 meters and as a doubler for the higher bands. Although there was sufficient drive on 40 meters, there was not



Interior of the 80 meter VFO. The FET oscillator and buffer are in front of the shield, the HEP 50 output stage to the rear. The planetary drive is attached to the front panel with spacers. The screws and spacers which hold the dial pass through a 1¼ inch hole in the panel.

enough output on 20, 15 and 10. In fact, at the higher frequencies, the transmitter operated at the resonant frequency of the doubler tank (or its harmonic) and refused to be varied by the FET oscillator!

In this VFO all the components were mounted on a ⅛ inch sheet of micarta 3½ inches wide and 5½ inches long. Soldering terminals were provided by drilling holes in the micarta with a number 42 drill and installing Vector pins (Vector T9.4). This m·th d is very rigid mechanically and is easy to duplicate. A narrow strip of thin copper was mounted in the center of the micarta terminal board to provide a common ground for the circuitry. At the front of the board a Hammarlund MC-50M variable capacitor was mounted with the copper strip between it and the board. Except for the toroids, all the components were mounted by simply soldering them to the Vector terminals. By placing the body of the component next to the board and by using short leads, almost all vibration and its effects were eliminated. In fact, the completed unit can be slammed down on the bench with no perceptible change in frequency.

When winding the toroids, the turns should be spaced to completely fill up the circumference of the core. In this way maximum Q is obtained. After the toroids are wound, a little polystyrene Q-dope will hold the turns in place. The completed toroids are mounted to the board with 3/16 inch flat head nylon screws that came in a nylon screw assortment from the local hardware emporium. If you can't find any nylon screws, brass screws and fiber washers will work as well.

The frequency of the oscillator is set to the proper range with the 47 pF trimmer capacitor. With the variable capacitor plates fully meshed, set the oscillator frequency to about 3480 kHz with the trimmer; when the variable is fully open, the output frequency should be approximately 4020 kHz. If a full 500 kHz range cannot be obtained, decrease the value of the 75 pF shunting capacitor to the next lowest value (68 pF) and try it again. If you want to reduce the frequency coverage, increase the value of the 75 pF capacitor. The power output stage is adjusted for maximum output at 3750 kHz.

Although this unit was designed specifically for 80 meters, it may be used on other frequency ranges by simply changing the number of turns on the toroid coils. Remember that the center frequency of the desired range is set with the inductance; the frequency spread is controlled by the size of the 75 pF shunting capacitor. In some cases a little juggling back and forth between the capacitor and inductor may be required to get the desired results, but it is not difficult nor time consuming.

Conventional bipolar transistors have eliminated many of the problems in stable VFO design, but they still have several minor disadvantages. The field effect transistor virtually eliminates these dsiadvantages and combines the low power and low heat of the bipolar device with the high impedance and predictable element capacitance characteristics of the vacuum tube. This VFO has provided such extremely stable results on 80 meters that an 8 MHz unit is under construction for use on six and two.

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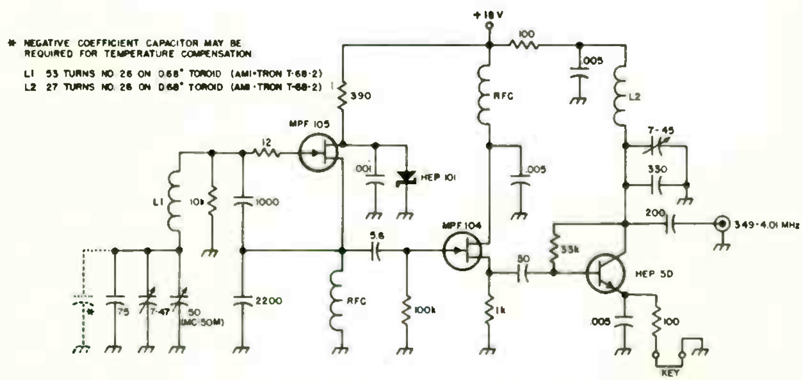
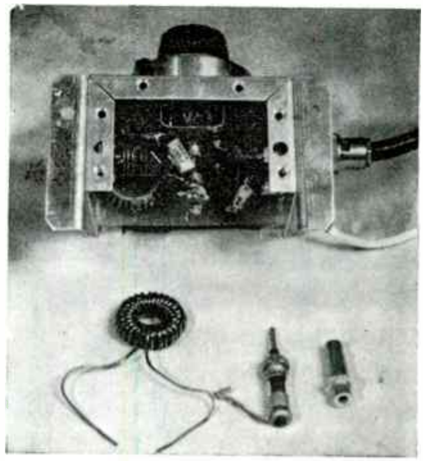


Fig. 2. The RFC's used in the FET 80 meter VFO are miniature 1 mH units available from J. W. Miller. The oscillator and output stage as shown here will run continuously. To key the output, remove the jumper marked "key" and connect the two terminals to your key. The two toroids are available postpaid anywhere in the U.S.A. for $1.00 from Ami-Tron Associates, 12033 Otsego Street, North Hollywood, California 91607.

CHAPTER 29

Stable VFO for 2- or 6-Meter Bands

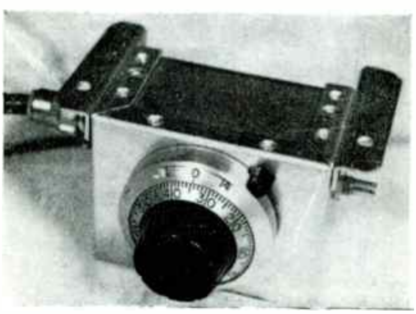
Like to build a small, simple, inexpensive, stable VFO for VHF or HF use? If you would, this article tells you how. The basic VFO described here drifts less than 80 Hz total in three hours, yet can be built very easily. The VHF model of the VFO is shown in Fig. 1. It can be used on six or two meters as it furnishes 24 or 25 MHz output with an oscillator in the 8 MHz range. The high frequency model shown in Fig. 2 is designed for use with an SSB mixer and operates at 5-6 MHz. Two methods of tuning are shown. One uses a conventional air variable capacitor. The other uses a piston trimmer capacitor which offers very small size, excellent stability and easy tuning.



Parts for the portable VFO. This VFO was made for two, but can be used on six with minor changes, or on 5-6 MHz.

Circuit description

The circuits of the two models are similar. Each starts with a modified high capacitance Clapp oscillator using a toroidal coil. The coil (L1 in each case) is wound on a small toroidal form and features a very high Q (around 250) on a Boonton Q Meter. The coil and the capacitors C1 through C6 form a resonant frequency at the VFO frequency. Series tuning is done with C1 through C4. C1 is used to set the basic frequency range, C2 is for temperature compensation, C3 is the calibration trimmer and C4 is the tuning capacitor. The range of C4 is determinded by the frequency coverage desired. On two meters, 10 pF is more than adequate, but this value only gives 1.8 MHz range on six. For the 5-6 MHz range, 70 pF of range is needed to cover 1 MHz. This means that the capacitor must have a maximum value of about 75 pF.



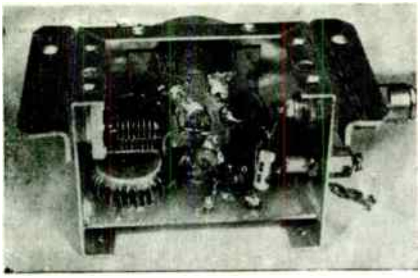
Front view of the six or two meter VFO using a piston tuning capacitor as shown in Fig. 3.

C5 and C6 form a capacitive divider for feedback voltage. Using the same value for both capacitors insures that the feedback circuit is balanced. Changing the supply voltage affects the VFO frequency very little so a voltage regulator is not required.

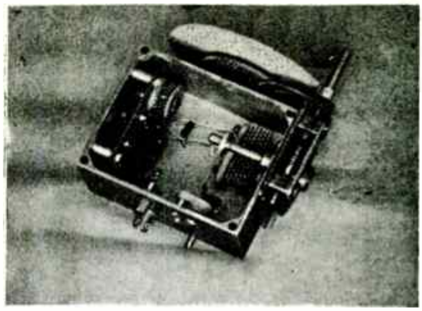
Output is coupled from the emitter of the oscillator with a 100 pF capacitor. Don't use more than 100 pF because of loading effects on the oscillator circuit. The lowest value that can be used is best.

The second stage of the circuit (Q2 and its circuitry) is used as an untuned buffer at the same frequency as the oscillator in the 5-6 MHz VFO. Its output is low impedance from the emitter. In the two or six meter model, the second stage is a tripler to about 24 MHz with an rf choke in a broadband collector circuit. The output is high impedance and used to drive a vacuum tube grid at this station.

The circuits in Figs. 1 and 2 both perform well. I've built the two meter and 5-6 MHz versions, but it should be easy to cover six with the two meter model by reducing the value of C1.



Inside of the portable VFO for six or two showing the construction.



Construction of VFO using parts from a command set transmitter dial.

Construction details

The VFO's shown in the photos are the results of using the same basic ciruit but using different construction techniques. The toroidal coil L1 used in both models is constructed by winding heavy gauge wire on the proper toroidal core. A good stable core with high permeability is a neccesity. Several manufacturers make suitable cores. I used a Micrometals core, which can be obtained from Micrometals, 72 E. Montecito Ave., Sierra Madre, California, or from one of their representatives. They have a minimum charge of $10 per order, but for $10, I was able to obtain a life-time supply of cores as each is very inexpensive when you buy a large number. I'd recommend that you write for their catalog and then order the cores.

If you'd rather not buy so many cores, Ami-Tron Associates, 12033 Otsego St., North Hollywood, California, will sell you an individual core for only 60¢ postpaid. You can also make up the proper coil inductance with the Ami-Tron RF Toroid Balun Kit available at many radio distributors.

After winding the wire on the core, the coil should be given a heavy coat of Hi-Q varnish or dope to prevent the wire from moving. All the capacitors in the circuit must be of good quality. A temperature-compensating capacitor is used to correct the minor drift in the circuit. All frequency-determining components must be securely mounted to prevent change in frequency due to movement of parts and wires.

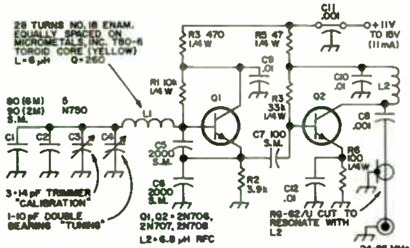


Fig. 1. Six or two meter transistor VFO with output on 24-25 MHz. The oscillator operates on 8-8.3 MHz.

Mobile VFO for two or six

The two meter VFO shown in the photos was built in a small package for mobile use. This VFO used a piston capacitor. Oscillator parts, transistors and the toroidal coil were mounted on a section of insulated board and cemented to the box with epoxy. This unit was built in a hurry for use during a trip so a few short cuts were taken. The box used was an LMB tight fit chassis box with self-tapping screws to fasten down the edges, but a sturdier box would be better. The tuning capacitor (C4, 1-10 pF) used in the two meter model came from surplus. This capacitor was mounted on a U-shaped bracket with its shaft moved by a stationary bushing (see Fig. 3). A hollow shaft was drilled out for a very tight fit over the bushing. This allows the capacitor to be turned with practically no backlash. A turn-count dial was fastened to the front of the VFO box and a calibration chart was made. With 22 turns, the capacitor has about 9 pF of travel, or from 8.0-8.3 MHz on the VFO. This resulted in a very stable VFO which could be tuned to zero beat while driving.

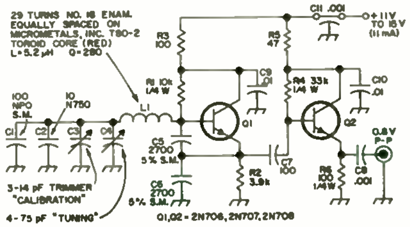


Fig. 2. Transistor VFO for 5-6 MHz for SSB mixing service.

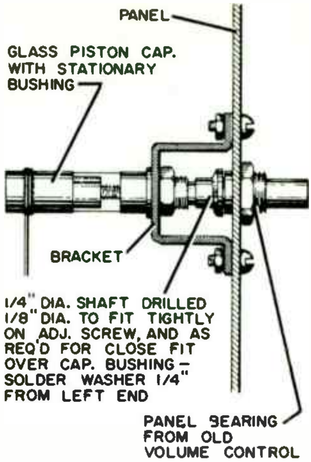


Fig. 3. Mounting the piston trimmer used in the portable VFO. A regular tuning capacitor can also be used.

5-6 MHz VFO

This VFO used different construction than the two meter one. The VFO parts were mounted on the insulated board as described before, but the box and tuning mechanism use a different technique. The box used was made from heavy cast aluminum found in a surplus store, with a cover fabricated from ⅛ inch aluminum. A sturdy double bearing capacitor was used for tuning. A gear drive and dial assembly from an ARC 5 transmitter was adapted to the VFO. This gear assembly gives plenty of bandspread with smooth tuning and very little backlash, though the mechanical work was a bit tedious. 48 turns are required to cover the 1 MHz range from 5-6 MHz, which is very nice for SSB tuning. A round plastic dial was cut out and installed in place of the original dial.

Both VFO's have given very good results. The two meter model in particular has provided much better stability reports than commercial VFO's.

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CHAPTER 30

VHF Parametric Multipliers

During the last year there has been considerable interest in the use of transistors as parametric multipliers. In fact, at least one semiconductor manufacturer has advertised a silicon power transistor that is specifically designed for parametric multiplication. R.C.A. has recently released its 2N4012 which, operating with 1 watt of 324 MHz drive, will produce 1.4 watts of 1296 MHz output.[1]

The parametric multiplier in Fig. 1, utilizes the emitter-base junction of a transistor in a ground-base Class-"C" amplifier. In this 3.5 MHz to 14 MHz quaddrupler the amplification takes place after multiplication at 14 MHz, since the idlers are across the emitter base junction.

The original method of parametric transistor multiplication has in its favor: an easy-to-follow development from a circuit composed of a separate varactor multiplier and class "C" amplifier, and good load isolation.

In the VHF multiplier, we will reverse the order, letting class "C" amplification take place at the input frequency, and do the multiplication thereafter. This means that, now, the idlers will be across the base-collector junction, which acts as a non-linear capacitance.

The advantage of this configuration is that transistors can be used to produce output power that exceeds their input power, even though operating above their f_T. The disadvantage is that the idlers and output circuitry become one big network, which makes understanding more difficult and does not give load isolation.

An elementary VHF parametric multiplier is presented in Fig 2, showing the idler-output circuitry on the base-collector side of the transistor. Since the input impedance to a grounded-base stage of this type is low, one can actually use such an untuned input circuit during initial tests.

In designing a parametric multiplier, we ought to restrict ourselves to driving frequencies of lower than 1/3 the frequency at which the transistor has unity power gain, f_T. Also, only multiplication factors ("n") of two, three, four, and possibly five should be tried. Multipliers with "n" greater than five require too many idlers, and the calculation and implementation of such circuits gets out of hand. As an example, a transistor with an f_t of 120 MHz would best be driven with 36 MHz input and designed as a quadrupler, if 144 MHz output is desired.

Bear in mind that *the transistor must be operated at an input frequency where it has some power gain,* since all amplification goes on at the fundamental frequency. This fact is clearly shown in a recent R.C.A. application note (SMA-40), wherein a 2N4012 will typically put out less than 1 watt at 1296 MHz as tripler (with 1 watt of 432 MHz drive) but will put out 1.4 watts at 1296 MHz as a quadrupler (with 1 watt of 324 MHz drive)! This relation is shown in Fig. 3.

As a start, let's design (and gain experience with) a circuit which is inexpensive and forgiving, yet which will demonstrate all the little subtleties of VHF parametric multipliers. A transistor costing less than a dollar is used; it is *normally* used "in small signal applica-

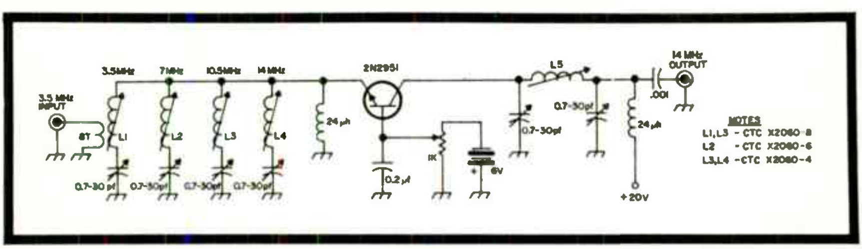


Fig. 1. High frequency parametric transistor multiplier. The multiplication (similar to that of a varactor) takes place in the emitter-base junction, then the transistor amplifies the signal at the multiplied frequency, 14 MHz.

tions up to 20 MHz." We will use a 2N3053 as a 48 MHz to 144 MHz tripler, and if there are any "semiconductor tragedies," at least we are not out much.

Looking at the 2N3053 spec. sheet, we find that it has a maximum gain-bandwidth product of 200 MHz, see **Fig. 4A**. The second fact that is apparent is that, unlike some of the "overlay" transistors, no curve of collector-base capacitance versus voltage is given. Since the collector-base capacitance (Cob) is a depletion-region capacitance, it should follow the same (exponential) curve as that for varactor diodes. We then can take the normalized capacitance-voltage curve from any varicap data sheet, plug in the one data point as given by our particular transistor spec. sheet for Cob; (15 pF at 10 volts for the 2N3053), and get the values of capacitance for any other voltage.

However, to simplify matters, suppose we make our first attempt at a collector of +10 volts, where Cob is known. This, also, gives an operation point that falls on the published Ec-Ic curves of the 2N3053, see **Fig. 4B**. If we decide on an average collector current of 30 mA, this puts our operating point approximately a third of the way between the 175 MHz and 200 MHz gain-bandwidth contours of figure 4a. This point is marked by an X, and gives us a gain-bandwidth product of approximately 180 MHz.

A 48 MHz to 144 MHz tripler easily fits within our criterion of 1/3 the f_T for a drive frequency. **Fig. 2** is such a tripler, simplified to make it easy to adjust initially. The effective output capacitance of the transistor is Cob plus Cs (a stray wiring capacitance of 5 pF), for a total of 20 pF.

L_1 and C_1, the 48 MHz idler, must be series resonant in combination with the 20 pf output capacitance. That is, **Fig. 5A** must resonate at 48 MHz. Similarly, **Fig. 5B** must resonate at 96 MHz, and **Fig. 5C** must resonate at 144 MHz. Or, at least, this is *nearly* true, since each idler's impedance at the other two frequencies modifies the design aim somewhat. Since we know that Cob+C_s is 20 pF, L_1 must be at least 0.6 μH to resonate at 48 MHz. Let's take L_1 to be 1.2 μH and make C_1=20 pF (a 3 to 30 pF trimmer in the actual case). In a similar way we make L_2=0.3 μH and C_2=20 pF. Then with the output series-circuit we must use a bit more caution, making L_3 relatively large in order that C_3 be small, so as not couple too much of the fundamental and 2nd harmonic to the output. L_3 is made 0.4 μH and C_3 is 3 pF, (a 3 to 12 pF trimmer). The output tap is deliberately placed very low on L_3, to cause small loading, during initial adjustment.

The 48 MHz, 50Ω output of a small (6 meter) exiter is coupled to the input of the multiplier, after the output of the multiplier is terminated in 50Ω. Increase the drive to no more than 10 volts rms and one should see the 2N3053 collector current climb from zero to our operating point of 30 mA. Using a grid dip meter as an absorbtion wave meter, couple it to L_3 and tune C_3 for the maximum 144 MHz output. Then tune C_2 for maximum 144 MHz output with the grid dip meter still coupled to L_3. And, similarly peak up C_1. The drive level will need readjusting (reducing) during this process. C_1, C_2, and C_3, should then be readjusted several more times for maximum 144 MHz output, until readjustment has a small effect.

At this point we will go back and put a matching transformer in the input, to optimize transfer of drive from the 50Ω driver to the 2N3053 emitter. Also, various taps on L_3 for optimum output loading are tried, each time retuning C_1, C_2 and C_3. Now, finally, that we've "juggled" and "tweaked" the whole thing up, the reader will begin to understand what he has in store if *large* multiplication ratios are involved.

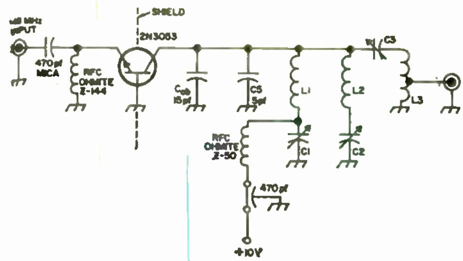


Fig. 2. 48 to 144 MHz parametric transistor multiplier. Note that this circuit is unlike the one in Fig. 1 in that the multiplication takes place in the base-collector junction. This is a preliminary circuit with untuned input. C5 is C_s, stray capacitance.

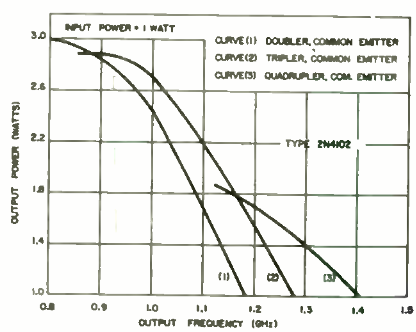
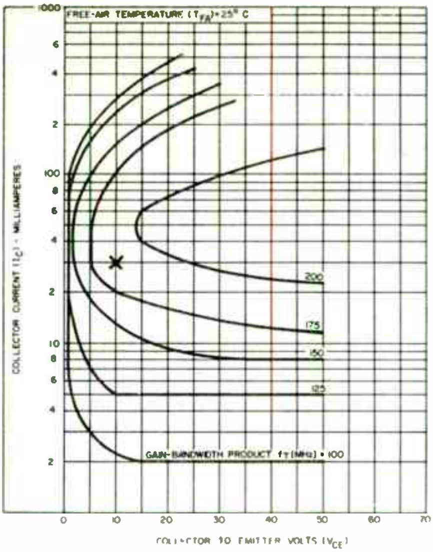
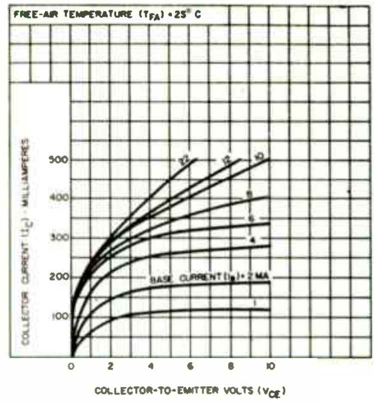


Fig. 3. Output of an RCA 2N4012 operated as a parametric transistor multiplier. Note that the 2N4012 will pu out more power at 1296 MHz as a quadrupler than as a tripler. The text explains this.



(A)



(B)

Fig. 4A. Gain-bandwidth curves for the RCA 2N3053 (a 96¢ transistor). Fig. 4B. Collector current (I_C) versus collector-emitter voltage (V_{CE}) for the 2N3053.

As to the performance of the unit, it required 1.5 volts rms of 48 MHz drive at 50Ω. The 144 MHz output was 2.25 volts rms across 50Ω. The collector efficiency, then, is 33% (300 mW dc power input, 100 mW of 144 MHz output). To asure ourselves that the RF output voltage that was being measured is really predominantly 144 MHz, a look at it was taken with a highly specialized oscilloscope. The oscilloscope used was a "storage" scope with "sampling" plug-ins. Such a device allows one to look at repetitive waveforms up to 1,000 MHz and "store" that waveform image on the 'scope face, for leisurely examination and sketching. The waveform of the above unit is shown in Fig. 6, as it was sketched, from storage 'scope face. Three cycles of 144 MHz RF are displayed so that the amount of 48 MHz present can be observed. The fact that a rather special 'scope was used for making Fig. 6 does not mean that it was essential, but its use only confirmed our other measurements. The grid dip meter was the only necessary piece of test equipment.

The final circuit is shown in Fig. 7. It was built inside a 5 X 7 X 2 aluminum chassis as shown in the photo. Note that the input circuit occupies occupies the 2½" compartment at one end of the chassis. The transistor is mechanically and electrically attached with an IERC TX 0507-1B heat sink to a 3 inch long piece of ¾" wide copper strap (the collector of a 2N3053 is connected to the case). This strap serves to dissipate heat, and also serves as a low inductance connection to the idlers. Note that the three coils for 48, 96, and 144 MHz are all spaced from each other and all at right angles to each other, to avoid inductive

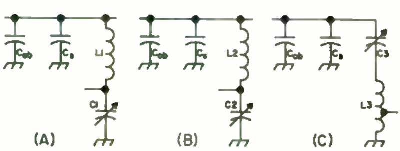


Fig. 5. Output circuit of the parametric transistor multiplier. In A, the resonance is at 48 MHz, in B, at 96 MHz, and in C, at 144 MHz.

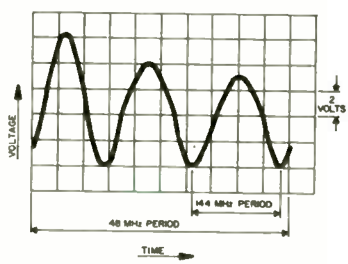
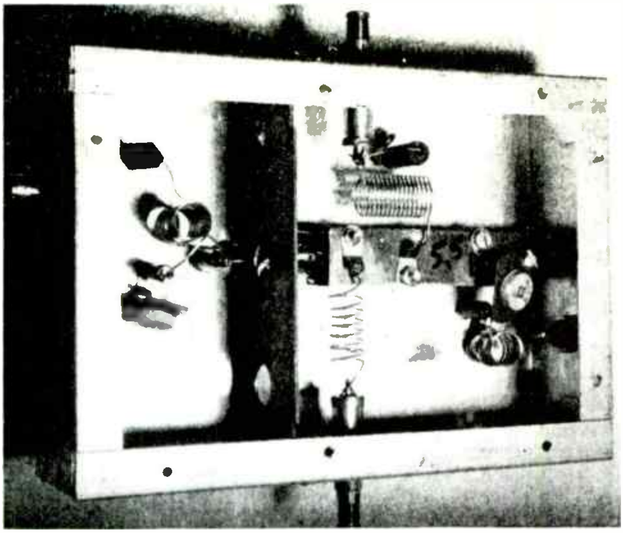


Fig. 6. Output waveform of the tripler in Fig. 7.



Bottom view of the 48 to 144 MHz tripler in Fig. 7.

coupling. A bottom plate is used which completes the shielding of the input and output compartments. It has a ⅜" hole in it for adjustment of C_3 and a 1⅛" hole immediately above L_3, so that the grid dip meter can be coupled to that coil during adjustments.

A second unit was constructed to try the principle as a 144 MHz to 432 MHz tripler (Fig. 8). The transistor used in this case was a 2N3553 (R.C.A.) which was about $8.00 when purchased. The newer 2N3866 should work as well and is only about $5.00. (the 2N3866 is now being manufactured by two other firms, which will tend to bring the price down).

The 432 MHz tripler was constructed in a smaller chassis, a Bud AC-431, measuring 4X6X2. It was partitioned, as before, and one will note the "wavelength scaling" throughout (ie: smaller inductances and smaller capacitances) in the photo of it. Note, also, that the copper heat sink tab had to be cut shorter than previously to reduce its inductance.

The output of the unit is 320 mW at 432 MHz with 180 mW of 144 MHz input. The DC input power was 12 volts at 50 mA, or 600 mW.

While the two parametric multipliers herein described are perhaps not pushing the state of the art, they do represent working models of a relatively new technique. No doubt the units shown can be driven harder, modified, etc.; but the experimentation and "smoke-testing" will be left to those interested in further work. As they stand, the circuits may be useful in handytalkies, or (offset a bit in frequency) as local oscillator chains.

What has been attempted above is to present an approximate method of designing a parametric-transistor-multiplier. The resultant

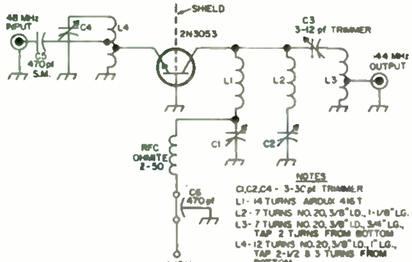


Fig. 7. Final circuit of the amplifier-tripler in Fig. 2. It puts out 100mW on two with 45 mW of 48 MHz drive and 300 mW dc input.

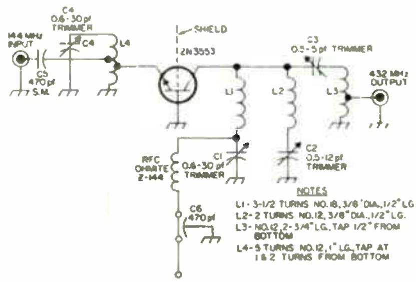
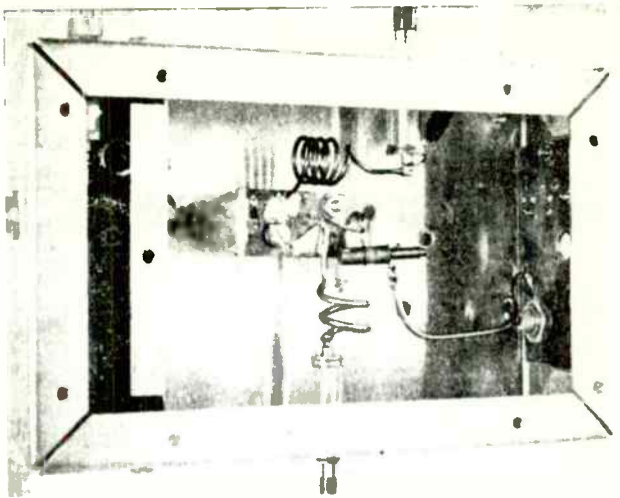


Fig. 8. Parametric transistor multiplier from 144 to 432 MHz. Output is 320 mW with 180 mW input on 144, with DC input of 600 mW (12 volts at 50 mA).



Bottom view of the 144 to 432 MHz tripler in Fig. 8.

circuits *do* work, though they are almost certain not to be optimum circuits. Since we can change drive level, operating voltage, C_1, L_1, C_2, L_2, C_3, L_3, and the tap on L_3; we can be said to be dealing with nine variables. The optimization of such an array is nearly hopeless theoretically, unless a digital computer is handy and you can set the nine equations in nine unknowns up for it.

However, this approximation method does work—give it a try!

The author wishes to thank Radio Corporation of America for permission to use Figs. 3 and 4.

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CHAPTER 31

RF Power Amplifier Design

Transistors capable of delivering several watts at frequencies up to several hundred megahertz are readily available at nominal cost, and development of improved devices, in regard to power output and operating frequency, is advancing rapidly. Because of transsistor voltage and current limits, and the need for proper matching, the tube oriented designer must slightly revise his design procedures and take certain cautions he would not ordinarily consider if he was designing with tubes.

These new design procedures, together with some of the precautions and other design dissimilarities are discussed here in sufficient detail to permit most readers to reap the benefits of technological advances being made by the semiconductor industry in RF power devices.

Voltage-current relationships

The majority of transistor RF power amplifiers fall into the Class B (zero base bias) category because this class of operation provides a greater power gain. (Some may want to call zero base bias Class C because it does take a few tenths of a volt to start the transistor conducting.) However, Class C operation (reverse base bias) with its higher collector efficiency is also suitable especially when efficiency is of greater importance than power gain. Moreover, it is practical, under certain conditions, for example, where a greater power gain than Class B provides is needed to operate a transistor RF power amplifier with a slight forward bias.

Darrell Thorpe, former WØPKB, and W9NYI, is the editor of the Motorola Military Electronics Division Engineering Bulletin.

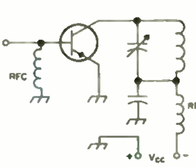

Fig. 1. Basic transistor RF power amplifier.

A basic transistor RF power amplifier circuit is shown in **Fig. 1**. The base has no bias applied to it, therefore, the circuit is operating Class B. With no signal and no bias applied to the base, the circuit is setting at the static operating point. That is, the voltage between the collector and emitter is equal to the supply voltage (V_{CC}) as shown in **Fig. 2**. Since the circuit illustrated employs a PNP transistor, the transistor conducts only when the voltage (applied drive signal) goes negative. Referring to **Fig. 3**, as the base voltage goes negative, there is a corresponding rise in collector current. Also, notice, from **Figs. 2 and 3**, that as collector current increases, the collector-emitter voltage drops to V_{min}.

Then, as the base voltage returns to zero, collector current goes to zero and collector voltage returns to V_{CC}. However, due to flywheel effect of the tank circuit, the collector voltage returns to V_{CC}. However, due to flywheel effect of the tank circuit, the collector voltage continues to increase until it reaches approximately $2V_{CC}$. The cycle then repeats.

Breakdown voltage and supply voltages

In vacuum tube circuits, breakdown voltage between the plate and other elements is seldom a consideration since it usually is much greater than the usual supply voltage. However, present day transistors are not so for-

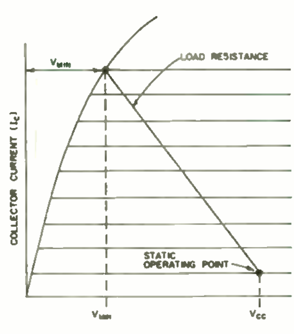

Fig. 2. Static operating point and load resistance curve.

tunate. Therefore, the breakdown voltage of the transistor must be considered when selecting devices and supply voltages.

As previously discussed, for an unmodulated transistor RF power amplifier, the collector voltage swings from V_{min} to approximately twice the source voltage. Therefore, the transistor must be able to withstand the peak-to-peak voltage which is $2V_{CC}-V_{min}$. Since V_{min} is usually only a few tenths of a volt. it can be neglected and the peak voltage can be considered as $2V_{CC}$.

Since the common-emitter configuration produces higher gain than a common-base configuration, the breakdown voltage of the transistor being considered will usually be BV_{CES}. Quite often BV_{CES} and BV_{CBO} are the same value. If the peak voltage happens to slightly exceed the breakdown voltage, the transistor will not be damaged provided the current and time duration are limited, but efficiency and gain will drop.

Thus, from the preceding discussion, it should be clear that a transistor should be selected with a BV_{CES} equal to or greater than $2V_{CC}$ or if this is not practical, V_{CC} should be set equal to or less than $BV_{CES}/2$.

For an AM power amplifier, these conditions must be modified to account for the increased peak-to-peak voltages which result from the modulating voltages. Fig. 4 illustrates unmodulated and modulated carriers. The m on the modulated carrier represents the modulation index (m = 1 for 100% modulation). From Fig. 4B, for a modulated transmitter, the transistor must have a voltage rating of

$$BV_{CES} \geqq 2V_{CC}(1 + m) \qquad (1)$$

or

$$V_{CC} \leqq \frac{BV_{CES}}{2(1+m)}.$$

Since m = 1 for 100% modulation,

$$V_{CC} \leqq \frac{BV_{CES}}{4}. \qquad (2)$$

Therefore, for CW or FM operation, where m = 0, the maximum collector voltage is approximately one-half the breakdown voltage, and for the final stage in an AM transmitter (m = 1) the maximum collector voltage must not exceed one-quarter of the breakdown voltage. Effects of slight clipping caused by breakdown on the upward modulation and saturation voltage on the downward modulation may generally be neglected.

Determining the optimum load resistance

In a transmitter, it is generally desirable to obtain a certain power output or the maximum power that a transistor is capable of delivering. Since the collector supply voltage is often limited either by breakdown voltage or the source i.e.; 6 or 12 volts in the case of mobile transmitters, the only variable is the effective load resistance at the collector.

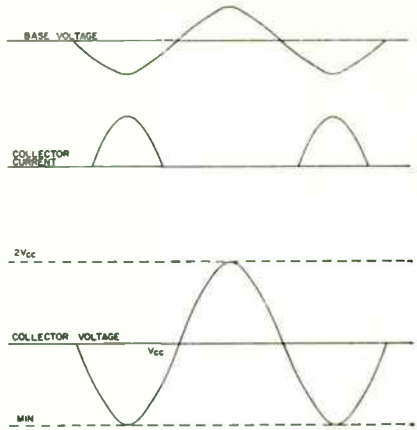

Fig. 3. Affect of base voltage on collector current and voltage.

To get a better understanding of how the load resistance (R_C) influences the maximum power output, it is necessary to examine the properties of an amplitude modulated signal more closely. The equations developed will easily reduce to the CW or FM case. Fig. 5 illustrates the output at the collector of an AM final stage. The voltage is given by the expression

$$e_c = E_c(1 + m \sin \omega_m t)(\sin \omega_c t), \qquad (3)$$

where

e_c = instantaneous carrier voltage
E_c = unmodulated carrier amplitude
$\omega_m = 2\pi$ (frequency of modulating signal)
$\omega_c = 2\pi$ (frequency of carrier signal)
t = time, and
m = modulation index = E_m/E_c, where E_m = peak modulation voltage and E_c = peak unmodulated carrier voltage.

The peak voltage, E_p, occurs at the crest of the sine waves and is easily determined to be

$$E_p = E_c(1 + m) \qquad (4)$$

Using the standard ohms law equations, the peak power (P_p) is given by

$$P_p = \frac{E_p^2}{R_C}. \qquad (5)$$

Substituting Equation 4 into Equation 5 we have

$$P_p = \frac{E_c^2(1+m)^2}{R_C}. \qquad (6)$$

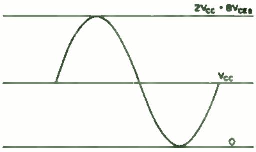

Fig. 4A. Unmodulated carrier.

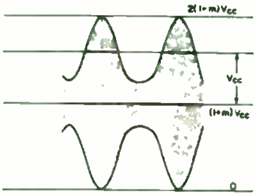

Fig. 4B. Modulated carrier.

By inspection of Equation 6 it is seen that the *peak* unmodulated (P_u) power (m = 0) is

$$P_u = \frac{E_o^2}{R_C}. \qquad (7)$$

and with modulation, the peak modulated power (P_m) is

$$P_m = P_u(1 + m)^2. \qquad (8)$$

Note, from Equation 8, that for a 100% modulated carrier (m = 1) that the peak modulated power is four times the unmodulated power. Thus, most of the power of an AM transmitter is contained within audio sidebands instead of the carrier. For this reason, when maximum talk range is desired from an AM transmitter, it should be designed and tuned so as to produce maximum demodulated audio signal with minimum distortion.

Getting back to determining the optimum load resistance for a desired power output, Equation 7 can be written

$$R_C = \frac{V_{CC}^2}{2P_u}. \qquad (9)$$

Where R_C is collector load resistance
V_{CC} is substituted for E_c
P_u is unmodulated power output

the factor 2 in Equation 9 comes from the conversion of peak power to rms power. R_C for a modulated transmitter is obtained by substituting Equation 8 into 9 which becomes

$$R_C = \frac{V_{CC}^2\left(1 + \frac{m^2}{2}\right)}{2P_u} \qquad (10)$$

Table I

	Max DC Supply Voltage (Vcc)	Max Load Resistance (Rc)	Peak Power
AM (100% mod)	$\frac{BV_{CES}}{4}$	$\frac{3V_{CC}^2}{4P_u}$	$8P_u$
FM/CW	$\frac{BV_{CES}}{2}$	$\frac{V_{CC}^2}{2P_u}$	$2P_u$

Note: The simplification of the equations for the modulated transmitter is arrived at by letting m = 1 which is the case for 100% modulation.

From Equation 9 or 10, *the maximum collector load resistance that can be used for a given output can be calculated.*

A summary of voltage and load resistance relations is given in Table I.

CW and modulated power output capabilities

At VHF, the factors which limit power output, in most cases, are other than device power dissipation. Usually, high frequency transistors are peak voltage or peak current limited.

The voltage limits have already been discussed; however, a few words about current limits are needed to aid in understanding why transistor circuits behave as they do.

A transistor can be compared to an emission limited tube. That is, the amount of peak instantaneous current available is determined by the transistor structure, and no reserve or space charge effect exists. If a transistor is operated with maximum allowable collector supply voltage and the drive is increased until there is no further increase in output the maximum peak current has been reached. This condition is very seldom encountered in a tube, because the power dissipation limit of the tube is usually reached first.

The peak voltage and current limits are important factors for amplitude modulated transmitters because a device which is already operating at its collector supply voltage and cur-

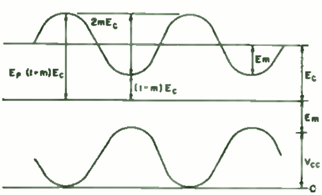

Fig. 5. Carrier with sine wave amplitude modulation.

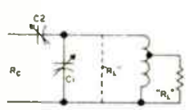

Fig. 6. One method of matching the load to the collector circuit.

rent limits can not be upmodulated from that power level. As discussed earlier, for collector modulation, the supply voltage must be limited to one-fourth of the maximum transistor voltage rating to prevent breakdown, and since the peak current must double in addition to the voltage, a carrier level of one-fourth maximum power output must be maintained if 100 percent up-modulation is desired.

And, while we are on the subject of modulation, it is worthwhile to mention that feed-through capacitance in transistors will allow a residual carrier to be passed from the driver through the final even if the down-modulating audio has reduced the collector-to-emitter voltage to zero. Hence, some modulation of the driver is needed to achieve good down-modulation of the final. Also, modulation of the driver will aid in achieving the higher peak current required by the final on up-modulation.

Design example

Up to this point, we have not considered matching networks to transform the actual load impedance, usually a 50-ohm or 300-ohm antenna, to the load the transistor needs to see so that the specified power output can be achieved. **Fig. 6** shows one method of coupling the collector circuit to the load. There are many other types of matching networks that can be used including the pi and L networks. If you desire to use one of these networks refer to one of the radio handbooks for equations to calculate component values. The matching circuit shown uses a parallel tuned circuit to couple the load to the collector circuit. The collector of the transistor is tapped down on the tank coil. Capacitor C_1 provides tuning for the fundamental frequency and C_2 matches R_C to the tank circuit.

Let's assume we are designing a 50 MHz final using the RCA, 2N3553 transistor. Characteristic curves on the data sheet (see **Figs.** 7 and 8) show that this device can typically provide 5 watts (CW) from a 13.5-volt source at 50 MHz and 10 watts with a 28-volt source. Specified BV_{CES} is 65 volts which is sufficient for either CW or AM from a 13.5-volt source (use the equations and check for yourself). Therefore, let's design for a peak modulated power output of 10 watts which corresponds to a 2.5-watt unmodulated carrier.

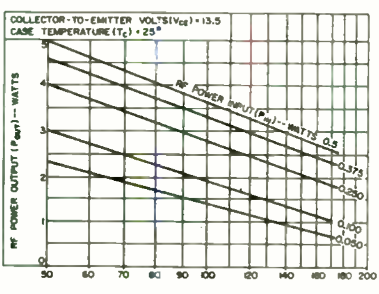

Fig. 7. Characteristic curves for the 2N3553 at 28 volt emitter to collector voltage.

The maximum load resistance is (see Table I)

$$R_C = \frac{3V_{CC}^2}{4P_u} = \frac{3(18.2)}{4(2.5)} = 55\Omega$$

Since at 50 MHz, the number of turns needed for L_1 will be rather small, and also, since the collector load impedance is very close to the 50Ω antenna impedance, we will assume a 4:1 turns ratio.

As shown in **Fig. 6**, coil L_1 transforms R_L to another resistance R_L''. This is given by the standard transformer impedance equation

$$\left(\frac{N_1}{N_2}\right)^2 = \frac{R_L''}{R_L} \qquad (11)$$

Since turns ratio is 4:1 and $R_L = 50\Omega$

$$\left(\frac{4N_1}{N_2}\right)^2 = \frac{R_L''}{50} = 16(50) = 800$$

$$R_L'' = 800$$

Let's assume that a loaded Q (Q_L) of approximate 8 or better is desired. Values of loaded Q in the range of 5 to 10 are practical to achieve.

Now, we can calculate C_2

$$C_2 = \frac{Q_L}{2\pi F R_L''}$$

$$= \frac{8}{6.28 \times 50 \times 10^6 \times 800} = 31.4\ \text{pF} \qquad (12)$$

and

$$L_1 = \frac{1}{(2\pi F)^2\, C_2}$$

$$= \frac{1}{(6.28 \times 50)^2 \times 10^{12} \times 31 \times 10^{-12}} = 0.333\ \mu\text{H} \qquad (13)$$

Using a coil nomograph, this turns out to be a 7-turn, one-half-inch diameter by one-inch long coil. Tap 1¾ turns from cold end.

Next, the value of coupling capacitor C_2 is calculated.

$$XC_2 = R_C\sqrt{\frac{R_L''}{R_C} - 1} = 55\sqrt{\frac{800}{55} - 1} = 55(13.8) = 210\Omega \qquad (14)$$

$$C_2 = \frac{1}{2\pi F X_{C2}} = \frac{1}{6.28 \times 50 \times 10^6 \times 210}$$

$$= \frac{1}{66 \times 10^9} = 15\ \text{pF} \qquad (15)$$

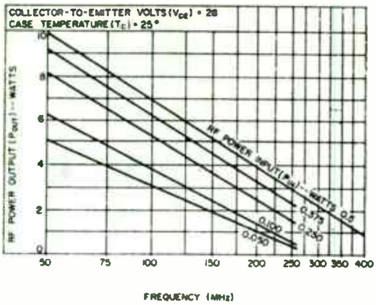

Fig. 8. Characteristic curves for the 2N3553 for 15 volt emitter to collector voltage.

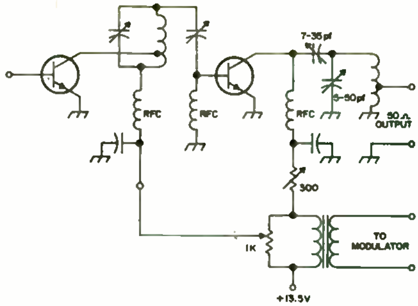

Fig. 9. Complete 50 MHz power amplifier stage and part of its driver. The driver must supply 75 mW for 2.5 W output.

The complete 50 MHz RF power amplifier and part of a driver stage is shown in Fig. 9. As shown, the driver stage can be designed using a similar matching network. The load for the driver will be the input impedance of the 2N3553 final transistor which, for all practical purposes, can be considered as the base spreading resistance (r_{bb}') of the transistor. For the 2N3553, r_{bb}' is typically 12 ohms. Therefore, the network must transform 12 ohms to the impedance the driver needs to see to develop the required drive power. The driver can be designed using the same procedure as given for the final. From Fig. 7, the driver must supply about 75 mW of power to the base of the 2N3553 to drive it to a 2.5 watts CW output. Since the driver should be modulated to improve down-modulation, the needed drive for peak power output of the final is provided by peak modulated power from the driver. Usually, modulating the driver between 25 and 35% is sufficient.

Potentiometers are shown in the modulation circuit to permit adjusting modulation for optimum performance. The pots can then be removed and replaced with fixed resistors after adjustment. Since the driver is being modulated the modulation to the final must be reduced, (i.e., driver modulation plus final modulation must not exceed 100%).

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CHAPTER 32

Diode-Controlled Break-In Switch

The search for the perfect T-R switch will continue for many years to come. The coax relay is hard to beat when one considers its broad bandwidth, its 100 db isolation, and its almost zero insertion loss. However, it is ponderously slow, it is noisy, and the best ones can cost a buck per db of isolation. What we require is a switch that will operate quietly and quickly, one which has low insertion loss and which is cheap and easy to make.

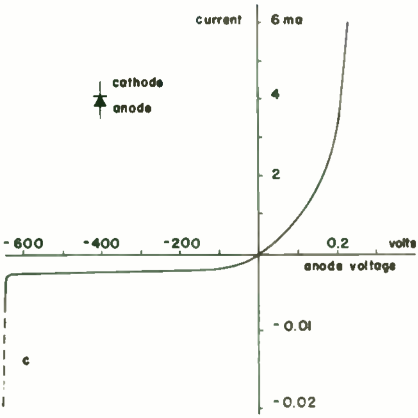


Fig. 1. Characteristics of silicon diode.

Electronic T-R switches are usually quite complex and seldom provide as much isolation as the coax relay. Some give a little gain for reception; most will degrade the receiver's performance. Some must be retuned as bands are changed. Many carry the hazard of TVI. A few are extremely tricky to adjust. The magic tee, ferrites, and coax and waveguide diode switches come into their own at UHF, but these techniques cannot easily be used in the HF bands.

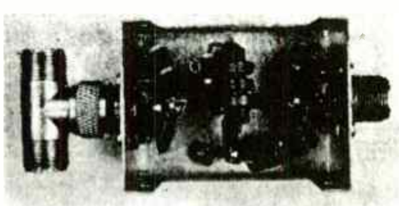
Let's keep it simple. The ideal T-R switch; Fast, low loss, simple, cheap—and no TVI.

The T-R switch described here will give about 80 db of isolation, less than 2 db insertion loss, and fast, silent operation; it is broadband, calls for a minimum of adjustment, and, above all, it is simple to build.

As the title tells, the active element is the diode. In fact we use the silicon diode. The characteristic (Fig. 1) shows that, depending upon the polarity of the applied voltage, the silicon diode either conducts heavily or (almost) not at all. In other words it is very much like a switch, where the switch's "handle" is replaced by the applied voltage. Whatever we do, we must be extremely careful not to operate the diode on the part 'C' of its characteristic, because the high dissipation in this so-called 'breakdown' region is destructive.

Let us investigate the silicon diode a little further. A forward biased diode (one with a positive potential applied to its anode) will conduct heavily. Under these conditions it presents a low impedance to a superimposed rf signal as well, providing the peak-to-peak amplitude of the rf is less than twice the available bias voltage. A reverse biased diode (one with a positive potential applied to its cathode) will pass hardly any current at all. It presents a very high impedance to dc. To an rf signal it will also present a high impedance, but the value of the impedance to rf will be lower than for dc. The reason for this difference is the fact that the diode possesses a certain amount of capacity, which passes the rf but blocks the dc. The impedance of a reverse biased diode therefore decreases with increasing frequency, but even at 30 mcs it will still be several thousand ohms if the bias is great enough.

These ideas may now be applied to the T-R switch, the circuit of which is shown in Figure 2. The resistors form dc circuits for the bias current. They have values that are



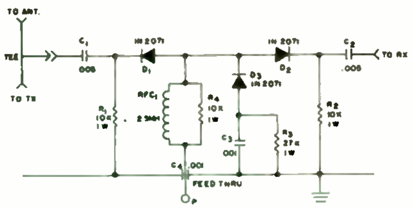


Fig. 2. T-R switch.

high in comparison to the impedance of the coax line, and so they do not shunt the signal excessively. The capacitors are there to block the dc and to pass the rf where necessary. Diodes D1 and D2 are biased in the same direction. D3 is biased in the opposite direction. Suppose we apply a positive potential at the point 'P'. Diodes D1 and D2 will conduct and diode D3 will be reverse biased and will not conduct. Thus diodes D1 and D2 present low impedances while D3 presents a high impedance. Under these conditions an rf signal from the antenna will be passed to the receiver with little attentuation.

Suppose now that we apply a negative voltage to the point 'P'. Diodes D1 and D2 will no longer conduct. D3 will be forward biased and will conduct. Therefore diodes D1 and D2 will present high impedances and D3, a low impedance. Rf from the antenna or from the transmitter will now be confronted by the high impedance of D1. A certain amount of the signal will leak past D1, only to be confronted by another high impedance, that from D2. A much easier path for any rf that does leak past D1 is to ground through the relatively low impedance of D3 and C3 in series. Accordingly, very little signal will now be able to reach the receiver.

Practical Considerations

It is now pertinent to decide upon the values of the components in the circuit and the voltages necessary for switching purposes. Let us first consider the operation of the device in its receiving mode. Signals will generally be quite small with voltages across the coax line seldom exceeding a few millivolts. Therefore, if the bias applied at the anodes of D1 and D2 is + 1v or more, the impedance through the unit will remain small for all signals.

During tests of the prototype, it was found that increasing the bias voltage above about + 10 volts at 'P' did not reduce the insertion loss materially. For general operation in the receiving mode, it is recommended that about + 30 volts be applied at 'P', but anywhere between + 10 and + 100 volts should prove satisfactory. From the point of view of heat, it is better to keep to the lower values of bias, since resistors R1 and R2 will be dissipating all their power in the close vicinity of the heat-sensitive diodes.

In exceptional cases, for instance, when a powerful transmitter is located nearby, cross modulation may be experienced, but an increase in the bias voltage should eliminate the problem. For a similar reason the bias supply must be well regulated, otherwise the slight change in transmission characteristics of the switch with changes of bias will modulate incoming signals, and hum will be apparent on them all. It should be remarked here that these effects have not been noticed under normal operating conditions. Only very strong signals cross-modulate (A closely coupled gdo will). Filtering on the author's transmitter, an RC filter at that, has proved adequate, and no hum has been detected during reception. However, with *no* bias the hum is severe.

Now we come to the bias requirements for the transmitting mode. It is possible to obtain the bias voltage directly from the rf output of the transmitter, but a number of disadvantages are associated with this method. TVI is one. Another is that large pulses of rf would reach the receiver at the start of each morse character, since there would be no provision for sequential keying. Also, for adequate isolation, a rather high voltage is necessary.

Tests were made with a signal generator having an output of a few volts. When a bias of - 6 volts was applied at 'P' the device had an isolation of 30 db. For increases of - 2 volts under these conditions there were increases of 1 db in the isolation. With - 120 volts applied at 'P' the attenuation reached 80 db. The available test gear was not sufficiently sensitive to measure any further increase in isolation.

We now know that the reverse voltage across the diode D1 must not be less than - 120 volts for proper isolation. Let us allow a safety factor of 30 volts, and make it our criterion that, whatever the signal from the transmitter, the voltage across D1 must be at least - 150 volts. A transmitter with an output of 100 watts will produce a peak-to-peak of 200 volts across a 50 ohm line. This peak-to-peak voltage will be centered on the bias voltage applied to D1 and will swing the voltage across D1 from 100 v greater than the bias to 100 v less than the bias at the rf rate. To meet our criterion, therefore, the bias

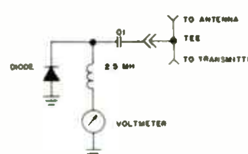


Fig. 3. RF voltmeter.

must be at least -250 volts in this case. To ensure that the diode is not operated in its breakdown region, its piv should be at least 350 volts. If we allow a 50 volt safety margin, then we must choose a diode with a piv of 400 volts.

The figures given above are correct if the coax line has negligible VSWR, and if the output of the transmitter is relatively free of harmonics. The voltage will vary from one place to another along a line with a high standing wave ratio, and if the switch is connected at a point of high voltage the diodes may be destroyed. It is simple to calculate the necessary PIV for any given transmitter power when the VSWR is low. Allowing for the safety margins mentioned above, we can calculate the PIV from the formula,

$$\mathrm{PIV} = 200 + 21\sqrt{P} \qquad \text{(50 ohm line)}$$

$$\mathrm{PIV} = 200 + 26\sqrt{P} \qquad \text{(75 ohm line)}$$

where P is the transmitter output power. Table 1 gives the value of piv necessary for a number of given output powers.

TABLE I

DIODE PIV	P-P XMTR VOLTAGE ACROSS COAX	XMTR PWR OUT (WATTS)* 50 ohm	75 ohm	OPTIMUM BIAS VOLTS
400	200	100	66	250
600**	400**	400**	280	350**
800	600	900	600	450
1000	800	1400	1050	550

* Power values for CW

** Author's unit uses 600 volt diodes, 350 volts bias, with a 150 watt transmitter. All other values untried

It is quite simple to measure the voltage across the coax, however, and eliminate the possibility of destroying the switch. All that is needed is a peak reading voltmeter like that illustrated in Fig. 3. The diode and capacitor should have voltage ratings greater than the voltage expected across the line. The meter should be as sensitive as possible (absorb little power) and it should be used on its highest range for greatest accuracy. However, even a meter with a basic movement of 1000 ohms per volt will introduce only a small error. Whatever meter you use, add 10% to the value you measure to be certain that you do not underestimate the voltage present—after all your meter may be reading low and the choke you use may not be too good. Be sure to measure at the point where you intend to connect the switch. This measurement will tell you the peak voltage across the line. Double it (to find the peak-to-peak), add 200 v, and that will be the minimum value PIV for your installation. The value of bias for a given transmitter is determined by adding 150 v to the value measured by the peak reading voltmeter. A higher bias may be used providing the diode you select has a high enough PIV.

Other components in the circuit are not so critical. D2 has to withstand the transmitting bias, and for convenience may be the

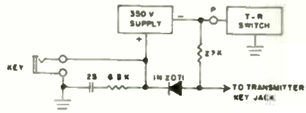


FIG 4

T-R switch hook-up.

same type of diode as D1. D3 has only to hold off the receiving bias and may have a lower rating than diodes D1 or D2. The capacitors should have reactances that are low compared with the line impedance, and the resistors should have values that are high compared to the line impedance. The choke should have no series resonance falling in or near a ham band, and it should present a reactance of several thousand ohms over the ranges on which it is to be used. R4 damps any tendency for ringing to occur due to the presence of the choke. Although C2 will normally have to withstand only the receiving bias, C1, C3, and C4 all must hold off the transmitting bias or the peak coax line voltage, and it may prove convenient to choose all the capacitors to stand the transmitting bias. Doing so when D2 has the same PIV as D1 might prove a useful feature, since the transmitter could be connected to either end without damage occurring. About 10 ma should flow through D3 in the transmitting mode, and one or two milliamps should flow through D1 and D2 during reception. The details given above should serve as guidelines for anyone who wishes to design his own switch.

Construction

The photograph shows the layout of the unit built by the author. It is housed in a 2¾ × 2⅛ × 1⅝" minibox. The heat-producing resistors are kept away from the diodes, and the imput and output circuits are arranged to minimize their mutual coupling as much as possible, thereby improving the isolation. The 2 watt resistors are not essential, and the 1 watt types specified in Fig. 2 will normally be satisfactory. However, for phone operation the dissipation of R3 should be checked to make sure that it is not being overrun.

The diodes are supported on low capacity ceramic standoffs arranged down the center of the box. Capacitors C3 and C4 are mount-

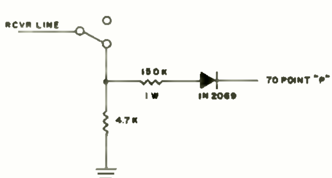


Fig. 5. AVC-MVC modification in receiver.

ed to the side. C4 is a ceramic feed-through and carries the bias current.

Control

There are many ways in which the biasing voltages can be applied. The method described below shows how another diode may be used to do some of the work.

Seventy volts appears across the contacts of the author's key when it is up. This voltage is used to bias the diodes in the receiving mode. A 27k resistor is connected in series with the 70 v line to the switch, as shown in Fig. 4. A diode is placed in series with the lead tó the key with its anode towards the transmitter. A 350 volt supply, with its output floating is connected from the switch to the cathode of the diode. When the key is up, the diode is back biased by the 350 volt supply. No current is drawn through the diode; the 350 volt supply is isolated. However, current can flow from the transmitter into the t-r switch and bias it for receiving. When the key is depressed, current is drawn through the t-r, through the 350 volt supply via the key, thus biasing the t-r switch for transmitting. At the same time the back bias is removed from the diode (Fig. 4) which becomes a low impedance and the transmitter is keyed.

The CR circuit across the key contacts is there to prevent the t-r switch from returning to the receiving condition before the rf from the transmitter has had a chance to die away. The capacitor has to charge to about 200 volts before the negative bias is removed from the switch, and this takes several milliseconds. As the switch switches over to isolation in less than one millisecond, the shaping circuit in the transmitter prevents too much rf from being developed before the switch is in the transmitting mode.

A further embellishment of the circuit may be noted. Again we can use a diode to do some switching for us. The negative bias applied to the switch during transmission can, at the same time, be used to bias the rf stage of the receiver and reduce its sensitivity. The diode isolates the receiver from the +70 volts applied to the switch during reception. Fig. 5 shows the arrangement used by the author. In the mvc position the avc/mvc switch normally grounds the avc line. The ground connection is removed and a 4.7k resistor soldered in its place. The switch side of the resistor connects to the 150k resistor and diode in series. A lead is taken from the cathode of the diode to the point 'P' of the t-r switch. In normal operation the 4.7k resistor makes no difference to the manual volume control, and, of course, it is out of circuit for avc. In other installations the value of the 150k resistor can be changed to suit individual requirements.

If *all three* diodes in the switch are reversed, a negative bias will be required for receiving, and a positive bias for transmitting. This alternative configuration may prove useful where the available voltages differ from those described above. There are very many ways in which biasing can be effected, and the reader will no doubt discover the method most easily adapted to his own needs. It is perhaps appropriate to include the circuit of a simple power supply that can be used with, say, the receiver power transformer, to supply the transmitting bias. It can also be used with a filament transformer and more filtering to provide the receiving bias. Fig. 6 shows a voltage doubler circuit that has a floating output. It can be used either for negative biasing as shown, or for positive biasing simply by reversing the connections to the output. The resistor R is adjusted to give the correct voltage at the point 'P' of the switch under operating conditions.

Conclusion

The diode T-R system described above has proved. extremely reliable and effective. The arrangement switches so quickly and cleanly that only a slight click is heard at the start of each morse character. The click is presumably caused by the sudden change in the receiver's parameters as the negative bias is applied to the rf and *if* stages. The recovery to full sensitivity, which, to the ear, seems to take place instantaneously, is clean and free from clicks. During transmission the receiver gives a pure S8 monitoring signal.

Loss of sensitivity during reception is negligible. The switch's insertion loss was measured as less than 2 db over the range from 10 to 30 mcs, and there is no reason to expect any great change down at 3.5 mcs. As 2 db is substantially less than one "S" point it would remain undetectable under most circumstances. The diodes add little noise to the circuit since they are not matched to the line.

Some people may suspect that diodes at the output of a transmitter would produce nothing but TVI. As the TVI from a diode is a result of operating it over a non-linear portion of its characteristic curve, we have made certain that this does not happen. No TVI has been introduced at the author's installation, although the TV antenna is only 6 feet from the transmitting antenna and the TV station is 130 miles away.

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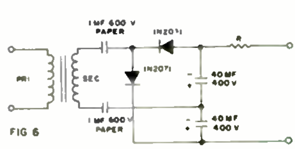

FIG 6

Voltage doubler power supply.

CHAPTER 33

A Low-Current, Slide-Bias Modulator

TRANSISTORS have been with us long enough now to have found their way into most AM mobile equipment, especially in the audio department. Although this has resulted in a major saving in battery current, no one is adverse to even greater savings. This is especially true when the increased saving is effected without further circuit complication.

The modulator described below delivers 40 watts of audio to the load. Yet, during the no signal condition, total current drawn from the battery is only 250 ma!

The microphone pre-amplifier consists ot a Motorola 2N1181 which is connected in the common emitter configuration. This stage should be driven by a dynamic microphone. If a ceramic or crystal microphone is preferred it will be necessary to insert a matching stage between the microphone and the pre-amplifier. Transistors other than the 2N1191 may be used without circuit alteration. Those tested and found satisfactory were the T.I. 2N1381, the Philco 2N1478 and the RCA 2N408.

The 2N1191 drives a class A Motorola 2N555. The 2N555 is a very inexpensive experimenter type transistor. Other equivalent transistors should give similar results. The 2N555 stage has its bias controlled by the following class B modulator stage. As the signal level increases so does the bias on the 2N555 stage increase. The 4.7K forward bias resistor allows a small bias current to flow so that the transistor is operating class A even for small signals.

Following the sliding bias stage are two Motorola 2N1554 transistors. These transistors are excellent for this purpose which is something that cannot be said for a good number of other transistors. A perusal of the 2N1554 I_c—E_c curves show that the transistor has very even spacing between the base lines which indicates that the 2N1554 is capable of excellent linearity. If other transistors are substituted, only those with good linearity characteristics should be used or distortion will result. Another Motorola transistor which gave excellent results (although a smaller transistor) was the 2N1540. If full output is required from the modulator, the builder is advised to stick with the 2N1554 even though it does cost more.

The required output transformer impedance is determined from the formula:

$$R_L = \frac{2\,(V_{ce})^2}{\text{Power out.}}$$

Where V_{ce} = collector to emitter voltage.

Example. $$R_L = \frac{2 \times 12^2}{50 \text{ watts.}} = 5.8 \text{ ohms.}$$

Note that the transformer impedance is calculated to allow 50 watts of audio to be developed. The reason for this is that approximately 10 watts is lost in the transformer. The closest available transformer was the Triad TY66A.

It is pointed out that it will not be possible to develop 50 watts of audio across the primary of the transformer unless the secondary is properly matched. The transistors "see" 6

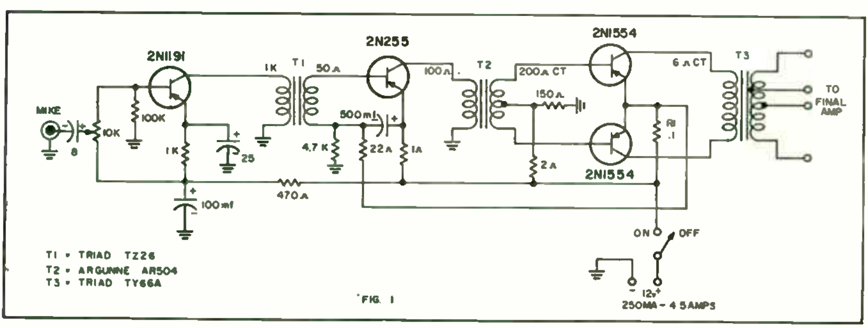

FIG. 1

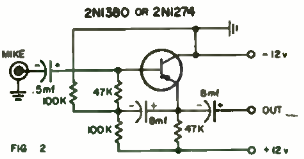

FIG 2

ohms only when the transformer secondary "sees" the correct load. An incorrect secondary load will reflect an incorrect primary load.

As the class B modulator is caused to draw current (simply by speaking into the microphone), so is a voltage dropped across the emitter resistor R1. This voltage is *negative going* in character and thus, when applied to the base of the driver through the 22 ohm resistor, causes an increase in 2N555 collector current. The 2N555 current increases from approximately 50 ma at no signal to approximately .5 amps at maximum signal.

Total current drawn by the modulator during standby is approximately 250 ma. This value will change 20% or so with changes in temperature. If necessary, the 150 ohm forward bias resistor may be raised in value to effect a decrease in the standby current. However, unless a compensating thermistor is added to the circuit, cross over distortion is likely to occur during cold weather.

Maximum current on voice peaks will be between 4 and 5 amps.

A 0.1 ohm resistor is not normally available from a supply store. However, the resistor is easily fabricated from a piece of resistance wire or even copper wire. If the latter is used the correct length may be obtained from a wire table.

The transistors should be firmly mounted to the chassis but insulated from it with mica washers. The chassis should have a reasonable area to dissipate heat generated when talking. Don't forget to remove the burrs from the holes before bolting the transistors down. A piece of fine emery paper wrapped around a flat ruler and rubbed over the area will effect the greatest amount of burr removal.

A suitable matching stage for use with the crystal or ceramic microphone is shown in Fig. 2. The stage may conveniently be mounted in the microphone case if desired.

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CHAPTER 34

Modulators for Solid-State Transmitters

Presented here are two unusual methods of modulating transistor transmitters that rival the commonly used transformer coupled collector modulation technique in performance. While neither circuit is really new (they both have basic vacuum tube histories), the implementation to transistor circuitry is unique. Both of the techniques discussed streamline the modulator by eliminating the bulky modulation transformer and its attendant matching problems. Hence, the primary advantages of reduced size, weight, and power consumption gain by utilizing transistors are further enhanced by streamlining the modulation amplifier.

DC-series modulation

The compact, high quality modulation technique shown in Fig. 1, is termed dc-series modulation because the modulating transistor is in series with the final dc supply voltage. In this circuit, the modulating transistor is shown in the emitter leg; however, it is effectively varying the collector supply voltage. The emitter is used because proper voltage polarities are available for coupling the low cost pnp germanium audio transistor to the NPN silicon rf power device.

The components shown in Fig. 1, and, in particular the 2N176 power transistor, are intended for use with a low-power (1 or 2-watt) final; however, a power transistor with a higher current handling capability could be used in place of the 2N176 to modulate higher power transmitters. In which case, select a transistor with a high current gain so that it can be easily driven by the audio driver transistor.

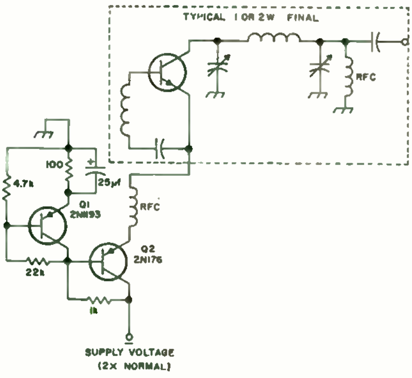


Fig. 1. DC series modulation of a transistor amplifier.

The circuit works somewhat like a controlled carrier circuit. With no audio drive, the 2N176 modulating transistor is biased to furnish about ½ of the supply voltage to the modulated stage. Then, audio peaks drive Q2 to full conduction furnishing the entire supply voltage to the final. Likewise, the audio valleys cutoff Q2 reducing the voltage to the final to zero. Hence, the carrier power output is being controlled by the modulating signal and is varied at an audio rate.

There is one minor disadvantage of this modulation technique. As indicated on Fig. 1, twice the normal supply voltage is needed. The reason for this comes to light, when the standard transformer coupled collector modulation circuit is examined. Here, the audio peak-to-peak voltage supplied by the modulation transformer effectively changes the instantaneous collector voltage from zero to twice the supply voltage at 100% modulation. However, in Fig. 1, the modulator is not supplying any additional audio voltage, it is only controlling the supply voltage. Therefore, to get the necessary voltage swing, the supply voltage must be double that normally used for transformer coupled collector modulation. Like transformer coupled collector modulation, the dc-series modulating technique is also plagued with driver feedthrough. That is, due to capacitance effects, the drive signal feeds through on downward modulation; hence, preventing full down modulation. Thus, it is difficult to obtain 100% modulation unless the driver is also modulated.

The audio driver Q_1 is a conventional common emitter amplifier that provides about 20 dB of gain. Approximately 0.14 volt applied to the base of Q_1 gives full modulation. This drive can be obtained from one or two stages of common emitter amplification depending upon the microphone used. Q_1 can be almost any type of good quality pnp transistor.

Performance wise, the dc-series technique is superior to the conventional method since it yields a greater modulated power output, lower distortion, and overall transmitter current is lower. Moreover, in a comparison test a larger detected audio voltage was achieved (with a diode demodulator) from the dc-series modulated transmitter than from the same final when it was transformer coupled collector modulated. This is an important criteria considering that what really counts, in an AM system, is the detected audio.

RC-coupled base modulation

A second circuit, RC-coupled base modulation, shown in Fig. 2, also has performance capabilities that are competitive with the usual collector modulation technique.

In this circuit, the modulation is injected to the base of the RF transistor using two resistors, R1 and R2. The effect of R1 is to linearize the waveform which is excellent for vallues of R1 between 10 and 20 ohms. Negligible improvement in linearity is achieved for larger values of R1. Also, R1 should not be bypassed for audio because bypassing introduces negative current feedback in the final RF stage at audio frequencies. Resistor R2 can range between 100 and 2,000 ohms. Ultimately, the upper value of both R1 and R2 is determined by the available rf drive power because greater drive power is needed for larger values of resistance.

The effective load resistance, presented to Q_1, is essentially equal to R2. Therefore, to reduce modulation power requirements it is desirable to make R2 large. Thus, with these conflicting requirements some type of a compromise is necessary. Hence, the values given in Fig. 2 represent a good compromise between linearity, rf drive power and modulation power requirements.

This rc-coupled base circuit yields excellent modulation characteristics because it has built-in feedback, and, it also prevents feedthrough which in turn permits 100% modulation to be easily achieved. A disadvantage is the higher power dissipation in the rf stage due to the

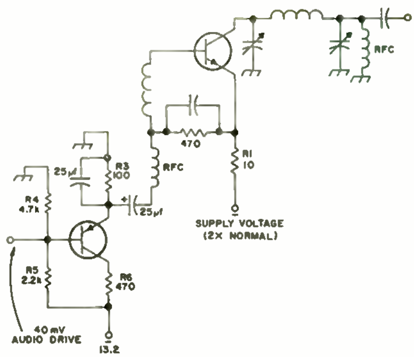


Fig. 2. RC coupled base modulation.

extra drive discussed previously. Therefore, you may want to add a little additional heat sinking to take care of the additional power dissipation.

The audio stage Q_1 uses a small-signal, general purpose audio transistor in a common emitter stage. From the circuit components shown, an undistorted output of about 2 volts can be supplied to the base of the final with 40 mV of audio drive signal. This is sufficient to provide 100% modulation for a transmitter in the 1 to 2 watt range. For higher power finals, an audio stage capable of supplying a greater voltage swing should be used. Since the audio signal is being applied to the base of the rf stage, it functions as a common emitter amplifier for both the RF and audio.

Why not try dc-series modulation or rc-coupled base modulation next time—they both perform well and do away with the bulky, expensive transformer that is difficult to match.

. . . Thorpe

CHAPTER 35

FET Audio Compressor Circuit

Presented here is an audio compressor circuit which makes use of two field effect transistors in conjunction with two conventional (bipolar) type transistors. Simplified circuitry has been employed in order to keep costs at a minimum, but without sacrificing the necessary requirements for good speech compression.

This compressor has a compression range greater than 20 dB, and is capable of handling up to 300 mV input before distortion occurs due to overload. After exceeding the compression threshold, an output change of less than 1 dB will take place for each 6 dB change in input. The high (2.2 megohm) input impedance allows the use of either crystal, ceramic, or high impedance dynamic microphones without the use of additional transformers.

A brief review of compressor circuits

Since this is primarily a construction article and not a treatise on audio compressors no attempt will be made here to analyze the numerous compressor circuits now available and in popular use. However, a brief review of compressor circuits in general may be of some help in understanding why some compressors work better than others. It may also explain why judicious substitution of components will sometimes improve the circuit, but indiscriminate substitution often results in the compressor being relegated to the junk box.

Audio compressor circuits, whether vacuum tube or transistor, make use of a dc control voltage, (either positive or negative) to control either the amplifier gain, or to attenuate the input or output signal of the amplifier by applying the voltage to a control element.

Transistorized compressor circuits can usually be classified into two types. Each type can further be classified into several methods.

In one type the control voltage varies the forward transfer characteristics of the controlled stage by either reducing the emitter current, or by reducing the collector voltage. While these two methods require only simple circuitry, and only a few components, they are subject to some serious disadvantages. For instance, large input signals will result in distortion because the transistor is driven into a non-linear portion of its characteristic curve in an effort to reduce stage gain.

In the second type the circuitry is usually a little more complex, but the results are also usually more rewarding. Here again, any one of several methods may be used.

In this type of compressor the dynamic resistance of a diode or a transistor is varied by the control voltage and made to act as a variable resistor. The voltage-variable resistor may be placed across the amplifier input circuit to shunt a portion of the input signal to ground, or it may be placed across the output circuit to shunt the output voltage to ground. The voltage-variable resistance element may also be connected either in series or parallel with other circuit components carrying signal voltages to either attenuate the signal, or to "switch" the component in or out of the circuit to reduce the stage gain.

Since in this type of compressor the variable-resistance element is usually isolated from the dc circuits, the operating voltages and currents are not effected by the application of the control voltage. However, there is a possibility that with some methods noise produced by the control element may be introduced into the signal path.

Of great importance to the proper operation of any compressor is the time constant circuit of the control voltage loop. Not only does this circuit determine the attack and release times of the controlled stage, but it must also filter the rectified ac voltage so that the control voltage will be reasonably free from ripple.

The attack and release times of some compressors may differ by as much as several hundred percent from the ideal. The ideal attack time should be about 1 msec or faster to prevent overshoot on steep wave front signals. In some circuits such a fast attack time could cause transient "thump" due to the inability of the amplifier to follow rapid changes in current or voltage.

Release times are usually made longer than the ideal time of 10 msec. If the release time is too fast the amplifier will recover before the lower audio frequencies have been filtered out. On the other hand, if the release time is too slow a weak syllable following a strong one

Bernard, former W2GRK and W2HPO, has been licensed since 1933. He's a research analyst with the Department of Defense in Washington.

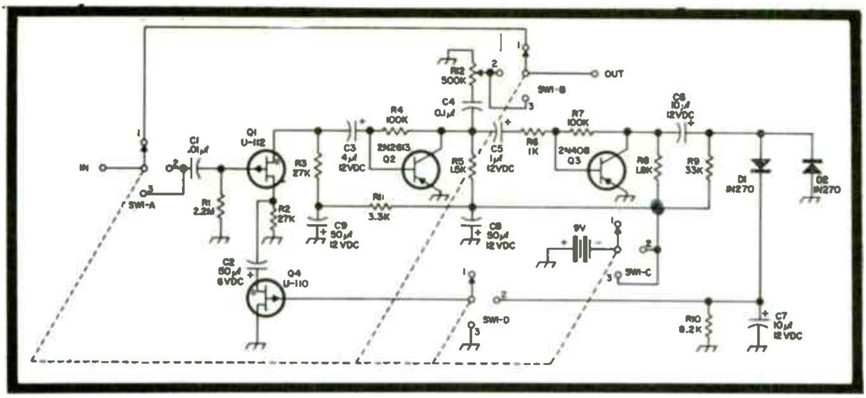


Fig. 1. Schematic of the FET compressor. There should be a dot indicating a connection at the cross-over between R8 and SW1-C

will be compressed just as much as the strong one.

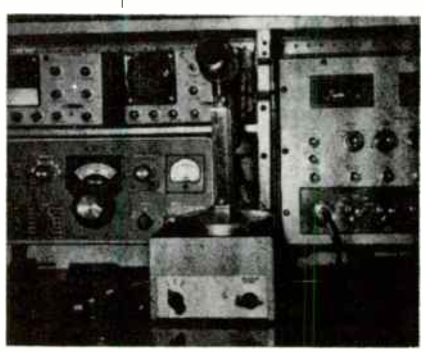
The compression range will depend on the circuit used and may vary from only a few dB to as much as 60 dB. The choice of compression range will depend upon the application of the compressor. A small amount of compression will limit audio peaks but will add little to the improvement in "talk power" Too much compression on the other hand will make the signal sound harsh. For speech compression a range of about 15 to 21 dB has been found to give the best results.

At this time perhaps one of the more popular misconceptions about audio compressors should be clarified. There seems to exist among many people the idea that the audio compressor will increase the peak power output of a transmitter. This is not quite true. Compression will improve the peak power output only if the low level audio stages of the transmitter have been deficient, or the microphone output level had been too low to supply sufficient audio to the transmitter, and then only because most compressors also act as audio preamplifiers. The primary purpose of the compressor however, is to improve the average power, or "talk power" of the transmitter by providing additional amplification to soft spoken syllables, and by reducing the amplification of loud syllables so that peak power output is attained over a greater percentage of the time. This higher average output will show up in SSB in on-the-air reports that the receiver S-meter appears to "hang" close to the peak with only a slight variation between syllables. It will also show up on the transmitter plate current as a higher average reading.

Circuit description

The heart of the audio compressor described here consists of a pair of field effect transistors, the Siliconix U-110, and U-112, which were recently made available for the price of $2.75. Replacing these FET's with bipolar transistors would require sophisticated circuitry to perform the same functions, and at a considerably greater cost. By using the U-112 as the controlled amplifier stage it was possible to achieve a high impedance input, (2.2 megohms) without the use of an input transformer or an additional emitter follower stage. The U-110 in this case functions as the voltage-variable resistor to control the amplifier stage gain through the application of negative feedback to the source of the controlled stage. In addition to the 20 dB of compression previously mentioned, the result is a low noise, low distortion preamplifier with high gain.

No noise measurements were made, but scores of on-the-air tests proved to be very gratifying in this area. Also, although no extensive distortion measurements were conducted, cyclogram tests indicated negligible distortion.



The FET compressor in use at K3VNR's shack.

The signal after being amplified by Q1 is further amplified by Q2, a 2N2613. This is a low noise transistor especially suitable for preamplifier circuits. The forward bias resistor R4 of this stage is connected between the base and collector in order to reduce the number of circuit components and still achieve some measure of stability.

After amplification by Q2 the audio signal to the transmitter is taken off through the level control R12. Since even after compression the output voltage will still show a gain of 6 to 9 dB the level control is used to reduce the output to a suitable level for the transmitter.

A portion of the signal from Q2 is also amplified by Q3 to a higher level which is then rectified by D1 and D2 to become the dc control voltage. No attempt should be made to take the audio signal from the output of Q3 since the signal at this point is in the form of a square wave and will sound highly distorted.

Notice that the forward bias resistor for this stage, like that of Q2, is also connected from collector to base. This means that a small amount of ac voltage will be fed back from the collector to the base and will also appear at R6. R6 will isolate this small amount of feedback so that it will not appear at the audio output as distortion. If the audio output is viewed on an oscilloscope, this distortion, if present, will appear as a bright spot at the baseline cross-over in mild cases, and in severe cases the baseline will actually show. Increasing the value of R6 will prevent this type of distortion but will also lower the available ac voltage at the base of Q3.

The output from Q3 is rectified by D1 and D2 and the resultant dc voltage is filtered by C7. This dc voltage is the control voltage which varies the drain-source resistance of Q4. Besides acting as the dc filter, C7 in combination with R10 sets the time constant of the control loop.

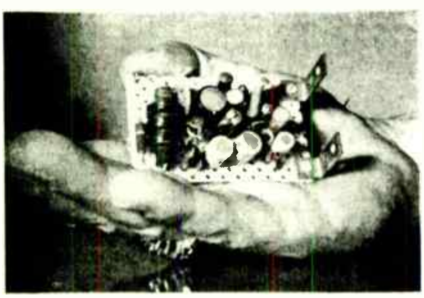
When a positive potential is applied to the gate of Q4 the drain-source resistance is increased, effectively switching off C2 which is the source bypass capacitor of Q1. When R2, the source resistor, is unbypassed an ac voltage drop is developed across the resistor and the gain of Q1 is reduced. With R2 in an unbypassed state negative feedback occurs and the percentage of harmonic distortion is reduced by an amount which is almost equal to the amount of compression.

When the amplifier gain is reduced the control voltage is also reduced and C7 discharges through R10. The bypass capacitor C2 is now "switched" back into the circuit and the amplifier recovers. Recovery time is determined by C7, R10 which has a time constant of approximately 82 msec.

In reality C2 is not switched in and out of the circuit by Q4 since Q4 has an irreducible amount of internal resistance between the source and the drain. However, since a switch is characterized by low resistance when it is closed, and high resistance when it is open, then for all practical purposes Q4 with its rise and fall in drain-source resistance is switching C2 on and off.

One of the advantages of this circuit is, that because Q4 is capacitor-coupled to the source of Q1 there is no change in bias current and therefore no distortion due to driving Q1 into the nonlinear region. Another advantage is that the sudden application of control voltage will not cause an objectionable transient thump.

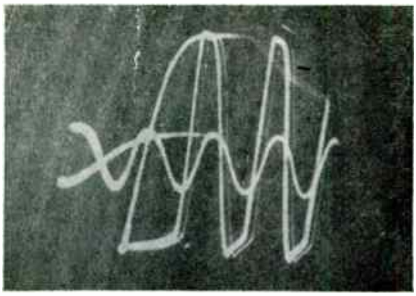
The circuit does have a minor fault, but one which will have no effect on normal operations. As the input voltage is increased the control voltage will approach pinchoff, and "remoting" will take place .This is because Q4 will be operating in the "remote" region of the characteristic curve. In this region, as the positive potential to the gate is increased there will be no increase in the compression range since the FET does not follow square law behavior at low drain currents and there is no sharp cutoff. This, incidentally, may illustrate the difference between "pinchoff" as applied to the FET, and "cutoff" as applied to a triode vacuum tube when the grid bias is increased. This small fault should cause no trouble when the compressor is used with a microphone input since the voltage developed by the microphone will never reach a level where "remoting" can take place, and is only mentioned here as a point of interest.



This photo shows the small size of the compressor board (1½" x 2½"). The RF filter mentioned in the text is mounted at the left.

Construction And Testing

No difficulties should be encountered in constructing this unit. With the exception of the switch and output level control, all components, including an RF filter not shown in the circuit diagram, were mounted on a piece of perf-board measuring 1½ x 2½ inches. An RF filter was included in the final construction as an added precaution against stray RF being



This oscillogram was made by driving the amplifier to overload distortion, then switching in compression. Note the lack of distortion when compression is applied to the signal. What appears to be phase shift is a result of the method used to get this photo and is not caused by the compressor.

picked up and rectified, but this filter may not be necessary in all cases.

Wherever possible the components were mounted in a vertical position in order to conserve space. With double-ended components, such as resistors, and some capacitors, the technique is to bend one lead back towards the body of the component so that both leads will face in the same direction, thus forming a single-ended component. The two leads are then placed in adjacent holes in the perf-board.

Although the unit is compact enough to be mounted in a small Minibox, one measuring 4 x 5 x 6 inches was used. This large size Minibox allows the controls to be mounted conveniently on the front panel without crowding, and also allows the use of a larger sized battery. Instead of chassis type of connectors, cable type connectors attached to short pieces of shielded mike cable were used. This eliminates the use of a patch-cord between the compressor and the transmitter, and permits easy changes to be made in the future in the event that a new transmitter might use a different type of mike connector.

The test procedure is quite simple. A VTVM, and oscilloscope (if available) is connected to the output. A 400 Hz audio signal is fed to the input. If a signal generator is not available a microphone picking up a beat note from a receiver will suffice.

The compressor switch is turned to position three and the audio input level is adjusted to give an output reading of 100 mV with the output level control turned full up. With the switch in this position no compression is being applied to the signal. Note the waveform and amplitude of the signal on the oscilloscope; the display should show a pure sine wave.

Without further adjustment of the input signal, turn the compressor switch to position two. With the switch in this position compression is now being applied to the signal. The output reading should drop to 18 mV for 15 dB of compression. The oscilloscope display should also indicate this drop in output, and waveform should remain a pure sine wave but lower in amplitude.

Switch the compressor back to position three and increase the input signal so that the output now reads 200 mV. This is an increase of 6 dB in output which also roughly corresponds to a 6 dB increase in input since at these signal levels the amplifier gain is quite linear. Again note the oscilloscope display.

Without changing the input signal switch the compressor to position two to apply compression. The output reading should now be about 20 mV indicating 20 dB of compression. Except for a lower amplitude the oscilloscope pattern should remain the same. Note that although the input signal had been increased by about 6 db with no compression, under compression the output signal change was about 1 dB.

This completes the testing, but an interesting little experiment can be conducted here by carrying these tests to the point where "remoting" takes place. It is also interesting to see that when the input is increased to a point where overload distortion takes place with no compression, by switching in compression the overload distortion will disappear. This is because negative feedback is now applied and a new input level is set for overload distortion.

Operation

Because there is only one control to adjust, the output level control, operation is virtually self-explanatory.

With the compression switch in position one, speak into the mike and adjust the transmitter audio gain control for proper operation. Since the mike is feeding straight through to the transmitter the audio gain control should be at the usual setting.

Turn down the compressor output level control and turn the switch to position two. Speak into the mike and slowly turn up the output control. Proper setting for this control will be indicated on the transmitter when full modulation is reached. If VOX is used, and if it had been marginal before, back off a little on the transmitter audio gain control as the compressor output level control is advanced. A few minutes of adjusting should show the proper settings for the audio gain control and VOX controls of the transmitter, and the output level control of audio compressor.

Note that when the compressor is first turned on an initial transient surge will render the compressor momentarily inoperative, but the amplifier will recover rapidly. Also note that position three of SW1 is not used during operation.

Conclusion

Although a 9 volt battery supply was used in the design of this compressor, voltages ranging from 6 to 12 volts have been used successfully. An ac pack could be used providing it is well filtered. The power source could also come from the transmitter filament transformer by using a pair of diodes in a voltage doubler circuit, and a proper filter. Bear in mind though, that any additional leads brought into the Minibox could introduce ac hum, and stray RF pickup.

If compressor is to be operated directly from a 12 volt car battery in a mobile installation, all 12 volt capacitors should be changed to 15 volt units. Also observe proper ground polarity; since the circuit as shown in the diagram has a positive ground.

Only the U-110, and U-112 FET's were tried in this circuit. Undoubtedly with minor component changes other FET's will work just as well. In any case Q4 should have a low pinchoff.

The diodes D1, D2 have been specified as 1N274's, but other diodes, including the "dollar a dozen" variety, were tried with satisfactory results.

Resistor R9 sets the compression threshold. A higher resistance will lower the threshold. Replacing R9 with a combination fixed and variable resistor will allow the threshold to be adjusted.

Several other refinements, such as a meter to read compression, could be made, but the additional cost would defeat one of the primary aims, a low cost audio compressor.

K3VNR

TEST EQUIPMENT

CHAPTER 36

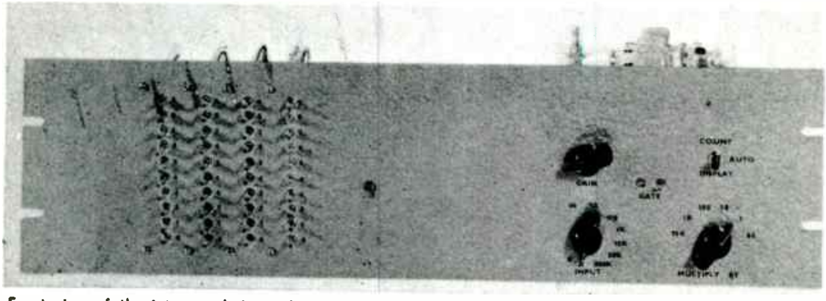
Integrated Circuit Frequency Counter

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 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 cycle 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



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

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.

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.

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

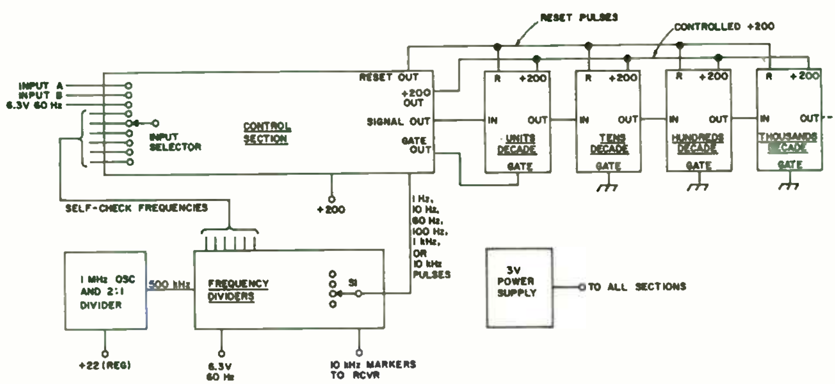


Fig. 1. Block diagram of the complete integrated circuit frequency counter. Any number of decades may be used, but for proper display, the units decade should be to the right, the tens decade to its left, etc.

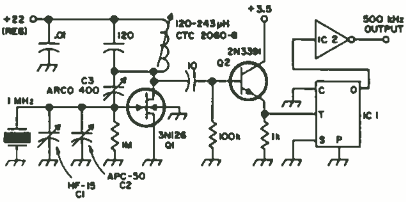


Fig. 2. The oscillator and 2:1 frequency divider used with the decimal counter. This unit was built into a separate box and may be used for obtaining markers as described in the text. Although the FET is a Motorola 3N126, an MPF-105 is less expensive and would probably work. IC1 is a Fairchild 923 or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

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 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

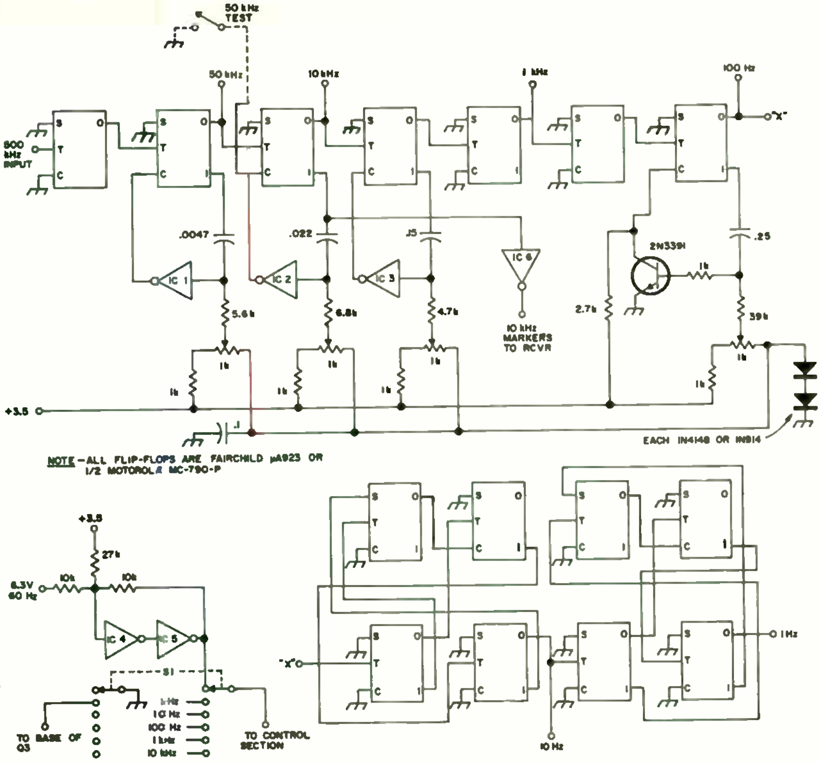


Fig. 3. The frequency dividers used in the IC counter. Integrated circuits IC1 through IC5 are one-half of Fairchild 914's or part of Motorola MC-789-P or MC-724-P; IC6 is a Fairchild 900 or one-half a Motorola MC-799-P.

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 mono-

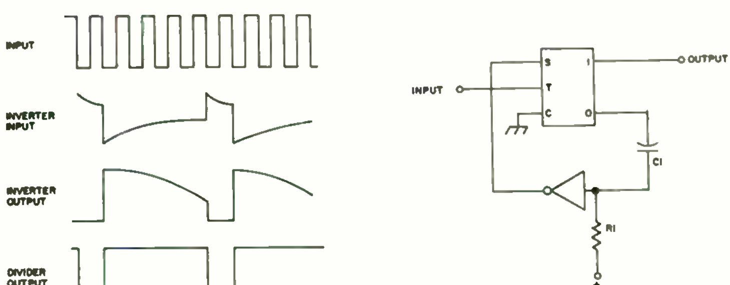


Fig. 4. The basic 5:1 frequency divider using a J-K flip-flop, and RC circuit and an inverter, along with the waveforms.

stable 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, 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 remembered 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.

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

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.

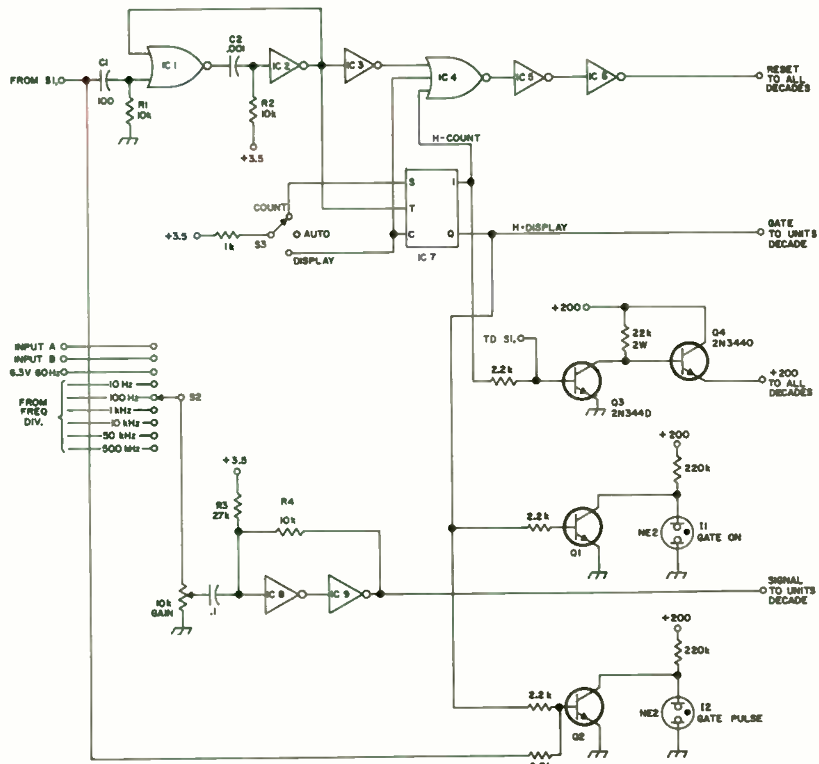
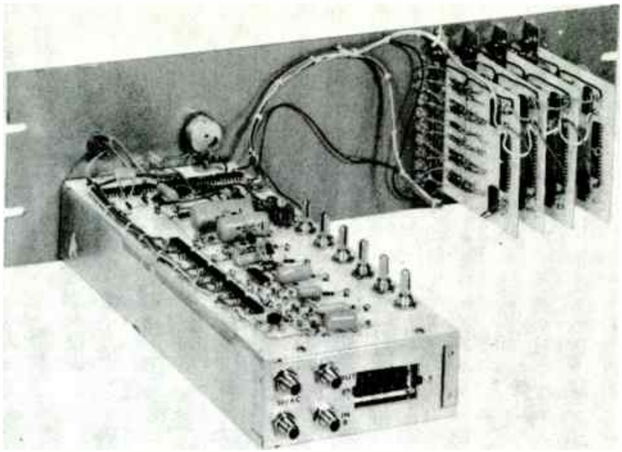


Fig. 5. The control section of the digital frequency counter. IC1 is a one-half a Fairchild 914, one-fourth a Motorola MC-724-P or one-third a Motorola MC-792-P. IC4 is one-third a Motorola MC-792-P. IC2, IC3, IC5, and IC9 are one-sixth of Motorola MC-789-P, one-fourth of Motorola MC-724-P, or one-half of Fairchild 914. IC6 is a Fairchild 900 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.

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.



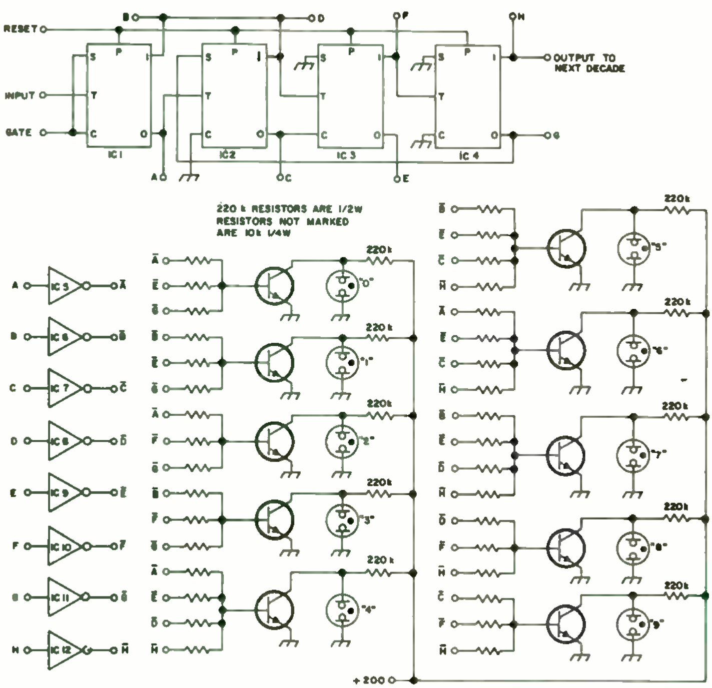


Fig. 6. A typical counting decade. In this circuit integrated circuits IC1 through IC4 are one-half Motorola MC-790-P's or Fairchild 923's. IC5 through IC12 are one-sixth Motorola MC-789-P's, one-fourth Motorola MC-724-P's or one-half Fairchild 914's. All transistors are 2N3877's or Poly Paks 2N1893's. All neon lamps are NE-2's.

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.

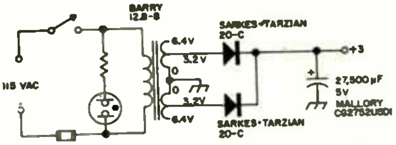


Fig 7. Three-volt power supply for use with the integrated circuit frequency counter. A truth table showing the proper levels on each logic line are shown in Table I.

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.

CHAPTER 37

Integrated Circuit Crystal Calibrator

The crystal calibrator ranks with the SWR bridge and code practice oscillator in coverage in the various ham magazines. The names have ranged from the obvious "crystal calibrator" to the less obvious "transistor secondary frequency standard" and "multical". The type of circuitry involved has likewise ranged from simple tube circuits to more complex solid state digital circuits.

The IC crystal calibrator could be considered one of the more complex of the above listed calibrators, providing 100 kHz and 10 kHz outputs. Because of the use of integrated circuits instead of transistors, it is, however, one of the simplest and cheapest. The one described was built for under $13.

The truth table shows all possible combinations of inputs (A and B) and the resulting output (Q) for the NOR gate. The Fairchild μL914 contains two of these gates in one package.

A	B	Q
L	L	H
H	L	L
L	H	L
H	H	L

H = high
L = low

Notice that a high output occurs only if both inputs are LOW.
Any other combination produces a LOW output.

The truth table for the flip-flop is as follows:

S	C	Q
H	L	H
L	H	L
L	L	Reverse
H	H	No change

The states S and C are considered before the trigger has occurred, and Q is the condition of the flip-flop after the trigger occurs with a given S-C combination. The flip-flop has two states. It is SET when Q is LOW. It is cleared when Q is HIGH. Another input, called PRESET, can be used to make Q LOW regardless of its state and *without a trigger*.

Binary number system

For the following discussion, forget that 1 and 1 is 2 and we can clear up the last bit of theory needed to understand the operation of this calibrator. To simplify large digital computers, a number system was evolved that contains only two digits (0 and 1) rather than ten digits (0 thru 9).

To see how this system works, think of the odometer on your car and imagine each wheel on this odometer using only two numbers—0 and 1. As the car sat on the showroom floor, the mileage might have read 0000. When you drove it off, registered the first mile (and lost

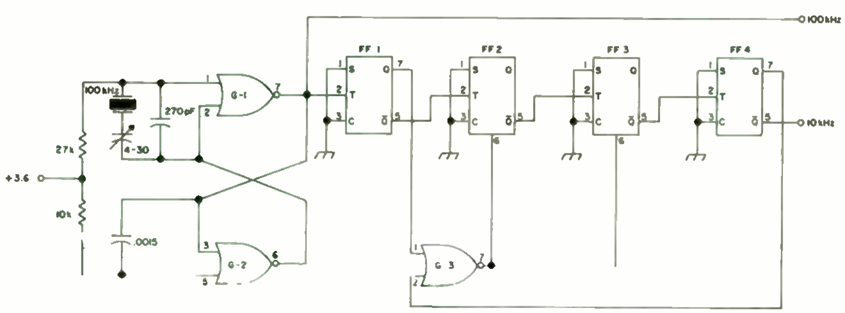


Fig. 1. Block diagram of K1DCK's crystal calibrator. Output is both 100 kHz and 10 kHz and total cost is under $13.

$1000 resale value), it would read 0001. Okay so far. But when you register the second mile and the right wheel on the odometer advances one more place, a zero appears again since that is the only other number! This second mile clears the first wheel back to zero and advances the second wheel to one. So now the odometer registers 0010. This represents the decimal number two in the binary system. The third mile would read 0011. The fourth mile turns the first wheel to zero, which turns the second wheel to zero, which advances the third wheel to one, and so on. With only four wheels on the odometer, we could register up to 1111 which would be fifteen miles. The sixteenth mile would register as 0000, so we would need a fifth wheel for this number (10000). A list of all the binary numbers from zero thru ten is shown below.

0	0000	6	0110
1	0001	7	0111
2	0010	8	1000
3	0011	9	1001
4	0100	10	1010
5	0101		

These binary numbers are called binary-coded decimal or simply BCD.

Pulse generator

The output of the pulse generator is a 100 kHz pulse. The high to low transition will be used to trigger the counter. The idea of using the µL914 as a crystal-controlled pulse generator is from a Fairchild application note. With the values shown in Fig. 1, the circuit works well with any of the surplus crystals in HC13/U holders. Another crystal in an HC6 case (also surplus) worked with different values of C1 and C3. Juggle them around; you can't hurt anything.

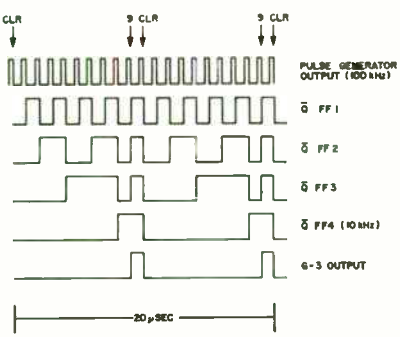


Fig. 2. Waveforms of various points in the BCD counter. Peak-to-peak voltages will vary from one to three volts.

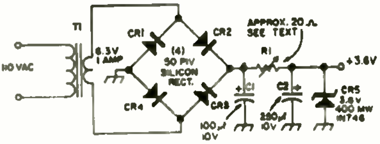


Fig. 3. W1CFW's power supply.

For those of you on a tight budget, stop here. You have just constructed the cheapest and most reliable 100 kHz calibrator anywhere! However, for those of you who aren't afraid of the XYL, we'll add on the counter/divider.

BCD counter

Sometimes called a divider, this circuit has the ability to count to nine and reset to zero on the tenth count. It is called a divider, because for every ten pulses at its input, it produces one pulse at the output—division by ten. Naturally, for every 100,000 pulses at the input, the output will be 10,000 pulses: the 10 kHz marker. Any inaccuracy in the 100 kHz pulse is divided by ten also, but the percentage of error remains the same.

Looking back at the odometer example, each wheel had two states—ZERO and ONE. A flip-flop also has two states—SET and CLEAR. So flip-flops can easily be substituted for wheels, making a counter which counts in exactly the same way the odometer did. But this counter counts to fifteen! The counter is fooled into thinking it has reached fifteen when, actually, it has only reached nine. The action of G-3 does this.

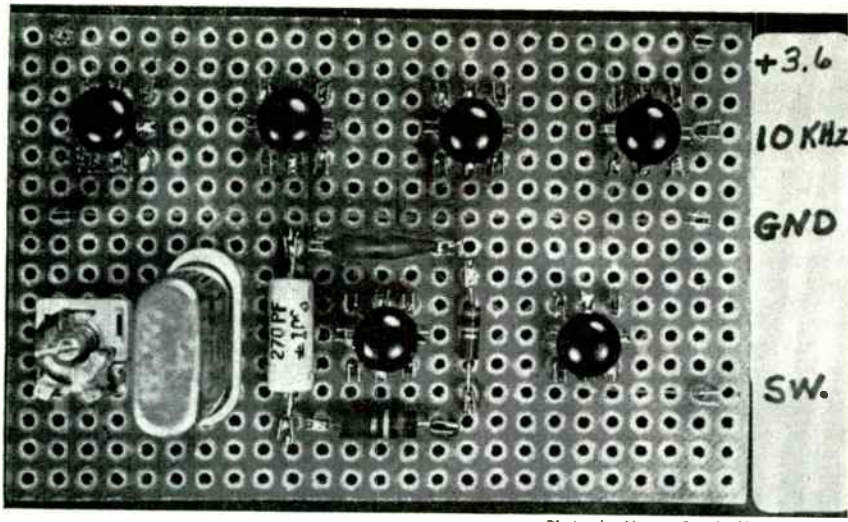
In the binary system, nine is 1001 and fifteen is 1111. The states of the flip-flop when it reaches nine are as follows:

FF	Q
FF1	H
FF2	L
FF3	L
FF4	H

To make the counter think it has reached fifteen, G-3 will go H and preset FF2 and FF3, changing Q to H. This is done by applying Q output of FF1 and FF4, both of which are LOW at this time, to the input of G-3. FF1 and FF4 are SET only when the counter has reached nine.

Power supply

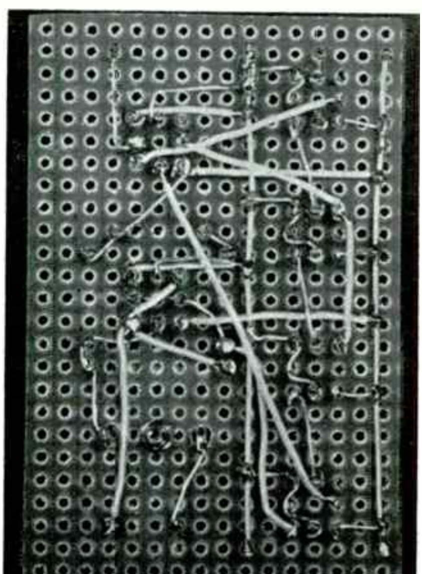
The calibrator draws 80 mA at 3 volts and 95 mA at 3.6 volts. A power supply similar to W1CFW's may be used and is reprinted in Fig. 3. Adjust R1 for 20 mA zener current, under load. The separate power supply is recommended since tube-type receivers will



Photos by Harvey Benoit, Haverhill, Mass.

Top, left to right—FF1, FF2, FF3, FF4. Bottom, G1/G2, G3/G4.

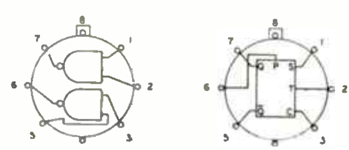
not have suitable voltages available. The power supply won't be very big; if you have a tube-type receiver, you're probably not that finicky about miniaturization anyway.



View of the underside of the board, showing wiring. The right bus is +3.6 V. The bus near the center is ground.

Construction

The details used in this project are clear enough in the photographs. Flea clips are mounted on a piece of Vectorboard about 2¼ x 3½ inches. No attempt at miniaturization was made, so you may want to try. The crystal socket and trimmer were mounted, then all the wiring underneath was done. The large bus across the middle is the ground bus; and the one at the top is the +3.6 voltage. Wire pin 4 of every can to ground and pin 8 to the +3.6 voltage. Pin connections are shown in Fig. 1. Wiring is point-to-point; and teflon-covered wire was used for crossovers. The IC's were then mounted using normal precautions with semiconductors. On some of the earlier epoxy units, pin 8 was identified by a red line; it is identified by a flat spot on present units. The Fairchild UX8991428X is used for G-1/G-2 and G-3/G-4. Fairchild UX8992328X's are used for FF-1 thru FF-4.



(a) 8991428 DUAL 2-INPUT GATE (b) 8992328 JK FLIP-FLOP

Fig. 4. Bottom view of Fairchild gate and flip-flop IC's. The right side of the gates should be curved.

Operation

The output of the calibrator may be connected thru a small capacitor, or "gimmick", to the antenna terminal of your receiver. As many receivers have provisions for a plug-in calibrator, the capacitor may already be in. My NCX-3 uses a 6-inch piece of coax. One end of the center conductor is connected to the output of the calibrator and the shield on the other end is connected to the antenna terminal. The small capacitance between the inner and outer conductors makes the small coupling capacitor.

There are two ways to turn the calibrator off when not in use. One is to put a switch in the primary of the power supply transformer. Don't switch the +3.6 voltage. Operating the power supply without a load may damage the zener. Another way is to make use of G-4. Fig. 5 shows how this may be done. With pin 5 always LOW and pin 3 open, the output of G-4 is HIGH. This HIGH has no effect on the operation of the 100 kHz oscillator. If this point goes LOW, then pin 6 will be clamped LOW and the oscillator stops. This point will be LOW if pin 3 is HIGH, which would be the case if the switch were in the OFF position.

Fig. 5. Gated on/off switch.

Harmonics were quite useable up to 30 MHz. It's possible they may be heard on six meters, but no receiver for this frequency was available. In the event the harmonics are not useable on six, this should present no problem, since the calibrator could be hooked into the antenna terminal of the HF receiver.

This calibrator points out several of the advantages of integrated circuits. It is small in size and cost, but big in performance. Since it "flew" right off the drawing board, and is easily duplicated, the ease of circuit design is apparent.

. . . K1DCK

CHAPTER 38

IC Pulse Generator

"What would a ham want a pulse generator for?" asks the amateur operator. "We're not in the radar business; pulses are what we want to eliminate, not generate!" Well, first let's find out what pulses are; and then find what hams can do with them.

Webster says that a pulse (as the word applies to "radio") is "an electromagnetic wave or modulation thereof, of brief duration". This definition allows us a great deal of latitude. However, the author feels that when most electronics-oriented people see the word "pulse" they think of a rectangular wave whose "on" time is short compared to its period.

A pulse train is shown in Fig. 1, where t_o is the "on" time, and t_p is the period, or distance between similar points on adjacent pulses. A pulse is *not* required, in general, to be rectangular; it *may* have any of a variety of shapes, like those shown in Fig. 2. These special-purpose pulses are sometimes used in pulse systems where it is desired to carefully control the bandwidth of the pulse signal. A pulse generator that will produce all of the pulses of Fig. 2 is not within the scope of this article; so we'll stick to the common rectangular pulses of Fig. 1.

Perhaps the simplest pulse generator known to solid-state-electronics is the unijunction relaxation oscillator, as shown in Fig. 3. It puts out a train of pulses whose repetition rate ($1/t_p$) is controlled by R_1 and C_1 and whose "on" time (t_o) is controlled by R_2 and C_1. This simple circuit can provide a timing pulse for everything from an electronic metronome to a "V.T." fuse. With a few slight modifications, the unijunction relaxation oscillator can be synchronized and the R-C charging of the emitter-capacitor modified to linear charging.

Since the circuit of Fig. 3 produces a pulse of rather short duration, we can use it to trigger a monostable multi-vibrator to obtain longer pulses. It is in such service that the monostable multivibrator is called a "pulse-stretcher". A realization of this simple unijunction-plus-monostable multivibrator type of pulse generator is shown in Fig. 4. A pulse amplifier was added between the unijunction section and the monostable multivibrator to provide both isolation and the level shift necessary. The values shown in Fig. 4 will give pulse repetition frequencies (PRF's) of 100 to 1000 Hz and pulse widths of 30 μsec to 500 μsec. The values of R_1C_1 and R_2C_2 could be switched to provide other PRF's and pulse lengths. In spite of the fact that this simple pulse generator has only five semiconductor packages and about a dozen other components, it is a very useful unit to have around the shack.

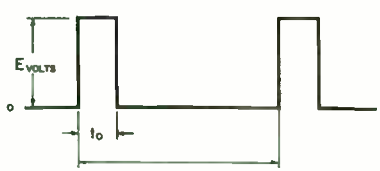


Fig. 1. A typical pulse train. The time t_o is the "on" time, and t_p is the period.

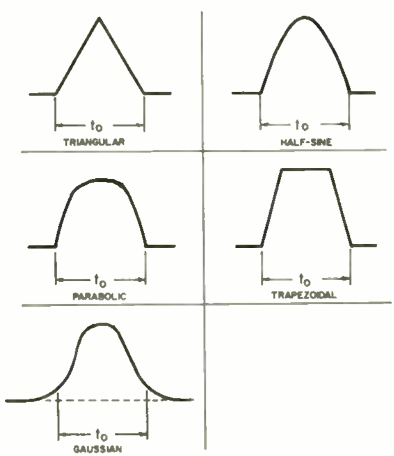


Fig. 2. Other pulse shapes which may be encountered—triangular, half-sine, parabolic, trapezoidal and gaussian.

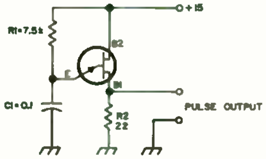


Fig. 3. A simple unijunction transistor relaxation oscillator which will generate the pulse chain. The pulse repetition rate is determined by R1 and C1; the on time is controlled by R2 and C1.

The "sync" connection to the circuit can be either used as a sync output or sync input. A negative-going wave, from a low impedance source, put *into* the sync port will synchronize the pulse generator.

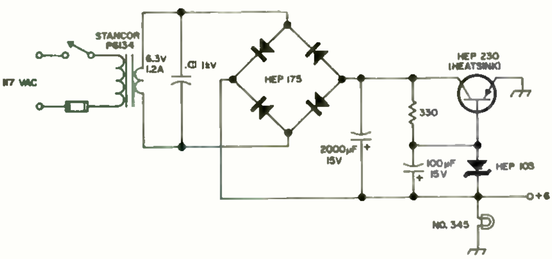
By switching the "pulse length" switch to "SQ", the monostable multivibrator, which forms the pulses, becomes a simple ÷2 flip-flop. In this mode, *nearly perfect* square waves are produced from 10 Hz to 100 kHz —just one half the normal PRF rate. This feature was added, because it was so simple; the monostable multivibrators which perform the "delay" and "pulse length" functions are basically flip-flops modified for monostable use.[3]

In the interest of simplicity, the four ports of the pulse generator are direct-coupled. Coupling capacitors, of proper size to accommodate the particular pulse one is negative-going pulse may be taken out from this same "sync" port. The pulse generator can be synchronized by waveforms which are multiples of its free-running frequency, and used as a divide-by-n unit.

The simple pulse generator of **Fig. 4** can be vastly improved upon, to create a general-use amateur model. Such a pulse generator is shown in **Fig. 5**. By replacing the unijunction oscillator with one using an HEP 556 integrated circuit, the need for both +15V *and* +6V as circuit supply voltages is eliminated. Only +6V must be supplied to this more versatile pulse generator.

Since the HEP 556 (a three-input ECL gate) oscillator puts out a rectangular waveform that has a logic level which is compatible with the HEP 558, no isolation amplifier is necessary. The first HEP 558 is used as a delay generator. This delay generator is a monostable multivibrator whose output pulse triggers the following stage at the *end* of its pulse. The second HEP 558, also connected as a monostable multivibrator, is the pulse generator.

To add versatility, several other transistors have been added: a split-load phase inverter, two output amplifiers, and a sync output emitter-follower.



A regulated power supply for the pulse generator shown in Fig. 5.

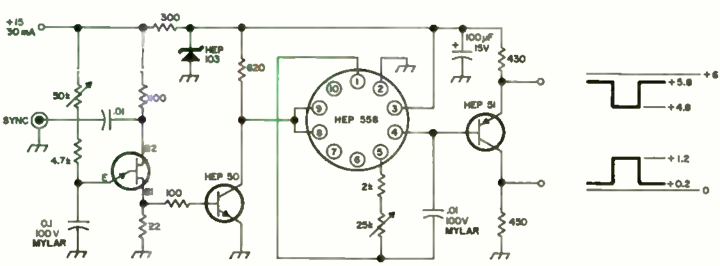


Fig. 4. In this pulse generator, a pulse amplifier has been added to the simple unijunction relaxation oscillator to provide isolation and the necessary shift level.

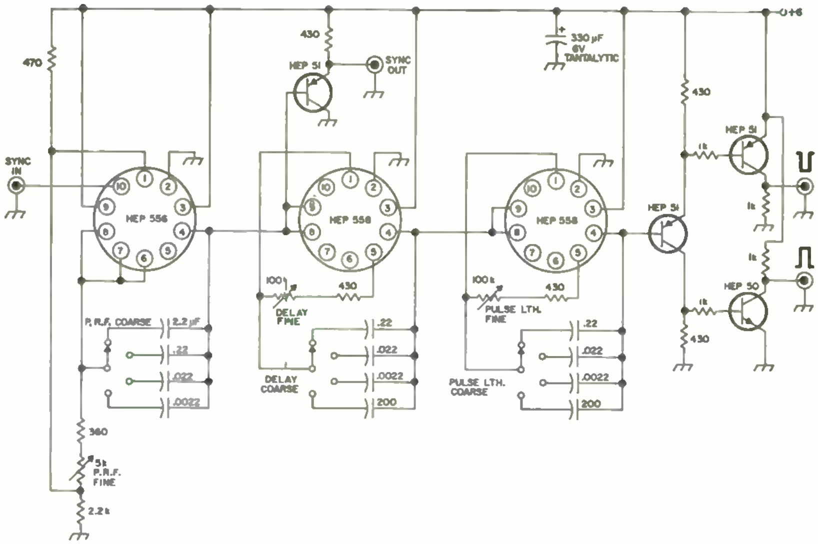


Fig. 5. The pulse generator of Fig. 4 can be vastly improved upon by replacing the unijunction transistor circuit with an integrated circuit. This pulse generator will provide complimentary outputs, variable pulse length, variable pulse repetition frequency and variable pulse delay. It will drive a capacitive-coupled 500-ohm load.

The waveforms in Fig. 6 show the two (complimentary) outputs when the generator is asked to produce its narrowest pulse. This narrow pulse clearly shows the rise times to be expected of our generator.

The output pulse is available either as a positive-going pulse starting near zero and going to nearly +6 volts, or the compliment of this. The compliment, of course, is a negative-going pulse starting near +6 volts and going to nearly zero.

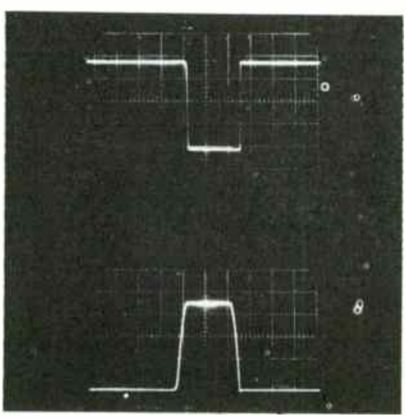


Fig. 6 Output waveform of the circuit in Fig. 5 for minimum pulse width. The sweep speed of the oscilloscope is 1 μsecond per division.

The finished pulse generator is shown in Fig. 7 and 8. The generator was built in a Bud cabinet (CD-1480) for two reasons. Firstly, this was the cabinet the author had on hand, and secondly, the 8" x 8" panel allows enough room to mount all the controls. The circuit board picture, Fig. 8, shows all the generator circuitry except the power-

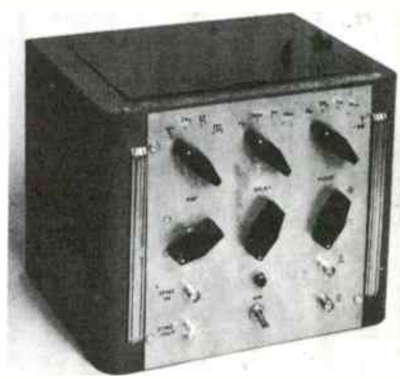


Fig. 7 The pulse generator. This unit uses integrated circuits and transistors to provide pulse repitition rates from 20 Hz to 200 kHz, pulse delays from 20 microseconds to 20 milliseconds and pulse widths from 20 microseconds to 20 milliseconds. A square wave output is also available.

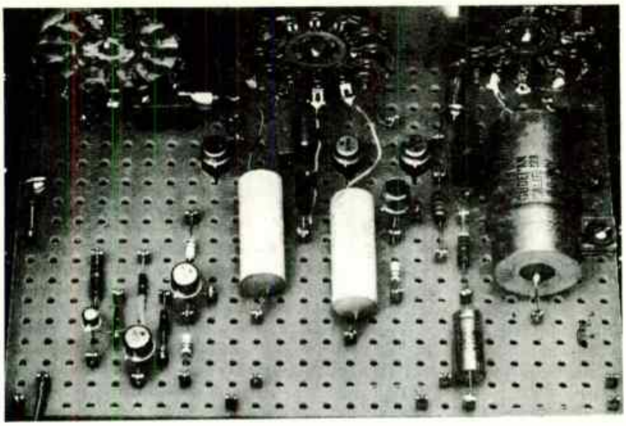


Fig. 8 Circuit board used in the pulse generator shown in Fig. 7. All components are mounted on this board except the power supply.

supply Obviously there is room to spare, and a much smaller unit could be built if miniature switches and pots were used.

Now that we've generated our pulses, let's have a look at their uses. The most commonly used "pulse" is the square wave, which the second pulse generator will produce.

Since the radio telegraph, radio teletype, and television modes of transmission are all based on pulse sequences, a pulse generator can be useful in the design, simulation, and testing of amateur equipment for these modes. The exact nature of the use of the pulse will depend on what the user is trying to do. If he wished to turn on a transmitter with the pulse, we'd call it "pulse (amplitude) modulation". If he wishes to turn a signal off with a pulse, we'd call it "blanking". Or turning a signal on for a desired time interval, after a desired interval would involve "delaying" and "gating". Of course, the uses of combinations of these functions are limited only by the imagination.

. . . W6GXN

CHAPTER 39

Wien Bridge Oscillator

Virtually every audio oscillator, that the author has ever seen in laboratory use, is of a type called the Wien Bridge. This type of oscillator is characterized by a particular configuration of R-C tuning network. The original circuit of the Wien Bridge Oscillator is shown in Fig. 1, as it was first constructed using vacuum tubes.

In Fig. 1, the two stage circuit sustains oscillations because of the phase-shift of the bridge (at a particular frequency) *and* the phase shifts of the two amplifiers (assumed constant over the frequency range of interest). Such an RC oscillator would produce a highly nonlinear waveform (like that of another R-C oscillator, the astable multivibrator), if it were not for the nonlinear resistance "r". The resistance r is variously called a positive-temperature-coefficient thermistor, a barretter, or a light bulb.

When the circuit of Fig. 1 is in its desired state, the tubes are running in class A, and the output is sinusoidal. A change in operating state toward class C (which would produce a much larger output of highly distorted waveforms) causes more current to flow in the R_2-r side of the bridge. This increases the temperature of the light bulb (r), which causes its resistance to rise. The increase in resistance of r causes the gain of the amplifier stage V_1 to decrease, which restores our original operating level.

To see how the lamp resistance varies with current, Fig. 2 depicts a (commonly used) 6 watts 120 V lamp E-I plot, with several lines of constant resistance drawn in for reference. The translation of the tube-type Wien Bridge circuit into a transistorized version has had many problems, and the solution of these problems has been so complicated that the basic simplicity of the Wien Bridge oscillator often has been lost. In many a transistorized Wien Bridge audio oscillator, when the problem areas have been designed around, the resultant circuit hardly resembles the original Wien Bridge at all. This is not bad, per se, and several good Wien Bridge audio oscillator designs have come forth using bipolar transistors.

Basically, the reason that the Wien Bridge oscillators using bipolar transistors are so hard to build is that ordinary transistors have a relatively low input impedance in the common emitter configuration.

In Fig. 3, we see a hypothetical Wien Bridge oscillator using bipolar transistors. Since the input impedance from base to ground is fairly low (approximately h_{fe}xf), this low impedance shunts R_2 and upsets the requirement that $R_1 = R_2$. Also, since h_{fe}xr is amplitude sensitive, frequency will be dependent on amplitude. These two problems generally force the designer to:

1. Use low values of R_1 and R_2, together with large values of C_1 and C_2. This means that resistance tuning *must* be employed.
2. Use some other negative feedback method for controlling amplitude, rather than

Front view of W6GXN's Wien Bridge audio oscillator.

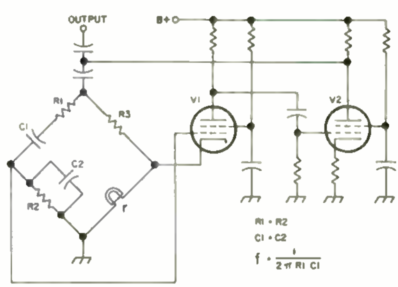


Fig. 1. Typical tube-type Wien Bridge audio oscillator.

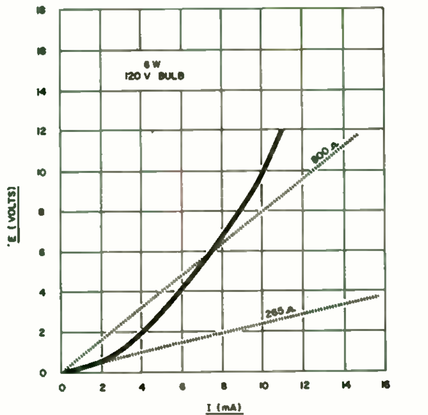


Fig. 2. E-I plot of a 6 W, 120 V pilot lamp. Two constant resistance load lines are also shown for reference.

the simple lamp-in-the-emitter method. Negative temperature coefficient thermistors and forward-biased diodes are two of the nonlinear elements used for this.

With the advent of field effect transistors, the design of simple solid-state Wien Bridge oscillators came within easy reach. The FET has an inherently high input impedance in the common source configuration. However, most of the designs that the author has seen using an FET as the input amplifier, have not used the same sort of lamp amplitude control as used in the older tube-type circuits.[3,4,5]

The circuits below were redesigned from the old vacuum tube Wien Bridge circuits, for simplicity and ease of understanding. The first attempt, Fig. 4, used the same type light bulbs as do many of the tube type oscillators, and also used capacitive tuning.

The circuit of Fig. 4 used one of the relatively new insulated gate FET's, the RCA 3N98. With a maximum design-capacitance in each section of the variable capacitor of 500 pF, at the minimum operating frequency,

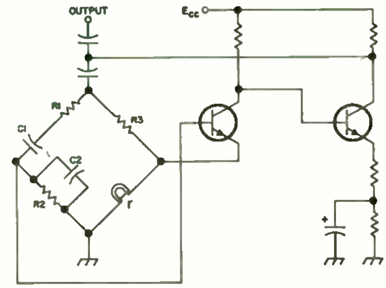


Fig. 3. Theoretical transistor version of the Wien Bridge shown in Fig. 1. Unfortunately, this simple adaptation isn't satisfactory because the low input impedance of the first transistor appears in parallel with R2 and loads it too much.

very high resistances (many megohms) were required for R_1 and R_2. At such a high impedance level, the circuit readily picked up 60-Hz ripple, and it was quite essential that it be enclosed in a shielded cabinent.

The bridge-sensing amplifier was the only FET in the circuit, since this was the only place where one was *needed*. A conventional bipolar voltage amplifier Q_2 and a complimentary emitter-follower completed the oscillator. The emitter-follower was used to provide a low output impedance. The circuit was powered by a separate +28 volt regulated supply.

Since the main frame of the dual variable capacitor was the common terminal, which was connected to the gate of Q_1, one would expect a fairly large stray capacitance to ground in shunt with C_2. This had to be equalized by a trimmer (C_3) across C_1 if oscillation was to be maintained near minimum C settings of the dual variable Also, of course, an insulated (ceramic) shaft coupling had to be used on the variable capacitor shaft and the capacitor frame had to be supported by ceramic or high-quality plastic insulators. C_4 and C_5, in parallel with C_1 and C_2 were simply to fix the minimum tuning capacitance.

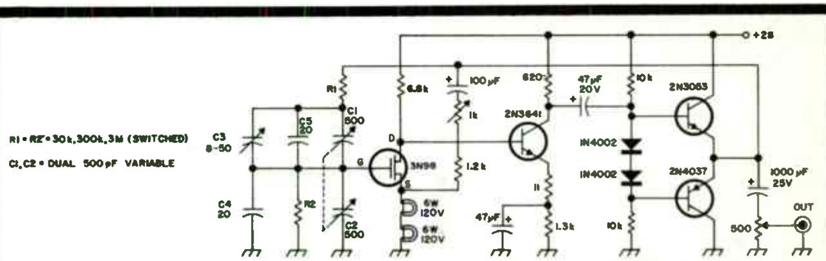


Fig. 4. First version of a moderately successful Wien Bridge oscillator. For low frequencies, R1 and R2 have to be so large that the circuit is very susceptible to noise and hum.

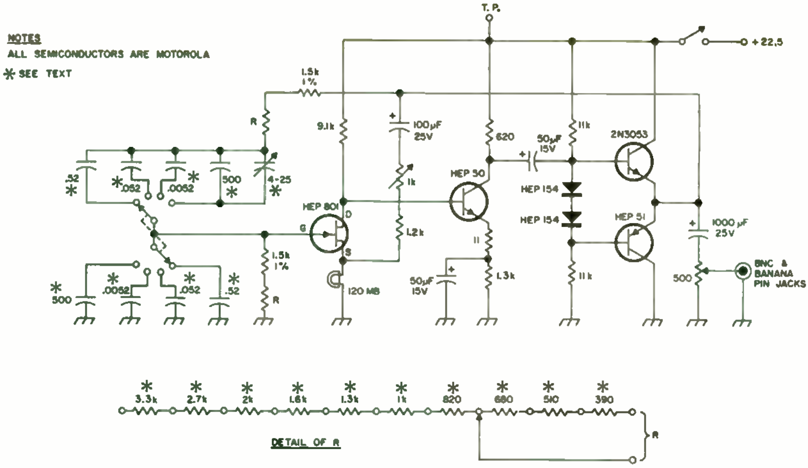
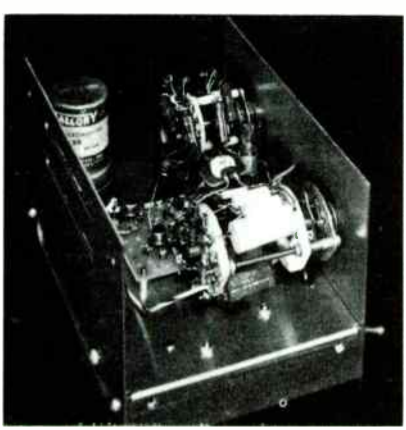


Fig. 5. The most satisfactory version of the Wien Bridge oscillator. This circuit is used in the oscillator shown in the photos.

The design worked quite well in the ranges above 100 Hz, but the lowermost range (10 Hz to 100 Hz), where the required resistance values were 30 megohms, was unreliable, as feared. At this point, capacitive tuning was abandoned in favor of a combination of capacitor and resistor switching.

The second and more successful Wien Bridge audio oscillator was built using a junction FET. By switching both R and C, bridge component values are more manageable (and available). The C values in this second version are 0.52 µF to 500 pF, and the R values are between 100 Ω and 3300 Ω. Fig. 5 shows the circuit of the oscillator; it is very similar to Fig. 4. The feedback control element used here is a Sylvania 120 MB lamp for which a typical E-I curve is shown in Fig. 6. Note that this lamp allows us to use a single bulb to operate at a source resistance of about 600 Ω. Also, the Sylvania 120 MB is physically smaller than most 120 V bulbs and fits a small bayonet pilot lamp socket, like that for a #47 or NE51. The lamp is available from Allied Radio for $0.46. The oscillator is constructed in a LMB-WIA cabinet, as shown in the photos.

The capacitors are switched only each decade, and the resistors are switched in ten increments between decades. The seemingly-nonsensical increments of frequency were chosen to give points that are approximately evenly-spaced on semilog graph paper—the type of paper usually used when plotting the



Interior of W6GXN's audio oscillator.

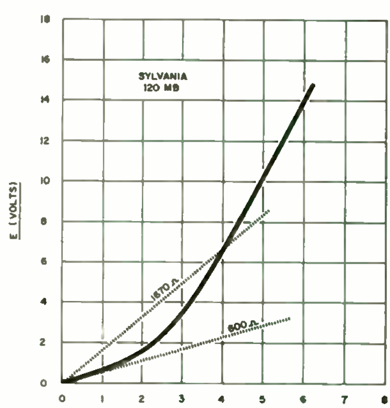


Fig. 6. E-I plot of the Sylvanis 120-MB pilot lamp.

frequency response of an audio amplifier. The four pairs of capacitors were "built-up" starting with 0.47 μF, 0.047 μF, 0.0047 μF, and 470 pF capacitors, by adding small capacitors in parallel; a bridge was used. The resistors were all 1% tolerance types from a local surplus emporium; Their marked values were trusted.

As in the first oscillator, a small trimmer capacitor was placed in parallel with the C in the series arm of the bridge to make up for stray capacitance to ground (and the input capacitance of the FET.) This trimmer was not necessary except on the high range, where 500 pF capacitors were used.

A burgess U15, 22½ volt, battery was used to power the oscillator. It is mounted under the chassis in an Austin #113 battery clip. In this mounting configuration, the battery cannot damage the circuit board if it leaks. A test point is provided on the rear of the cabinet to test the battery voltage *under load.*

A quick check at 1000 Hz revealed that second harmonic content of the waveform was 48 dB below the fundamental. Higher harmonic content was greater than 50 dB down, with the even harmonics being the strongest. The output amplitude was within 1 dB across the entire frequency range.

The author wishes to thank Gene Howell, WB6JOV, for the photographs of the audio oscillator.

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CHAPTER 40

UHF Dipmeter

Paul Franson, WAICCH, had a simple UHF dip meter. It worked well, but was difficult to package satisfactorily. I did some playing around with the circuit and found that it wasn't hard to modify it slightly, put it on an etched circuit board, and make a nice looking, convenient set of VHF-UHF dippers. The etched circuit board construction makes the dippers easy to duplicate, and the flat, U-shaped inductors are convenient to fit into tight places that conventional coils can't reach.

For a discussion of the circuit, see Fig. 1, the oscillator should be described as a grounded base rather than as grounded emitter oscillator.

I made three models of the dipper. They cover 130-175 MHz, 175-250 MHz, and 250-480 MHz. The first two models use the same size inductor, with the 130-175 MHz model using a larger capacitor for tuning with a ceramic trimmer across it. This trimmer is not shown in the schematic; its value is 7-45 pF and it should be adjusted to cover the proper range.

The 250-480 MHz dipper uses a smaller inductor than the others. It also has a copper jumper (shown in the layout) that the lower frequency dippers don't have.

Each dipper is complete (including the battery) except for the meter. The meters were omitted to save space and money, but can be included if you wish to use a slightly larger case.

Each dipper uses one RCA 2N3478 NPN silicon transistor. These transistors cost only $1.90 apiece, but it's likely that other transistors that are even cheaper could be used. The 2N3478 has odd basing—the only reference is the short case-shield lead—so don't shorten any leads until you're sure that you can keep track of the connections.

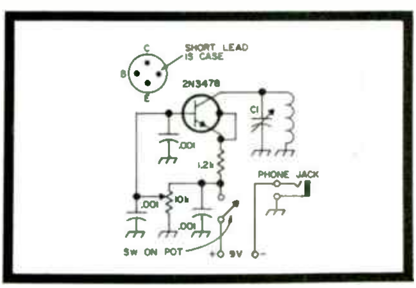


Fig. 1. The etched circuit dipper is very simple. C1 is Johnson 160-104 (9 pF) for the two higher frequency dippers, and 160-107 (14 pF) for the 130-180 MHz model. There is a trimmer across C1 in the 130-180 MHz model; see text.

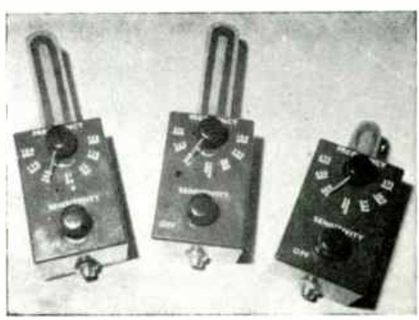
The copper side of the board for the dipper is shown in Fig. 2 with the component side in Fig. 3.

Use a glass based board. Paper or bakelite board probably wouldn't be satisfactory. Trim the board to the proper shape with a nibbler. The inductor should be coated with coil dope to keep from shorting it when you use it.

To mount the boards, you'll have to cut a thin slot near the edge of the Minibox used as a case. One way to do it is to drill a number of holes of the proper size in a row, then use a file to finish the slot. You'll have to bend that side of the Minibox out to get the board in. It's held in place by an extra set of nuts on the shafts of the potentiometer and the tuning capacitor. Be sure to trim the leads projecting from the copper side of the board so that they won't touch the metal of the case. The battery is held in place with a simple clip made from scrap metal.

The dipper is very easy to use. But before we get to that, let's check it out and calibrate it.

Plug a 2 to 5 mA meter into the meter jack. You can use a more sensitive meter if you shunt it with a resistor that gives the proper scale. Put the resistor across the meter jack terminals in the dipper if you use the meter for other things.



The three dippers shown here cover 130-480 MHz.

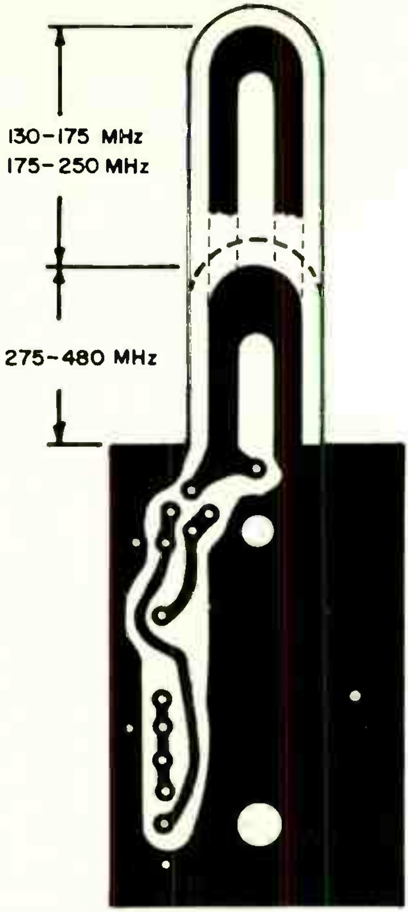


Fig. 2. The copper side of the etched circuit board used in the dipper. This layout is full size. Use board suitable for these frequencies: fiber glass or Teflon.

Turn on the dipper by twisting the potentiometer knob clockwise until it clicks. The meter should show very low current. As you turn the pot, the current should suddenly jump to about 1 mA. That means that the transistor is oscillating. If you touch the coil, the meter reading should drop and the dipper may stop oscillating completely. Now tune the capacitor through its range. There should be a little variation in current, but not too much.

Now you're ready to calibrate the dipper. The easiest method is a sensitive wave meter that covers the range, but it's quite easy to do the job with a TV set. A TV set covers 176 to 216 MHz (channels 7-13) for the low calibration. Then the second harmonics of the dipper tuning 235 to 445 MHz can be received on a UHF TV set (470-890 MHz). If you have a two meter receiver, that gives you another maker at 146 MHz. You can put on the panel markings with Ami-Tron or Datak rub-on lettering.

The dipper should be complete now, and ready for use. Bring the dipper near a resonant circuit in the dipper's range and tune the frequency control. You should get a prominent dip in current when both circuits are tuned to the same frequency. The amount of dip depends on the setting of the pot in the dipper, the distance from the tuned circuit, the Q of the circuit, and the type of coupling. In many cases it's easiest to leave the dipper stationary and tune the other circuit.

The dipper can also be used for monitoring AM transmitters by plugging a set of headphones in the meter jack and adjusting the tuning and pot. You can also use the dipper for determining the frequency of another oscillator. Simply tune the oscillating dipper with headphones plugged in until you hear a slight

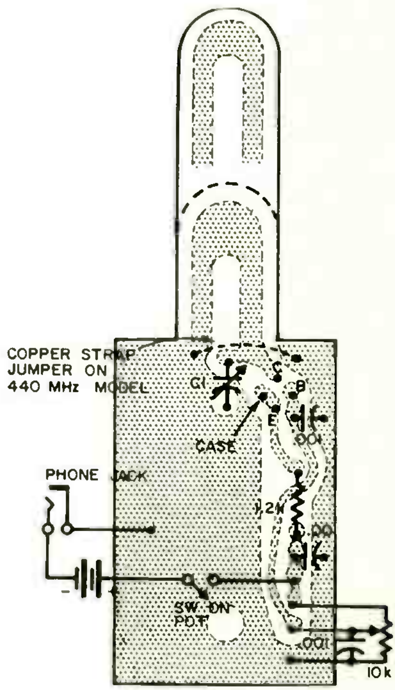
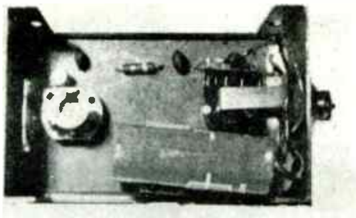


Fig. 3. Component side of the dippers. There is a 7-45 pF ceramic trimmer across C1 in the lowest frequency model. See the text.



Here's the inside of one of the dippers.

click. You probably won't be able to get a zero beat at these frequencies.

Be careful that you don't use the dipper around an energized transmitter of more than a few watts output or the dipper may be damaged.

These dippers are simple, inexpensive and non-critical to build. After you've build them, you'll wonder how you ever tried to build UHF equipment without a good dipper.

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CHAPTER 41

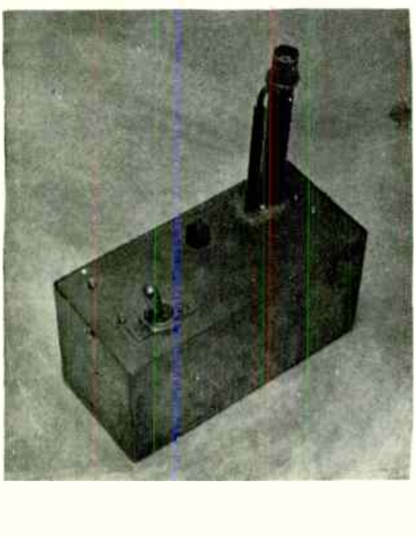
Solid-State Beacon Source

Often during the alignment of receivers and the general check out of equipment and antennas, a signal source is required to make sure you've located yourself in the exact part of the spectrum you intended. The beacon source described here is designed to fulfill that requirement with a minimum of trouble and a maximum of usefulness. To begin with, it should be readily portable with a self-contained power supply, small, operate on a fixed frequency, have an adjustable output level and be inexpensive and easy to build.

Satisfying these requirements posed some rather interesting problems and resulted in several hours of construction time that has proven to be well spent.

Circuit

The crystal oscillator circuit uses a Philco 2N502 transistor in an overtone circuit with a 72 mc crystal. A McCoy miniature crystal unit was used to conserve space, but there is more than enough room to use a standard size unit. The choice of frequency gives a second harmonic marker at 144 mc, a sixth harmonic marker at 432 mc and a multiplied final output at 1296 mc. If you are more interested in 6 meters than 2 meters, use a 54 mc crystal—this gives you output on the fundamental 432 mc and 1296 mc. The 2N502 transistor was used because it was available here. It is a little more expensive than some other types that would perform equally well. The circuit is straightforward and subsequent substitution of other types has yielded essentially the same performance.

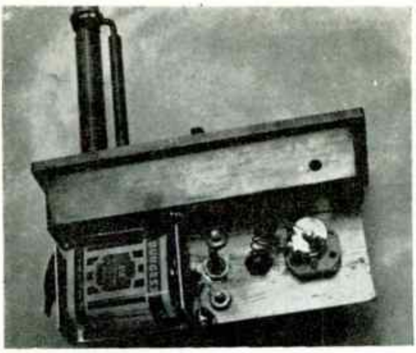


The oscillator circuit supply voltage is regulated with a small zener diode through a dropping resistor to provide a bit more in the way of frequency stability. The output of the oscillator is fed directly to a diode multiplier and tuned by a $\frac{\lambda}{2}$ line tank tuned to 1296 mc. Even though the multiplication factor is 18, there is sufficient output at 1296 mc to completely quiet a 1296 mc diode mixer type converter and a 432 mc receiver. The multiplier diode, an OK 733, is one that was salvaged from a standard coil UHF TV plug-in strip that was picked up for next to nothing in a local surplus store. The DR 404 that is commercially available for about a dollar works just as well.

The entire unit is powered by a 15 volt dry battery (Burgess U10) and has a relatively long life in this service provided you remember to turn it off when you are through.

The only controls and adjustments that appear on the outside are the 1296 mc tank tuning, the on/off switch and the output jack and attenuator. The attenuator is the device that makes the unit extremely flexible and useful. Therefore, a bit of careful attention to its construction will really pay off in the long run. Fundamentally, it is a simple coaxial attenuator using loop pick-up. It is designed for a characteristic line impedance of 50 ohms and the appropriate values of telescopic tubing were chosen to give a snug fit consistent with the proper diameter required for the center conductor.

An essential part of the attentuator is the guide that maintains the probe coupling in a fixed plane. The guide rod and sleeve technique used here is satisfactory and other



methods are left to the ingenuity of the constructor. The pick-up sleeve is made long enough so that when it hits the guide stop, the pick-up loop is almost in contact with the center conductor of the line tank and parallel to it.

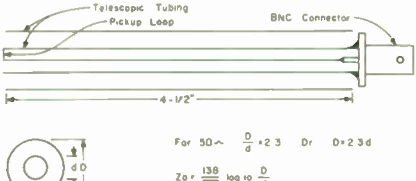
You can use whatever size telescopic tubing that is available or pick some up at the local hobby shop provided care is taken to keep the ratios of the inside diameter of the outer conductor to the outside diameter of the inner conductor as close to 2.3 as possible. (See Fig. 1.) If other values of impedance are desired, they can be calculated from the formula given in Fig. 1.

If good laboratory equipment is available, the movable part of the attenuator may be calibrated in db as a further refinement, but it is not necessary for completely satisfactory use.

Construction

The entire unit is housed in a brass box 2″ wide by 2¼″ high by 4 ⅜″ long. This can be readily bent up from .03 sheet stock in a moderate sized vise. There is no strict requirement on the type of material used as long as reasonable care is taken to shield the energy that is generated. When the box is formed to proper shape, solder the corners carefully and a shielded strong unit results.

The top plate mounts all the hardware and was made from a piece of sheet brass 2″



COAXIAL ATTENUATOR

FIGURE 1

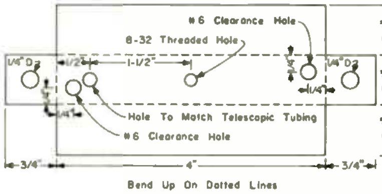


FIGURE 2

wide by 4 3/8″ long by 1/8″ thick. It could just as well have been made from other material since the only requirement is that it be rigid enough to support the components and attenuator. The only purpose for using such relatively thick stock for the top plate was to provide enough material to drill and tap this plate to anchor the shield can. Thinner stock could readily be used if ¼″ lips were bent on the four sides. See Fig. 4.

The entire unit is designed to be hand held and operable leaving the other hand free for adjustment of other equipment.

The next piece to fabricate is the $\frac{\lambda}{2}$ line tank for 1296 mc. This was also bent from .030 sheet brass stock so that an open trough measuring ¾″ by ¾″ by 4″ resulted. The corners are soldered for rigidity and holes drilled before bending as indicated in Fig. 2.

Tuning for the tank is accomplished by using an 8-32 bolt with a small disc 3/8″ in diameter soldered on the end that comes in close proximity to the center conductor. The top plate is threaded and tension is provided by a nylon nut at the top. Be very careful when soldering the disc on the end of the bolt to make sure the nylon nut is flush up against the bolt head and do not disturb the nut until the unit has completely cooled.

Next insert the center conductor. This is cut from ¼″ round tubing and just long enough to pass beyond the end plates of the tank. The center conductor is soldered in place and the additional material left sticking out is carefully filed off to make a neat flat surface.

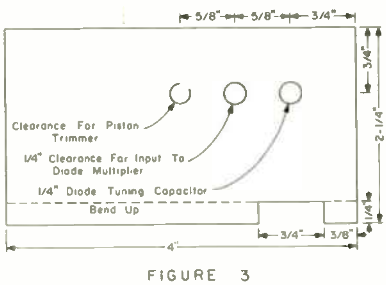


FIGURE 3

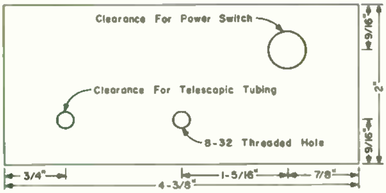


FIGURE 4

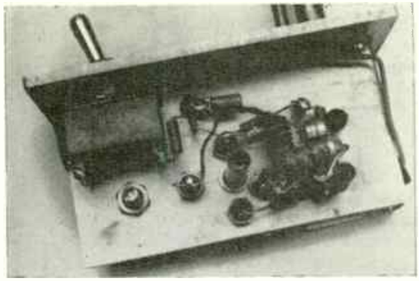
A small L bracket was used to mount the battery and the tuning capacitors and provide shielding between the circuits. The dimensions for this are shown in Fig. 3.

After all the pieces are fabricated, mount the 1296 tank to the top plate using 6-32 screws. Use short 6-32 binding head screws with the head side in the tank. Next mount the outer conductor of the attenuator from the top of the plate so that it is flush with the inside of the tank. Solder this in place on the *top* side. Next mount the L bracket with battery clip and tuning capacitors, make sure that the on/off switch has proper hole clearance and bolt it down.

The unit is now ready for wiring. To aid in the placement of parts miniature tie points were used in conjunction with stud mounted by-pass capacitors to support all components.

The original unit constructed used a ferrite slug coil form. This was subsequently replaced by air wound coil when it was determined that a significant amount of frequency instability resulted from its use.

Placement of parts is not critical with the single exception that the oscillator and multiplier circuits are shielded from each other. Care should be taken to anchor all compo-



nents securely and ground the crystal case. The schematic diagram is shown in Fig. 5.

Adjustment

After the oscillator circuit is wired completely check all connections before applying supply voltage. Make sure the correct battery polarity has been observed. Apply voltage and with the unit lightly coupled to a receiver on 54, 72 or 144 mc, depending on your choice of crystals, tune C1 for maximum output. When this point is found, turn the supply voltage off and then on again to see if the unit is still oscillating. If not, retune C1 slightly higher in frequency until the oscillator starts up instantly on application of supply voltage.

Next solder the multiplier diode from C2 to the center conductor of the 1296 mc trough line one inch from the end plate opposite from the side of the output pick up. Couple the output of the beacon to a 1296 mc receiver using maximum coupling to the line tank and with C2 set for approximately half capacity, adjust C3 carefully for maximum output and tighten the nylon nut. Next adjust C2 for maximum output. This adjust-

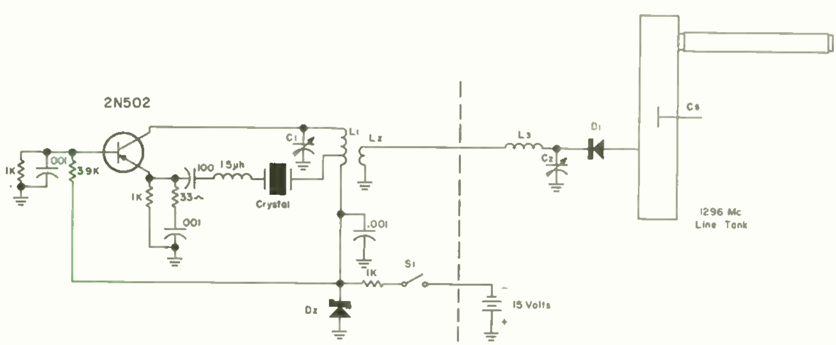


FIGURE 5

D1—CK733 or DR 404
D2—Hoffman IN 1314 Zener
C1—1-7 Piston Trimmer
C2—Johnson 1-7 mmfd miniture variable
C3—Disc

L1—6T #20, 1/4" dia, 3/8" long tapped 1T from cold end
L2—2T #20 1/4" dia, at cold end of L1
L3—9T #20 1/4" dia close spaced

ment will be quite broad and once set does not require additional setting. Now vary the coupling loop from the oscillator and readjust C1 for maximum output until you find the optimum coupling point. Place the shield can in place and bolt it down. You now have a completed beacon source.

After many months of use here in the shack, it was found that ready access to C1 was desirable to make minor adjustments in the exact output frequency. To facilitate this a small hole was drilled in the case opposite the C1 tuning screw. A small snap-in hole plug was used to cover the hole when the adjustment was complete.

The unit here has been used for more than one year and a half on the same battery and is still going strong. It gets more than its fair share of work, particularly when tuning up 1296 antennas. It can often be seen on top of a step ladder in the back yard with a 1296 mc ground plane plugged directly in the output jack. . . . W6GGV

CHAPTER 42

Meter Amplifiers

Sensitive meter movements are the backbone of all accurate electronic measurements but their expense limits their use to all but the most costly instruments. If you have priced a 0-25 or 0-50 microampere meter in the past few years you know why inexpensive meter amplifiers are useful. At one time sensitive meters could be obtained quite inexpensively on the surplus market, but nowadays even surplus meter prices are prohibitive.

With transistors it is quite easy to build a meter amplifier that is inexpensive, portable, and extremely sensitive. Depending on the planned usage, several types of amplifiers are available which are suitable for this purpose. These circuits vary all the way from the most simple to the quite exotic. Of course, the simple circuits are limited; the more sophisticated circuits allow for such things as gain control, linearity and meter zeroing. In most cases drift in a properly designed unit will be negligible.

Most amateurs are familiar with the simple meter amplifier circuit shown in Fig. 1. This arrangement has been used extensively in transistorized "grid-dip" meters and other units where only relative readings are of interest. In this circuit the only components are a transistor and its associated battery; you can't get much simpler than that! Admittedly, this circuit is very limited and suffers from several serious disadvantages. Probably foremost among these limitations is that the gain is dictated completely by the transistor. Furthermore, if the transistor is chosen without a little forethought, meter drift may become very serious. For this reason, a silicon transistor with extremely small reverse leakage should be used. Linearity may also be a problem in some applications, but where only relative readings are of interest, this may usually be neglected.

A more generally useful transistorized meter amplifier is shown schematically in Fig. 2. This bridge type circuit overcomes the major disadvantages of the simple circuit by the addition of three resistors. The gain control and the transistor form one leg of the bridge while the zero control forms the other. Since the circuit is essentially a voltage amplifier, gain is easily controlled as is linearity. In the simple circuit previously described, the gain was dictated by the current gain of the transistor, but in the bridge arrangement almost any desired amount of gain may be obtained by the proper choice of resistances.

In the bridge circuit a small current at the transistor base is amplified by the transistor to a larger current in the collector; the increase in collector current reduces the voltage at the collector and unbalances the bridge. The amount of unbalance is indicated on the milliammeter. Since the voltage gain of a transistor circuit is proportional to the collector load resistance, this circuit may be adjusted to nearly any specified gain. Because of its shunting effect, the optimum value of zero adjusting resistance is determined by circuit gain.

It is not absolutely necessary to use the 2N3392 transistor shown in the diagram, but other transistor types will likely require different resistance values. This particular transistor was chosen on the basis of cost (69¢) and low leakage current (0.1 microampere at room temperature). Silicon transistors are recommended for this circuit because they are less susceptible than germanium to the effects of temperature.

As shown in Fig. 2, the gain of this circuit may be varied to almost any desired amount. In the most sensitive arrangement (gain of 75-100) a one milliampere meter movement may be used to measure a full scale current of ten microamperes. This is sensitive enough for nearly any application; if used in a voltmeter, this corresponds to 100,000 ohms per volt.

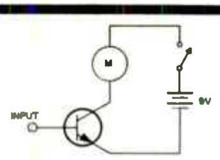
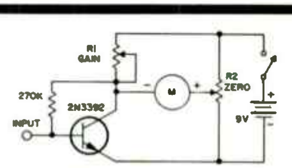


Fig. 1. Simple meter amplifier.



GAIN	R1	R2
5-10	1K	50K
10-40	5K	20K
40-80	5K	10K
75-100	10K	5K

Fig. 2. Bridge type meter amplifier.

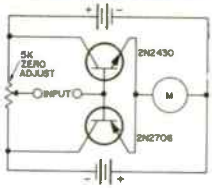


Fig. 3. Null amplifier.

One small word of caution should be extended concerning the choice of a meter movement. A good name brand meter should be used if at all possible. Most of the cases of nonlinearity associated with this circuit have been traced to inexpensive meters. The low cost Japanese meters are particularly bad in this respect; scale nonlinearities on the order of 20% are not unusual, at least on the units I have tested. Of course, it's like anything else, you pay your money and take your choice.

Adjustment of this circuit is simplicity itself. With no input, turn the circuit on and adjust the zero gain control so that the needle is off the pin. Use a little care during this initial operation because the circuit will probably be extremely unbalanced. The use of a small shunting resistance or "meter protector" is recommended in the initial phases of zeroing to prevent damage to the instrument. After the initial zero is obtained, each time the gain potentiometer is changed it is necessary to rezero the circuit.

The meter is calibrated by applying a known current to the base of the transistor and alternately adjusting the gain and zero controls until the desired full scale reading is obtained. For instance, 10 microampere amplifier could be calibrated with a 1 megohm resistor in series with the transistor base and a 10 volt battery. This arrangement would provide 10 microamperes full scale. To check linearity, a 2 megohm resistor (5 microamperes) should provide a half-scale reading.

Incidentally, one advantage of the bridge type meter amplifier is that zero current may be set at any point on the meter. In other words, a regualr meter could be used as a zero center instrument. If the 10 microampere circuit were adjusted to zero center, a 5-0-5 microampere device would result; just try pricing an instrument with that sensitivity! Use of very large currents with a zero center may not provide useable results however, because the circuit does not have much negative range.

A simple meter amplifier designed for use with zero center meters and useful in null networks is illustrated in Fig. 3. It is particularly useful in instruments where high-accuracy measurements are necessary but a relatively rugged inexpensive meter movement is required.

In this circuit a pair of complimentary transistors are operated in push-pull. The current gain of the amplifier with the transistors shown varies between 50 and 100, the nominal range of the transistors shown in the circuit. If more gain is desired, the transistors may be cascaded. Due to the complimentary nature of the circuit, it is extremely stable with temperature; drift of between 0.2 and 0.1 microamperes has been experienced over a three hour period.

The resistance of the potentiometer is not at all critical; it is used simply to center the needle of the meter. Almost any junk-box potentiometer will work.

Although the use of matched transistors is not required, the unit will be considerably more stable if matched devices are used. The Amperex 2N2430 and 2N2706 may be obtained as a matched pair under the part number 2N2707. The matched pair is only about 10¢ more than the individual transistors so is well worth it in terms of the excellent results obtained. The main disadvantage of this circuit is that both negative and positive batteries are required, but considering the simplicity of the circuit, this is not considered to be too serious.

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CHAPTER 43

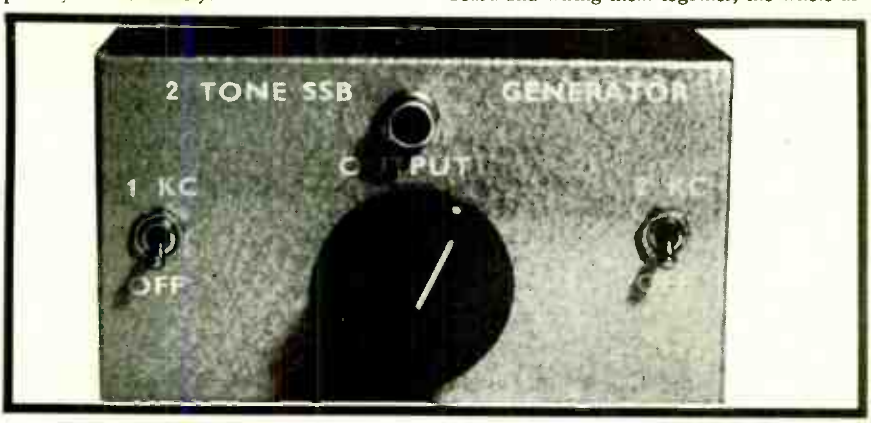
Two-Tone Test Generator

As most single sideband enthusiasts know, the simplest and easiest way to properly adjust a linear rf amplifier is with a two-tone signal generator. With an oscilloscope and two audio frequencies about 1000 cycles apart it becomes a relatively simple task to adjust loading, drive and grid bias for maximum linearity.

Two-tone audio generators that have been described in the past have used bulky inductors and capacitors in an LC circuit, but the straight-forward phase-shift circuits used in this unit maintain good stability and low distortion with two 50¢ transistors and the simplest of bias arrangements. The secret to this phase-shift oscillator's stability is the low value of collector load resistance used; this resistance effectively swamps out any changes in forward current gain that may exist between transistors of the same type. This means that the amplitude of the output signal will remain relatively constant regardless of the transistor that is used. Actually, the only requirement for the transistors is that they have high beta (h_{fe}). Many different types of PNP transistors have been tried in this circuit; some work and some don't, but both 2N2953's and 2N2613's (50¢ varieties) have been used successfully. Some of the older types—such as the 2N107—will not oscillate in this circuit because they don't have quite enough gain. This circuit is not limited to PNP types; silicon NPN transistors like the 2N3391 work quite well if you reverse the polarity of the battery.

With a new nine volt battery, the audio output from this unit is adjustable from about five volts peak-to-peak down to several millivolts. This is more than sufficient for nearly any application requiring this type of a signal. Since the total current drain for this unit is only three milliamps, even with both oscillators going, the life of the battery is just about its normal shelf life. The output decreases accordingly, but this circuit will continue to oscillate with as little as 1½ volts applied. This means that this unit will provide a useable output even when the nine volt battery is four or five years old!

Construction and layout of the circuit is anything but critical. In the unit in the photographs, the active circuitry was laid out on a piece of Vector board (32AA18), 2 inches wide and 3½ inches long. This is quite a bit larger than necessary, but in the author's model, junk box parts were used. If one-quarter watt resistors and miniature capacitors are used, the total size could be easily cut in half. However, miniaturization can go *too* far. If an instrument of this type is made as small as physically possible, there's no room on the front panel for the switches, knob, and output jack! The 2¼ X 2¼ X 4 inch LMB type J-875 Jiffy-Box seems to be a good compromise. It is small enough not to be obtrusive, but large enough so you can operate the controls. After laying the components out on the punched board and wiring them together, the whole as-



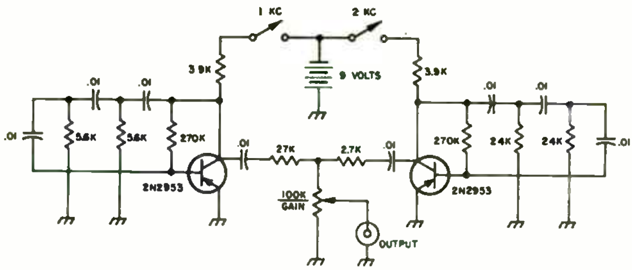


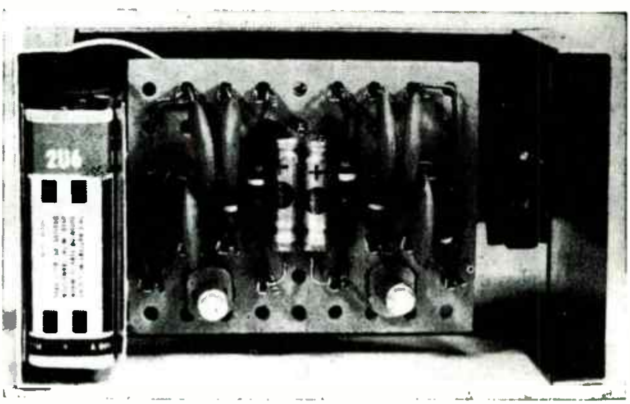
Fig. 1. Schematic of the two tone test generator.

sembly is glued to the back-end of the output potentiometer with epoxy glue. The end result is an integral, easily installed package. Granted, it's pretty tough to replace any of the components when everything is glued together, but after all, transistors last nearly forever and the other components in the circuit are operated so far below their ratings that they should last nearly as long. The battery clip is formed from thin Reynold's aluminum sheet and epoxied to the side of the Jiffy-box as shown in the photograph. The toggle switches are miniature Japanese types that are available for 29¢ apiece.

As was previously mentioned, the two-tone audio generator is especially useful when adjusting linear rf amplifiers for maximum linearity and minimum intermodulation (IM) distortion. Since the procedure for making the necessary adjustments is quite simple and has expressed as follows:

$$I_{pk} = I_{dc}\,(1.57 - 0.363\, I_o)$$

Where: I_{pk} = Peak plate current
I_{dc} = Plate current with two-tone signal
I_o = Zero signal plate current



Inside view of the two tone test generator. Note the position of the battery bracket on the left side.

CHAPTER 44

Scope for RTTY

Teleprinter circuits are admirably suited for transistorization, as indicated by the number of recent transistor TU designs. These TU's have had to use meters or tuning eyes for tuning indicators due to the difficulty of designing scope deflection amplifiers.

This article will describe a simple (nearly) all transistor X-Y oscilloscope, designed for RTTY tuning, but suitable for general X-Y plotting and frequency comparison work. It incorporates a unique blanking circuit to dim any undeflected spot.

Cathode ray tubes require such large deflection voltages that transistor drivers have been uneconomical and unpractical. Recently several 300 volt silicon transistors have become available to reverse this situation. They are ideal for scope deflection amplifiers. The RCA 40264 ($1.21) uses a small diamond case, with non-standard leads. The Industro Transistor TRS-301-LC uses a standard TO-5 case and was used here. Either will work.

Amplifier

One of the photos shows an excellent 300 volt p-p sine wave at just below the amplifier's clipping level. This is enough voltage to deflect almost any 2, 3, or 5 inch CRT operated at moderate accelerating voltages.

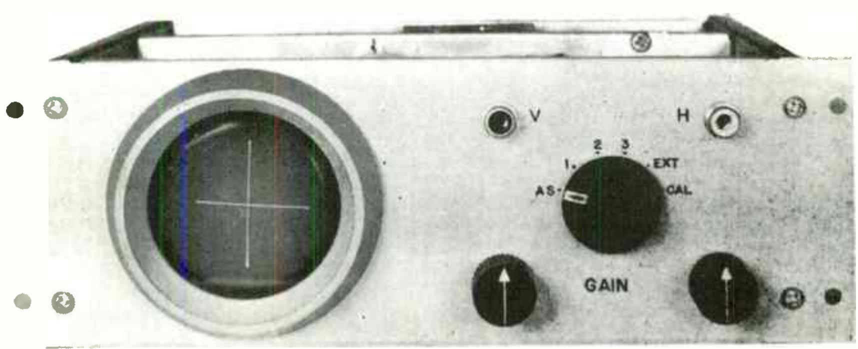
The amplifiers are quite conventional except for the 300 volt B+. My particular scope obtained full deflection with inputs of only 0.3 volts (RMS) vertical and 0.5 volts horizontal. The input impedance varies with the gain setting from 20 k to 100 k. Higher impedances may be obtained by adding series input resistance. The bandwidth is from 50 Hz to over 20 kHz. The lower response can be improved by increasing the values of the 0.1 μf input condensers. Heat radiating fins on the deflection transistors are desirable but not essential.

Blanker

A particular problem with RTTY scopes is that often an undeflected spot occurs, which burns the screen. With the blanker, the undeflected spot is adjusted for very low intensity with the "spot int." control. Any input signal drives Q_3 into conduction, shunting the intensity control with the second, "trace int." control. This is adjusted for proper operating brilliance. This circuit completely prevents screen burning.

Power supply

The supply is unusual in that only two diodes are needed to supply both 300 volts for the amplifiers and 850 volts in a multiplying circuit for the CRT. Any small 500 volt transformer will work. If the B+ is much over 300 volts, increase the 100 ohm surge resistor. If you need more voltage for a 3 or 5 inch tube, replace the 0.5 M resistor with another 1800 volt diode, and raise the voltage rating of the two 0.5 μf filters to 1000.



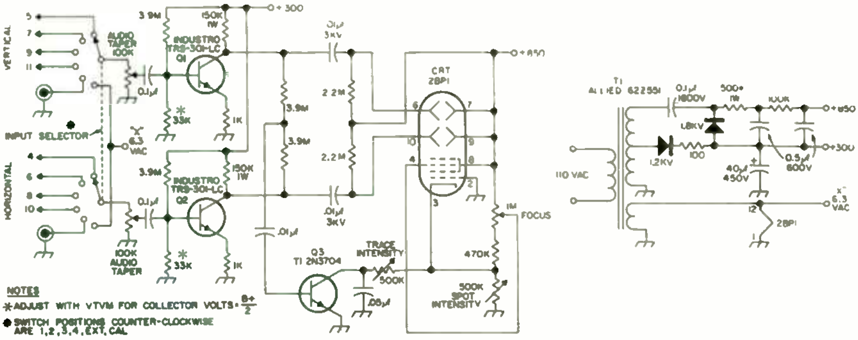
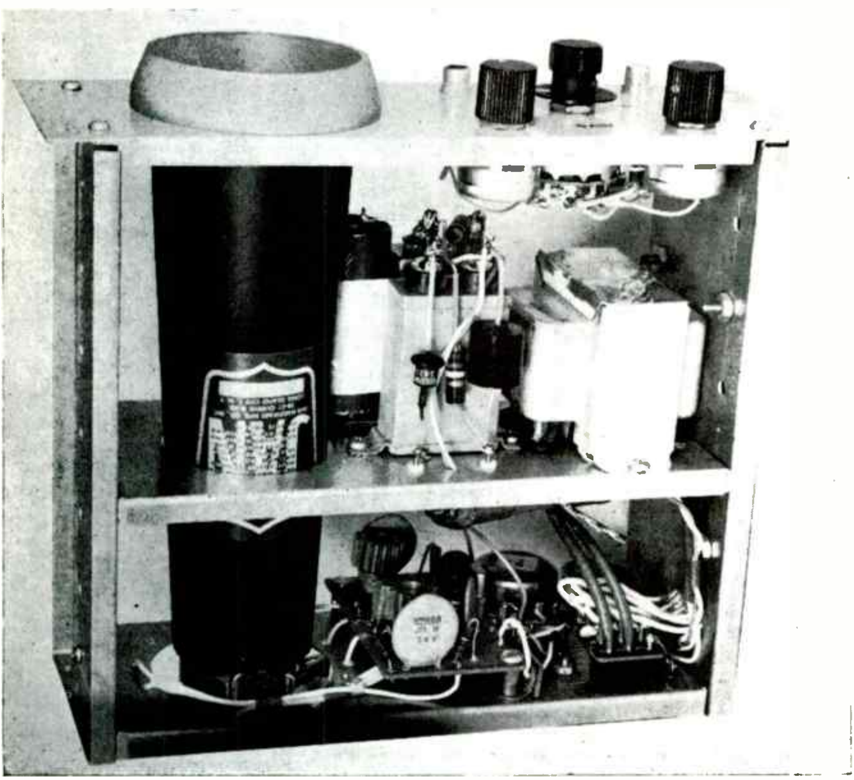


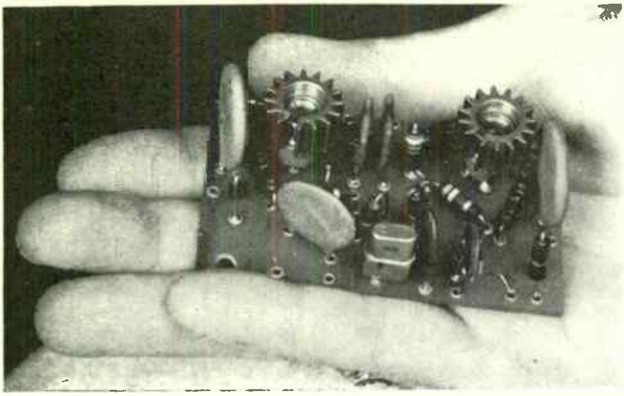
Fig. 1. Schematic of K8ERV's RTTY scope using transistor deflection amplifiers.

Adjustments

Remove Q_3 from its socket then plug in the scope. Check the B+ for about 300 volts, using a Variac if necessary to keep it down. Connect a VTVM or 20,000 OPV meter on the 500 volt range to the collector of Q_1. Select or pad the base to ground resistor to obtain a voltage of ½ B+. Do the same with Q_2. Now adjust the power supply resistor for



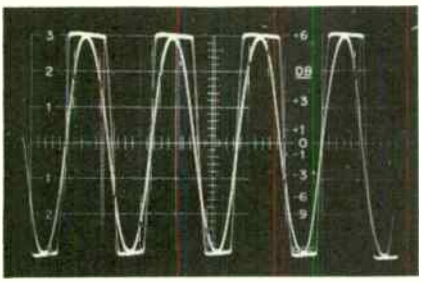
Internal parts layout showing the location of the amplifier board, controls, power supply and CRT with shield.



Board containing the vertical and horizontal amplifiers and the blanking circuit.

a 300 volt B+ with normal line voltages.

Now adjust the focus and spot intensity controls for a small dim spot. Unplug the scope, wait ten seconds, plug in Q_3, and plug in the scope. The spot should be the same as before. If it is brighter, either the transistor or its base coupling condenser is leaky, or the transistor has too low a breakdown voltage.



300 volt output from one of the deflection amplifiers.

Obtain a full deflection by switching the input selector to "Cal." and adjust the "trace int." control for a bright sharp pattern. Removing the signal will cause the remaining spot to dim to the intensity set by the "spot int." control.

This scope was built as a separate unit with input selector so that it could be used to monitor several different TU's. It is small enough to be built into compact TU's or other equipment. If you aren't yet using transistors, this is an excellent first project, with practically nothing to go wrong?

. . . K8ERV

CHAPTER 45

Mobile Monitor Scope

DURING the course of experimenting with transistorized mobile modulation systems, the desire for a means of monitoring performance of the rig on the road led to the adoption of the miniature scope shown here. The 1" screen is mounted out of the way on the dash and can be seen without turning your head from the road. It can be used for either AM or SB. The power requirements are so small the scope can be run from flashlight batteries.

The scope tube is an RCA 913. The high voltage requirement is supplied by a small transistor oscillator putting out 300 to 400 volts at approximately 100 micro amps.

The 913 filament is turned on by the transmitter on-off switch. To prevent a bright spot from burning the screen during stand-by the power supply is turned on with the push-to-talk system, thus the screen is dark during stand-by.

Construction

The power supply and controls are built into a 2¼ x 2¼ x 5" Bud Mini box which is mounted out of sight under the dash. Only the tube and it's socket are visible. The only important thing to remember about contruction is that the oscillator must be shielded from the other components to prevent undesirable traces. The 913 can be replaced by a 2 BP1, but it is somewhat large for dash mounting. The 2 BP1 requires a different value filament dropping resistor.

Photographed by S/Sgt. John Matheson, USAF

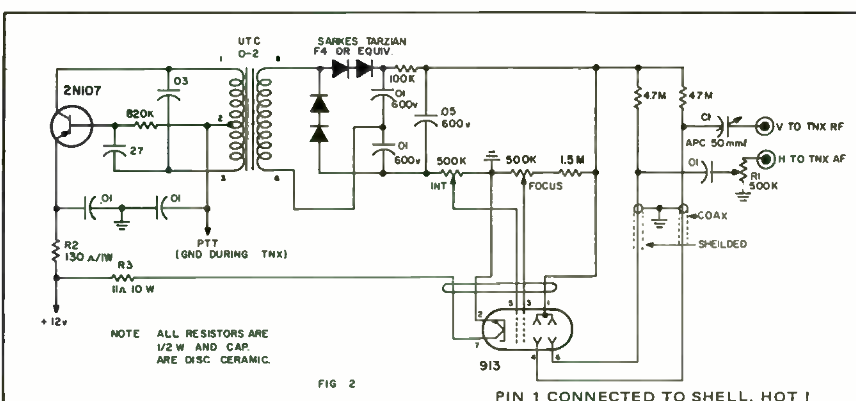


FIG 2

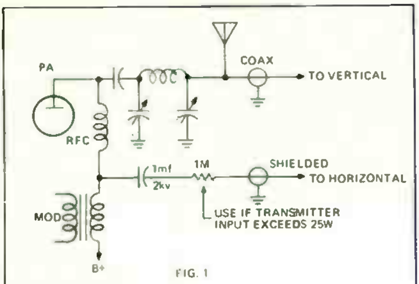


FIG. 1

The audio is brought into the scope by a shielded lead from a capacitor connected to the PA side of the modulation transformer. Rf is obtained from a tee connector on the transmitter's antenna plug. Coax cable is used for the rf leads. Connections shown in Fig. 1 are for a trapezoidal pattern.

The aluminum bracket, holding the CRT, was made from scraps around the shack and is held in place by one of the bolts in the speaker grill.

Operations

If it is desired to make the scope completely self contained for portable use only a few changes are required. Increase the chassis size to accommodate the 913 and a 6 volt battery, 4 size D cells or a 6 volt lantern battery such as RCA VS040s. Remove the filament dropping resistor R3, for the 913 and remove R2 in the power supply. The brightness, although decreased, should still be sufficient.

After tuning the transmitter, adjust C1 with an insulated screwdriver (the rotor of C1 is hot) for a narrow line that fills about ¾ of the 913 screen. This is done without modulation and indicates carrier magnitude. Next loosen the CRT clamp and rotate the tube until the line is vertical.

Now apply modulation and adjust R1 for proper horizontal size and you are in business.

. . . K9DYS

CHAPTER 46

Transistor Testers

Two simple transistor testers are shown here both having the same basic circuit. The one shown in Fig. 1 is about as simple as can be made for measuring the relative dc beta of either an NPN or PNP transistor. It was built into an aluminum box 3 x 5 x 2 inches with an old 0-200 Am meter. The latter had the internal shunt removed, giving a 0-2 Am meter with a 0 to 200 scale reading. A small half ohm resistor was shunted across it to make it read somewhere between 5 and 10 milliamperes full scale since most small transistors operate within this value of collector current.

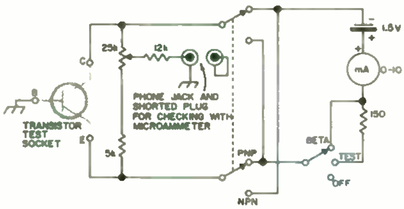
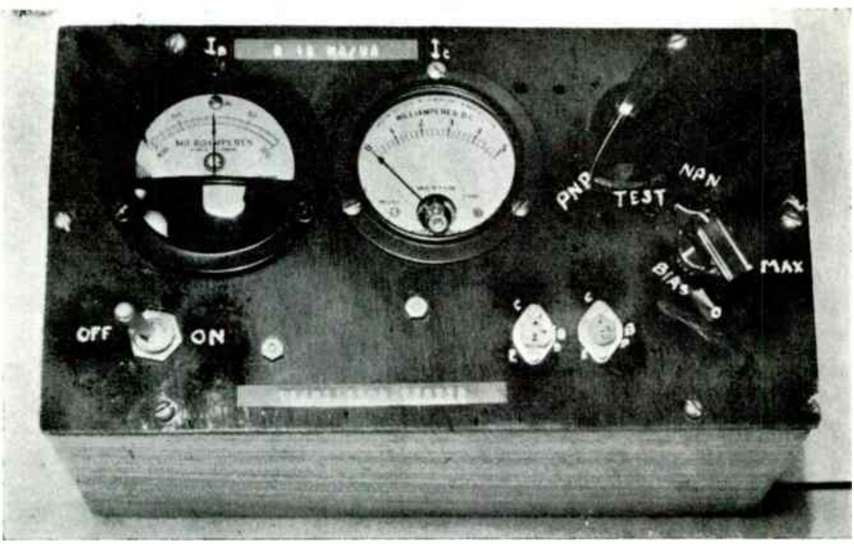


Fig. 1. The simpler of the two tronsistor checkers described by W6AJF in this orticle. The proper scole con be set by odjusting the potentiometer.

The exact reading is not important since the beta reading can be set to use the 0-200 division scale on the meter by adjusting the potentiometer in the bias circuit.

A battery and meter polarity reversing switch in Fig. 1 is a DPDT toggle switch labeled NPN and PNP. By having a "test" position on the other switch, an unknown type of transistor can be plugged in for test without damaging it or the meter. The protective resistor should be 150 ohms for a 0 to 10 Am meter, or 300 ohms for a 0 to 5 Am meter in order to keep the meter reading to within range even with a short-circuited transistor. If no reading is obtained with the NPN-PNP switch in either position, it indicates a very weak transistor or one with an open lead. Once these tests have been made, the dc beta can be read on the meter in the third position of the "test" and "off" switch.

The calibrating potentiometer can be set to read correct beta for a known type of transistor which has been measured on a more accurate transistor tester. The battery voltage affects the beta reading which means it should be checked occasionally to be sure it is near the 1.5 or 1.4 volt reading. The ordinary pen-



Here's the front ponel of the tronsistor tester shown in Fig. 2. Note the use of o zero-center meter for meosuring bose current. This ovoids the necessity of using o meter-reversing switch there.

lite sized cell should measure 1.5 volts and a single mercury battery cell should read 1.4 volts. Either type is suitable in this tester.

A more accurate type of dc beta tester is shown in **Fig. 2**. This tester was built into a larger box with two meters, one a zero center microammeter for reading the transistor base current for either NPN or PNP transistors without need of a reversing switch. The other meter, a 0 to 5 mA unit, reads the collector current for any particular value of base bias voltage and current. The milliampere reading can be set to any desired value such as 2 mA by means of the bias potentiometer knob. The reading multiplied by 1000 gives the collector current in microamperes. This value is then divided by the base current reading to give the dc beta of the transistor being tested. If the latter reading was 20 microamperes then for our example the beta is 2000/20 = 100.

Front of the simple transistor tester in Fig. 1.

Inside of the transistor tested shown in Fig. 1.

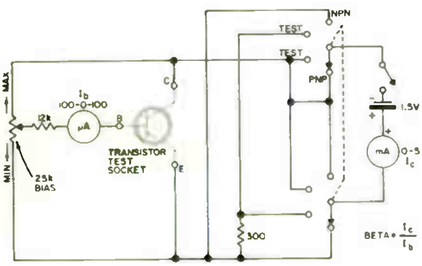
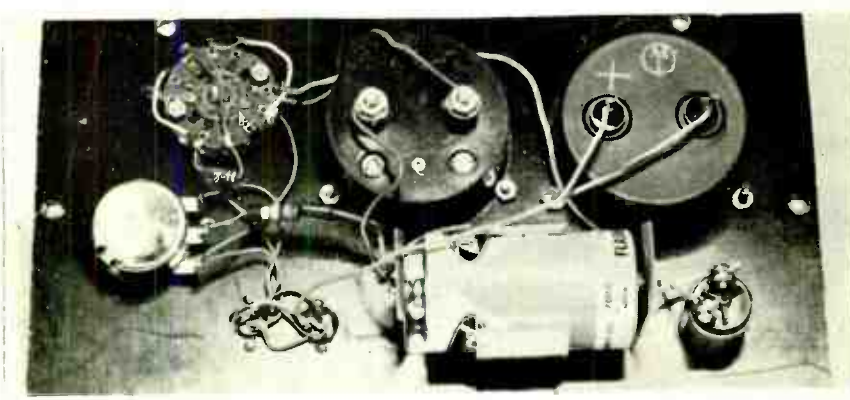


Fig. 2. The more complex of the two simple testers. You con figure the beto of the tronsistor under test more occurotely with this tester thon the one in Fig. 2 since you get o specific collector current for eoch volue of bose current you use.

The PNP-test-test-NPN switch is a DP4T wafer switch. Two test positions were used with the 300 ohm protective resistor inserted series with the meter to prevent burnout for the case of a short-circuited transistor. Another protective resistor of 12,000 ohms was connected in series with the base circuit microammeter in case of a faulty transistor. A single flashlight battery was used to power the tester.

In testing either NPN or PNP transistors the current of both meters should increase simul-



Bock view of W6AJF's tronsistor tester os shown in Fig. 2. The circuit is very simple ond construction is completely non-criticol. Either o mercury cell or o regulor floshlight bottery (shown) con be used.

taneously as the bias is increased from 0 towards maximum. If such is not the case, try the other switch position PNP instead of NPN or vice versa. If the beta reading is too much lower than transistor handbook values listed for "hfe" for a given type of transistor, it should be discarded. Higher valeus generally mean that you are in luck, as the transistor has a higher dc beta and hfe than the average units.

These testers do not measure anything except the relative efficiency as a dc device. It does not show up noisy transistors or give any indication of the operating frequency range. However, if it tests good on dc values, the transistor will probably work well in the frequency ranges listed in transistor handbooks.

. . . W6AJF

CHAPTER 47

VOM Transistor Test Adapter

Like most of you, I have always desired and half-way needed a reasonably good transistor checker. The problem has been that the ones available in the $7.00 to $10.00 price bracket are practically useless, and the actual need never seemed to warrant the investment for a decent checker. The checker described here provides a reasonably good solution to this dilemma.

If you have any kind of a junk box, this transistor checker adapter for your VOM or VTVM will cost considerably less than the ready-made cheapies, and it will give more test data. Getting down to details, you can check current gain, determine which junction is open or shorted, and check for leakage.

The circuit

The adapter tester utilizes a standard common-emitter circuit as shown in Fig. 1. Hence, devices are being checked under typical working conditions. Here is how it works. With R4 set at zero resistance, the base voltage will be equal to the voltage V_B which is 6 volts (less a tiny drop across R3). Then, depending upon if the transistor is germanium or silicon there will be from 0.3 to 0.7-volt drop across the base-emitter junction. Thus, with a good transistor in the circuit the remainder of the voltage will be dropped across R5 giving a meter reading of about 5.5 volts.

Checking transistor defects

With S2 in the test position and R4 set at zero, faults are checked as follows:

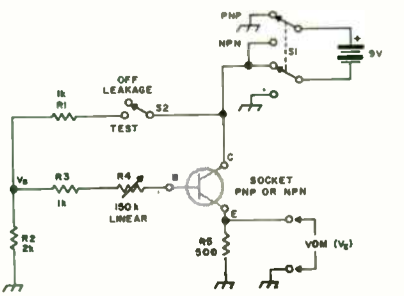


Fig. 1. Transistor tester adapter for use with a VOM.

1. A collector-emitter short will give a meter reading equal to the full battery voltage (9 volts).
2. A collector-base short will also yield a full supply voltage reading.
3. A base-emitter short will produce a 2-volt reading on the meter. There is no transistor action, hence, R_3 and R5 function as a voltage divider.
4. An open collector will also give a 2-volt reading.
5. An open base or open emitter—no reading.

All of these meter readings are fairly widely spaced for easy recognition of a particular transistor defect.

Accurate beta measurements

By now, you are probably wondering why R4, and 150 kΩ linear potentiometer, is in the tester. Well, this is used to make rather accurate beta measurements.

By performing a circuit analysis, it can be shown that the voltage V_E across R5 is a function of V_B, R3, R4, R5 and the gain of the transistor. We won't bore you with the math details, just take our word for it. Actually, the equation boils down to $\beta = \frac{R4}{2R5}$ if $V_E = 2$, and, since we have made R5 = 500 ohms, transistor current gain is equal to R4 in kilohms. Get the idea, all that is necessary to get accurate gain measurements is to calibrate R4 in steps of say 5 kΩ. Then, to measure current gain, simply adjust R4 so that the meter reads 2 volts, and read gain from the calibrated pot. For example, a 50 kΩ reading on the pot (meter set at 2 volts) is equivalent to a current gain of 50.

To check leakage, open S2. A low leakage device will give a very low or zero meter reading. Also, a pair of headphones connected in place of the VOM will give a fairly good indication of transistor noise.

Depending upon the type of meter you own, this adapter checker could be built in a small case outfitted with appropriate plugs spaced to match the jacks on the meter.

. . . Thorpe

CHAPTER 48

Transconductance Tester

With the recent advances made in the manufacture of field effect transistors (FET's), the resultant price reductions and their extended usage by industry, greater and greater quantities of FET's have been dumped on the surplus market. Many of these surplus FET's are perfectly good, but some are defective; this isn't usually the fault of the dealer, he just doesn't have an easy way to check them.

Unfortunately, you can't check out an FET with an ohmmeter like you can a conventional junction transistor; some other method must be used. The most obvious approach is to use a transconductance tester similar to that used in testing vacuum tubes. However, since the FET requires no filament power and a relatively low value of B+, the entire test unit may be made quite compact and portable.

You will remember from your school days that the transconductance of a vacuum tube was the ratio of the change in plate current to the change in grid voltage with the plate voltage held constant. For the field effect transistor, the definition is almost the same; just change the name of the respective electrodes and you have it. FET transconductance is the ratio of the change in drain current to the change in gate voltage with the drain voltage held constant. In practice all you have to do is apply the proper bias voltages to the FET, apply a small measured amount of gate drive voltage and measure the amount of resultant ac drain current. Then the transconductance in mhos[1] may be calculated from the following formula:

$$g_m = \frac{I_d}{E_g}$$

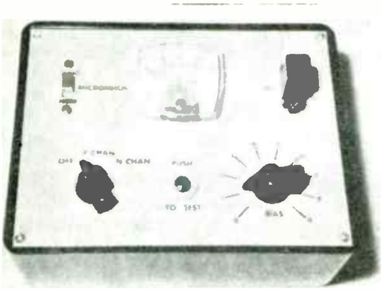
Where: g_m = Transconductance in mhos
I_d = Change in drain current
E_g = Change in gate voltage

When the drain and gate measurements are given respectively in microamperes and volts, the transductance is given in millionths of a mho or micromhos, the conventional term.

The FET transconductance tester described here combines the essential necessities with some operating conveniences that make it more versatile for all around FET tests. Basically, a small amount of 1000 Hz voltage is applied to the gate of the FET; the resultant 1000 Hz drain current is rectified by the full wave diode bridge and measured on the meter. The bridge circuit is capacitively coupled to the drain of the FET so the dc supply component will not affect the meter reading. Likewise, a large choke is included in dc supply lead to prevent the 1000 Hz signal from being bypassed to ground through the power supply.

A potentiometer is connected in the gate bias circuit so that the gate bias may be varied from zero to nine volts. This is very helpful in determining the effect of various bias levels and in measuring the gate-cutoff or pinchoff voltage of the device. Since the polarity of the dc supply voltage may be conveniently switched from the front panel, this unit will accommodate either P- or N-channel FET's. In addition, two ranges of transconductance, 2000 and 20,000 μmhos, are provided by placing appropriate shunts across the meter.

The entire transconductance tester is built on a $4\frac{15}{16}$ by $6\frac{7}{16}$ inch aluminum panel laid out as shown in Fig. 2. After the unit is completed, this panel is mounted in a standard bakelite instrument box. All of the active circuitry, including the 1 kHz audio oscillator, is laid out on a piece of perforated Vero board $1\frac{3}{8}$ inches wide by $4\frac{3}{8}$ inches long. This board is mounted to the front panel with an alumi-



The FET transconductance tester. The push-to-test switch is a Grayhill model 35-1; the transistor test socket a Pomona TS-187.

[1] The new term for mho is siemens. Unlike hertz, siemens is not widely used yet.

num angle bracket. No screws are used to hold this bracket to the panel; the battery polarity switch, push to test button and bias poteniometer do the job. There is no crowding of the board. The layout is not at all critical and any convenient arrangement is suitable. The two batteries are mounted on one side of the unit in an aluminum bracket which is epoxied to the Vero board.

The toughest part of the whole construction lies in the meter shunts. The meter used in the author's tester, a 50 microampere unit from Radio Shack, required two shunts, one for 100 microamperes and one for one milliampere full scale. These currents correspond respectively to 2000 and 20,000 micromhos full scale. The required values were calculated from the standard formula and then made up from standard carbon composition resistors. In each case a carbon resistor with a resistance value *less* than the desired shunt resistance was chosen. Then a small amount of the resistor was filed away with a rat-tail file until the resistance was raised to the desired value. Initial resistance checks were made with an ohmmeter; final tests were made by comparing the shunted meter to an accurate VOM. Except for the nonlinearities which seem to be inherent in low cost meters, the results have been encouraging.

After the shunt resistors are completed, they should be completely covered by a coat of epoxy cement. Since the protective composition cover is destroyed during the filing, the epoxy coating will prevent the ingress of moisture. Moisture will change the value of the shunt resistance and affect the accuracy of the meter.

Interior of the transconductance tester. The 1000 Hz oscillator is mounted in the upper left hand corner of the Varo board. The batteries are installed in a metal clip on the far right.

Although it is not strictly necessary to duplicate the parts used in the author's transconductance tester, it does help in acquiring the necessary components. All of the parts are available from the large mail order houses such as Allied, Lafayette or Newark. This is a help because several of the parts, notably the 2PST pushbutton switch, are not normally available through neighborhood distributors. The 50 microampere meter was chosen because of its modern appearance, availability (locally) and low cost. However, any small instrument with a sensitivity of 50 or 100 microamperes would be suitable. A 100 microampere meter would have the additional advantage of requiring only one shunt (for 20,000 micromhos full scale).

After the tester is completed, there are several adjustments which must be made before plugging in an FET. First of all, connect a

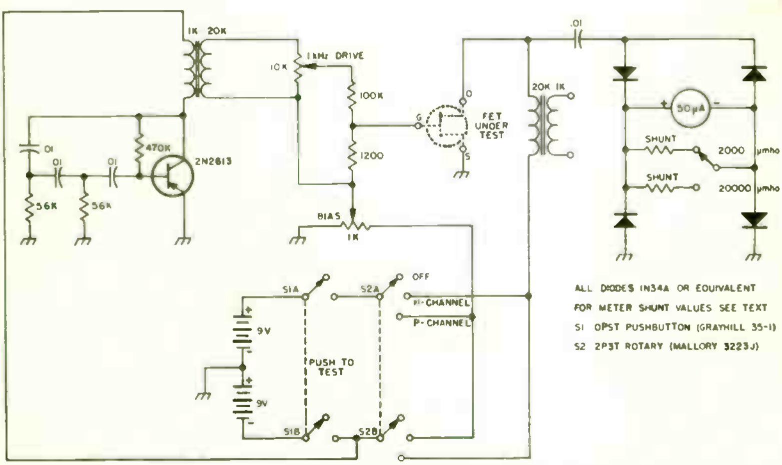
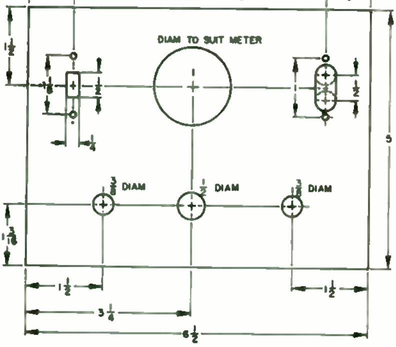


Fig. 1. Schematic of the FET transconductance tester. Although a 2N2613 was used in the 1000 Hz oscillator in the original model of this tester, almost any high gain transistor may be used in this circuit.

Fig. 2. Panel layout



set of ear phones across the secondary of the transformer in the output of the audio oscillator. When the push to test button is depressed, a 1000 Hz note should be heard. Switching from P-channel to N-channel should have no effect on the tone. Next, measure the amount of 1000 Hz signal available on the high end of the drive potentiometer with a VTVM; 8 volts RMS is about right. Now, place the VTVM test probe on the wiper of the driver pot and adjust for 1 volt RMS. This will put 100 millivolts of 1000 Hz drive on the gate of the FET being tested.

Now connect the dc probe of the VTVM to the gate connection on the test socket and adjust the bias potentiometer for 5 volts. Loosen the knob and set it opposite the 5 on the bias scale. If you don't have a VTVM for these initial setups, don't worry about it; a good VOM will work just about as well.

The discerning readers among you have probably figured out that 100 millivolts of drive (0.1 volt) and 100 microamperes of drain current *do not* add up to 2000 micromhos. You're right, they don't—1000 micromhos is more like it. That is, if *all* of the 1000 Hz drain current were flowing through the bridge. However, in this circuit, the 1000 Hz drain current divides just about equally between the audio choke in the drain power supply lead and the capacitively coupled meter circuit. So, twice as much drive must be applied to obtain accurate readings. This would be circumvented by using a larger value of coupling capacitance. However, when the push to test button is depressed, there is a large surge of current through the capacitor as it charges through the diode bridge. In the original model of this tester, the large voltage spike from the charging of a 0.47 coupling capacitor (since replaced by the 0.01 μF) destroyed a couple of $13 FET's.

Now you're all set to test those new FET's. Set the P- or N-channel selector switch, put the meter on 20,000 micromhos, and set the bias pot to zero. Push the test button—if the FET is a good one, the meter should swing up scale. If it doesn't, try adjusting the amount of gate bias. If you still don't get a reading, change the setting of the P- and N-channel switch; the device may have been mismarked.

When a good FET is being tested, note that the transconductance varies with the amount of gate bias voltage. Normally, as the bias is increased from zero, the transconductance will increase and then decrease. The point where it starts to decrease after reaching a peak approximates the gate cutoff-voltage or pinchoff voltage. This is directly analogous to the grid-cutoff voltage in a vacuum tube. With knowledge of the cutoff voltage and transconductance curve, it becomes quite easy to optimally bias the device in a circuit. Remember though that these are the characteristics of the FET with a 9 volt drain supply; other supply voltages will change the operating characteristics slightly. However, this does not negate the usefulness of the transconductance tester; on my bench it has proven to be extremely useful in determining bias levels and in sorting out defective FET's. . . . WIDTY

CHAPTER 49

Diode Tester

I think that semiconductor diodes are almost as useful in circuit construction as resistors and capacitors. Used properly, diodes are good for all sorts of tricks beyond detecting rf and rectifying ac. Well, I recently saw a chance to acquire a huge batch of assorted computer types at an irresistible price (five dollars), and my resistance being what it is, I bought them all. But when I got them into my lab, a new perspective emerged: which ones are good? As I was sorting out the color coded varieties, I developed an idea.

Like Topsy, the idea growed up! It became a schematic and some simple calculations. It developed into a mess of clip leads and components attached to a Heathkit oscilloscope. And finally I built . . . a James Dandy Diode Tester.

This simple circuit tells which end of the diode is which, what its reverse breakdown characteristics are, and it gives you a rough indication of quality. You'll have to try something else if you're interested in determining rf performance or pulse risetime and turnoff characteristics, but you can tell if it's worth further attention. The Tester also checks zeners and transistors by observing the properties of their inherent diodes. And mavbe there are one or two other uses we can find for it.

Theory

If we pare all the trimmings off the James Dandy Tester schematic, we end up with Fig. 1. This shows a high-voltage transformer in series with a resistor and a diode, and an output terminal added across the diode. Note that the diode points up. A second winding which provides the scope sweep voltage is not needed for a basic explanation. So let's work out what happens when the circuit is turned on. The key lies in the diode properties of reverse breakdown, forward conduction, and internal resistance.

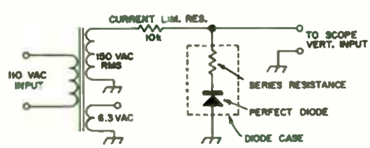


Fig. 1. Basic circuit of the Tester.

The dotted box in Fig. 1 represents the shell of the real diode. Electronically we can never open up this shell and find something inside that visibly accounts for what the diode does. But we can suppose there's a perfect diode inside the shell, and a resistor that somewhat spoils the diode's properties. Then we can describe the real diode's behavior in terms of this model. My diode-resistor model is very simple but it'll get by for now. So let's say the perfect diode goes into reverse breakdown at 20 volts, forward conduction at 0.7 volts (appropriate for sli-
resistor has the rather high value of 100
con, choose 0.2 volts for germanium) and the
ohms.

Fig. 2 illustrates the resulting situation with two superimposed curves. The upper curve represents the 150 volt RMS sine wave, always seen at the transformer terminals. The lower curve shows what we see at the diode terminals, generally a much lower voltage. Let's follow this through a complete cycle.

Starting at zero volts and going in the positive direction, we follow the sine wave along its natural course until it reaches 20 volts. At this level the diode goes into conduction, and the circuit sees the 100-ohm resistor as a heavy load with its bottom end held at 20 volts. This state continues until the transformer's sine wave returns to the 20 volt level on its downward swing. Then the diode goes off, we return to the sine curve, and follow its natural course back to zero.

The 150 volt RMS wave goes to 212 volts peak at the center of the half-cycle. We see roughly 200 volts across 10 Kilohms, or about 20 mA at this instant. Passing through the diode's 100 ohms, this current adds 2 volts

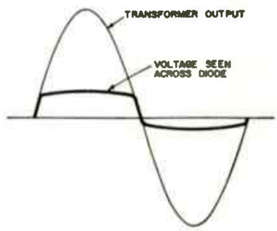


Fig. 2. Where the diode characteristics curve comes from.

SWITCH POSITIONS

POS.	S1	S2
1	30 mA	10V CALIBRATE
2	10 mA	3V CALIBRATE
3	3 mA	OFF
4	1 mA	DIODE TEST
5	300 µA	TRANSISTOR BC DIODE
6	100 µA	OFF
7	—	TRANSISTOR BE DIODE

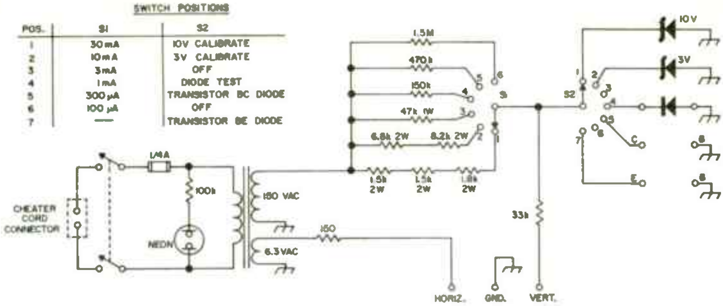


Fig. 3. Complete schematic of the James Dandy Diode Tester.

to the perfect diode's 20 volts. We will have to push the top of the diode voltage curve up a little bit, and we should round off the corners since that's what we expect to find in a real circuit. This is how we get Fig. 2, which very closely resembles the real curves you will observe using a triggered or sawtooth sweep.

The negative half-cycle closely resembles the positive curve, but the break points are very much closer to zero. The (silicon) diode takes over at 0.7 volts rather than 20 volts, and the curve bulges in the opposite direction because the current flow is reversed.

My transformer has a 6-volt heater winding which I put to use as a horizontal sweep source. This gives a linear presentation. That is, starting at the center of the trace, which should rise towards the right, percentage of distance to the end equals percentage of peak applied voltage. This eliminates using a simple trig equation if you want to know the diode current at any part of the curve. And it gives a presentation closely resembling the manual and textbook illustrations. By changing some output connections you can get an exact correspondence.

Depending upon conditions of operation, 200 volts or more can appear at the Tester output terminals. If you're looking at fine detail in the diode characteristics, this could be applied directly to your scope's input tube. The 33k resistor in series with the vertical output terminal limits current flow under these and short-circuits conditions to 5 mA or so at the price of a slight loss in signal amplitude. A much larger current is available at the diode test terminals, so watch your fingers! Turn the Tester off when changing diodes.

Construction

Fig. 3 shows a complete schematic of the Tester. Those protective resistors and the two-pole power switch might seem a little elaborate to you. But I've been in this field for some time and I think I've blown as big fuses as anybody, and got bit a few times too. The lots of little precautions like these tucked away in everything I build add up to a pretty fair insurance policy for me as well as the gear.

A 5x7x2 chassis serves as case and panel, and a bottom plate makes a worthwhile im-

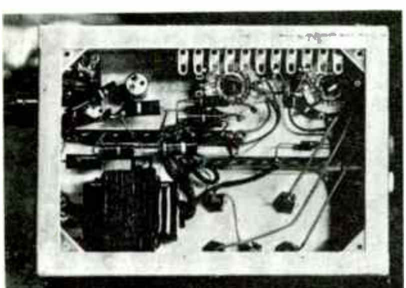


Fig. 4. Bottom view of the Tester. The calibration zeners are on the lug strip at the upper right hand corner of the chassis.

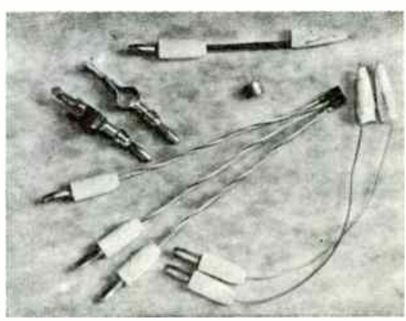


Fig. 6. Assorted leads for the tester. They go well with the Heathkit transistor tester too.

Fig. 5. Top view. I finished the Tester with slow-drying enamel and freehand India ink lettering.

provement. All wiring is point-to-point, and three 11-lug solder strips provide additional useful tie points. About half the lugs actually got used. Fig. 4 shows a bottom view of the Tester.

You can see the transformer in the lower left hand corner of the chassis. If your transformer won't go in upside down there is lots of room on the back wall. The AC cheater-cord connector goes in the LH side wall beside the transformer, with a half-inch of clearance around its solder lugs. The fuseholder is in the same wall perhaps two inches forward. There wasn't enough room for it on the top, and fuseholders aren't very interesting anyway. I might have used a TV solder-in fuse and saved cutting a hole.

On the top surface, three rotary switches and a neon pilot lamp are mounted on the same line slightly more than one inch from the front wall. See Fig. 5. With the transistor and diode terminals toward the rear, there is a clear area across the inside of the chassis which takes two of the three 11-lug strips.

I used banana jacks for all test and output connections. They seem to be more convenient than anything else. Fig. 6 shows a collection of connecting adapters made up for the Tester. The ones on the left are made up of Grayhill #2-0 breadboarding terminals soldered onto banana plugs, and they are particularly handy when testing diodes. The others are make up of banana plugs and some light and some heavy wire, with Mueller's micro-gator clips. The more common alligator clips just don't get a grip on fine wires and transistors leads. Somewhere in there is a transistor socket with short leads, which I might have color-coded emitter yellow, base green, and collector blue. These assorted adapter leads tend to congregate on my Heathkit transistor tester when I'm not checking diodes.

A rotary switch turns the power on and off. I always use a rotary switch in this critical location. A toggle switch could collapse someday, accidentally turning on the circuit. A rotary switch can't possibly do that, and its general health is immediately apparent just by looking at it. I like that.

The other two switches are single-pole non-shorting (make after break) rotary switches, and any of several varieties are usable.

When you are finishing up the circuit, leave the transformer heater leads a little loose. You may want to reverse them. Before you finalize things, hook up the tester to a scope, set the scope to very low vertical sensitivity, and see which way the trace goes. It should be a straight line, rising to the right. That is, if a positive voltage to the scope's

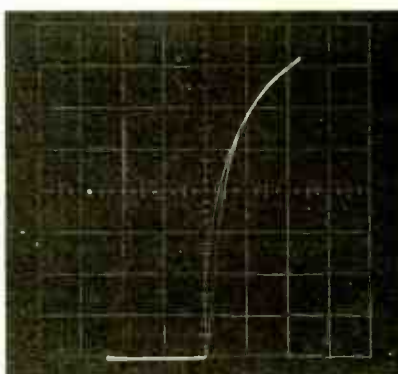


Fig. 7A. Germanium diode characteristics, showing gradual breakdown with increasing reverse voltage, and low forward resistance.

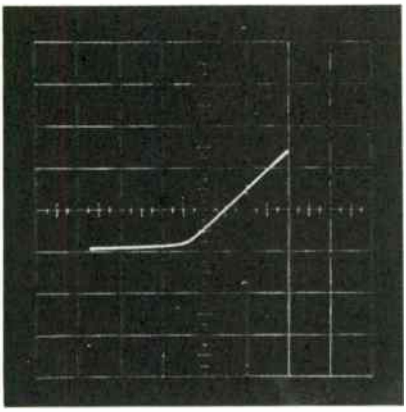


Fig. 7C. A germanium diode after overheating. The scope gain is very high, so we see that its diode characteristics are nearly gone.

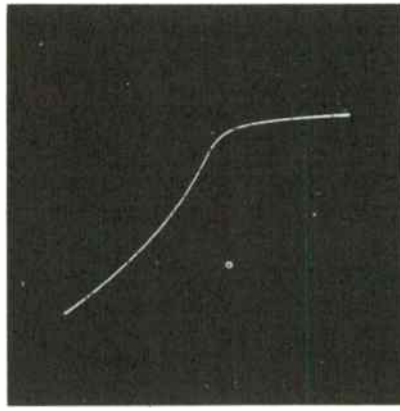


Fig. 8A. BE diode of a germanium transistor. Downward curve indicates a PNP transistor, and rather vague conduction and reverse characteristics suggest high leakage.

vertical input deflects the spot upwards, and to the horizontal input deflects the spot to the right. Otherwise you may have to redraw the curves shown in the illustrations.

The calibrating diodes go in last. Finish up everything else, and use the Tester to choose them. They'll be zeners or other diodes that show good zener characteristics. Details follow shortly.

Component values in this circuit are not critical because I don't expect too much from it. If I need exact measurements I get them somewhere else. I've chosen properly sized resistors so you can leave it on all night without anything roasting. If you want to change those resistors, it's easy. Ohm's Law: RMS voltage over resistance equals RMS current, and you can see in Fig. 5 which values I chose. If you can't find a 150 volt transformer, compute new resistances for what you have available. I wouldn't use a lower voltage because some transistors and small diodes show breakdown voltages in the 100-volt range.

The case is finished off with good enamel and careful hand lettering.

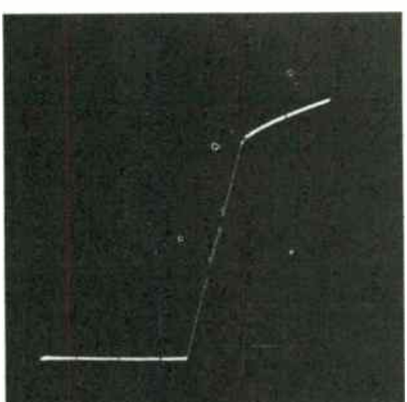


Fig. 7B. Another germanium diode, showing a sharp knee but poor dynamic resistance.

The calibrating zeners

If you have a scope with fixed voltage ranges, you probably aren't interested in the calibrating zeners. If not, you need them, but how are you going to find out what their values are?

Perhaps you have some zeners of known characteristics, but the usual 10%, 20% or greater tolerances seem rather excessive. If you're familiar with your VTVM, you may have guessed the answer already: use its ability to indicate peak-to-peak AC voltages.

A review of the meter manual should answer any questions that may arise. So far as I know, all inexpensive VTVM's use a peak-reading circuit, with a meter scale that is labeled for sine-wave readings. We'll just convert those estimated sine-wave figures right back to P-P, by multiplying by 2.82. Or perhaps, like my Paco, P-P scales are included on the meter face.

Set up the Tester and your oscilloscope. Attach your meter ground lead to the Tester ground return, and the meter probe to the scope Vertical Input terminal. Set the VTVM for AC measurements and start testing diodes. When you come to a diode that has nice sharp corners and flat top and bottom, make

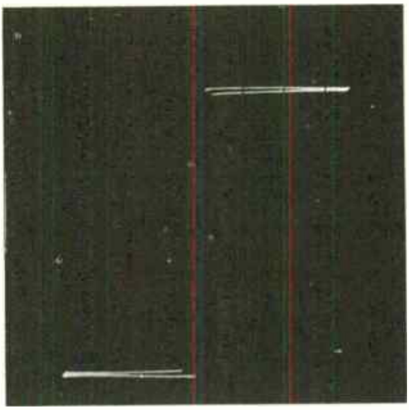


Fig. 8B. BC characteristics of the same transistor. This curve is also downward, and it shows a very sharp conduction and breakdown knee.

an RMS reading, convert to P-P, and you have that diode calibrated. I think 3 volts is a little low, because I went through nearly a hundred diodes and transistors before I found one of this value; you might try 5 volts and you'll find one quickly. Three more of them would add up to 15 volts, and these are probably better choices than 3 and 10 volts.

Remember to make your measurements at the same current you will use when calibrating the scope. My zeners give true readings at 1 mA; you'll get sharp corners more easily at a higher current.

Testing diodes

The quickest way to understand the Tester indications is to put a diode in it and then

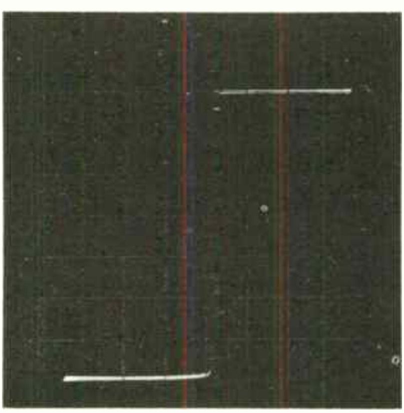


Fig. 9A. It's not obvious here, but this silicon transistor BE diode curve turns upwards.

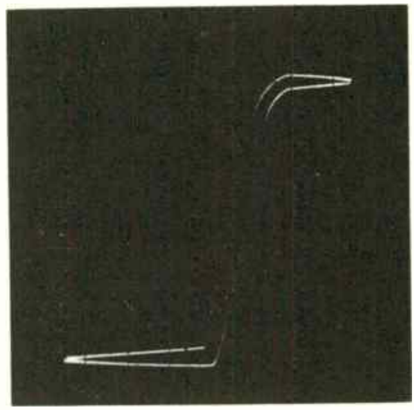


Fig. 9B. BC characteristics of the same transistor. I don't know what causes the very noticeable phase shift. Can somebody tell me?

work out the meaning of the different parts of the curve. Repeat with several different diodes. Most everything you need to know is in the theory section, and in several widely distributed handbooks. Just take a little bit at a time and ask, how did it get that way? I've included some illustrative photos and brief explanations.

All bipolar transistors have two inherent diodes. One is the base-emitter diode, and the other is the base-collector diode. The Tester checks these diodes one at a time, and it doesn't tell you anything about how the transistor will work. But if one of the diodes is bad, the transistor won't work. And the direction the curve goes indicates whether you have a PNP or an NPN transistor. See Figs. 8 and 9.

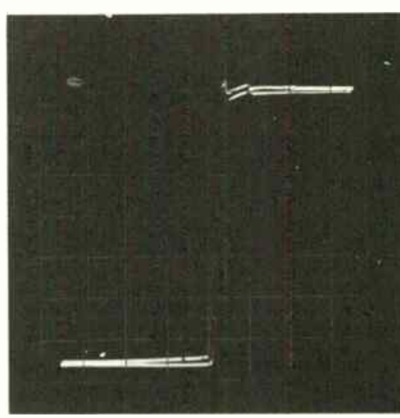


Fig. 10. A very close look at a perfectly good GE Z4XL6.2 zener diode. It shows some zener noise under 200 microamps, and low dynamic resistance.

Why do many diodes show a double line in the vertical parts of the pattern? These lines merge at higher currents but are very distinctly separate for small currents and large diodes. I think this is phase shift of the applied voltage through the RC network of series resistor and reverse-biased diode capacitance before it goes into breakdown. In that case, the LH line would be the rightward-going trace (phase retarded).

Zener regulators

Do you have trouble finding zener regulators? The Tester will find lots of them, and tell you how they'll work in your circuits.

It turns out that not only specially built silicon diodes will serves as zener regulators, but some unspecial diodes and even germanium transistors! The Tester finds the ones that can regulate, and some of you out there working with low-power circuits can now find very low-power zeners to go with them. Wish I'd found out about this sooner! I haven't done any work in the matter, but I expect germanium zeners aren't going to show as good temperature stability as silicon zeners. Well, that is another problem. Fig. 11 shows the base-emitter breakdown characteristics of an unlabeled germanium computer transistor from somebody's printed circuit board.

Closing comments

The Tester can supply lots more power than is required to roast a good diode into complete uselessness. Fortunately, this is harder to do than you might think. The power dissipated in the device is the usual product of voltage times current, and since most transistors and diodes break down at under 20 volts, and can take over 100 milliwatts ,the average danger line lies around 5 mA. But some transistor may unexpectedly show a base-collector breakdown at 50 volts or more, and if it's rated at 60 mW you may easily overdo things. I hope you'll try to roast a few semiconductors to get a feel for what you can and cannot get away with . . . just watch the scope and you'll see the curve begin to slump off towards a straight line. Then heave the poor thing into a nearby wastebasket so it can't end up in one of your circuits.

If you're careful to use things only for their intended purposes, you are missing a lot of fun. What can you do with the Tester? Try a little neon lamp at low current. Another thought that comes right to mind is that perhaps it can be used in some way to check computer switching cores. I'm sure you can work up some new ideas. And as you puzzle them out you'll pick up a few pointers enabling you to make better use of this simple but surprisingly handy James Dandy Tester.

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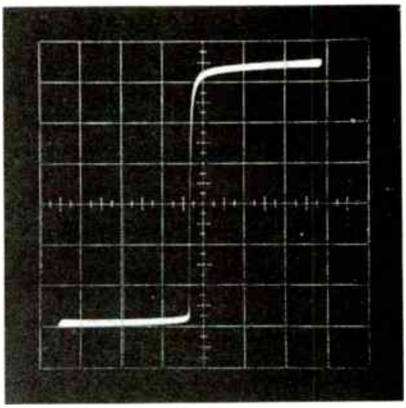


Fig. 11. BC reverse breakdown curve of an unknown computer-board germanium transistor. This one would make a good low-power zener.

CHAPTER 50

"Multical" Crystal Calibrator

What is the "Multical"? As the name implies, "multi" would suggest several uses, and "cal" might infer a calibrator of some sort.

Well, that's right, but there is slightly more significance to the name. "Multi" is also a short form term used to describe flip-flop circuits known as multivibrators.

By combining the basic characteristics of a free-running multivibrator (astable) with crystal control, you have a simple, stable, virtually-insensitive-to-temperature-changes, crystal calibrator for that receiver you have been wondering about.

The circuit uses no inductors and depends upon the crystal for the proper feedback for oscillations. Temperature stability is partially due to the absence of capacitors.

Transistor stage Q_2 operates with unity gain, whereas transistor Q_1 operates at considerably more gain. Both stages are operating as feedback amplifiers. The harmonic generator diode D_1 is a 1N128. Any general purpose diode may be used.

By using the multivibrator circuit, the waveform obtained is comparatively rich in harmonics and could be used without any further refinements. However, to insure useful harmonics through 30 MHz starting from a 100 kHz crystal, a harmonic generator consisting of R_6 and D_1 shown in Fig. 1 was added. The capacitors C_1 and C_2 are used strictly for coupling and have no effect on frequency stability.

Crystals from 100 kHz up to 1 MHz may be used in the Multical with no changes. The circuit will oscillate from voltages as low as 2 volts and can be operated safely from voltages as high as 20 volts. This wide range of voltage operation allows the source to be obtained from virtually any place.

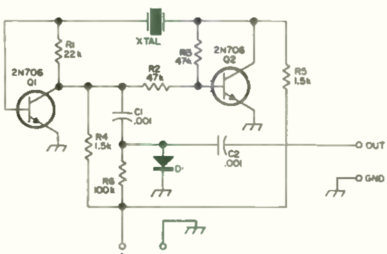


Fig. 1. Schematic of the Multical.

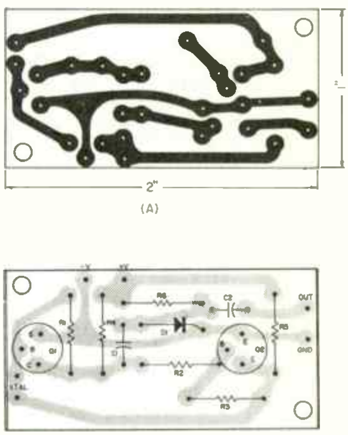


Fig. 2. Suggested printed circuit board layout for the Multical. A gives the copper side, B the component side.

Output from the calibrator may be fed directly into the receiver's input, or may be coupled to a short whip antenna. With a whip antenna, close coupling to the receiver's input may be required at higher frequencies. (Especially at the lower voltage levels.)

For the more ambitious builders, Fig. 2 shows the printed circuit board layout for the Multical. Due to its small physical size (1" x 2"), room can probably be found even in the most compact of receivers. Fig. 2A shows the foil side, and 2B shows the parts placement.

So the next time you wonder about the accuracy of your receiver calibration, give this simple circuit a try and you'll know for sure.

. . . K9VXL

CHAPTER 51

Audio Frequency Meter

WANT to measure frequency down to the last cycle per second? Or maybe find the exact resonance point of that AF filter you've just whipped together?

Here's a gadget that can help you do either of those, and more besides. RTTYers will find it handy for checking frequency shift. Experimenters can use it for measuring drift in a VFO. And you will find many uses for it, too, around your own shack.

It's an audio frequency meter, costing approximately five dollars if all parts are purchased new and taking only about an hour to put together (less time than that if you're used to homebrew techniques).

While many AF frequency meters have been described in previous articles, none have all the advantages of this pocket-sized unit. Designed around the peculiar properties of most transistors, it uses only nine components (aside from range-switching circuitry), is rugged, and features high accuracy.

Before getting into the construction of the little gem, let's take a look at how it works.

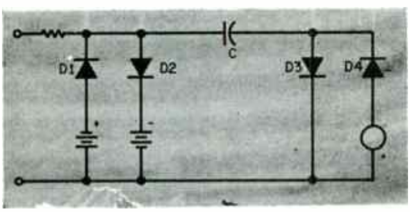
The basic principles of the direct-reading audio frequency meter have been with us for at least 15 years (see the references). However, the limitations of vacuum tubes and later of vacuum-tube-directed design techniques have kept the beastie complicated enough to prevent many hams from building it.

The block diagram in Fig. 1, adapted from Terman's *Electronic and Radio Engineering*, shows the conventional circuit. Diodes D1 and D2 limit the signal to a definite peak value. Capacitor C1 differentiates the limited signal into positive- and negative-going spikes. Diode D3 shunts the negative-going spikes around the meter circuit, while D4 allows the positive-going spikes to pass through the meter. The deflection of the meter is directly proportional to the number of spikes which pass through it within a given time.

As shown, the circuit is simple enough. However, the usual input signal is small—and this circuit requires spikes some 45 volts high to give an accurate indication.

Previous designs have solved the problem by first amplifying the signal, then passing

Fig. 1—Simplified Diagram



it through a limiter stage, and finally differentiating and measuring it. See any of the construction references for further details.

The five-buck special, on the other hand, uses the switching properties of transistors to accomplish the same purpose.

Looking at the schematic diagram, Fig. 2, you will see that input signals go directly to the base of Q1 through the 470-ohm current-limiting resistor. Normally, since the base is not forward-biased, Q1 is cut off and passes little or no current. As a result, there is no voltage drop in the 5600-ohm collector resistor and the voltage at the collector is -9 volts.

When the negative half-cycle of an input signal comes along, however, the picture changes. The negative input places forward bias on the base, and when this bias becomes large enough, the transistor switches to saturation. Resistance from collector to emitter becomes less than one ohm, and the entire supply voltage is dropped in the collector resistor. Collector voltage drops to zero.

Approximately 0.2 volts is the crossover point for the 2N107 used in this circuit. This makes 200 millivolts the smallest signal which can be measured. Upper limit is determined by punchthrough voltage rating, and is about 10 volts for this unit.

We have seen how a square wave is developed at the collector of Q1 from a sine-wave input. Now let's look at the rest of the circuit.

Capacitor C, the timing-reference unit, differentiates the square wave into spikes exactly as in previous circuits. These spikes are applied directly to the base of Q2.

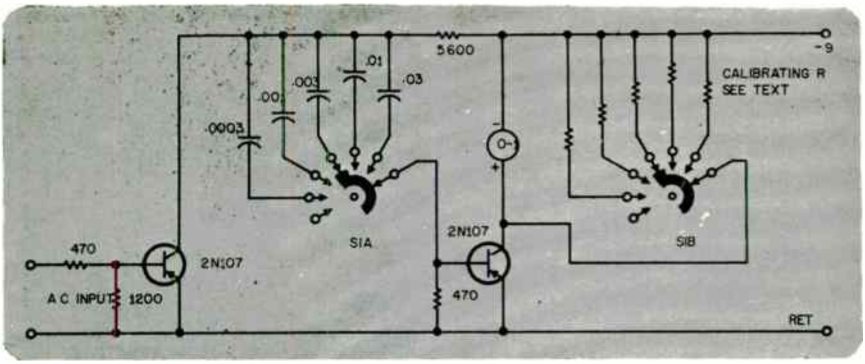


Fig. 2—This is the diagram of the deluxe model. For the five-dollar special, omit all switches and components associated with them. Connect a capacitor of proper value in place of SIA. Ranges are: OFF, 30 KC, 10 KC, 3 KC, 1 KC, and 300 CPS. Meter is 0-1 ma.

Q2 also acts as a switch. Positive-going spikes simply reverse-bias the base and have no effect on the collector circuit. However, negative-going spikes turn the transistor "on" for the duration of the spike and allow pulses of current to flow through the meter.

Since the amplitude of the spikes is increased through Q2's switching action, an inexpensive meter is highly satisfactory. The circuit provides linear operation up to approximately 10 milliamps current flow through the meter. Earlier designs required movements in the 1 ma to 100 microamp range for linear operation.

That's how it works. Now, to construction.

Perforated phenolic board makes a fine "chassis" for the two transistors and four resistors used. I built the prototype on a salvaged printed-circuit board given away at an electronics parts show. If you don't want to strive for the ultimate in miniaturization, use three-terminal tie points.

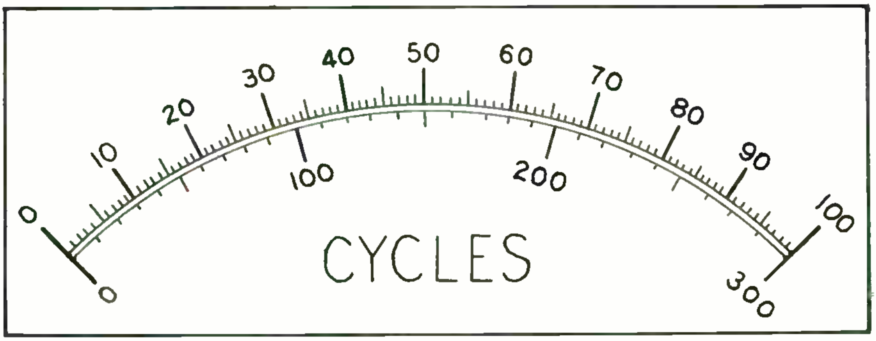
All the usual heat precautions applicable to any transistorized construction apply here. Leave the leads long or use long-nose pliers as a heat sink between the transistor and the solder joint. Aside from that, wiring is not critical.

If you're building a single-range frequency meter, timing capacitor C can be placed on the circuit board also, as can calibration resistor Rc (when its value is determined as described later). For a multi-range meter, these components should be mounted on the range switch.

Whether you are building a single-range or a multi-range meter, leave the calibration resistance out of the circuit at first. It will be permanently connected later, after its exact value is determined.

If you're building a multi-range meter, you can use the meter face shown in Fig. 3. Simply have a photocopy made, the proper size to fit your meter. For a single-range meter, pick the basic movement to show the same values as the frequency range you're interested in—that is, for a 0-3 kc meter range, use an 0-3 milliammeter, etc.

Fig. 3—To dress up the frequency meter, use this meter face. Have a photocopy made in the exact size to fit your meter dial (tell the photographer to print it on Type A paper) and glue the copy to the meter dial with rubber cement.



Once the meter is built, calibration comes next. WWV provides a handy source of 440- and 600-cycle tones, but be wary of frequency distortion caused by multipath transmission of the signal. A reliable 60-cycle calibration note can be obtained from the secondary of a filament transformer.

However, the method used on the original provides a number of tones in the range of interest, at very low cost. A hi-fi frequency range test record, such as those distributed by Cook Records or RCA Victor, is placed on a convenient record player and the freq meter is hooked to the speaker leads. The result is a large number of calibration points from 15 kc down to 50 cps.

Since the scale is completely linear on each range, only one calibration point per range is necessary. If all capacitors in a multi-range unit are within 1 percent of the marked value, only one calibration is necessary for the entire instrument.

However, since you can buy five 5-percent resistors far cheaper than the difference in cost between 20-percent and 1-percent capacitors in the range needed, let's use the 20-percent capacitors and calibrate each range individually. If you're building a single-range meter, simply stop when you've calibrated your single range.

To calibrate the unit, jumper in a 500-ohm rheostat across the meter using test leads. Connect the frequency meter to the calibration source you're using. Adjust the rheostat until the meter needle indicates the proper frequency (such as .6 if you're using a O-1 milliammeter for a freq meter on the O-1 kc range, with 600-cycle calibration tone).

Disconnect the rheostat without disturbing its setting and measure its resistance with an ohmmeter or bridge, if you have one. Select a ½-watt resistor with the same resistance and connect it in the unit.

Repeat this procedure for each range of the meter. That's all there is to it.

On a multi-range meter, you may find it easier to get the exact resistance value needed by connecting a number of small resistors in series. When you do this, you may be able to cut down the number of resistors used by noting the resistance necessary for each range, then making up the smallest resistance first. Add just enough to it to reach the next higher value, then enough more for the next, and so forth. Bring out taps to the range switch. A glance at Fig. 4 may make this clearer.

Operation of the completed and calibrated instrument is simplicity itself. Simply connect its input to the unknown signal source (making certain that no dc is present; if in doubt, use a transformer or a coupling capacitor), crank the gain up until the reading becomes steady, and read the meter.

Gain must be cranked up for this because the input transistor acts like an amplifier instead of a limiter for signals smaller than the 200-millivolt turnover point. You will also find that the reading increases as gain goes up, until it reaches a point at which it comes back down. If you increase the gain still more, the needle backs off to a point and holds steady. This is the proper indication.

Reason for the variation in reading is this: When the gadget first starts limiting the input signal, it merely clips off the top of the negative half-cycle. This produces a pulse-type output instead of a true square wave. The pulse output acts, to the meter circuit, like a combination of a high and a low frequency. The needle responds by wavering. When proper limiting level is reached, the needle is steady as a rock.

Since this frequency meter is current-operated rather than being voltage-driven, be sure the input can supply a little power. A half-watt is enough, but purely-voltage sources (such as the output of a hi-fi preamplifier) simply won't operate it, even though the voltage is far above the 200-millivolt level. If this seems a disadvantage, it can be overcome by putting an emitter-follower amplifier ahead of the limiter circuit as shown in Fig. 5. This amplifier was omitted in the prototype because it was felt unnecessary.

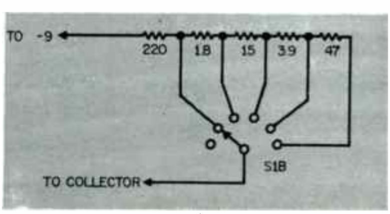


Fig. 4—Alternate calibrating-resistor circuit discussed in text. Values shown for resistors are for example only—the exact value to be used in each position must be determined after the instrument is built as described in text. In this example, position 1 gives 220 ohms; 2 gives 221.8; 3 gives 236.8; 4 gives 240.7, and 5 gives 287.7. If necessary, switch contacts can be jumpered to give identical resistances or the sequence reversed to give lower resistance on higher range.

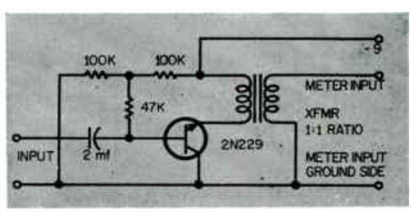


Fig. 5—This tiny preamplifier may be added to the input of the frequency meter to increase the meter's ability to indicate frequency of voltage-only sources. The emitter-follower circuit has an input impedance of nearly half a megohm, reducing loading effects, and provides more than 10 db current gain, which will more than fully drive the frequency meter from any ordinary signal. Any small transistor-type interstage transformer may be used; its prime purpose is to keep all dc out of the frequency-meter input.

CHAPTER 52

Capacity Meter

Of all the various methods of capacity measurement, the direct reading capacity meter has the greatest appeal from the standpoint of operating convenience and rapidity of measurement. The capacity meter is of course a great help to those who have difficulty remembering the myriad of color codes, as well as those of us who may have color perception deficiencies. Equally useful is the ability to measure the capacity of a length of coax cable to determine if there is a break close to the end where it is easily accessible for repair. Odd lengths of antenna can also be readily measured to enable calculating the amount of series inductance needed for resonance at lower than the natural resonant frequency of the antenna. Many other odd jobs can be quickly accomplished with the aid of a portable capacity meter, such as measuring stray wiring capacitance, locating breaks close to the surface in coils, breaks in line cords, etc. An ohmmeter will tell you a cord or cable is open, but a capacity meter will tell you where.

The instrument was transistorized to add to the convenience of operation and eliminate the need for power cords, or waiting for it to warm up and settle down. Along this same line, the meter is large and easy to read accurately, and the small case with a carrying handle compares favorably with most standard multimeters.

Ranges

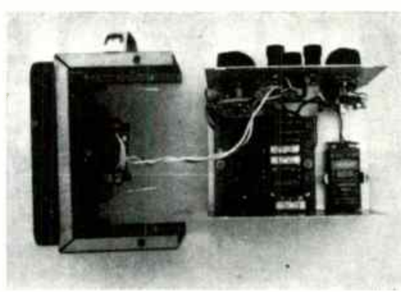
Four basic ranges were provided, calibrated at full scale by means of the built-in standard capacitors of 100 mmfd, 1000 mmfd, .01 mfd, and .1 mfd. By means of the built in standards and the calibration control other ranges may be used to increase the ease of measurement. For example, a capacity which read just off scale on one range would be slightly above 1/10 scale on the next range. Instead, by re-adjusting the calibration control, so the calibrating capacitor read ½ of full scale, an unknown capacitor just slightly larger can be readily determined, using a mental multiplier of two. Although the author's instrument has an apparent residual capacity of about 0.8 mmfd, capacitors as low as 1 mmfd can be measured if this residual capacity is allowed for and subtracted from the indicated reading.

Accuracy

The accuracy, as well as the cost of the instrument will depend mainly on the basic meter selected and the four standard or selected capacitors. The transistors are fairly by the meter adjustment screw. The error was greatest near the center of the scale.

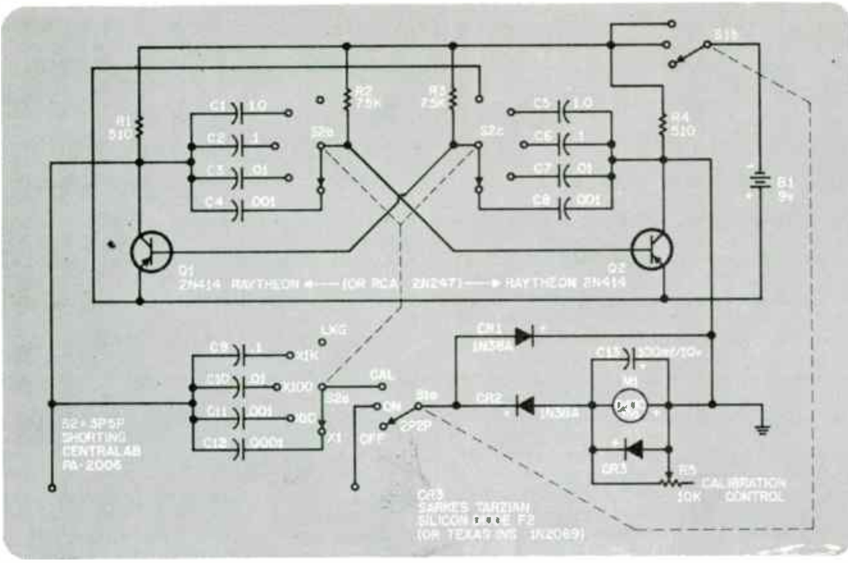
Theory of Operation

The circuit operates by measuring the amount of charge which the capacitor under test receives by the application of a square-wave from the multivibrator. A pair of diodes in a simple rectifier circuit enable a microammeter to be used as the indicator. Since the amount of charge on a capacitor, with a given voltage applied, is directly proportional to the inexpensive rf or if type PNP units. The mercury battery shown in the photographs is not really a necessity for any dry-cell type can be used if it has enough voltage and can maintain a steady full scale reading on all operating ranges. The multivibrator used in the circuit



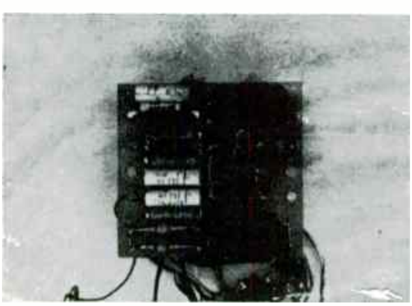
Inside view of the capacity meter shows the compact construction made possible by use of miniature components.

Fig. 1. By proper wiring the residual capacity indication is easily reduced below 1 mmfd.



is quite stable, hardly changing frequency over wide supply voltage variations, and in any case each range is normally calibrated by the built-in standards before use. The author obtained 1.0% of full scale accuracy on the three higher ranges, and 3 to 4% of full scale accuracy on the 100 mmfd range. This does not mean that a very low capacity may be off 3 or 4 mmfd. since zero is mechanically set capacitance, the scale is linear and there is no need for other than full-scale calibration. Hence, a variable resistor, R5 in Fig. 1, is used to shunt a bit of extra current around the meter, to allow for battery aging and also to eliminate the necessity of setting the multivibrator exactly on frequency.

By operating the MV at four fixed frequencies, in decades, the range of operation covers



Use of a component board greatly simplifies wiring, with connections made to the underside of the board.

practically all small color-coded capacitors. The frequencies used are 100 cycles, 1 kc, 10 kc and 100 kc.

The MV, in Fig. 1, has three possible states of normal operation. They are: Q1 conducting and Q2 cut off, or Q2 on and Q1 cut off, or a transitional state where both conduct. When the power switch, S1, is first turned on either Q1 or Q2 starts to conduct more readily than the other due to inherent circuit unbalance. Due to the regenerative action of the cross coupled amplifiers one or the other soon is driven to saturation, with the opposite amplifier cut off by the large positive bias developed by the charge on the coupling capacitor. The plus charge drains off toward the B— thru the base bias resistor, and at about —0.1 volts of base bias, the cut off transistor then conducts, and regeneration quickly causes this transistor to become saturated, with the opposite one cut off. This process repeats itself at a rate governed mainly by the base bias resistors, R2 and R3 in Fig. 1, and the intercoupling capacitors, C1 thru C8. The result is more or less a square-wave. A capacitor, Cx, is connected across the terminals J1 and J2 with the instrument turned on. When Q1 is conducting and Q2 is cut off, Cx is charged to practically the full battery voltage thru CR1. On the next half cycle Cx discharges thru CR2 and the meter, M1, and recharges again in the opposite polarity, to the supply potential. The result is, of course, a current indication on the meter, the exact value depending upon the supply voltage, the capacity of Cx, and the rate at which this charge and discharge effect takes place.

In order to protect the meter from damage due to shorted capacitors, CR3, a silicon diode biased in its forward conduction direction, was included. This limits the maximum voltage across the meter to about 0.5 or 0.6 volts. The meter movement itself was thus protected, but spikes due to the capacitor charge caused erroneous readings when the diode conducted prematurely. To prevent this, C13 was included, and in addition C13 provides damping which further prevents meter damage.

In actual use, the range switch is set to a position which gives an on-scale reading. The power switch is then set to CALIBRATE, and R5 is set to give a reading of 100. Then switch back to ON, read the meter, and use the appropriate multiplier indicated on the range switch. Since leaky capacitors would give erroneous indications, a leakage test position was included on the range switch. For this test, Q1 is cut off, Q2 conducting, and the full battery voltage connected across J1 and J2. Any indication of course means a defective capacitor.

Construction

The entire circuit including the battery and a 4½ inch panel meter was built into a 3 x 4 x 5 inch LMB chassis box. Most of the components were mounted on a piece of 1/16 inch epoxy glass board, using eyelets and jumper wires. Use of 100 volt rating capacitors helped cut down the size. The 9 volt mercury battery was mounted in a clip from a cabinet latch. The current drain is in the order of 20 ma, and the required voltage slightly above 6 volts, so a 7.5 or 9.0 volt dry battery could be substituted if desired.

The range switch, S2, should be of the shorting type to prevent the multivibrator from stopping when switching ranges, which would require turning the power switch off, then on again. Aside from the usual precautions to observe polarity of the diodes and battery, the only critical wiring is in the need for short direct leads to J2 from the power switch, and from the power switch, S2, to CR1 and CR2. The capacity of the lead to the standard 100 mmfd capacitor, C12, was about 3 mmfd in the authors instrument, and was allowed for in the selection process. The standard capacitors (C9-10-11-12) were measured on an accurate bridge, and were padded where necessary to obtain the correct value. All other capacitors can be whatever is on hand in the range required, either paper or mylar dielectric type being suitable. For reduction of stray capacity effects a ground from the meter circuit to the case is quite essential.

POWER SUPPLIES

CHAPTER 53

12/24 Volt Power Supply

In the old days when vacuum tubes were in vogue, it was an easy matter to obtain operating voltages for experiments and testing equipment. This was done by tapping off power supplies in whatever gear was at hand. Then came a wonderful little device known as the transistor (sometimes known as a "three legged fuse"). Most everyone was happy with these semiconducting devices (except for tube manufacturers), especially battery manufacturers. This was a shot in the arm for new and better batteries. Battery power was fine for the old breed of transistors that only consumed milliamperes, but the new breed has some mighty hungry units that consume AMPS. The old dry cell doesn't last long with this kind of power drawn from it. And the wet cell—the less said about this messy device the better.

A short while back I started a project on a solid state UHF exciter. My first thoughts were on how best to power the unit. Since the total power consumption of the unit was to be about one-and-a-half amps, batteries were out of the question. A transformer and bridge rectifier was OK, except with a small load (the oscillator and buffer in my case) the voltage was high. As the load increased, the voltage of course came down. Since the junction capacitance of a transistor will change somewhat with varying voltage, this makes it difficult to keep a stage tuned properly at the higher frequencies. The only solution to the problem is to stabilize the voltage.

There are many ways to make a stable voltage source, some of which get rather involved and expensive. Since my needs were not too critical, I chose the simplest method which is the series regulator, which will hold within 2 volts from no load to full load. In my case I needed 24 volts for the exciter under normal operating conditions; however, it was deemed desirable to be able to reduce this voltage for initial testing and experimentation. What finally evolved was a 12/24 volt supply capable of delivering 3 amps at 24 volts and 1½ amps at 12 volts. The current limitation at 12 volts is set by the dissipation of the transistor and heat sink. Power dissipation of the transistor is approximately equal to supply voltage in minus voltage out multiplied by the load current. Since the supply voltage is approximately 35 volts, it can be seen that 34.5 watts is dissipated with a load of 1.5 amps with the output set at 12 volts. This is close to the maximum safe limit for the heat sink I used and keeping within the temperature rating of the regulator transistor.

Since my rig will use 24 volts under normal circumstances, I will still be dissipating 33 watts while drawing 3 amps. Construction is not critical and general layout can be seen from the photos. The only precaution is to use heavy gauge wire in the unit to prevent IR losses.

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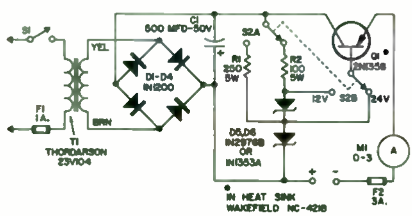
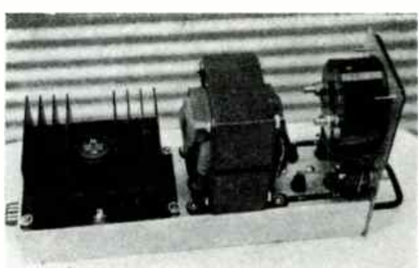


Fig. 1. Schematic of the supply.



Side view of the regulated power supply.

CHAPTER 54

Lab-Type Power Supply

A typical laboratory power-supply will have variable voltage output, low internal impedance, good voltage regulation with a variety of loads, freedom from output changes with line-voltage fluctuations, adjustable current limiting, low ripple and noise voltage in the output, and accurate metering of output voltage and current. Clearly all these features are not needed for *every* system, but several of them will be suited to each particular circuit under test. Having tested the circuit with our laboratory supply, we can then vary the supply voltage around to test the circuit sensitivity to input voltage change. We can also, put resist-

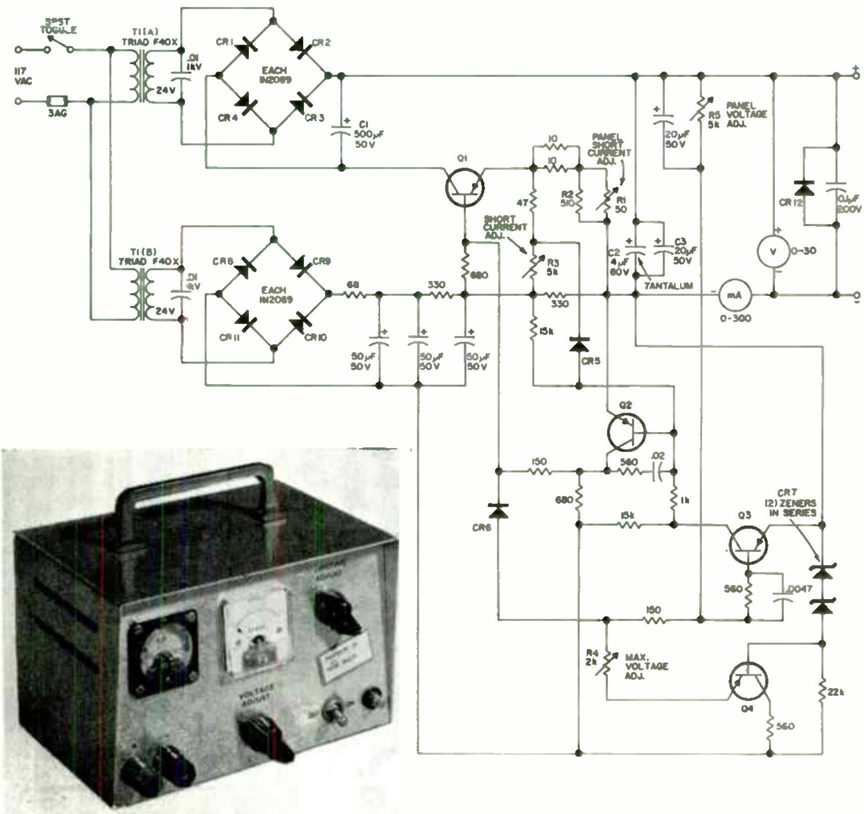


Fig. 1. Laboratory-type power supply. Q1 is a 2N375 or 2N1542A; Q2, Q3 and Q4, 2N508; CR1, 2, 3, 4, 8, 9, 10, 11, 12, 1N4002 or 1N2069; CR5 and CR6, HB5 or FD1135; CR7, 1N755 or two 1N468's in series.

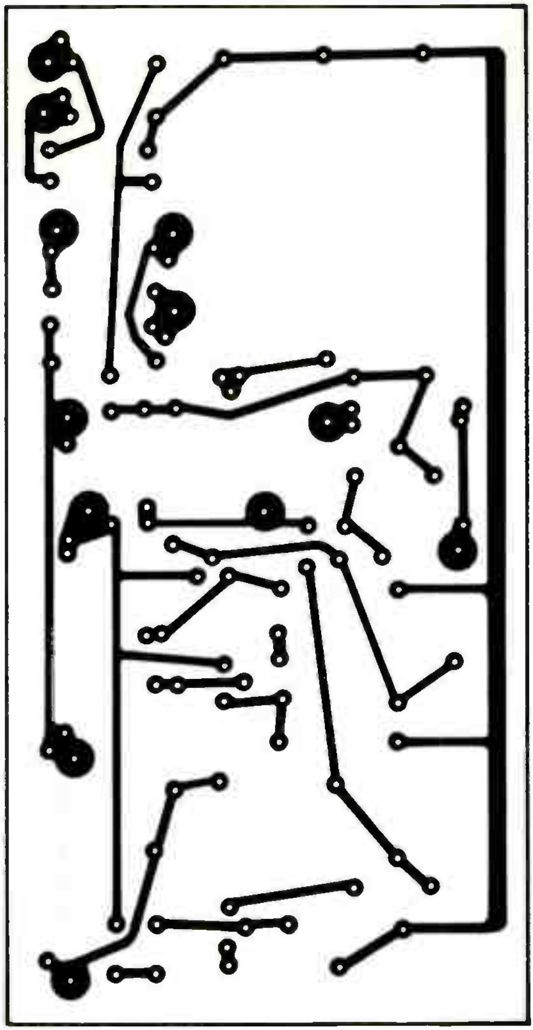


Fig. 2. Circuit board layout for the laboratory power supply shown in Fig. 1. This board is available from the Harris Co., 56 E. Main Street, Torrington, Conn., for $2.

ors in series with the lab power-supply to check the circuit dependence on the power-supply internal impedance (has your high gain audio amplifier ever 'motorboated'?)

Then with complete knowledge of the voltage, current, internal impedance, and voltage stability requirements, we can proceed to build a simpler power-supply for our circuit.

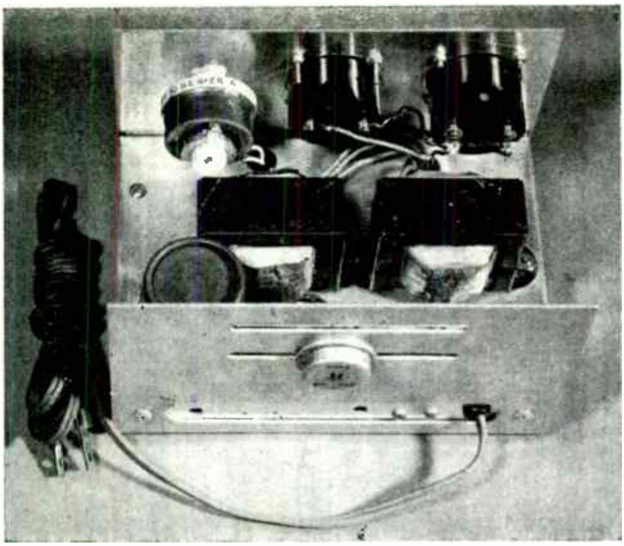
One very satisfactory laboratory power-supply has been described in the *Handbook of Selected Semiconductor Circuits.*[1] Although not so identified, this circuit appears to be that of an early model of the Hewlett-Packard 721. The laboratory power-supply circuit, as modified to make it possible to build with readily obtainable parts, is shown in **Fig. 1.** As presented in the *Handbook of Selected Semiconductor Circuits,* it had several special components in it, and equivalents had to be found for these. The power transformer T_1 was replaced by two Triad F40X's, both operated in the full-wave bridge configuration. The series of three resistors and a switch, comprising the current-limiting selector, was replaced with a 50-Ω rheostat having a 510-Ω fixed resistor across it, which gives continuous current-limiting adjustment. Also, the meter switching and associated resistors were done away with, and separate meters used. CR5 and CR6 were originally of a type not readily available and can be replaced by Hoffman HB5 silicon diodes, or any similar general purpose silicon types. CR7, another obscure type, was replaced with either two 1N468's or one 1N755 (but nearly *any* zener of about 7 volts will do). The series transistor (Q1) is listed as a 2N375, but a number of other 80 volt power transistors have been used in its stead, including the 2N174 and 2N1542.

Most of the circuitry is laid out in a simple etched circuit board as shown in **Figs.** 2 and 3, and so wiring is very fast. The supply is housed in a standard LMB-W1A cabinet. The series regulator (power) transistor is heat-sink mounted on the back plate (with a mica insulator, of course).

One of the units the author built was made entirely of MARS-supplied semi-conductors; a large percentage of the other small components were also MARS-supplied.

A silicon rectifier was added across the output of the supply in addition to the 0.1 μF capacitor to help kill any transients (from the equipment being run by the supply) that are opposite to the supply polarity.

The meters used are the inexpensive Japanese-made miniature types available under



Top-rear view of the lab supply with cover removed. A Triad F45X has been substituted for one of the F40X transformers.

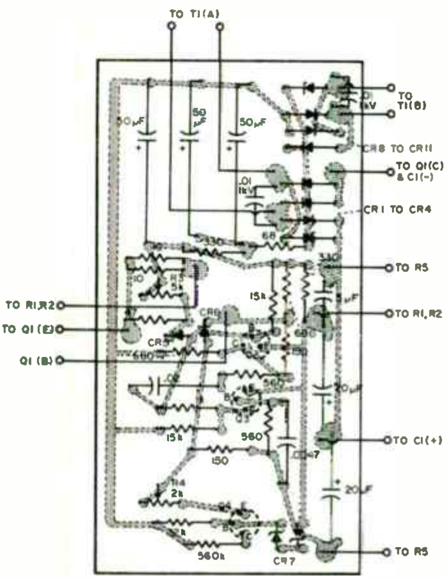


Fig. 3. Component side of the circuit board shown in Fig. 2.

many U.S. names including "Calrad", "Monarch", and "Lafayette". Their accuracy seems adequate for this application.

One will note that C1 and C2 are in parallel. The original purpose of this apparent duplication was to assure that the 20-μF electrolytic was a low impedance for low frequencies, and the 5-μF tantalytic was a low impedance for higher frequencies. However, if one has a 25-μF to 50-μF tantalytic, the single unit will do by itself.

When the power supply is finished, just plug it in, adjust the two screw-drive adjustments: "max. voltage adj." and "short circuit current adj." for the 0-30 V range and 25 to 225 mA range of the front-panel controls.

. . . W6GXN

[1]Sherrill, P. N., "Small Lab Supply with Current Limiting." in *A Handbook of Selected Semiconductor Circuits* (NObsr 73231); Available from U. S. Govt. Printing Office.

CHAPTER 55

Comments on Power Supplies

The power-supply is usually the very last part of the circuit to be built into an electronic design. This is perhaps as it should be, since the circuit designer often doesn't *really* know the exact voltage and current requirements until the design is done. But, because the power-supply *is* last, it often is a victim of the "dollar-short and day-late" effect. Also, the power-supply may suffer at the hands of a designer too well-steeped in the ac-dc broadcast receiver practice, to the point where "power-supply" and "a rectifier and a capacitor" are synonymous.

The author's point of view is that the power-supply should not be an after-thought, but rather should be well designed; since a good power-supply can effect many simplifications in the associated circuit design. Also, in this age of very inexpensive diodes and transistors, the difference between a rough power-supply and quite a nice power-supply is only a matter of a couple dollars worth of parts.

If one accepts the idea that a good power-supply design is desirable, will result in a better overall circuit, and will save design time to boot; how does he approach the total design? Or, which comes first, the chicken or the egg? If one first builds himself a laboratory power supply which has *all* the features of any supply he could want, he can use it to try out each circuit, as it is designed. Then, having a circuit that works well with the lab power-supply, one can design a simpler power-supply that encorporates only the *needed* aspects of the laboratory unit.

Transformers can be inexpensive

The design of a power-supply almost invariably requires the use of a transformer. This is for two reasons: isolation from the powerline and obtaining a voltage different from the line voltage. The transformer is likely to be the highest priced single item in the power-supply, so resourcefulness in choosing a power transformer can really save money. Utilization of transformers that are available everywhere avoids buying the special types with their high price tags.

It is almost a truism that any electronic part we can use that is made for an automobile or a TV set is at the rock-bottom price for an item of its type. This is simply because the TV set and the auto represent "essentials" to the American public—everyone has at least one. Further, junk TV's and autos abound, and the price of a part is even lower after its carrier's demise. The large number of amateur transmitters powered by old TV set transformers is witness to this fact. One can find 6.3 Vac transformers in some of the junk TV sets; they were used to provide heater voltage to some or all of the tubes, while plate voltage was obtained using a semiconductor doubled directly from the power line. These 6.3 volt heater transformers can be useful in transistor power supplies used in the full wave bridge or conventional doubler circuits.

The early 12 V auto radios (with tubes that utilized +100 to +150 volts for plate voltage) are another source of a power transformer. If we simply connect the 117 Vac line across half the HV secondary (center-tap to one side), we will get about 12 Vac each side of the primary center-tap.

A surplus 24 V vibrator transformer represents an even more useful find. If the HV secondary is about 100 V each side of center-tap, such a transformer will put out about 24 V each side of the primary centertap. This higher voltage is more desirable in most instances.

Oh yes, don't forget to dig the silicon or germanium rectifiers out of the old TV set; they work just as well for low voltage transistor power supplies as they did supplying several hundred millamperes of B+ to the TV tubes. Although large by modern standards, Sarkes-Tarzian M500's and the like have worked very well in the low-voltage power supplies built by this author.

Rectifier circuits

In the area of 60-Hz single-phase rectifiers, there are five types of circuits that we'll be concerned with. These are "half-wave," "full-wave," "bridge," "conventional doubler," and "cascade doubler." These are illustrated in Fig. 1.

The "full-wave," "bridge," and "conventional doubler" circuits all charge their filter capacitors continuously throughout the 60-Hz period. This continuous type of rectifier output waveform contains no 60-Hz components, but only 120-Hz which is more easily filtered. The "half-wave" and "cascade doubler" circuits, having considerable 60-Hz energy, in their output waveforms, will prove to be the most difficult to filter and will have poorer regulation than the others. However, these relatively less desirable circuits are the only circuits one can use, if one side of the transformer secondary *must* be grounded for some reason.

With all these rectifier circuits at our disposal, *and* with the additional variation of being able to choose either "capacitor input" or "choke input" to our filter section, in full wave or bridge circuits, a given transformer can provide us a variety of dc output voltages.

To illustrate the variations of performance available with one transformer, a Triad F40X 24-volt center-tapped transformer was used in all the seven variations of Fig. 1. Then each circuit's voltage-current characteristic was tested; these are shown in Fig. 2. To make the circuits comparable, the "capacitor-bulk" (μF x voltage) was kept constant throughout. That is, the constant bulk concept would equate 500 μF at 50 V with 1000 μF at 25 V. Note that in Figs. 2f and 2g (the choke input cases) that a pilot lamp was added as a "bleeder resistor." This is in line with the advice given

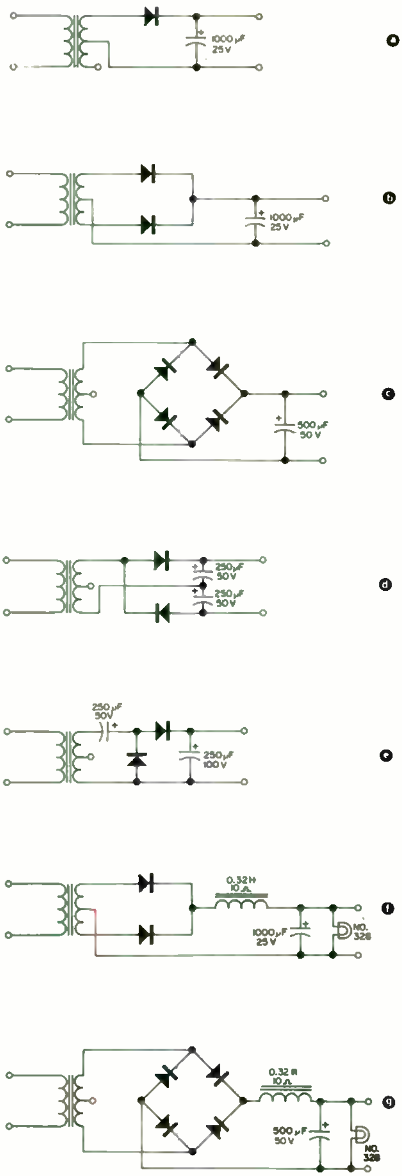


Fig. 1. The six basic types of rectifiers. a is a simple half-wave rectifier; b, full wave; c, bridge; d, conventional (full-wave) doubler; e, cascade (half-wave) doubler. The preceding are all capacitor input circuits. f is a full-wave rectifier with choke input and g a bridge rectifier with choke input. Performance of these rectifier circuits can be seen from curves in Fig. 2.

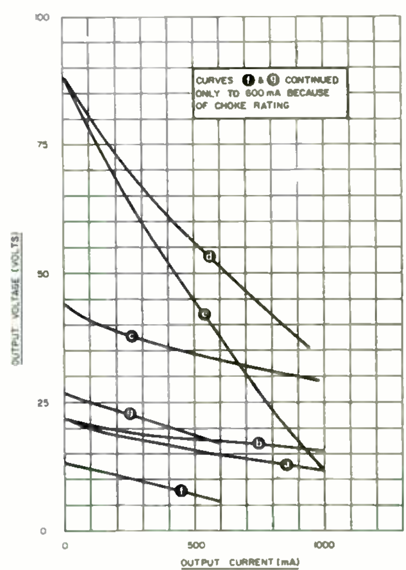


Fig. 2. Performance of the rectifier circuits in Fig. 1.

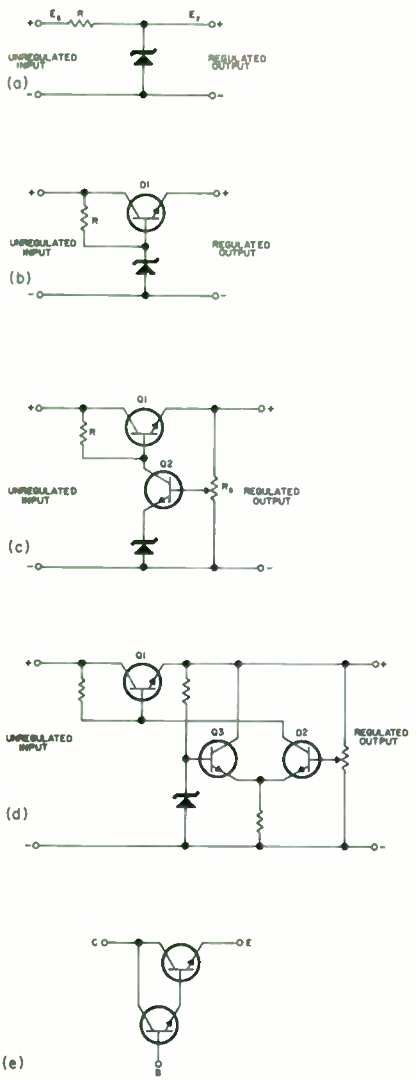


Fig. 3. Regulator circuits. a shows a zener regulator; b, an emitter follower regulator; c, an emitter follower with amplifier; d, emitter follower with full differential amplifier. The circuit in e is a Darlington Pair of transistors which acts as a single transistor with higher gain.

by Terman, where he states that a minimum load must be furnished a choke input system of 1130L. That is, for full-wave rectifiers operating from 60-Hz, R should be less than or equal to 1130L (where L is in henries). For our example, L was 0.32 henries so R should be 362 ohms or less. The number 327 bulb is approximately such a resistance at the voltages used, and provides a "free" pilot lamp for the supply.

Regulators

Having developed a number of rectifier-filter circuits, it is clear that they *all* have some lack of regulation. If we wish to provide our associated electronic circuit with a very constant voltage, some form of voltage regulation must be added. There are a number of methods of regulation, but the one this author has had the most experience with, is series regulation.

The series regulator can be easily developed, for understanding, as follows. Consider first of all a simple zener-regulator, as in Fig 3a. The output voltage characteristic is that of the zener diode, providing that the load current doesn't exceed $(Es - Ez)/R$. A capacitor can be added in shunt with the zener diode, to afford additional ripple filtering. Note that the addition of this capacitor *will* work in a zener diode circuit whereas it will often *not* work in a VR tube circuit,* since the zener diode has no region of negative resistance. One serious drawback of the zener regulator is that the zener voltage *does* change somewhat with zener current.

A better regulator is the "emitter-follower" type, wherein a transistor is used as a series resistance, as in Fig. 3b. The base of the transistor is held at a constant voltage by a zener diode which derives its current through R from the unregulated input. Since the current flowing into the base is Ic/h_{FE}, the zener current can be much less than in the simple zener regulator. If we make R small so that the zener current is large compared to the maximum base current the transistor will ever draw, the percent variation in zener voltage can be made fairly small. Again, a capacitor can be put in shunt with the zener, giving the ripple reduction of the well-known "capacitive multiplier."

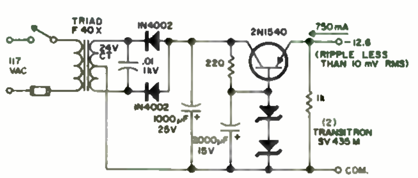


Fig. 4. Simple emitter follower regulated power supply for the heaters of five 12AX7's.

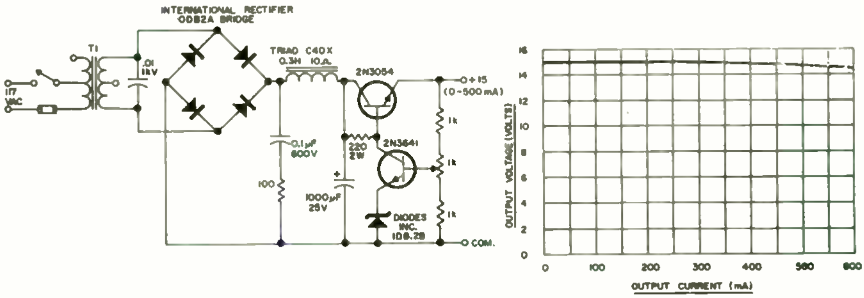


Fig. 5. Circuit and performance of a regulated power supply using the basic circuit of Fig. 3c. T1 is a surplus vibrator transformer marked 25.2 Vdc input and 135 Vdc, 118 mA output, vibrator frequency 115 Hz. Half of the high voltage winding is used in this circuit.

Perhaps the next step, in sophistication, is to add dc gain to our series-regulator as in **Fig. 3c.** In this circuit one is not tied down to the zener voltage to determine the output voltage. In fact, the output voltage is adjustable, over a few percent, with the output sample-pot, Rs. The circuit still suffers somewhat from variations in zener current with load current, however.

Finally, then, we add the full differential amplifier for our gain stage as in **Fig. 3d.** Now, since the zener diode derives its current from the output side of the regulator, the zener current is nearly constant and so is the zener voltage. As in **Fig. 3c,** the output is adjustable over a few percent.

In all the circuits 3b, 3c, and 3d, it is possible to replace the series transistor (Q_1) with a "Darlington Pair"; That is, replace one transistor with two as in **Fig. 3e.** This combination yields the equivalent of a transistor with the product of the h_{FE}'s of the two component transistors. This will allow us to use higher dc gain in the regulator system, affording better performance. The Darlington pair method is a way of obtaining a high-h_{FE} power-transistor for use as Q_1.

Examples of practical circuits

Having looked at a variety of series-regulators, a selection of real-life examples to illustrate them may be in order. These are presented in **Figs. 4, 5,** and **6.** The circuits are not so-designed to make value judgments between types, but are simply examples from the author's back file.

Another differential regulated supply is presented in **Fig. 7** to illustrate a trick in substituting the base-emitter junctions of inexpensive silicon transistors, for zener diodes. Note that in both **Figs. 6** and **7** a "pre-regulator" zener diode is used to supply half the differential amplifier with collector voltage.

One will also notice that in the examples of **Figs. 4, 5, 6** and **7** that there is added, across the transformer secondary, a 0.01 µF 1 kV disc ceramic capacitor. This is a line transient suppressor which is to "kill" any line spikes that may otherwise exceed the PIV of our rectifiers. Another type of transient suppressor is useful in choke-input filter systems; a 100-Ω resistor and 0.1 µF capacitor, (in series) are put ahead of the choke, as if to

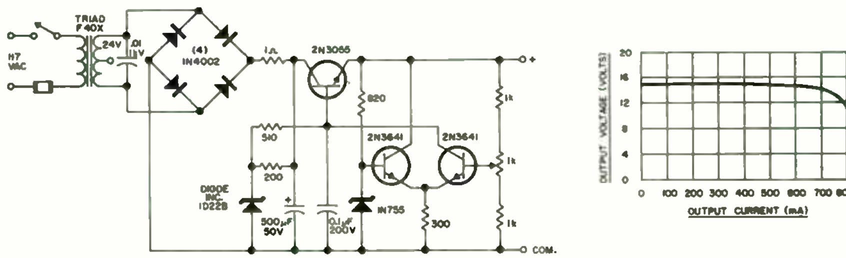


Fig. 6. Practical supply using a differential pair amplifier.

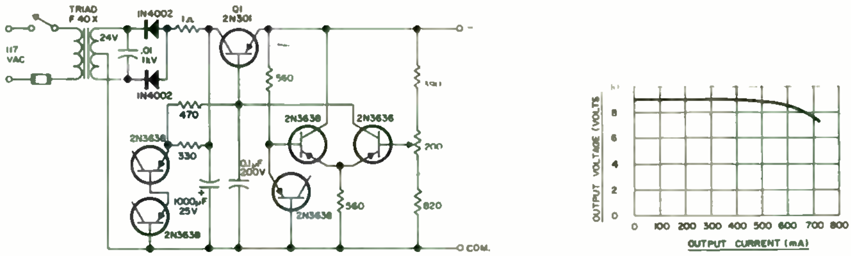


Fig. 7. Another differential pair regulated supply. Here the base-emitter junctions of inexpensive transistors are used as zener diodes.

provide capacitor input. This serves to dampen the voltage created by the choke field-collapse when the supply is turned off. This technique is used in **Fig. 5.**

Fig. 8 is a regulated power-supply for use with the inexpensive epoxy-encapulated Fairchild integrated circuits. These epoxy IC's are specified for +3.5 to −4.5 volts. This regulator utilizes one of the old 2.5 V filament transformers that was put in the junk-box when you replaced the 816's or 866A's with silicon HV rectifiers. The Fairchild µL923 (a J-K Flip Flop) takes about 20 mA, so this power supply will run up to 15 of them.

This design, a low-voltage one, brings out several of the worst points in series-regulator design. Since the difference between rectifier voltage (point A) and the regulator output voltage (point C) is only a few volts at most, the (small) voltage drop of the base-emitter junction becomes appreciable percentage-wise. Since germanium transistors have lower emitter-base drop, a germanium unit is used here; a high h_{FE} germanium type was chosen, the 2N2147.

Fig. 8 is similar to the regulator of **Fig.** 3b, except that R has been replaced by a #49 (2 V, 60 mA) pilot lamp. This trick was necessary to help hold the zener current more nearly constant. The bulb acts as a nonlinear resistor, increasing its resistance as the voltage across it increases. The 1N658 diode in series with the 1N4728A was simply used to "jack-up" the reference voltage about 0.7 V, over what the zener alone turned out to be. In this circuit, this forward-biased diode *does not* effect a temperature compensation.

Also, low-voltage brings out the worst in zener diodes. This is illustrated in **Fig. 9.** Notice that for zener diodes with voltages below about 6 volts, the "knees" do not have a sharp "break," but are quite rounded. This phenomenon is associated with the different mechanism of breakdown at the lower voltages. (To be really correct, one should only call regulator diodes that break down below about 6 volts "zeners" since this is in the zener region. Similarly regulator diodes that break down above 6 volts should be called "avalanche" diodes. However, the technology has come to call them *all* zeners.) The "soft-knee," as this rounding-over at breakdown is called, can be coped with by running more zener diode current, within the dissipation limits of the diode.

This "soft-knee" behavior of low-voltage zeners is not all as bad as it might be. It turns out that the "zener" and "avalanche" regions of breakdown have opposite temperature characteristics. So that at about 5 volts (see **Fig.** 9b) the two opposite temperature coefficients cancel, yielding a ready-made temperature-compensated reference diode.

At breakdown voltages above 5 or 6 volts, we are operating with a positive temperature coefficient, in the avalanche region. Since the

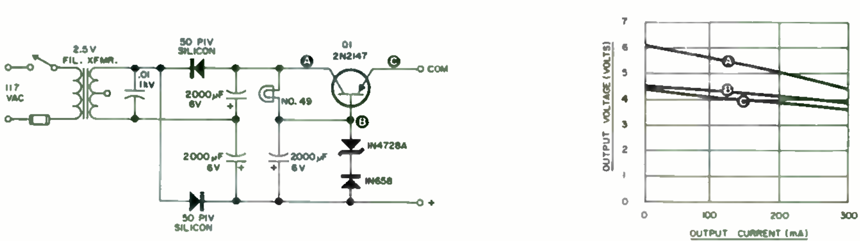


Fig. 8. Regulated power supply designed for use with inexpensive Fairchild integrated circuits. The letters on the graph refer to the voltages at the points shown on the schematic.

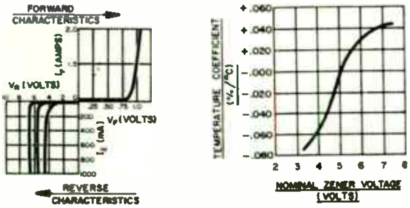


Fig. 9. a, on the left, gives the reverse characteristics of low-voltage zener diodes. b, on the right, shows how the temperature coefficient of zeners is lowest at about 5 volts.

temperature coefficient of an ordinary silicon diode or rectifier, when forward-biased, is negative; the arrangement of Fig. 10 can be

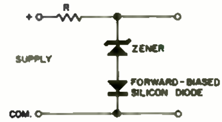


Fig. 10. Temperature compensation of a zener diode (for use with zener diodes of greater than 5 or 6 volts).

used to help temperature-compensate higher voltage breakdown diodes.

There is certainly a lot more that could be said on power-supplies and regulators, and the author knows he has only touched on a vast subject.

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CHAPTER 56

The Mini-Vidicon

The availability of low-cost high-quality semi-conductors has permitted the design and construction of a small, efficient, stable CCTV vidicon type camera at moderate cost.

The camera to be described has been used primarily for ATV, however, many other uses suggest themselves. Since this camera requires only 12 volts, it can be used in mobile or remote work where no 115 V 60 Hz power is available. This unit features field effect transistor imput for low-noise video, unijunction sweep oscillators for reliability and simplicity, sync inputs, 6 MHz band-width, built-in mechanical focus, and low input power requirements. The total power input is only 3 watts!

To aid in construction and operation of the camera, and to get an idea what makes it work, we will delve lightly into the circuit operation within the camera.

Circuit Description

Video amplifiers

The vidicon target current fluctuations, which comprise the video signal, are impressed across R 1, which serves as the vidicon load. Since the gate circuit of Q 1 is a very high impedance network, the gate receives a relatively high voltage signal (approximately 200 mV). Q 1 is connected as a source follower, which provides low impedance drive for the base of Q 2.

Due to the stray capacitance associated with the target of V 1 and input to Q 1, the video response will start to fall off around 15 KHz and continue to drop at an rc rate (6 dB/octave). Q 2 has a frequency response which is complimentary to the input response, thereby giving flat over-all video response. This is accomplished by controlled emitter degeneration at low frequencies. C 3 sets the breakpoint of the peaking stage (Q 2).

Q 3 is an emitter follower, which prevents the output stages from loading the peaking stage. R 6 adjusts overall video gain. Q 4 is the keyed clamp. When Q 4 receives a pulse from the vidicon blanking circuit, Q 4 conducts, clamping the input of Q 5 to a level determined primarily by the setting of R 27. R 27 therefore controls the blanking level.

Q 5, Q 6, and Q 7 comprise an output feedback amplifier with a gain of approximately 35 and low output impedance to drive the 75 ohm video line.

Vertical deflection

The vertical sweep system utilizes a unijunction transistor Q 8, as the frequency generator. This type of device is very useful as a relaxation oscillator due to the fact that when the emitter voltage rises to a fixed percentage of the voltage between the two bases, the unit will suddenly conduct. All that is necessary to construct a relaxation oscillator is to provide an rc charging circuit for the emitter. In our camera, this circuit consists of R 14, R 15, C11 and C 12.

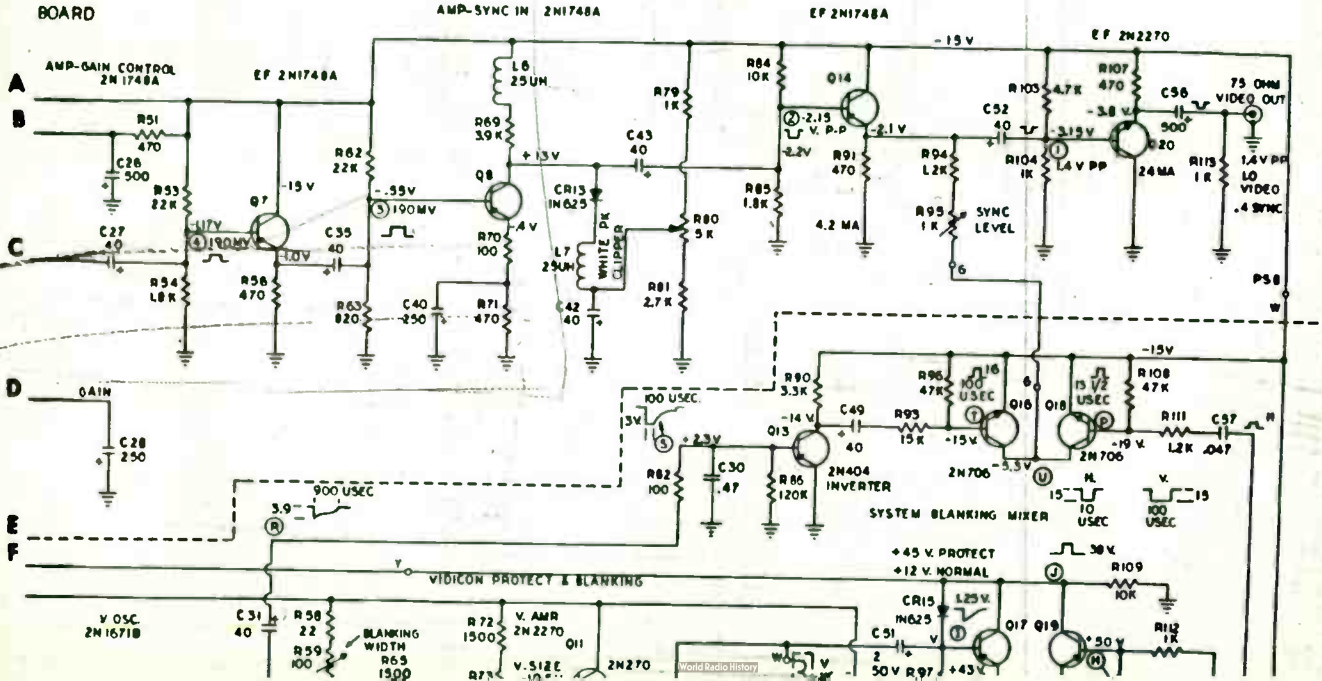
C 11 and C 12 charge toward the supply voltage and when the unijunction firing voltage is reached, the unijunction conducts, discharging the capacitors. This generates a sawtooth at the emitter, a positive pulse at B 1, and a negative pulse at B 2. As we will see later, these waveforms will be very useful throughout the camera.

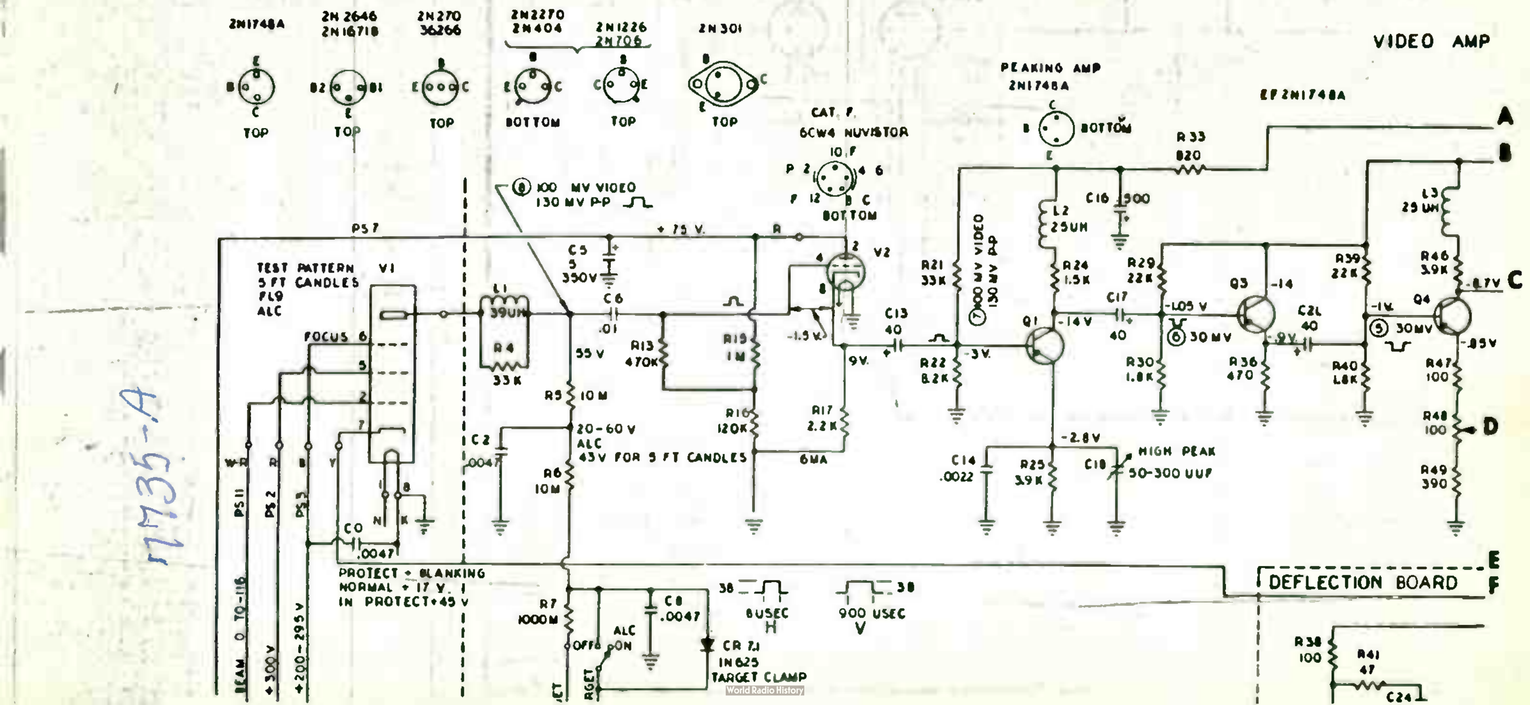
R 14 controls the charging current to the oscillator capacitors and therefore controls the frequency.

The saw waveform from Q 8 emitter is coupled to the base of Q 9, which amplifies and inverts the signal. A portion of the signal is taken from the emitter of Q 9

The completed camera

BOARD


AMP-GAIN CONTROL
2N1748A
EF 2N1748A
AMP-SYNC IN 2N1748A
EF 2N1748A
EF 2N2270
-15 V
75 OHM
VIDEO OUT
A
B
C
D
E
F
GAIN
R51 470
C26 500
R53 22K
R54 1.8K
C27 40
Q7
R56 470
C35 40
R62 22K
R63 820
-.55V
190MV
Q8
L6 25UH
R69 3.9K
+13V
R70 100
.4V
C40 250
R71 470
CR13 1N625
WHITE PK CLIPPER
L7 25UH
C42 40
C43 40
R79 1K
R80 5K
R81 2.7K
R84 10K
R85 1.8K
Q14
-2.15 V. P-P
-2.1 V
-2.2V
R91 470
4.2 MA
R94 1.2K
R95 1K
SYNC LEVEL
C52 40
-3.15V
1.4 V PP
R103 4.7K
R104 1K
R107 470
-3.8 V
Q20
24MA
C56 500
R113 1K
1.4 V PP
LO VIDEO
.4 SYNC
PS8
C28 250
100 USEC
3V
R90 3.3K
-14 V
C49 40
R93 15K
Q13
+2.3V
R82 100
C30 .47
R86 120K
2N404
INVERTER
R96 47K
100 USEC
Q16
Q18
15 V/2 USEC
R108 47K
-15V
2N706
-5.3V
-19 V.
R111 1.2K
C57 .047
SYSTEM BLANKING MIXER
H.
V.
10 USEC
100 USEC
900 USEC
3.9
+45 V. PROTECT
+12 V. NORMAL
30 V.
R109 10K
VIDICON PROTECT & BLANKING
V OSC.
2N1671B
C31 40
R58 22
R59 100
BLANKING WIDTH
R65 1500
R72 1500
V. AMP
2N2270
Q11
V. SIZE
2N270
CR15
1N625
1.25V
C51
2
50 V
R97
+43V
Q17
Q19
+50 V
R112 1K



VIDEO AMP
2N1748A
2N2646
2N1671B
2N270
36266
2N2270
2N404
2N1226
2N706
2N301
TOP
BOTTOM
PEAKING AMP
2N1748A
EF 2N1748A
CAT. F
6CW4 NUVISTOR
TEST PATTERN
5 FT CANDLES
FL9
ALC
FOCUS
100 MV VIDEO
130 MV P-P
+75 V.
PS7
V1
V2
L1
39UH
R4
33 K
C5
5
350V
C6
.01
55 V
R13
470K
R19
1M
R5
10 M
20-60 V
ALC
43V FOR 5 FT CANDLES
R6
10M
C2
.0047
R16
120K
R17
2.2 K
6MA
C13
40
9V.
R21
33K
R22
8.2K
700 MV VIDEO
1.30 MV P-P
Q1
-3V.
-14 V
R24
1.5K
L2
25UH
C16
500
R33
820
C17
40
R29
22K
R30
1.8K
-1.05 V
30 MV
Q3
-14
R36
470
C2L
40
-9V
R39
22K
R40
1.8K
-1V.
30MV
Q4
L3
25 UH
R46
3.9K
-.87V
-.85V
R47
100
R48
100
R49
390
A
B
C
D
E
F
-2.8V
C14
.0022
R25
3.9K
C18
HIGH PEAK
50-300 UUF
CO
.0047
PROTECT + BLANKING
NORMAL + 17 V.
IN PROTECT +45 V
R7
1000M
OFF
ON
ALC
C8
.0047
CR 7.1
IN625
TARGET CLAMP
38
8USEC
H
900 USEC
V
DEFLECTION BOARD
R38
100
R41
47
C24
BEAM 0 TO -115
+300 V
+200-295 V
PS11
PS2
PS3
W-R
R
B
Y

R 27. R 27 therefore
level.
7 comprise an output
with a gain of approxi-
output inpedance to
to line.

p system utilizes a uni-
S, as the frequency
of device is very use-
oscillator due to the
emitter voltage rises
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the unit will suddenly
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or is to provide an rc
the emitter. In our cam-
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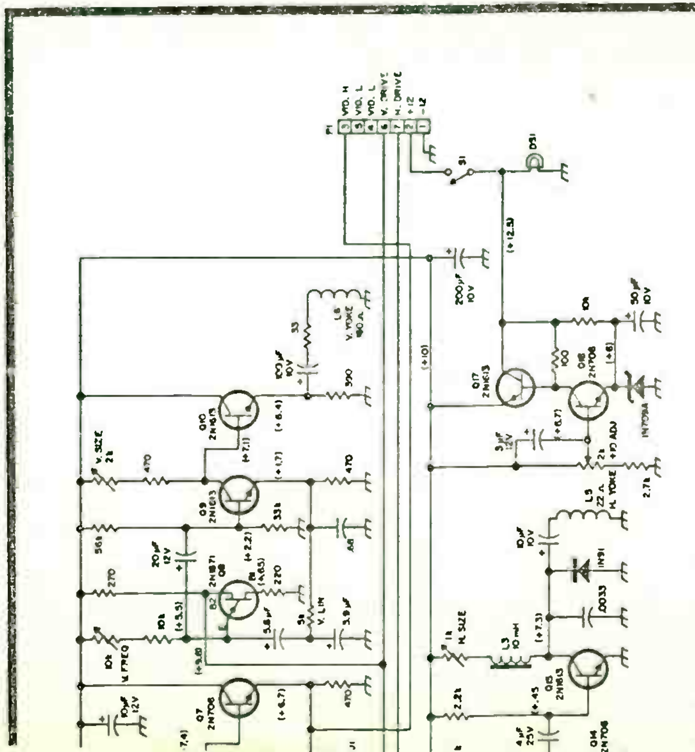
the unijunction firing



NOTE
VOLTAGE READINGS TAKEN WITH SIMPSON 260 VOM

l schematic for the Vidicon Minicamera.

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signal. A portion of the
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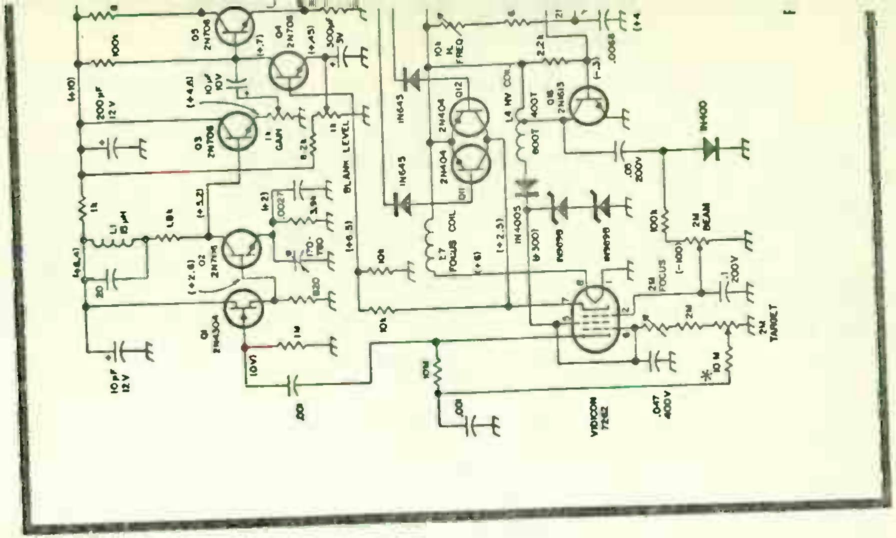


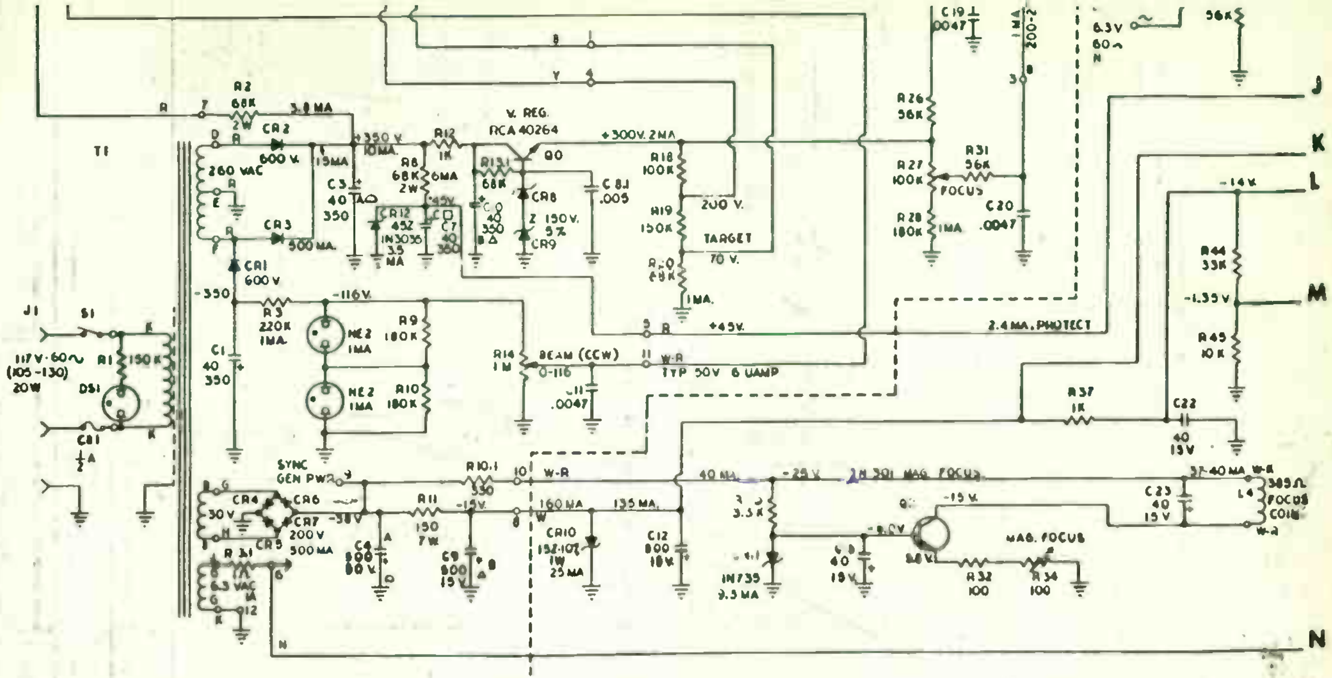
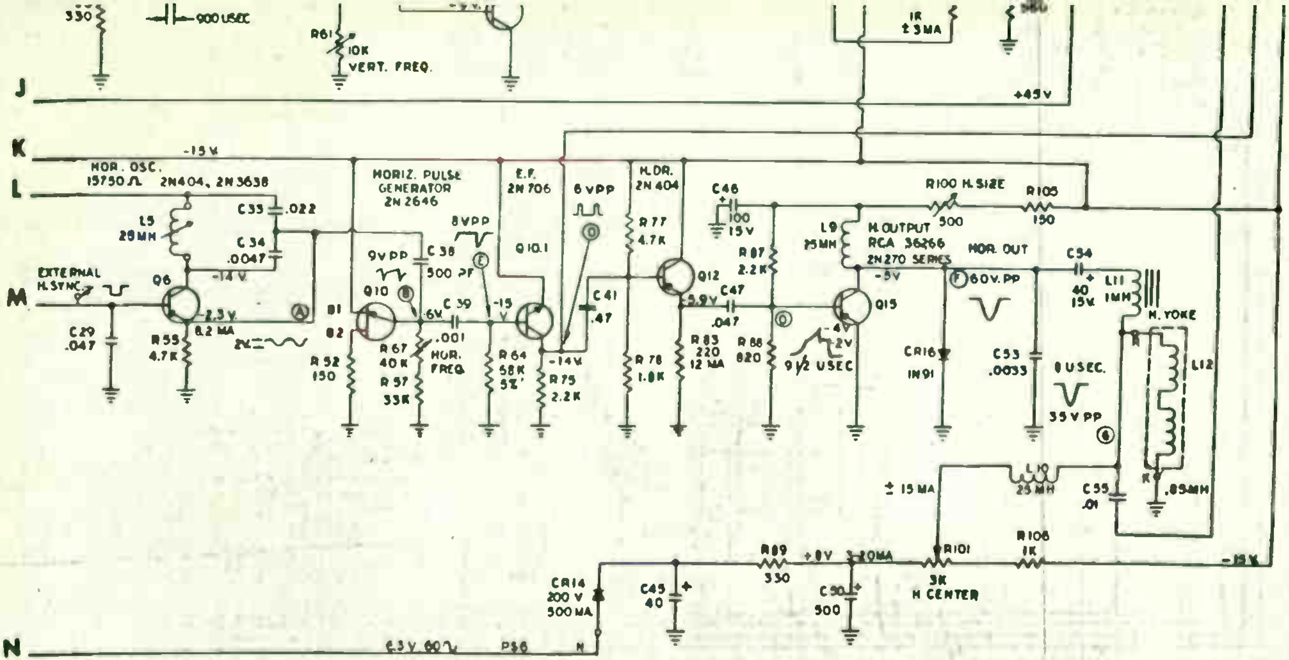


Fig. A38. Packard Bell



Model 920 camera.

Courtesy Packard Bell Electronics

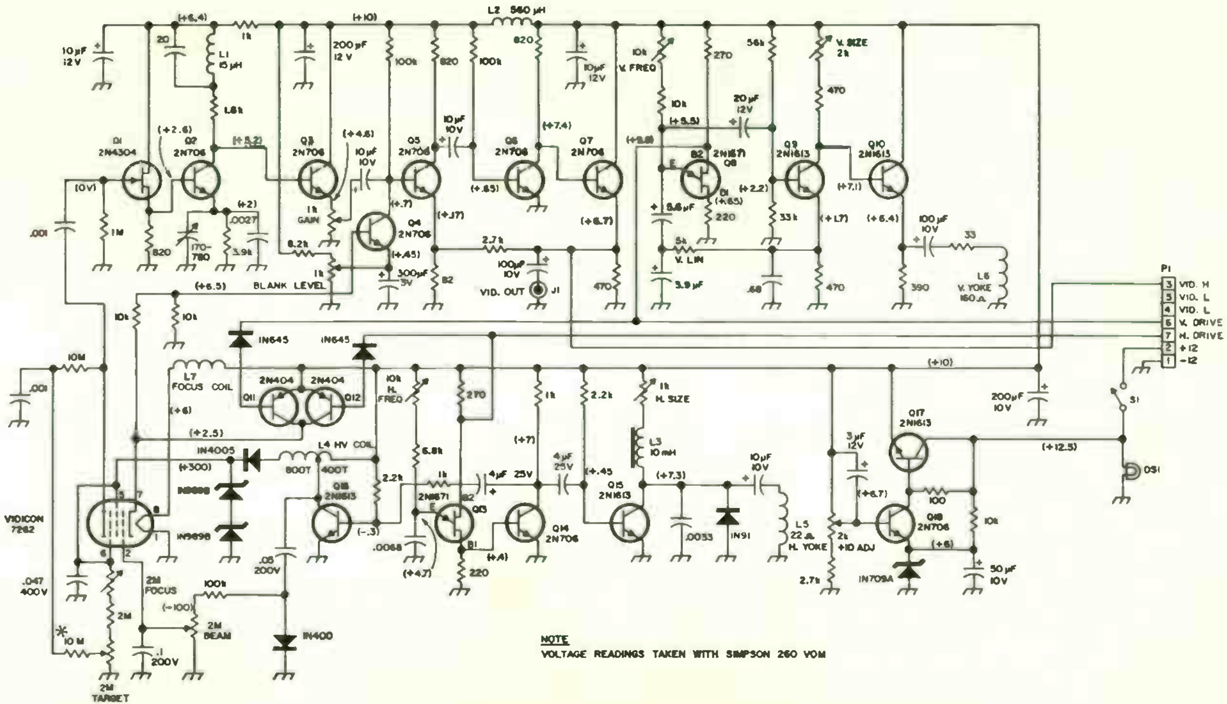


Fig. 1. Electrical schematic for the Vidicon Minicamera.

and fed back to the oscillator through the linearity pot, R 16, for improved vertical linearity. R 21 controls the gain of this stage and so controls vertical size.

The collector of Q 9 is direct coupled to emitter follower Q 10, which provides low impedance drive to the vertical deflection coils through a network consisting of C 14 and R 25.

Horizontal deflection

The horizontal deflection chain starts with an oscillator similar to the vertical oscillator, with R 28, R 29, and C 15 as the frequency determining elements. R 28 controls horizontal frequency in the same way as R 14 controls vertical frequency. The positive pulse appearing at B 1 of Q 13 is direct coupled to pulse amplifier Q 14. The inverted pulse is then capacitively coupled to the horizontal deflection amplifier, Q 15.

Due to the predominance of inductive reactance in the horizontal deflection coils at the line rate, the waveform supplied to the yoke must be a pulse of voltage to obtain a sawtooth of current in the yoke. This is why the horizontal deflection chain is a pulse type amplifier.

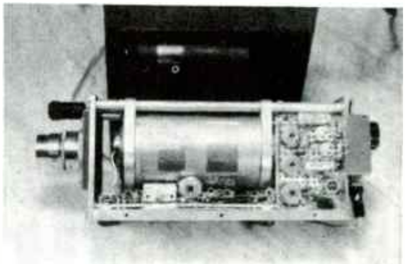
When Q 15 is turned off by the pulse from Q 14, the stored energy in L 3 is released, providing a large spike of voltage to drive the yoke. R 34 controls the size of this spike, and hence the horizontal size. C 17 tunes the output for maximum efficiency, and D 3 serves as the damper diode.

High voltage power supply

Since the vidicon requires operational voltages in the +300 and −100 volt regions, and only 12 volt input power is available, some type of dc-dc converter is required. The converter used is a simple stored-energy type which is driven from the horizontal deflection circuit. The power supply receives a pulse from the horizontal pulse amplifier, which turns Q 16 off. When Q 16 is cut off, a very large spike appears on the collector due to the release of energy which is stored in L 4.

The pulse at the collector of Q 16 is rectified by D 8 and filtered by C 24 to provide the −100 V for vidicon beam control. This same pulse is multiplied by the ratios on L 4, and is then rectified by D 7 to provide +300 for anode and mesh power within the vidicon.

Diodes D 5 and D 6 are 150 V zeners which regulate the output of the power supply to 300 volts.



Looking inside the Vidicon Minicamera.

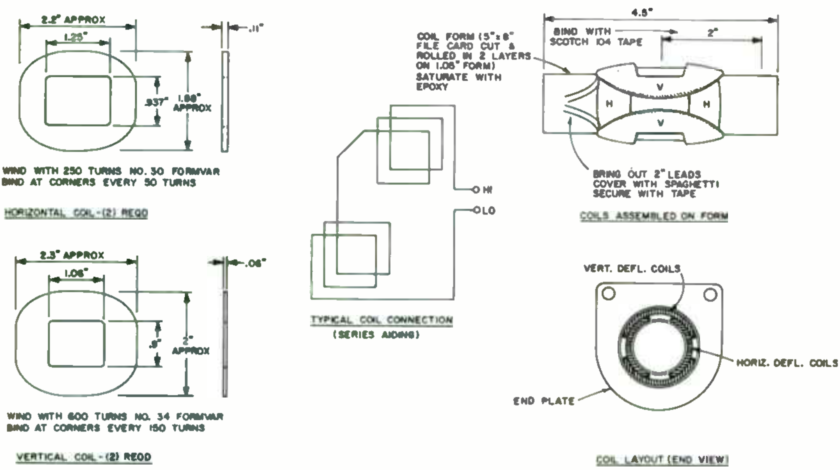


Fig. 2. Deflection coil data.

Vidicon circuits

The only external controls on the camera are the beam, focus, and target controls. These 3 controls adjust the operating parameters of the vidicon. The beam control sets the grid bias on the electron gun in the vidicon. Optimum setting of this control will allow just enough beam current to land on the photosensitive surface to replace the highlight current lost to the load resistor and target supply. Keep in mind that the beam control has an effect on vidicon life, and should always be kept at the minimum beam (maximum negative voltage on G 1) position when any abnormal condition could occur in the vidicon parameters. This would include camera turn-on, turn-off and especially during any kind of sweep failure, as the photosensitive surface is easily damaged. During normal operation, the beam should be turned up just enough to discharge the highlight whites in the picture.

While the major part of the beam focussing is accomplished by the electromagnetic focus coil around the vidicon, a small vernier focus is available electrostatically by varying the voltage supplied to the vidicon focus anode. This voltage is supplied through R 39, the focus pot on the camera, which is adjusted for maximum picture detail.

The vidicon target requires a voltage of from +5 to +65 volts depending on tube characteristics, light level, and ambient temperature. This voltage is supplied through isolating resistor R 43 from R 41, the target control on the rear of the camera.

Since the vidicon is a storage type tube, the beam must be prevented from landing on the photosensitive surface during retrace time. This is accomplished by the blankers Q 11 and Q 12.

When vertical blanker Q 11 receives the negative pulse from the vertical oscillator during retrace, this stage conducts, pulling the cathode of V 1 up to the +10 V buss, thereby cutting V 1 off.

The horizontal blanker, Q 12, works in the same manner. It should be noted that either Q 11 or Q 12 can control beam cut off.

Diodes D 1 and D 2 prevent blanker conduction due to the small drop across R 17 and R 30 during scanning time.

Low voltage regulator

Since the camera is designed to operate from storage type batteries whose potentials may vary widely during charge and discharge cycles, and camera stability is considered important, a regulator circuit is included to hold the +10 buss reasonably constant over input fluctuations from approximately 10.8 to 14 volts.

Q 18 is connected as a difference amplifier which compares a portion of the 10 volt regulator output to a fixed reference, supplied by D 4. This amplifier controls Q 17, which is operated as an emitter follower pass transistor, R 35 controls the percentage of feedback and therefore, the potential on the + 10 V buss.

Power for the camera

In as much as the camera is designed for extreme portability, storage batteries are a natural choice for power.

Two types of cells have been found satisfactory, rechargeable alkaline, and nickle-cadmium, 3 Eveready #563 connected in series will do nicely, at moderate cost. 10 surplus nicad cells have been used very successfully.

Deflection and focus assembly construction

The deflection and focus assembly is designed to operate with low power input, and allow complete compatibility with the camera design philosophy, both mechanically and electrically.

Construction details for the deflection assembly are outlined in **Fig. 2**. The focus coil consists of 2600 turns of #25 Formvar wound on the focus coil core shown in **Fig. 3**. After the focus coil is wound, insulate with Scotch # 104 tape, and cover with a double thickness of conetic foil.

Camera construction

The electronic circuitry for the camera is divided into 5 basic sections, which are built up on the No. 85G24EP Vectorboard. Since the component density is high within the camera, neat and precise layout is essential.

The complete camera system. The black box is the battery pack using NICAD cells. The two rectangular pieces are coil forms for winding the deflection coils.

Smaller and cheaper equivalent semiconductors have become available since this camera was built, and some consideration might be given to these, but keep in mind that the camera has been well proven in its present form.

The cases of most transistors used are connected to the collector, so be sure no shorts exist between cases and nearby components.

As the boards are completed, they should be tested in the manner outlined in the test procedure.

After the boards are tested, camera assembly can be started. The use of the .250 inch baseplate results in a very solid unit. The .250 diameter rods at the top serve two purposes, providing extra rigidity, and also providing a stable track for the moveable vidicon assembly to slide upon.

The board containing video, blanking, and sync circuitry is mounted on the baseplate below the focus coil area.

L 1	15 Microhenry	RF choke, Miller	9310-40
L 2	560 Microhenry	RF choke, Miller	9350-26
L 3	10 Millihenry	RF choke, National	R-40-10
L 4	HV XFMR	1200 T #34 on ¼" iron-slug form. tap at 400 T.	

[illegible]

One sheet 16x8x.004" conetic foil, conetic AA, made by Magnetic Shield Div., Perfection Mica Co.

Lens, Cinepar 25.4MM (1 in.) FL F/2.5 Edmund Scientific, # 40, 724

Lens mount, #SCM, available from ATV Research, PO Box 396, So. Sioux City, Neb.

First, an insulating sheet is cut from a file card to fit between the video board and the baseplate. The video board is then mounted to the baseplate, using spacers drilled and cut from Vectorboard, and 4-40 screws through the drilled holes in the video board corners. It will be necessary to clip all T 28 terminals at a point just below the board to clear the baseplate.

The vertical and horizontal deflection boards can now be mounted using cut-down grounding lugs soldered to the lower corner terminals, and 4-40 screws.

LV regulator and HV power supply installation can now be accomplished. These boards slide between the vertical and horizontal deflection boards in the positions indicated, and are supported by the projecting ends of the pins in these boards. Any interfering terminal ends should be clipped. All possible board interconnections should now be made.

The vidicon socket can now be wired. This is a two-piece assembly, and the rear part should be removed and discarded to provide clearance. The socket should be wired to the previously installed boards with lead lengths appropriate to allow freedom of motions from the rear cover to approximately .750 inch forward of this position. C 22 is wired directly from pin 6 to the ground end of C 29 on the video board.

All leads from the boards and vidicon socket to the control panel should now be pigtailed out and the rear endplate installed with all pigtails brought through the large hole.

All connections should now be made to the control panel, leaving enough slack for accessibility.

P 1 is made by modifying a cinch 8S M socket. This is done by soldering leads into the socket to make a mating plug. Scrap transistor leads will do nicely. First tin in the leads, force them down into the plug,

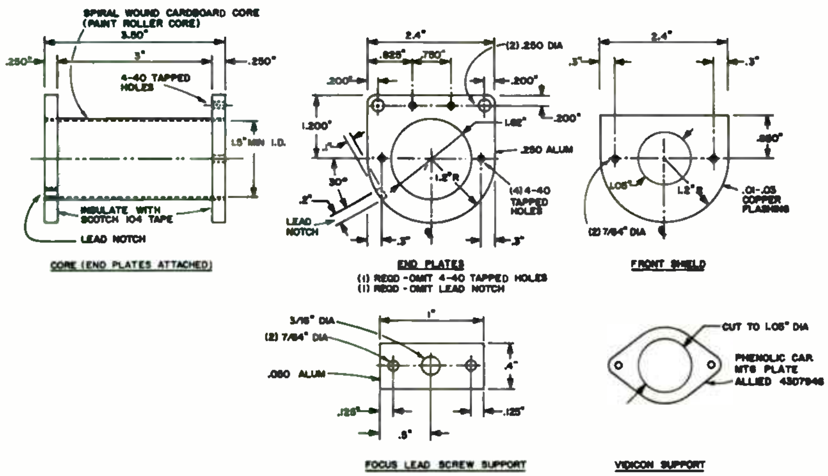


Fig. 3. Detail for the focus coil core.

and heat the plug pins from the rear to fuse the leads into the socket. Cut the leads off evenly, add a piece of # 16 wire to the center of the plug, and P 1 is complete. P 1 is now cemented to the control panel.

Now install the front endplate, vidicon, focus, and deflection assembly, and slide rods. The front endplate should be painted flat black before assembly.

Check and finish all interconnections, and install the lens mount and lens.

The lens mount shown in the picture was turned from .250 aluminum stock on a thread-cutting lathe. An easier solution to the lens mount problem is to use the unit shown on the parts lists, with the front-plate drilled and tapped to match.

The camera is now ready for the tests described in the system test procedure.

Test Procedure

Board checkout

To avoid vidicon damage, the scanning circuits must be thoroughly checked out before vidicon installation.

To make checkout less complex, each board should be tested as soon as it is completed. Assemble the LV reg. board first, so it can be used to power the other modules.

To test the LV reg. board, first parallel C 19 with a 50 ohm 5 W load resistor. Next apply + 12 V from battery to Q 17 collector through a 1 amp fast-acting fuse. Battery minus goes to the minus side of C 21.

Now, with a meter connected across C 19, R 35 should vary the voltage from less than 9 V to more than 10.75 volts. Leave final setting at +10 V. Remove load resistor.

After constructing the vertical deflection board, check with ohmmeter to be sure no shorts exist on the +12 volt line. After this test, apply +10 V power from LV reg. board, and apply scope probe to Q 8 emitter. Adjust R 14 and R 16 to obtain trace like Fig. 4.

Temporarily connect V yoke and set R 21 and reset R 16 and R 14 for trace (Fig. 4) at junction R 25 and yoke lead.

When the horizontal deflection board is completed, check for shorts, apply 10 V, and connect scope to Q 13 emitter. Adjust R 28 for Fig. 4 trace. Temporarily connect H yoke, L 5, and set R 34 for trace similar to Fig. 4 at yoke lead.

To check out the HV supply it will be necessary to have the horizontal deflection circuit operating. Check for shorts, temporarily connect a wire from C 27 to Q 14 collector, and fire up the horizontal deflection circuit.

Apply 10 V to the HV board and check for a pulse similar to Fig. 4 at Q 16 base. Unless a low-capacitance probe is available, the scope will load any readings taken on the HV transformer, so it probably is best to take output at the dc terminals with a VTVM. Check for +300 ± 30 V and −100 + 40 −10 V with VTVM.

Video board tests should now be held to voltage checks, with signal readings taken during camera system checkout.

Camera system checkout and adjustment

After all board inter-connections have been thoroughly checked, test across C 19 for shorts. The resistance here should be approximately 20 ohms, using a Simpson 260 VOM with the common lead to ground. Make sure the vidicon is disconnected. Connect the 12 V source to P 1, turn power on with S 1, and adjust R 35 for + 10 V at Q 17 emitter.

Now test all points listed in the board checkout procedure for similar voltages and scope waveforms.

We are now ready to connect the video output to a monitor, either directly or through a pretested "jeep", (A unit used to convert video to one of the TV channels).

A rough check of video board operation can be made by holding a finger near the vidicon target lead. Now, vary R 6 to obtain noise on the monitor. When the above tests are satisfactorily completed, turn the power off and plug in the vidicon. Preset the beam

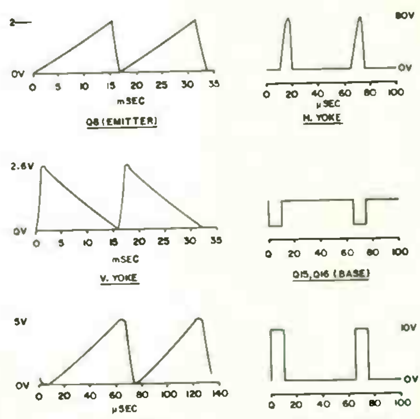


Fig. 4. Scope waveforms used in making the various adjustments in test procedure.

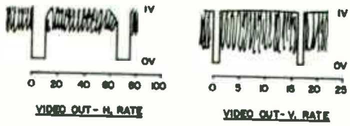


Fig. 5. Horizontal and vertical output forms.

and target controls to minimum and the focus control to midrange. Now reapply power, adjust vertical and horizontal frequency controls to sync the raster, and advance the target and beam controls until the monitor flashes with the beam control.

Set up a test pattern, open the lens iris, and it should be possible to see the first glimmerings of a picture on the monitor screen.

Set up the test pattern 18 inches from the lens, and a 60 watt bulb behind the camera. Put a dark cloth over the top of the camera to prevent stray light from falling on the vidicon faceplate.

Now adjust mechanical (optical) focus, electrical focus, beam, and target controls to obtain the best picture. The image will probably be distorted with some streaking at this time.

We are now ready to make final adjustments. Connect the scope vertical channel to the junction of R 25 and the yoke lead, and supply a 60 Hz signal to the horizontal channel.

Adjust R 14, R 16, and R 21 to obtain full vertical height and good linearity on the monitor, with a single, nearly stationary pattern showing on the scope.

Now move the vertical scope input lead to the junction of C 18 and the yoke. A stable 15.75 kHz reference for the scope horizontal can be provided by a well-insulated lead draped over the horizontal output tube of a nearby TV receiver synced to a local station.

Adjust R 28 and R 34 for a full picture on monitor, and the single, stationary pattern on the scope.

Now connect the scope to the video line, and adjust R 6 and R 27 for a waveform as shown in **Fig. 5.** Rotate the deflection yoke inside the focus coil to obtain the proper image orientation. Reverse the yoke leads to correct for image reversal. Set C 3 for minimum streaking following the heavy horizontal bars on the test pattern. Readjust R 6 for **Fig. 5.**

This completes camera checkout.

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Appendix

The following tables list semiconductors recommended for use in amateur equipment. The list is obviously not complete, but the devices listed are inexpensive and readily available from Newark Electronics as well as from many other leading industrial distributors. Prices given are approximate current single-quantity prices from most of these distributors, but they are subject to change. All of these devices are made by Motorola, the most widely available semiconductor line, but the recommendations are the editor's. All devices are silicon except two power amplifiers.

DIODES

Rectifiers

	PRV	RMS	Price	Case
1N4001	50 V	35	$.45	59
1N4002	100 V	70	.49	59
1N4004	400 V	280	.67	59
1N4007	1000 V	700	.90	59

These tiny Surmetic* silicon rectifiers are excellent for applications requiring up to 1 ampere or so. They have been standardized for general use by many hams (and manufacturers). Their cases are insulated plastic about the size of quarter-watt resistors. (*Trademark of Motorola Inc.)

Dual silicon detector diodes

	VR	IF	Price	
MSD6102	70 V	200 ma	common cathode	$.75 TO-92 (F)
MSD6150	70 V	200 ma	common anode	.82 TO-92 (G)

	VR	IF		Price	
MSD6102	70 V	200 ma	common cathode	$.75	TO-92 (F)
MSD6150	70 V	200 ma	common anode	.82	TO-92 (G)

These epitaxial dual diodes can be used conveniently in many detectors, balanced modulators and other applications. They are packaged in the Unibloc* plastic package used for most of the transistors in this listing.
(*Trademark of Motorola Inc.)

Zener Diodes

	V_Z	Max Z_{ZT}	Max DC current	Price 1-99	Lower Cost Version
1N4729	3.6	10 Ω	252	$1.26	MZ1000-2
1N4736	6.8	3.5Ω	133	1.26	MZ1000-9
1N4739	9.1	5.0Ω	100	1.26	MZ1000-12
1N4742	12	9 Ω	76	1.26	MZ1000-15
1N4749	24	25 Ω	38	1.26	MZ1000-22

These 1-watt ±10% Surmetic zener diodes are packaged in the same subminiature plastic case used for the rectifiers listed above. This list is a sampling of a line running from 3.3 to 200 volts. A suggestion—mount the diodes with a short cathode lead for best heat dissipation. Heat is conducted primarily thru this lead. DO-7 package. The lower cost versions numbered MZ1000-1 to MZ1000-37 are similar, but cost only $.67.

THYRISTORS AND TRIGGERS

Silicon Controlled Rectifiers (SCR)

	Reverse Blocking Voltage	RMS Current	(Max) Gate Trigger Current	(Max) Hold Current	Price	Case
2N4168	50 V	0.8 A	20 ma	25 ma	$1.75	86
2N4170	200 V	0.8 A	20 ma	25 ma	2.10	86
2N4172	400 V	0.8 A	20 ma	25 ma	2.95	86
2N5060	30 V	0.8 A	0.2ma	5 ma	.77	TO-92 (A)

Triac. (Bi-directional Thyristor)

	Peak Blocking Voltage	RMS Current	Gate Trigger Current (Max)	Holding Current (Max)	Price	Case
MAC2-2	50 V	8 A	30 ma	30 ma	$2.55	86
MAC2-4	200 V	8 A	30 ma	30 ma	3.45	86

3-Layer Diode (Bilateral Trigger Diode)

	Breakover Voltage	Switchback Voltage	Price	
MPT20	20 V	7 V (Typical)	$.67	TO-92 (H)

Three-layer diodes act much like unidirectional low-voltage, reliable neon bulbs. They simplify many SCR triggering circuits.

Unijunction Transistor

	η	Valley point current	Power Dissipation	Interbase Voltage	Price
2N4870	.56-.75	2ma min.	300 mw	35	$.75 TO-92 (E)

This inexpensive Unibloc, plastic-encapsulated unijunction is useful in SCR triggering and in any other circuits.

CONVENTIONAL (BIPOLAR) TRANSISTORS

Small Signal

	V_{CB}	Pd (mw)	F_T	h_{FE}	Polarity	Price	Case
Cheap replacements for popular devices:							
MPS404	25	310	–	30-400	PNP	$ .42	TO-92 (B)
MPS706	15	310	–	20-	NPN	.45	TO-92 (B)
MPS918	30	200	600	20-	NPN	1.30	TO-92 (B)
MPS3638	25	310	100	20-	PNP	.46	TO-92 (B)
HF, VHF, RF, IF amplifier and mixer:							
MPS6570	20	310		20-200	NPN	.96	TO-92 (D)
HF, VHF, UHF oscillator:							
MPS6507	30	310	700	–	NPN	1.12	TO-92 (B)

General Purpose and Audio Amplifier:

	V_{CB}	Pd (mw)	F_T	h_{FE}	Polarity	Price	Case
2N4123	40	310	250	50-150	NPN	$.52	TO-92 (B)
2N4125	30	310	200	50-150	PNP	.52	TO-92 (B)
MPS3394	25	310	–	55-110	NPN	.29	TO-92 (B)
MPSA10	40V_{CEO}	210	20	40-400	NPN	.29	TO-92 (B)
Nixie tube driver (high voltage, low current):							
2N4409,10	80, 120	310	60 mhz	66-400	NPN	.67, .96	TO-92 (B)

FIELD EFFECT TRANSISTORS

Field effect transistors (FETs) are high impedance semiconductors combining the best features of tubes and transistors. FETs are recommended for all applications where high impedance and/or resistance to crossmodulation and overloading is needed, i.e., all receiver rf amplifiers and mixers. FETs are also excellent as oscillators.

Type	V_{OS}	Polarity & Type	Y_{FS}	Typical Input Cap.	Typical Feedback Cap.	Price	Case
General RF amplification and mixing through VHF:							
MPF102	25	N-JFET	1600	7	3	$ .90	TO-92 (c)
General LF amplification and switching:							
2N5458 (MPF.104)	25	N-JFET	4000	7	3	.82	TO-92 (c)
2N5461 (MPF152)	40	P-JFET	3500	7	2	.75	TO-92 (c)
UHF and VHF amplification (Plastic 2N4416):							
2N5486 MPF107)	25	N-JFET		5	1	1.20	TO-92 (c)
100 mhz: min. gain 18 db max. NF 2 db; 400 mhz: min. gain 10 db; max. NF 4 db							
Dual-gate VHF and HF amplification: Excellent mixer and for AGC control.							
MFE3006	25	N-MOSFET	10-20K	7	0.02	1.50	TO-72
100 mhz typ. gain 24 db, typ. NF 3 db							

POWER AMPLIFIERS

Complementary Audio Power Amplifiers:

	V_{CE}	Pd-W	F_T	h_{FE}	Polarity	Price	Case
MPSU02	40	6	50	150	NPN	$ .99	152
MPSU52	40	6	150	50	PNP	1.15	152
2N4919	60	30	3	20-100	PNP	1.90	77
2N4922	60	30	3	20-100	NPN	1.48	77
2N4919 and 2N4922 excellent for 15 W output							
High-voltage (line-operated) audio amplifier, etc:							
MJE340	300	20.8	10	30-240	NPN	1.06	77
MPSU04	180	5	100	40	NPN	1.62	152

General Purpose:

	V_{CE}	Pd-W	F_T	h_{FE}	Polarity	Price	Case
2N3055	60	115	2	20-70	NPN	$2.05*	TO-3
2N3611	30	77	.7	35-70	Ge PNP	1.15	TO-3
2N1553	30	106	.7	30-60	Ge PNP	1.85	TO-36
MJE3055	60	90	2	20-70	NPN	1.41	90

TRANSMITTER RF

	V_{CEO}	F_T	Output	Price	Case
2N3553	65	500	2.5 W out at 175 mhz	$4.37	TO-39
2N3866	55	800	1 W out at 400 mhz	$2.97	TO-39
2N4072	20	550	250 in W at 175 mhz	$2.25	TO-18
2N4427	20	500	1 W out at 175 mhz	$2.25	TO-39

LINEAR INTEGRATED CIRCUITS

	Price	Case
MC1439 G open loop gain 100,000	$3.60	71
MC1709 CG open loop gain 45,000	3.90	71

The MC1439 G is a super operational amplifier (op amp); the MC1709 CG is the least expensive of the popular 709-type op amps. Both are very useful in amplifying, filtering and many other circuits.

MC1550 G 26 db gain at 60 mhz. 5 db NF at 60 mhz 1.50

The MC1550 G is a very inexpensive rf–i-f amplifier with constant impedance over entire AGC range.

	Price	Case
MC1460R	$6.75	70-66 type
MC1460G	5.25	71

The MC1460 is an IC voltage regulator. The R version is in a TO-66 type case and can put out 1/2 ampere. The G version is in a TO-5 type case and is good for 200 ma.

DIGITAL INTEGRATED CIRCUITS

Resistor-transistor logic (RTL) in the plastic dual in line package (MC700P series) is recommended for general ham use. It is a relatively fast, easy-to-use, versatile and inexpensive. Three especially useful parts are:

MC724P Quad 2-input gate	$1.08	93
MC789P Hex inverter	1.08	93
MC790P Dual J-K Flip-flop	2.00	93

CROSS REFERENCE GUIDE

Since many of the semiconductors listed in this book are either obsolete or hard to find, this cross-reference list may be helpful. It lists devices with their equivalents in the widely available (1200 distributors) Motorola HEP line. HEP publishes a more complete cross reference guide that is available from its distributors or from Motorola Semiconductor Products Inc., Box 20924, Phoenix Ariz. Zip 85036.

Diodes	HEP
1N134	134
1N91	156
1N128	134
1N295	134
1N457	157
1N459	156
1N540	157
1N547	158
1N645	157
1N658	156
1N1200	153
1N2069	156
1N2071	158
1N4002	156
1N4005	158
10DB2A	176
F2	156
F4	157
M500	158
0A79	134

Zener Diodes	
1N468	602
1N709A	103
1N1314	101
1N3024	607
1N3039	611
1N4733	602
1N4735	103
1N4739	104
1N4752	609
Z4XL6.2	103

Transistors	HEP
2N107	253
2N168	641
2N174	233
2N176	232
2N214	641
2N218	254
2N229	641
2N234	230
2N241	632
2N247	2
2N251	232
2N255	230
2N256	230
2N269	3
2N270	632
2N301	230
2N336	53
2N338	53
2N371	3
2N375	232
2N384	3
2N396	254
2N404	739
2N408	636
2N414	635
2N442	233
2N443	233
2N465	253
2N495	52
2N502	2
2N508	254
2N549	53
2N555	230
2N597	254
2N599	254
2N634	641
2N696	53
2N697	53
2N706	50
2N708	50
2N711	3
2N718	50
2N918	56
2N1178	3
2N1180	3
2N1191	631
2N1193	633
2N1265	254
2N1274	253
2N1305	253
2N1309	2
2N1358	233
2N1360	232
2N1380	253
2N1499	2
2N1517	3
2N1526	3
2N1527	3
2N1534	230
2N1540	232
2N1542	232
2N1554	232
2N1638	639
2N1742	253
2N1743	253
2N1744	253
2N1746	2
2N1749	2
2N1893	53
2N1907	232
2N20842N2084	3
2N2089	3
2N2092	3
2N2147	232
2N2160	310
2N2374	254
2N2613	254
2N2430	641
2N2706	254
2N2613	254
2N2951	53
2N2953	254
2N3054	241
2N3392	725
2N3402	54
2N3404	54
2N3414	55
2N3416	55
2N3478	56
2N3563	56
2N3567	54
2N3638	716
2N3641	50
2N3704	55
2N3819	802
2N3823	802
2N3845	723
2N3877	50
2N4122	57
40237	56
40250	241
AC151	254
AF114	637
AF139	3
MPF103	801
MPF105	801
0C72	254
SB100	2
T2364	253
TF 78/30	230
TIS34	802
TRS3016LC	240

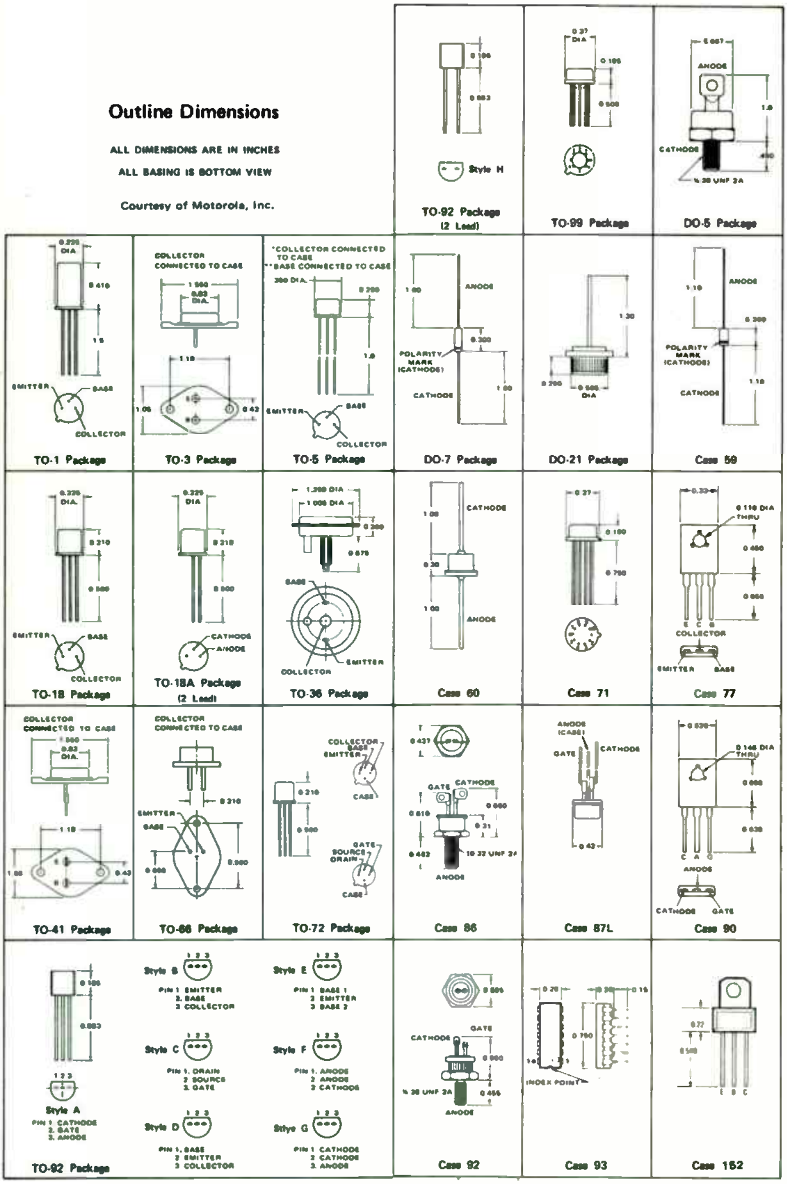
Integrated Circuits	HEP
MC790P	572
MC799P	571
MC1550	590
μA703E	590
μL900	1/2 571
μL914	584, 1/2 570
μL923	1/2 572
UX8991428X	584, 1/2 870
UX8992328X	1/2 572

Outline Dimensions

ALL DIMENSIONS ARE IN INCHES

ALL BASING IS BOTTOM VIEW

Courtesy of Motorola, Inc.



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