PRACTICAL TRANSISTOR CIRCUITS FOR MODERN TEST EQUIPMENT

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BY BERNARD B. BABANI

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Although every care is taken with the preparation of this book the publishers will not be responsible for any errors that might occur.

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Test Components and Appliances with an INSULATION CHECKER

Have you checked the insulation of your electrical appliances lately? At high voltage? This handy checker lets you do this sort of test easily. It also checks transformer and cable insulation, capacitors and high value resistors.

There are many occasions when an insulation tester is useful in the laboratory or home workshop. For example, whenever an electrical appliance is repaired, an important check is to test the power cord. There should be a low resistance connection of the earth lead to the appliance case, and also adequate insulation resistance between the active and neutral conductors and the case.

Often these tests are performed with a multimeter switched to an appropriate "Ohms" range; the low range for earth connection and the highest for the insulation resistance. The insulation resistance should be as high as possible, ie, many megohms.

While the multimeter check is better than none, the insulation of the appliance should really be tested at a high voltage–many insulation breakdowns will not show up at the low applied voltage from a multimeter. The usual test voltage for electrical appliances is 500 volts DC.

The time-honoured method of checking the insulation resistance of electrical wiring and appliances is to use a "megger" tester, which incorporates a hand-driven generator that produces a nominal 500V DC across the load. The insulation resistance measurement is then read on the megger's dynamometer movement (an RMS indicating meter with fixed and moving coils) which takes into account the applied voltage and the current flowing. This is the method used on factory assembly lines and by electricians.

Unfortunately, most home repairmen and hobbyists do not have access to the hand-driven type of insulation tester. Indeed, many electricians now use more compact, battery-driven insulation testers similar to the one presented here. The 500V insulation checker described here satisfies the requirements of the home repairman and hobbyist. It is compact, economical to build and quickly shows up insulation defects.

Besides the appliance testing mentioned above, the insulation checker is handy for checking the insulation of older transformers and filter chokes. These should have high resistance between windings and also between windings and core. It can also check for insulation breakdowns in automotive generators and starter motors. Besides these, it can give a rough-and-ready check on the insulation resistance of peper and plastic dielectric capacitors, and also the value of resistors of more than 1 meg-ohm which are not easily checked with a multimeter.

As the circuit diagram shows, the high voltage is generated by a two-transistor converter. The two transistors operate as a transformer coupled multivibrator. The transistors alternately apply almost the full 9V supply across each main primary winding, which voltage is stepped up by the multi-turn secondary winding. A small centre-tapped feedback winding provides the bias voltage for the transistors.

The square-wave AC voltage from the secondary winding is rectified by a half-save voltage-tripler consisting of three diodes and three capacitors. With no load connected and depending on the condition of the battery, the voltage developed can be anywhere from 500 to 600V.

To enable the same open-circuit voltage to appear at the Checker's terminals at each test, we have an adjustable bleed resistor across the high voltage supply consisting of 470k 1W resistor and a 5M potentiometer connected as a rheostat (variable resistor). This enables the Checker to be used at battery voltages down to 8V.

Adjustment of the Checker for 500V open-circuit voltage is achieved by connecting a jumper lead across the output terminals and setting the 5M potentiometer for full-scale deflection of the meter. This ensures consistent results as the battery ages. When the meter pointer can no longer be grought to FSD (zero), the battery should be replaced.

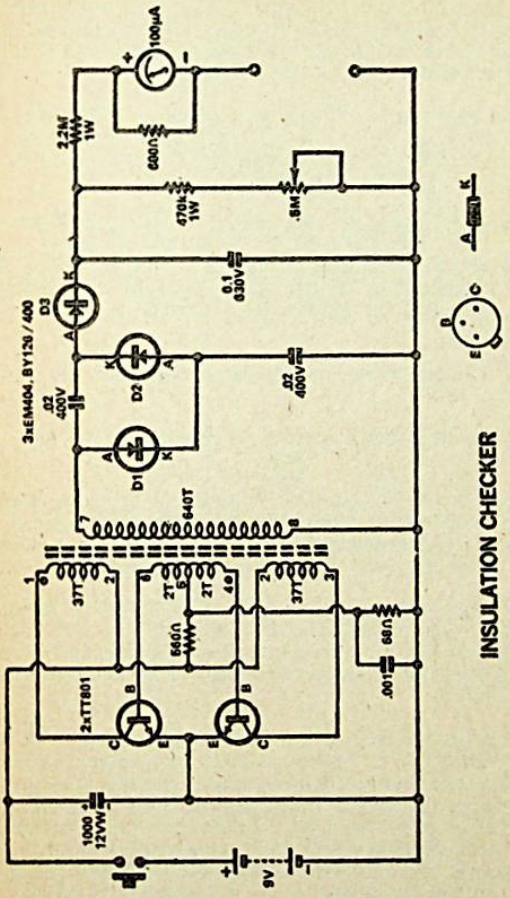
The mode of measurement is as follows: The 500V from the converter is applied to the load terminals via a 2.2M 1W resistor. The current which flows through the load is monitored by a 100uA FSD meter shunted by a 680 ohm resistor. Maximum current flows when a short circuit is placed across the output terminals and this is indicated by the meter as zero ohms. Higher resistances are indicated accordingly, up to and beyond 30 megohms.

Note that unless the insulation resistance of the device being measured is extremely high, the voltage across the output terminals will be less than 500V. Also if you attempt to measure the Checker's output voltage with a VTVM with input resistance of 10 megohms, or a 20,000 ohm/volt meter on the 500V range (which also results in a 10 megohm load) this will yield a voltage reading of only just over 400 volts. Both effects are because the effective output resistance of the Checker is 2.2M.

This order of resistance is a compromise and is required to protect the meter and converter against short-circuits across the output. Note that if the resistance was reduced, the meter scale would be more cramped at one end–at present the centre-scale reading is approximately equal to the output resistance, ie, 2.2M.

Thus, a 2.2M resistor connected across the output terminals of the Checker will have only half the available supply voltage impressed across it, 250V approximately. Similarly, a 1 megohm resistor will have approximately only 150 volts impressed across it.

While the lower voltage across loads of a few megohms is not necessarily ideal, the voltage is still sufficient to show any insulation defects. After all, an electrical appliance should have an insulation resistance of many megohms. The only likely exceptions to this will be found with immersion heaters and washing machines, which may be only 1 megohm or so.



An inverter driving a voltage-tripler rectifier provides a nominal 500V DC for the Checker.

Current drain of the Checker is of the order of 70 milliamps, but since its use will be intermittent, the battery can be expected to last for almost its "shelf life"

A standard diecast box measuring 120 x 95 x 55mm is used to house the Checker. We used one which has internal slots to mount component boards. All the necessary holes are drilled in the case and it is painted and labelled as required. We used Letraset rub-on lettering.

Most of the components are mounted on a piece of Veroboard measuring 90 x 48mm with 0.15in hole spacing. This slides neatly into the slots in the case, dividing it into two compartments. The meter and board components occupy one compartment while the battery and the remaining components occupy the other.

The two silicon transistors on the circuit diagram are 2N3053 or equivalents as shown on the parts list. The transistors run at light load and do not require heatsinks. Note that one transistor has its base lead bent to fit into the Veroboard.

All the capacitors are mounted "standing up" to save space. Take care to mount them exactly as indicated by the wiring diagram, otherwise they may interfere with the back of the meter. The 470k and 2.2M resistors should be 1W types since they have high voltage applied across them, and ½W resistors are not rated for operation in excess of 350V.

The converter transformer can be wound from the details shown in the data panel.

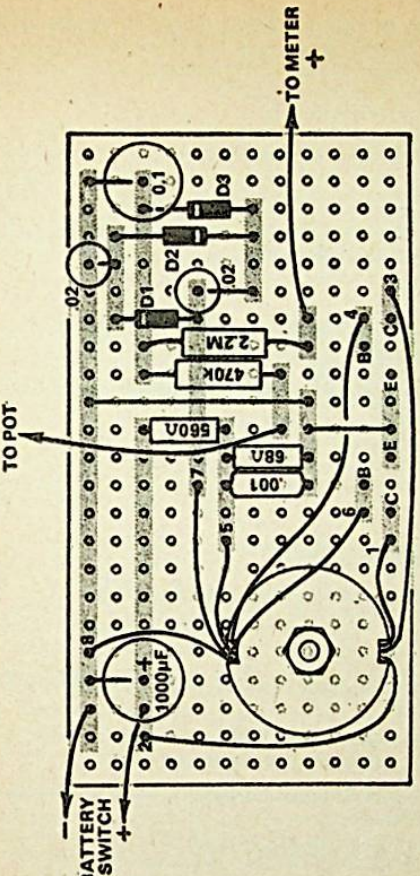
Cut away the corners of the Veroboard so that they do not foul the lid of the case.

A battery clamp should be made from 16SWG aluminium to secure the battery to the lid of the diecast box. The ends of the battery should be covered with insulating tape after the connectors have been pushed on, to avoid the connectors shorting to the case.

If the converter does not operate correctly at switch-on, try reversing the feedback connections–this usually brings forth a welcome high-pitched whine from the transformer. Watch out for the 0.1uF 630V reservoir capacitor–it charges to a high voltage and can deliver quite a nasty "bite".

A new scale will have to be made for the meter movement. This can be done after the rest of the unit is complete and working. Remove the front of the meter movement and carefully unscrew the two screws which secure the scale to the movement. Spray the back of the scale with white paint. After it has dried, draw a suitable arc on the blank face with drawing ink and compass, with same radius as the original scale. Place the scale back on the meter and adjust the potentiometer so that the pointer moves to the end of the scale when the output terminals are shorted together. Now connect 1, 3, 10 and 30 (3 x 10 in series) megohm resistors to the terminals and mark the scale accordingly. Letraset can be used to obtain a neat job.

This completes the description of our solid state Insulation Checker.



Most of the components are mounted on a piece of Veroboard which slides into the case.

PARTS LIST

1 diecast box, 120 x 95 x 55mm, with internal slots
1 moving coil meter movement, 100uA sensitivity
1 SPST miniature push-button switch
2 banana plug sockets
1 metre of miniature figure-8 flex
2 banana plugs
2 alligator clips
1 9V battery plus snap-on connectors
2 FX 2240 Ferroxcube half-cups, with DT2179 bobbin to suit half-cups, or converter transformer

2 2N3053, BSX45, BSY44, 2N2243A, SK3024, 2N3498, RS276-2018
or similar silicon NPN transistors.
3 BY126/400 or similar silicon 400V diodes
1 1000uF/12VW electrolytic capacitor
1 0.1uF/630V polyester capacitor
2 .022uF/400V polyester capacitor
1 .001uF/50V polystyrene or ceramic capacitor

RESISTORS
(¼ or ½W, 5pc tolerance unless otherwise noted)
1 x 68 ohm, 1 x 560 ohm, 1 x 680 ohm, 1 x 470K 1W, 1 x 2.2M 1W, 1 x 5M (lin) potentiometer

MISCELLANEOUS
4 rubber feet
1 piece of Veroboard, 0.15in hole spacing, 90 x 48mm
28 and 42 SWG double enamel copper wire
Aluminium for battery clamp
Varnished cambric sleeving (spaghetti), Hook-up wire, screws, nuts, washers, solder, paint

WINDING THE TRANSFORMER

Order of windings: Secondary, Primary, Feedback.

Secondary : 640 turns, 38 B & S or 42 S.W.G. enamel. Code start with knot.

Interleave : One wrap of electrical tape or polythene film between secondary and primary.

Primary : 37 turns plus 37 turns, bifilar (wound together), 26 B & S or 28 S.W.G. enamel. Place two knots in start.

Feedback : 2 turns plus 2 turns, bifilar. 26 B & S or 28 S.W.G. enamel. Place three knots in start.

Outer wrap : One wrap of electrical tape.

When winding coils of this type, counting and handling is made easier by climping the bobbin between two large washers, using a long bolt and nut and rotating this assembly in the chuck of a small hand drill held in a bench vice. This leaves one hand free to guide and tension the wire to obtain an even winding. The washers prevent collapse of the side cheeks of the bobbin as winding proceeds and assist clamping.

Count the number of times the drill chuck rotates for one turn of the handle. Divide this ration into the number of secondary turns. If your drill ration is 5 : 1 it will require 128 turns of the handle to wind the necessary 640 turns onto the bobbin.

Take the 38 B & S or 42 S.W.G. enamel copper wire, tie a knot in it about two inches from the end and lay the wire in the bobbin so that the end passes through a deep cheek notch with about three inches to spare. Anchor the start with small piece of electrical tape. Secure the free end of the wire on the drill chuck with ordinary celluloid tape or a rubber band

Make sure that the wire passes through a deep cheek notch and not a half-depth one, otherwise the wire will build up against this lead for about half the winding and place too high a voltage stress on the wire insulation when the converter is operated.

Proceed to wind the secondary, guiding the wire carefully to keep an even build-up. Finish off the winding with a small piece of tape so that the last turn will come out on the same side as the start. Place one layer of polythene film or electrical tape over the complete winding.

The primary winding is of thicker wire and cannot be conveniently wound with a hand drill. A tighter, more even winding can be obtained with hand winding.

Take two 150cm lengths of 26 B & S or 28 S.W.G. enamel wire and lay them side by side. Tie two knots in the start end one and twist lightly with the other for about two inches. With the bobbin assembly still mounted on the bolt but placed in the vice instead of the drill chuck, wind on the required number of turns in the same direction as the secondary winding. Anchor the start in the same way. Avoid twising the two wires and keep the turns as close together as possible. Count off the turns as though the two wires were one conductor. Anchor the finish with electrical tape.

The feedback winding consists of two turns bifilar wound in the same way as the primary using 26 B & S or 28 S.W.G. enamel wire. One start should be designated with three knots. Place the final wrap of electrical insulation tape over the outside of the complete winding.

Before assembling the transformer, the following points should be observed : (1) that there are no traces of foreign matter on the core faces, otherwise the cores can be cracked when they are tightened together : (2) that the secondary leadout wires are laid in the notches provided inside the cores. Inspection of the inside of each core will reveal these leadout notches.

Before finally placing the core halves together, place two pieces of coloured PVC sleeving on each lead of the secondary winding, as close as possible to the bobbin. Use different colours for the start and finish. Suitable small diameter sleeving can be stripped from scrap lengths of thin hook-up wire.

Press the cup-core halves together with the fingers and anchor them with electrical tape around the outside. Clamp the transformer with a 1" long x 1/8" Whitworth screw and nut through the core assembly. Use an appropriate washer on either side of the core to prevent undue mechanical stress on the ferrite material.

A SOLID STATE VOLT-OHM METER

Our latest test instrument design is a solid-state equivalent to the familiar 'VTVM', and offers all the advantages of the latter together with increased sensitivity, virtually instant warm-up, and complete independence from the power mains. These features, together with simplicity and low cost, should make it a popular project.

Specification:

A portable high input impedance volt-ohm meter, employing two silicon junction field-effect transistors (JFETs) and five diodes. Input impedance on all DC and AC voltage ranges is approximately 10.9 Megohms shunted by a few picofarads. Seven DC ranges covering the range 1V–1000V FSD, and six AC ranges covering the range 3V–1000V RMS, both in 10dB range steps; seven resistance ranges covering from 10 ohms–10Megohms centre-scale. Power consumption is less than 150 milliwatts, supplied by either an internal 18V battery or an external supply.

The familiar VTVM is widely used for making circuit voltage measurements in electronics workshop and laboratory situations, and is particularly suited for this task by virtue of its ruggedness and high input impedance–typically around 10M. However, the usual VTVM is a mains-powered instrument, and cannot be used easily either in situations remote from the power mains, or to measure potential differences which are "floating" with respect to earth. It also requires some 5 to 10 minutes "warm-up" following switch-on, before the internal thermionic valves and associated circuitry stabilise sufficiently to permit accurate and reliable measurements.

The solid state instrument to be described in this article offers virtually all the advantages of a VTVM, together with some important new performance features. It has an extended DC sensitivity–down to 1V FSD, an effective "warm-up" time of but a few seconds, and complete independence from the power mains. The last-named feature permits the instrument to be used not only for making measurements "in the field" remote from the power mains, but also allows it to make measurements which are fully "floating" with respect to earth.

Despite these added features the new instrument is no more complex than the usual VTVM, and if anything should involve a slightly lower initial cost. It should thus prove a most popular project, and a useful addition to either the home workshop, the service shop or van, or the development lab.

At the heart of the new instrument is a balanced-bridge DC amplifier circuit rather similar to that of a conventional VTVM, but using two junction field-effect transistors (JFETs) instead of a double triode valve. As with the VTVM, the DC amplifier effectively converts a standard 0-1mA/100 ohm meter movement into DC voltmeter having an extremely high input resistance.

The JFET is a relatively new semi-conductor device, and one whose operation is actually closer to that of a thermionic valve than to a conventional "bipolar" transistor. In contrast to the bipolar transistor, it is a voltage-controlled device rather than one which is current-controlled, and has the very high input resistance of a reverse-biased semiconductor P-N junction rather than the relatively low input resistance of a forward-biased junction.

Briefly, its operation involves the control or modulation of the conductance of a relatively narrow strip of semiconductor material called the channel, by means of a transverse electric field produced by an adjacent semiconductor electrode called the gate.

At the ends of the channel are the source and drain electrodes, corresponding roughly to the cathode and plate respectively of a thermionic valve. An input voltage applied to the gate electrode is accordingly able to control any channel current flowing from source to drain, in a rather similar fashion to that whereby the bias voltage at the grid of a valve is able to control the cathode-plate current.

It has been possible for some time to produce a solid-state instrument equivalent in performance to the VTVM, using bipolar junction transistors. However, almost of necessity such instruments must be rather complex and expensive, mainly due to the measures required if the inherently low input resistance of bipolar devices is to be overcome without prejudice to stability or reliability.

The very high input resistance of the JFET device renders it considerably better suited for this type of application, so that in theory the release of modestly priced JFETs some two years ago should have enabled designers to produce solid state volt-ohm meters which compared favourably with the VTVM, not only in terms of relative simplicity and low cost. But in general this was not the case, mainly because the principal behaviour parameters of the JFET devices initially released were subject to production spreads which were embarrassingly wide for this and similar applications.

For the present, it is perhaps sufficient to note that the first economy devices released had a zero-bias source-drain current (Idss) spread range of a full dacade (3-30mA), together with a slightly greater range for the nominal pinch-off voltage Vp (0.75-10v).

Until recently, then,the would-be designer of a solid state volt-ohm meter faced the frustrating situation wherein he was effectively prevented from producing a simple and economical design largely because the very devices which were in theory almost ideal for the job were subject to "incidental", but nevertheless very real, spread problems.

Few situations remain static in contemporary electronics, particularly those concerning semiconductor devices, and as one might infer correctly from the appearance of the present article, the foregoing situation has happily changed for the better in recent months. In short, there are JFET devices now available at economy prices whose parameter spread range is considerably narrower than those previously available.

The particular devide which has been used in the new instrument is the Motorola type MPF105, an N-channel device which has an Idss spread range of only 4:1 (2-8V). While still not small, even by semiconductor standards, these ranges represent a significant improvement over those of the devices available earlier.

The MPF105 is also significantly superior to earlier devices in terms of the gate reverse leakage current Igss, which parameter plays a big part in determining the DC stability of the device in voltmeters, and similar applications wherein the gate circuit resistance must be kept high. Whereas earlier JFETs had a maximum Igss of some 10nA at 25 degrees C, the MPF105 has a maximum Igss – at the same temperature – of only one-tenth this figure.

Using the MPF105 device, and by careful design, we have been able to produce a solid state volt-ohm meter which compares very favourably with the conventional VTVM, in terms of both performance and cost.

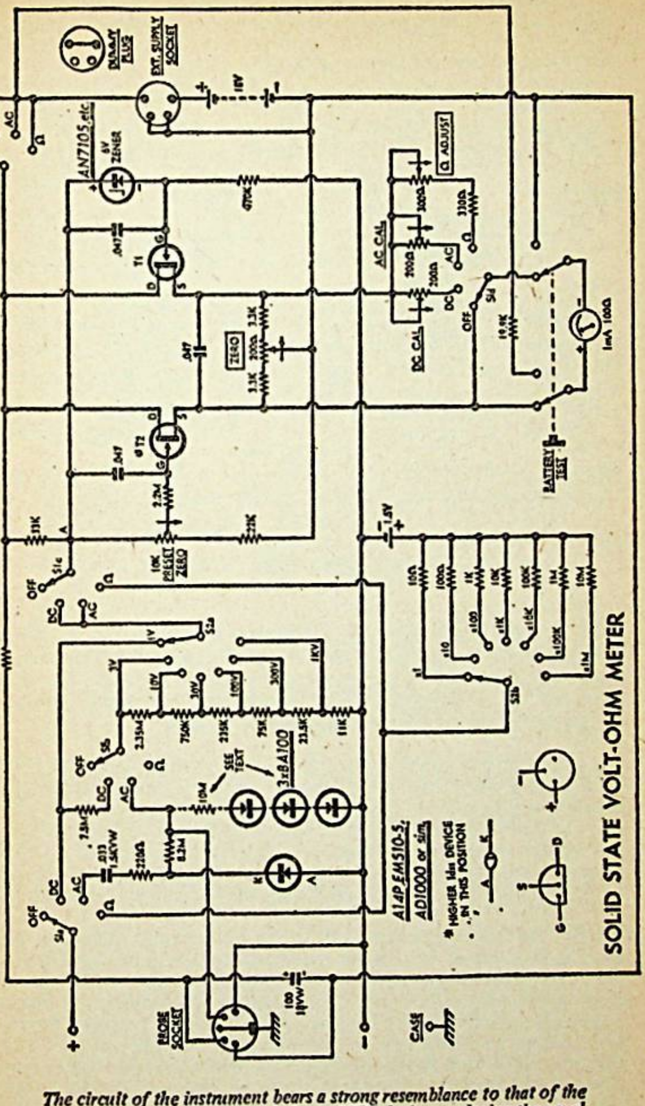
The instrument has an input impedance on both the DC and AC measuring ranges of greater than 10 Megohms shunted by a few pF, while the amplifier linearity is considerably better than any meter movement likely to be used. Zero drift and basic calibration drift from switch-on are virtually negligible, and are each stable typically to better than 1 per cent to 40 degrees C (104 degrees F) ambient. Even at the rather unrealistic ambient temperature of 50 degrees C (122 degrees F) they will each typically be within approximately 3 per cent, which should be more than acceptable for the majority of applications.

Power consumption of the instrument is very modest, at around 150 milliwatts: it draws a mere 7-8 milliamps from an 18V supply, and is therefore capable of very economical battery operation. Using the 18V battery, the useful service life of the battery should be more than 200 hours. In addition to the main internal bettery the instrument also employs a nominal 1.5V cell for the resistance ranges, as does the usual VTVM.

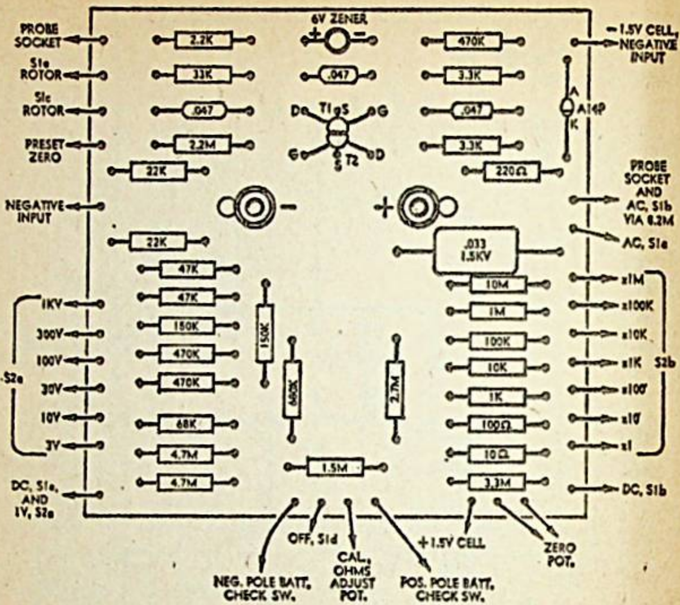
Provision has been made for alternative operation of the instrument from an external (floating) 18V supply, so that the internal battery may be conserved if desired when the instrument is used on the workbench or in similar situations wherein mains power is available. Connection of an external supply is facilitated by a simple plug and socket at the rear of the instrument case.

Reference to the main circuit diagram should assist in understanding the operation of the new instrument. The heart of the circuit is the balanced-bridge DC amplifier centred around JFETs T1 and T2, which may be recognised as rather similar to the equivalent section of a VTVM.

T1 and T2 are each connected in the "source-follower" configuration across the 18V supply, with the meter movement connected between the two sources via the usual calibrating multiplier resistors. Each JFET has a source load, consisting of a 3.3K resistor together with part of a shared 200 ohm potentiometer which functions as a differential load vernier for quiescent balancing or "zero" adjustment.



The circuit of the instrument bears a strong resemblance to that of the familiar VTVM, with a pair of silicon JFET devices replacing the usual double triode valve, and a silicon diode in place of the usual valve rectifier. The zener is for overload protection of the active transistor T1, and must be a low-leakage type to avoid input shunting.



Device T1 is actually the "input" or active arm of the bridge, although this may not be immediately apparent from the circuit. The quiescent operating current of T1 is determined by a fixed forward bias derived from tap "A" on a resistive divider across the power supply, and applied to the gate of T1 via the input attenuator system.

A similar forward bias is applied to the gate of the balancing device T2, via a 2.2M series resistor which approximately matches the average attenuator resistance in the gate circuit of T1. As may be seen the bias for T2 is not fixed, but is variable over a range of some 3V extending negative from divider point "A". This has been done to enable the circuit to accommodate even worst-case differences between the devices used for T1 and T2, in terms of the spread variations in Idss and Vp.

Briefly, the reasoning behind this is that for accurate balance of the circuit, as indicated by a meter reading of zero with no input signal applied and the vernier load pot in mid-position, the two JFETs should draw equal channel currents. This will occur automatically if the two devices are matched and if they are presented with the same forward bias voltage; however, with devices purchased as "unmatched" it is far more likely that for equal channel currents one will require a lower forward bias than the other. In other words, the device with the higher Idss figure must be effectively "biased back" with respect to the other if the two are to draw equal quiescent currents.

When the two devices are operating at equal channel currents Id, the ratio Id/Idss for the device with the lower Idss will naturally be greater than for the device with the higher Idss. It follows that the former device will tend to display a higher transconductance, because the transconductance of a JFET is proportional to the square root of the ratio Id/Idss. Hence since the linearity of the voltmeter amplifier will be at least a partial function of the transconductance of the "input" device T1, it is the device with the lower Idss which should be used in this position.

Conversely it is the device with the higher Idss which should be used in the T2 position, and therefore the T2 device for which adjustable gate bias should be provided to allow coarse circuit balancing.

Although it might seem from the foregoing that it would be necessary to measure the Idss currents of the two devices concerned before they could be connected into circuit, this is not necessary. As will be described in greater detail later, the correct positions for any two MPF105s can be found quite simply and rapidly by trial and error: there are only two possible combinations so that if balance cannot be achieved with one it is simply a matter of reversing to the other.

In order that the amplifier circuit should be capable of driving the meter movement linearly to full-scale deflection with any MPF105 device in the T1 position, the source load resistances of T1 and T2 and the forward bias on T1 have been chosen to give the highest quiescent channel current Id compatible with correct operation of all devices at temperatures to greater than 50 degrees C. Consequently the input voltage of the instrument is arranged to reduce the current of T1 and this explains why the gate of T1 connects to the negative input polarity while the positive input polarity unlimately connects to the bias tap "A"

The 470K resistor and 6V zener diode associated with the gate of T1 are designed to protect the JFET from damage if the instrument is connected to excessive and/or reverse polarity input voltages. The zener may to some extent be regarded as optional "added insurance", as the current limiting provided by the resistor is possibly sufficient to prevent damage to T1 in almost all likely overload situations; however, the additional protection provided by the zener is recommended as worthwhile in view of the modest outlay involved.

Three 0.047uF capacitors associated with the gates and sources of T1 and T2 perform high-frequency bypassing, and render the instrument virtually insensitive to AC signals superimposed upon the input voltages being measured.

The input switching and attenuator system of the new instrument is very similar to that of a conventional VTVM. There are only two control switches, a function/power switch S1 and a range switch S2. The former has four positions, Off-DC-AC-Ohms; while the latter has seven positions covering the range 1V-1KV for DC voltage measurement, and the range 10 ohms–10M (centre scale) for ohms measurement. Only six of the seven positions provided by S2 are used for the AC voltage ranges, the 1V position becoming inoperative for this function.

It may be noted that in contrast with a conventional VTVM, there is a single "DC" function rather than two of opposite polarity. The reason for this is that the instrument is capable of "floating" with either or neither input terminal earthed, and therefore has the same flexibility as a passive meter. Polarity switching is accordingly not required.

The provision of a 1V FSD range for the DC voltage function is something of a "bonus" as the majority of conventional VTVMs have a maximum sensitivity of 3V FSD. This range has been made possible by the increased sensitivity of the JFET meter amplifier, and should prove very useful for making measurements in low-voltage semi-conductor circuitry and similar situations.

An equivalent range has not been provided for the AC voltage measuring function, for the reason that linearity problems are created at low AC input voltage levels by the forward conduction characteristic of the silicon diode used in place of the usual diode valve as the peak rectifier. It has been found possible to obtain an order of linearity adequate for ranges down to 3V RMS full scale, using a non-linear shunting technique, but this technique would not really be capable of providing a worthwhile 1V range. In fact, to provide AC measurement ranges of higher sensitivity than 3V with a solid-state instrument it is necessary either to employ an active rectifier system, or to employ a rectifier preamplifier system as used in AC millivoltmeters.

The non-linear shunt used to linearise the 3V AC range consists of a 10M resistor in series with three low-voltage silicon diodes, type BA100 or similar, and connected across the main voltage attenuator divider string. In effect, the shunt varies the sensitivity of the DC metering amplifier in a matter which compensates for the nonlinearity of the main rectifier diode.

Note that the non-linear shunt is connected permanently to the rectifier output, and therefore remains operative on all AC ranges. For input voltages of higher than about 3V RMS it acts purely as a 10M resistor, and the circuitry has been arranged to allow for the slight shunting effect involved.

In the interests of extreme economy the resistor and diodes of the shunt may simply be omitted, but this is not recommended because both the linearity and the calibration of the 3V AC range will suffer as a result. If the shunt is omitted, the AC calibration of the instrument should be set up using a range other than the 3V range, otherwise the accuracy on all AC ranges will suffer.

As is the case with conventional VTVMs, the highest (1KV) voltage measuring ranges are best regarded as "over-range" facilities rather than full measuring ranges, because although the instrument is suitably calibrated for measurements on these ranges it is doubtful whether the components associated with the input circuitry would be capable of withstanding the potentials corresponding to full-scale deflection. This particularly applies to the 1KV AC range, where the peak voltage present at the input of the instrument for FSD would be almost 1500 volts; at this figure most input connectors and rotary switches will tend to suffer damage, as one might imagine.

In fact, it is best to use the 1KV ranges only to perform measurements up to about 600V, employing a suitable high-voltage divider probe if the instrument is to be used to measure voltages higher than this figure. A suitable probe for this purpose may be described in a following article.

The silicon diode of the AC peak rectifier should ideally be capable of withstanding at least the peak value of 1000V RMS, i.e., 1414V. However, in view of the probable damage which would be sustained by the input connectors, switches and other components in the event of this voltage being applied, and also in view of the relative scarcity and high cost of diodes having a rated P.I.V. of higher than 1000V, it seems almost fatuous to specify a diode with the required 1500V+ rating.

Instead we are simply recommending that the constructor should use a diode of at least 1000V rating; suitable types are shown in the circuit. Of the types shown it may be noted that the A14P is a General Electric type featuring voltage transient overload protection, and should accordingly offer increased reliability.

It should also, perhaps, be noted that a surge-limiting resistor (220 ohms) has been included in the peak rectifier circuit to protect the diode from current transient overload. While not required in a VTVM because of the relatively high internal resistance—and ruggedness—of a diode valve, this resistor is worthwhile "insurance" when a semiconductor diode is used.

The peak rectifier reservoir capacitor (.033uF) is in a more-or-less parallel situation to that of the diode; ideally it should have a voltage rating well above 1500V, to allow for any DC superimposed upon the AC voltage being measured. However the specified rating of 1500V should be sufficient to cope with most situations, and should be adequate if the precautions noted earlier are observed.

The resistance measuring ranges are exactly the same as those in most VTVMs, both in circuitry and in operation. A standard 1.5V cell is used, as noted earlier, although other types such as the manganese-alkaline cell or nickel-cadmium cell might prove more suitable in some circumstances.

The switching associated with the meter movement is quite straightforward, and its operation should be almost self-evident. The function switch S1(d) is arranged to apply a short-circuit across the movement in the "off" position, which will subject it to heavy electrical damping and reduce the risk of mechanical damage during transit.

To allow the meter to be used to check the condition of the internal 18V battery, a microswitch button is used to disconnect the movement from the DC amplifier and connect it across the battery via a multiplier which converts it into a 0-20V passive voltmeter. As the battery may be tested when the instrument is either "on" or "off", it is therefore possible to observe the battery on and off-load.

The "battery test" may also be used to check the voltage fed to the instrument when it is operated from an external power supply in place of the internal battery.

Provision has been made for connection to the instrument of both active and passive AC probes, by means of a 5-pin "DIN" socket on the front panel. An RC filter provides de-coupling for the DC supply available at the probe socket, to simplify power supply problems and to ensure stability with active probes.

It may be noted that the circuitry of the instrument is fully insulated from its metal case, so that the latter may be either earthed, left unconnected or tied to any appropriate point to ensure both reading accuracy and operator safety for "floating" measurements. The probe socket body is connected to the case, so that by arranging that the cable braid and casing of any probes used are connected to the instrument case via the socket, the case-probe system will function as an electrostatic "guard" system. Naturally this will involve insulation of both the "active" and "return" probe input connections from the probe casing, and this is easily achieved.

The main external differences are actually at the rear of the case, with the three preset controls along the top and the "case" connector and external supply socket at centre right. A final difference is the possibly conspicuous absence of a mains cord!

Incidentally it may be noted that the preset controls of the prototype instrument are mounted directly on the rear of the case. This has the disadvantage that the settings may easily be altered by accidental contact, and accordingly it is recommended that constructors do not mount the controls in this fashion. A far better plan would be to mound them on a small sub-panel bolted just inside the case; this will prevent the settings from being disturbed, while still allowing convenient screwdriver adjustment.

Inside the case, most of the minor components are supported by a small printed wiring board which connects to and is supported by the meter connection studs. We have prepared a wiring diagram showing the position of the components on the board, and the destination of each of the leads connecting to it; as the remainder of the wiring is fairly straightforward, assembly of the instrument should therefore presnet few problems.

Note that two sets of meter mounting holes are provided on the board, to suit most commonly available movements. Also that the two MPF105 transistors are arranged to be close together with their "flats" adjacent, allowing them to be clamped together by a small strip of sheet copper or brass. This will ensure that the two always remain at much the same temperature, and will reduce any possible drift to a minimum.

Those resistors in the circuit which have non-preferred values are in fact made up using two or more units of standard value. In one case, this has been done to protect individual resistors from excessive applied voltage; however, in the remaining instances, multiple units are used purely to obtain the required special values.

The 19.9K meter multiplier used for battery checking is made up from two in parallel, one a 22K 5 per cent type and the other a 220K whose tolerance is not critical.

The remaining non-standard values are in the input voltage attenuator, and these are made up as follows: The 7.5M value consists of three in parallel, with values 3.3M, 2.7M and 1.5M. The 2.35M value consists of two 4.7M units in parallel, while the 750K value consists of a 680K-68K series combination. The lower values are all made up from parallel combinations of preferred values double the specified value: 2 x 470K giving 235K, 2 x 150K giving 75K, 2 x 47K giving 23.5K, and 2 x 22K to give 11K.

The type of resistor used in the input attenuator and ohms reference circuits will depend largely upon the extent to which the instrument is to be taken seriously, and also upon the allowable cost. Ideally the resistors concerned should all be precision high-stability types, but these are admittedly quite costly and would scarcely be justified except for serious development laboratory or field work.

A compromise would be to use high-stability types of wider tolerance, such as cracked-carbon components with 2 per cent tolerance. This approach will probably give the instrument an overall accuracy and reliability adequate for most general applications. In fact this was the approach adopted with the prototype instrument

If the outlay involved with the foregoing approach is still regarded as excessive and inappropriate, there is no reason why the constructor should not use standard carbon composition resistors of 5 or 10 per cent tolerance, perhaps selecting the values closest to those required with the aid of a resistance bridge. With care this approach can still give adequate overall accuracy for servicing and amateur work, although the long-term reliability may leave something to be desired.

The two batteries which are included in the instrument are clamped firmly in place in the rear lower right of the case by a small bracket of sheet aluminium. To ensure that there is no risk of breakdown between the battery electrodes and the instrument case, both batteries are wrapped tightly with a few layers of polythene sheeting before being clamped in position.

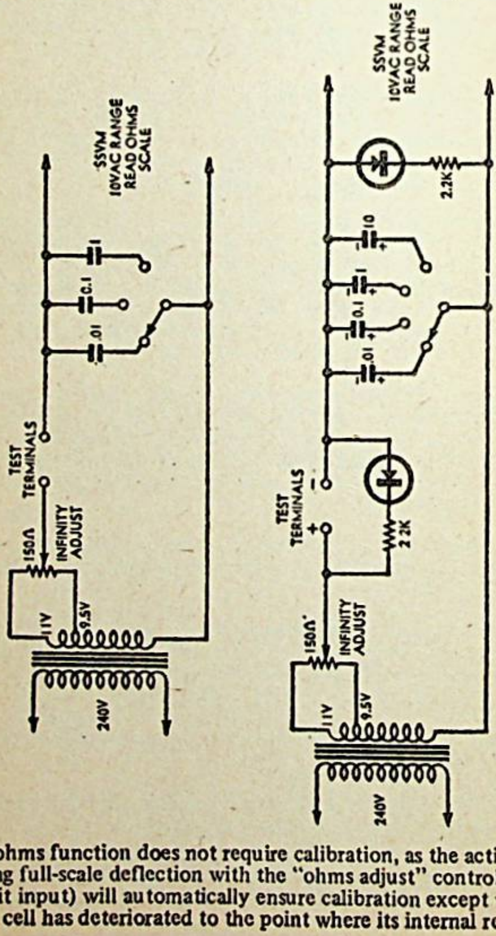
When the instrument is completed, the metering amplifier must be balanced before calibration can be performed, and this may be done in the following manner.

The first step is to ensure that the meter movement itself is correctly adjusted to indicate "zero" when no power is applied. If there is any residual reading, this should be removed using the usual screwdriver adjustment. The instrument may then be turned on, to the "DC" function and the 3V range, with no input applied and the front-panel zero control set to mid-position. The input connectors may be either left open or shorted together.

At this point there will most probably be a significant meter reading, indicating that the circuit is unbalanced and the two JFETs are drawing unequal currents. There is a 50 per cent chance that this state of affairs may be corrected simply by adjustment of the "preset zero" pot at the rear of the case, the pot simply being turned until the meter reading falls to zero.

If it proves impossible to reduce the meter reading to zero, this will be because the device in the T1 position is actually that with the higher Idss. In this case it will be necessary simply to reverse the position of the two devices, whereupon the problem should dissappear.

With the metering amplifier balanced the remaining set-up operation is to perform DC and AC calibration. This may be done with the instrument set to any of the appropriate ranges which proves convenient, as single calibration controls are used for the two functions. The calibration operation will normally involve a suitable source of variable voltage (preferably regulated), together with a voltmeter of known calibration against which the new instrument can be compared.



The upper circuit is for a capacitor tester suitable for all non-electrolytic capacitor types. The lower circuit shows the modifications needed to accommodate electrolytic types.

The ohms function does not require calibration, as the action of producing full-scale deflection with the "ohms adjust" control (with open circuit input) will automatically ensure calibration except when the 1.5V cell has deteriorated to the point where its internal resistance has become significant relative to the reference resistors selected by S2(b) Before this point is reached the battery should normally have been replaced; we have deliberately restricted the sensitivity available on the ohms function, by means of the 330 ohm resistor in series with the "ohms adjust" pot, in an effort to ensure that this is the case.

It was noted earlier that the service life of the main battery used in the instrument should be in excess of 200 hours; with careful use this should correspond to a considerable period of time. One factor which should aid in obtaining economical operation is the virtually "instant warm-up" of the instrument, which will enable it to be turned off between readings in many situations.

The battery should normally be replaced when its voltage, as read under load using the "battery test" button, has falled to about 15V corresponding to approximately 75 per cent of full scale. With supply voltages lower than 15V the accuracy and linearity of the instrument will be impaired.

List of Components

1 Instrument case, 7½in x 5in x 4in, with flanged front panel
1 wiring board, 4½ins square
1 Set case hardward (handle, rubber feet, screws, etc.)
1 0-1mA meter movement, 4in rectangular, 100 ohms, with standard VTVM scales
1 Rotary switch, 3 sections 2-pole 4-positions
1 Rotary switch, 2 sections 1-pole 7-positions
1 Microswitch button, DPDT spring return
1 18V battery
1 1.5V cell

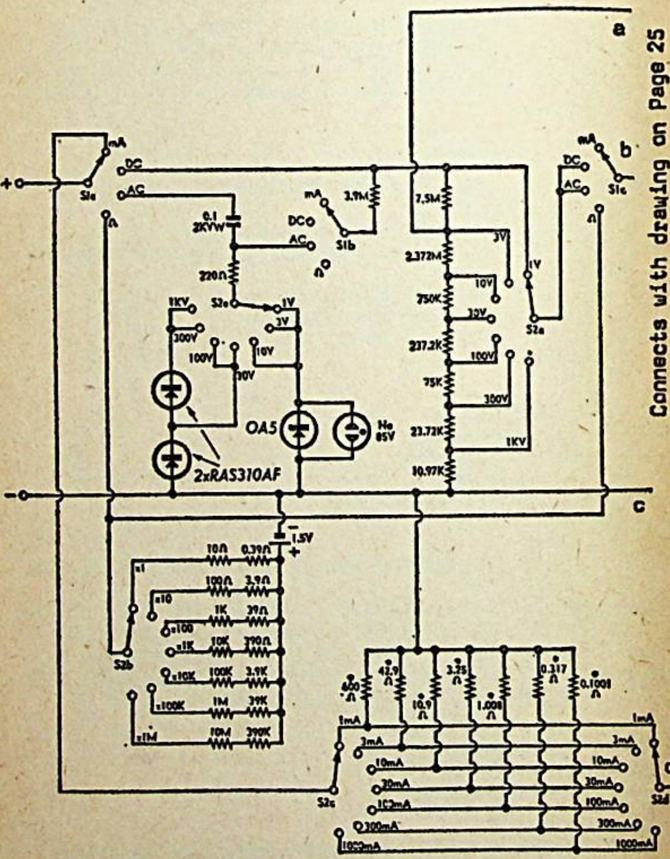
SEMICONDUCTORS
2 MPF105 n-channel JFETs
1 low leakage 6V zener diode
1 1000V rectifier diode
3 BA100 or similar silicon diodes

RESISTORS

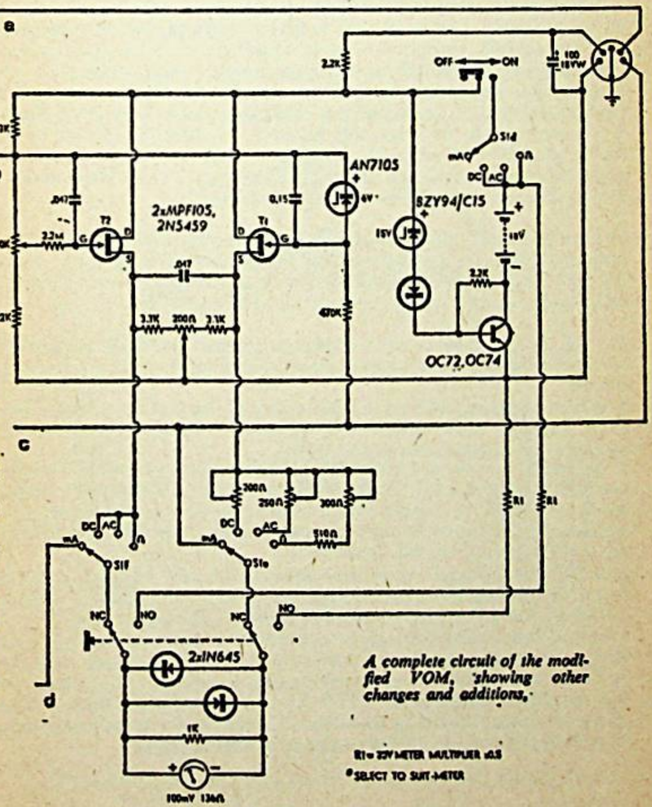
5% ½-watt type: 220 ohms, 330 ohms, 2.2K, 2 x 3.3K, 2 x 22K, 33K, 220K, 470K, 2.2M
High stability close tolerance (see text): 10 ohms, 100 ohms, 1K, 10K, 2 x 22K, 2 x 47K, 68K, 110K, 2 x 150K, 2 x 470K, 680K, 1M, 1.5M, 2.7M, 3.3M, 2 x 4.7M, 8.2M, 10M
Potentiometers: 1 x 200 ohms lin. or WW; 2 x 200 ohms preset, lin. or WW; 1 x 300 ohms lin, or WW; 1 x 10K preset, lin. or WW

CAPACITORS
1 .033uF 1.5KV plastic
3 .047uF LV plastic
1 100uF 18VW electrolytic

MISCELLANEOUS
5-pin DIN socket, polarised 4-pin socket and plug, 3 x banana jacks (red, black, green), 4 x control knobs, scrap aluminium for battery clamp and preset pot bracket, connecting wire, solder, etc.



Connects with drawing on Page 25



Connects with drawing on Page 24

A complete circuit of the modified VOM, showing other changes and additions.

R1 = 25V METER MULTIPLIER x0.5

*SELECT TO SUIT METER

CAPACITOR MEASUREMENT WITH THE SOLID STATE VOM

This section describes a number of modifications to the Solid State Volt-Ohm Mèter including circuitry to provide direct reading of capacitance. This facility can even be extended to electrolytic types.

Recently having the need to measure several paper capacitors (value unknown) I devised the enclosed circuit to be used in conjunction with my modified version of the Solid State Volt-Ohm Meter. The scale law it follows is the same as the existing ohms scale, and the capacitance range selector is identified with the centre scale value of the range selected.

The procedure to use it is the same as for any ohm meter. First, short the leads together and adjust the "infinity set" control for FSD with the meter selected to 10VAC. Then select a suitable capacitance range and connect the capacitor to be tested. In practice this works very well and makes a useful accessory to the meter.

As shown by the circuit, I have altered the meter AC voltage rectifier circuit to give a greater margin of safety on the higher AC ranges and included a 1VAC range in the meter itself by using a separate germanium diode for the low AC ranges and a separate 1VAC scale.

To overcome an appreciable calibration drift with battery voltage drop (approximately 1 per cent per volt change), a zener controlled regulator was incorporated. This not only cured the calibration drift but by running the circuit at 15 volts, the standing current is reduced by an amount which more than compensates for the slight zener diode bleed current, thus giving a marginal increase in battery life.

One fringe benefit of the battery switching circuit is that in the "OFF" position, operating the Battery Test switch gives battery volts, while in the "ON" position the zener regulated voltage is monitored. This enables both battery condition and zener action to be verified.

Finally, current ranges were added as a logical extra and resulted in a 1 per cent accuracy meter with comprehensive volts, amps, and ohms ranges.

It is also possible, with little additional cost, to test electrolytic capacitors. The only modification required is the addition of two diodes and resistors as shown in the basic circuit.

In operation the test and reference capacitors form a voltage divider, the one with the highest voltage developed across it will cause its respective diode to conduct and charge BOTH condensers to the peak value, after which both diodes are reverse biased and play no further part in the divider ratio action of the two capacitors.

The DC bias developed is at least equal to the peak AC applied and prevents reverse current flow through the electrolytic capacitor. The 2.2K resistors act as current limiters in short circuit conditions.

It is possible to use electrolytic capacitors in the reference position also provided the correct polarity is observed. In practice 10uF seems to be the upper limit for reasonable scale accuracy. As the bias voltage developed varies from 7.07 to 14.14 volts DC depending on the ratio of the two capacitors it is unsuitable for very low voltage rating units. A complete practical modified circuit is shown on pages

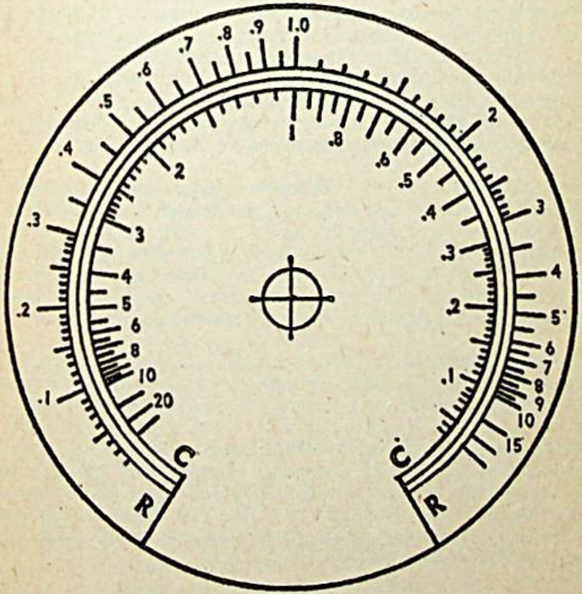
SIGNAL INJECTOR AND R-C BRIDGE

A signal injector, built around a single micro-circuit, doubles as a signal source for an equally simple R-C bridge.

This article is intended as a further introduction of micro-circuits to those readers who have not, as yet, begun to familiarise themselves with them. As you have probably already realised, micro-circuits are very much here to stay and lend themselves to a seemingly endless number of uses.

The major part of this article is centred on a Signal Injector which is built around a digital IC. Such an instrument may be described as a crude form of signal generator which will operate over a very wide band of frequencies. Basically, it consists of a low audio frequency which produces a waveform rich in harmonics. Usually, the waveform is rectangular or saw-tooth.

Such a device therefore becomes an ideal signal source for testing amplifiers, radios and perhaps even television circuitry. For reasons which are outlined later we decided to take a different approach, although



The above is a reproduction of the scale for direct reading of resistance and capacitance.

readers who wanted to could fit this particular circuitry into a Penlite torch case, with a little ingenuity. One advantage of this approach is that one can obtain a professional finish with a minimum of effort. The disadvantages will become apparent as the article progresses.

The integrated circuit used for the injector is a Fairchild Ful 914 dual 2-input gate. Each gate consists of two transistors, as can be seen from the diagram of the internal circuitry. In the circuit of the Injector one transistor of each gate is disabled by connecting its base (pins 2 and 3) to the negative rail. The remaining transistors are connected in an asatable multi-vibrator configuration. The output can be taken via a capacitor from either pin 6 or 7. The waveform is rectangular, frequency and mark/space ratio being determined mainly by the two 22K resistors and their associated .047uF capacitors. The frequency using these components, was 630Hz in the prototype and the mark/ space ratio was unity.

The circuit can, in fact, be rigged up as an elementary variable frequency square wave generator simply by substituting a 25K potentiometer for one of the 22K resistors. However, the mark/space ratio would vary with the frequency and for best results a 25K dual ganged potentiometer would have to be substituted for the two 22K resistors.

The injector can be used with a single 1.5 volt cell but we decided that a 3-volt supply was desirable to obtain the maximum possible output signal. This makes it difficult to assemble inside a Penlite torch case as mentioned above unless one uses more compact cells such as mercury or alkaline-manganese. Assuming that only a simple device, without benefit of a level control, was required, then it might be possible to fit it in the torch case, using the compact cells.

However, we decided that a level control was necessary, otherwise equipment would be overloaded by approximately 2 volts peak-to-peak of signal available. With this addition, unless one is to use a miniature potentiometer as well as the mercury or alkaline batteries, or is prepared to compromise on performance, it is not feasible to assemble the device in a miniature torch case. The same reasons, plus the one of expense, ruled against assembly in a conventional probe case.

We therefore decided to assemble it in a 'pill case'. These are used for packaging headache pills. For this reason they should be easily available to everyone and are very economical. The one we used was aluminium although plastic will do just as well. Plastic also has the advantage that it is an insulator and one does not have to worry about components 'shorting' inside the case. These pill cases (plastic) can also be used as coil formers, but that is another story. The diameter of the case was 1¼in and the length was 2½in. The limited length meant that the components were tightly packed and a case with a slightly larger diameter would be an advantage.

To obtain a nominal 3-volt supply we used two alkaline-manganese cells made by Mallory. This was necessary as we found that it was not quite possible to assemble the injector with two Penlite cells. Their capacity is comparable to the Penlite cell, their shelf life is considerably longer and the size is almost half that of the Penlite cell.

The current drain with a 3-volt supply is 5mA which falls to 2mA with a 1.5 volt supply. With intermittent use the life of the batteries will almost equal their shelf life.

The batteries in themselves deserve more than a passing mention. In actual construction the cells are quite different from conventional carbon zinc cells. A typical alkaline-manganese cell uses a cylindrical depolariser (main constituent manganese dioxide) in contact with the cell container which, in this case, is nickel-plated steel. Since the steel is not acted upon by the electrolyte it can be the cathode as well as being a strong leak-free case. The depolariser surrounds the cylindrical granular zinc anode and electrolyte mix, the two electrochemical components being separated by a porous sheet.

The polarity is reversed from the conventional carbon-zinc cell in that the case is the positive terminal and the centre button is negative. However, some versions are made with the conventional polarity. The open-circuit voltage is around 1.4 volts and the cell 'end point' is, typically, 0.8 volt.

There is a popular misconception that these cells – or any cells which are not carbon-zinc – are rechargeable. This fallacy is probably based on confusion with the nickel-cadmium cell, but we must emphasise that it is a fallacy. Any attempt to charge these cells would generate gas and possibly cause them to explode.

Since the cells in the prototype were not supplied with solder tags it was necessary to make a battery holder for them. The holder was made from a piece of miniature tag-board with four tags on each side. The tags were bent perpendicular to the board and small brass strips soldered to them to make the necessary contacts. Two supply wires were soldered to the appropriate contacts, the batteries inserted and the whole assembly wrapped in tape as mentioned above.

The injector circuitry itself, apart from the potentiometer/switch and 400VW capacitor was assembled on a piece of Veroboard 1-3/8in x 1-1/8 in, with six copper strips running lengthwise. The wiring layout is as shown in the diagram. The three capacitors are rated at 25 volts but lower ratings can be used. We do not recommend the use of very low voltages, e.g., 3VW, as ceramic capacitors in this range can have low insulation resistance, by normal standards, and this may prejudice the operation of the multivibrator.

We suggest a minimum rating of 400VW for the output coupling capacitor so that equipment operating at high voltages may be tested without damaging the components of the injector and without danger of electric shock to the user.

The probe itself, which must be insulated from the case, was fashioned from a 1/8 whit. screw and nut and an old meter probe. Well there it is – quite an economical project. As mentioned above, plastic cases can be used, but care is required when drilling.

There are quite a number of uses to which this device can be put. Apart from trouble-shooting in audio amplifiers the harmonic components can be used to trouble-shoot the RF stages of radios etc. We found useful output up to 30MHz in a full coverage communications receiver with the signal injector applied to the aerial. This high-frequency harmonic content is due to the very fast rise time of the rectangular waveform.

Readers may well ask how the high-frequency harmonics of the signal become audible at the output stage of a radio. The tuned circuits into which the signal is fed will respond only to those harmonics which fall within their bandpass (frequency range). These are detected in the normal way. The harmonics are amplitude modulated by the fundamental frequency and its audible harmonics so that the output from the detector is in the audible range.

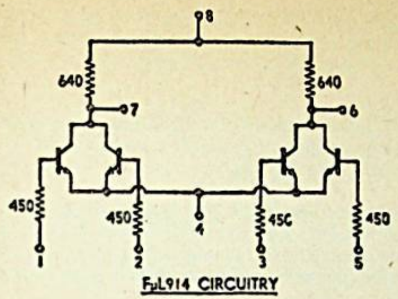
Another use for the signal injector forms the second part of this article. We decided to use it as the source of AC for a simple R-C Bridge. In normal use, the injector is connected to the AC input terminals as shown in the diagram. The level control is fully advanced for maximum signal level.

In the past all RC bridges we have described have used the 50Hz AC from the mains supply. This simplifies the construction of the bridge and is also economical. Many inexpensive commercial bridges have used the same economical approach. In commercial laboratory instruments the frequency used is usually much higher, typically 1000Hz or 1MHz. There are several reasons for this which we will not discuss here. As far as normal capacitance or resistance measurements are concerned the frequency is not critical provided it is between about 50Hz and 5KHz. Above the latter frequency the inductance of resistors and capacitors may add significantly to the impedance.

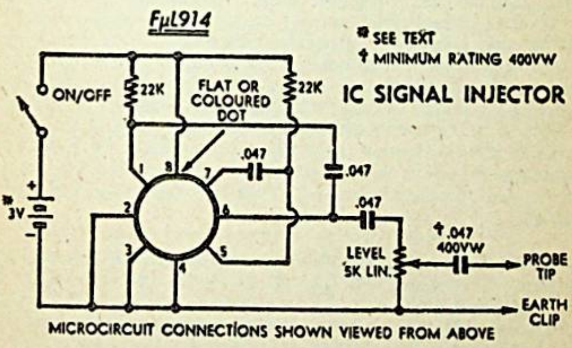
The principle of operation of RC bridges has been discussed many times before in previous articles but since this article is aimed specifically at beginners we will reiterate. The circuit is an AC version of the Wheatstone bridge. The bridge is used to find the ratio of a known resistance to that of an unknown resistance. When the ratio potentiometer has been adjusted so that its two arms (shown as AB and BC in the Wheatstone bridge diagram) have the same ratio as that of the know to the unknown resistance there will be no voltage across the null indicator (i.e. the null indicator will indicate zero).

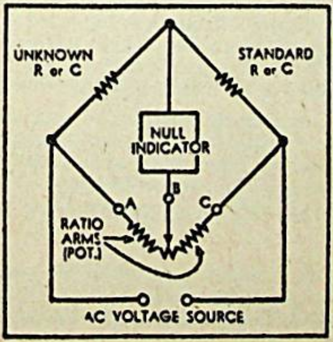
The unknown resistance can then be expected as the product of the known resistance and the ratio of the two potentiometer arms AB and BC. As far as resistance measurements are concerned the voltage source can be DC. In this case the null indicator is usually a centre-reading galvanometer. However, since this bridge is to be used for measuring impedance of capacitors, the input voltage to the bridte must be AC.

It should be noted that the accuracy of the Wheatstone bridge is limited only by the accuracy of the standards used and the sensitivity of the null indicator, the more expensive commercial instruments can measure a wide range of resistance from approximately 1 to 10 megohms with an accuracy of _.003 – using high precision substitution boxes and a mirror galvanometer. For resistance measurements above 10 megohms sensitivity can be a problem. This can be overcome by using a higher input voltage to the bridge or very high-gain DC amplifiers ahead of the null indicator.



The contents of the dual gate micro-circuit. Its operation is explained in the text.





The basic circuit of AC Wheat-stone bridge.

In the case of bridges used for measurement of resistance, capacitance and inductance the null indicator can take several forms, the most common being a centre-reading galvanometer or a 'magic-eye' thermionic indicator tube.

In our case the null indicator is a pair of phones. The accuracy of the 'null' will depend on the sensitivity of the phones and of the user's ears. More will be said about this later.

The bridge was built on a 'chassis' of plywood and hardboard, painted white. The dimensions were 8 inches by 5½inches by 4¼inches deep. All the components for the range switch were mounted on two sections of miniature tagstrip. The ratio potentiometer is a standard 5K linear taper type.

We have included six ranges for the bridge plus a switch position for an external standard. This requires a seven-position switch which would constitute a 'special' so we used a standard single pole eleven-position switch. This includes a tolerance scale. This (tolerance scale) is usually used on the external standard range and is brought into operation with a double-pole, two-position switch which adds a series resistor to each end of the ratio potentiometer. We did not incorporate this feature in our simple unit but for those who wish to do so the value of the resistors to be added to each end of the potentiometer is 2.2K. We found the scale to be 'close enough' for a simple instrument of this type expecially if low tolerance range standards are used. We do not think an instrument such as this justifies the purchase of high tolerance resistors and capacitors. We used 10 per cent capacitors and 5 per cent resistors.

The simple single arc scale we used on our experimental model has one limitation. It can be used to indicate resistance directly – if the balance pot is wired one way round – or capacitance directly – if the pot is wired in opposite sense – but cannot accommodate both. This is because the impedance of resistors i proportional to the resistance value, but that of capacitors inversely proportional to the capacitance value. This will mean that if the potentiometer is connected so that it indicates resistance directly, capacitance readings will be the reciprocal of the scale reading. Thus if the scale reading is 0.25 and the range setting is 1uF then the actual value of the unknown capacitor will be 4uF.

To enable such a scale to indicate resistance or capacitance directly, without the necessity of reciprocal calculations, it is necessary to use a double-pole multi-position range switch so that standard and unknown are switched to opposite sides of the bridge. Of course, the more complicated switch will be more expensive than a single pole type.

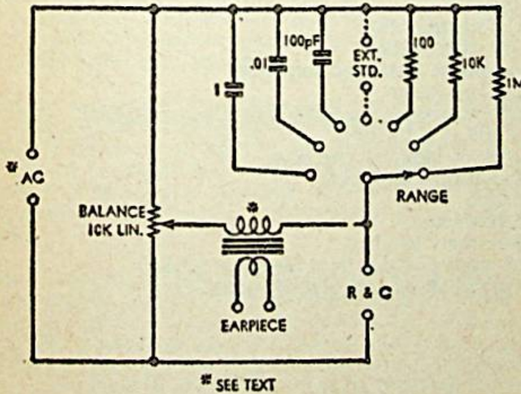
We decided to make up a calibrated scale with two ranges, one the reciprocal of the other, to enable readers to make the bridge with direct reading scale for both capacitance and resistance.

The range switch could be replaced by a flying lead with a crocodile clip and a row of banana sockets, each socket connected to a separate range. This would be a very economical approach.

readers will notice that we have shown an external standard position on the range switch. However, we did not provide terminals for an external standard on the prototype. Readers, who duplicate the prototype, can still connect an external standard, however. The external standard would be connected between the upper terminal for the AC input signal as shown on the circuit diagram and the 'Unknown' terminal which connects to the range switch rotor. For use with external standards, in this mode, the range switch should be on the appropriate setting (i.e. EXT. STD.) or an unused position.

The external standard position will enable the measurement of inductance but there is a catch. Most inductors have some resistance and if this is not small compared with the inductive reactance the measurements will be inaccurate. This is normally overcome in more elaborate bridges by having a variable resistance in series with the inductance standard to balance the resistance of the unknown. The same procedure is used for the power factor measurement of capacitors.

As mentioned above the null indicator is a pair of headphones. High impedance crystal types were found to be so insensitive as to be unusable. The best results were obtained from a pair of 2000-ohm dynamic phones. We also found that the low impedance earpiece as used in transistor radios is quite usable, especially if used with an audio output transformer to provide a better impedance match. We found that a 5000 ohms to 15 ohms transformer used with an 8-ohm earpiece gave good results.



R–C BRIDGE

The circuit of the practical R-C Bridge.

Instead of using headphones as a null indicator one could use an audio amplifier and speaker as null indicator, with the volume control used as a sensitivity control. Alternatively, one could rig up a simple amplifier, rectify the output, and feed it to a milliameter.

Other AC sources may be used to power the bridge. Any audio generator would be ideal. The frequency could be set to give the most output from the headphones used. Similarly, a small filament power transformer could be used. The disadvantage of this approach is that most headphones are very inefficient at 50Hz. A solution to this would be full wave rectification of the AC to obtain frequency doubling plus higher harmonics. The resulting waveform will be more audible with headphones.

PARTS LIST

1 FuL914 dual gate micro-circuit
3 .047uF/low voltage polyester or ceramic
1 .047 uF/400VW polyester
2 22K/½ or ¾ watt resistors
1 5K (1in) potentiometer with rotary switch
2 alkaline-manganese cells or 2 1.5v cells (see text)
1 Piece of Veroboard, dimensions as per text
1 Pill case or suitable probe case (see text)
Miscellaneous: Tagboard for battery-holder, insulation tape, hook-up wire, meter probe, alligator clip, solder lugs, etc.

R-C BRIDGE
1 Wooden chassis, 8in wide x 5¼in high x 4¼in deep
1 Single pole, 11-position switch
1 Pointer know
1 10K (1in) potentiometer
1 Large know with pointer
1 1.0uF/low voltage metallised polyester
1 0.01uF/L.V. polyester
1 100pF/L.V. polystyrene or mica (½ or ¼ watt resistors)
1 x 1M, 1 x 10K, 1 x 100 ohm
1 Low impedance earphone (see text)
1 Output transformer (see text)
2 7-lug tagstrips
4 Banana-type sockets
1 Jack socket to suit earpiece
1 Calibrated scale (available from the query service)
Miscellaneous: Hook-up wire, screws, solder, etc.

A NEW TRANSISTOR RF TEST OSCILLATOR

A further addition to our popular range of compact test instruments. It delivers a clean, stable CW or modulated signal, tunable from 360KHz to 18MHz in four contiguous bands, and is very suitable for general testing and alignment.

The instrument described in this article employs modern components in up-to-date circuitry and, while simple and economical in construction, it is capable of quite impressive performance; in fact the performance of the basic circuit is such that it could well form the basis of a more pretentious laboratory-type generator. These features should make the new design eminently suitable for use by service technicians, amateur radio enthusiasts and experimenters.

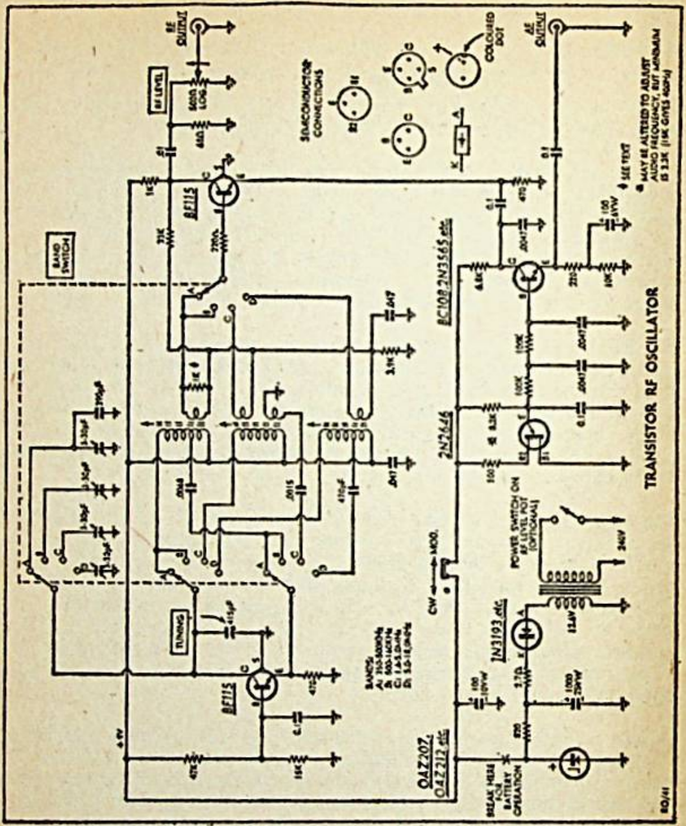
The RF section of the instrument uses two silicon planar transistors type BF115 or similar, one transistor being employed in the oscillator stage and the other in a buffer-modulator stage. The use of a buffer-modulator stage separate from the oscillator itself gives the instrument a high order of frequency stability, reduces loading effects to a low order and results in a very low degree of spurious frequency modulation of the carrier when it is modulated.

The oscillator stage is a fairly conventional transistorised circuit with the transistor operated in the common-base configuration. Coverage is continuous from 360KHz-18MHz, provided by four switched bands: however only three oscillator coils are used, the lowest (IF) band being provided "free" by switching fixed shunt capacitance across the coil used for the band next above. While primarily an economy measure, this technique gives the IF band a highly desirable "bandspread" characteristic—permitting easier scale resolution and a potentially higher reset accuracy.

Tuning is by means of a standard small 10-415pF single gang capacitor. Calibration of the bands to conform with pre-marked dial scales is performed by means of iron-dust and ferrite cores in the coils, for the low ends of the bands, and trimmer capacitors for the high ends. The trimmers are arranged so that they may be adjusted when the instrument is within its case, to compensate for stray capacitance changes.

Small secondary windings on the oscillator coils provide drive signal for the untuned buffer/modulator stage. The latter is basically a common-emitter stage arranged for light loading of the oscillator stage, low output impedance and substantially flat gain over the operating frequency range. The effective output impedance is 68 ohms, determined by the RF load resistor shunting the output level control.

In the present instance the output control is a simple 500 ohm potentiometer. However, those desiring a more refined circuit could easily replace both the 68 ohm resistor and the potentiometer with a full 70 ohm ladder attenuator system. Naturally if this course is



Although the new instrument employs but four transistors in relatively simple circuitry, its performance is impressive.

taken it will also be necessary to provide appropriately improved shielding of the instrument itself.

The basic performance of the RF section of the instrument should be sufficient to make this sort of elaboration quite feasible. As a result of the buffer between the oscillator and the output circuit, frequency disturbance due to loading variations is virtually negligible at approximately .001%. The stability is particularly good, drift amounting to approximately .07% maximum absolute from switch-on; as may be seen from the drift plots of figure 1, this compares favourably with typical valve-type signal generators of modest laboratory standard. As may be seen from the lower curve, drift after approximately 3 hours is very low, at less than .003%.

The AF section of the instrument employs a further two transistors, one a unijunction type 2N2646 (or 2N2160) and the other a general-purpose silicon planar NPN type BC108, 2N3565 or similar. The unijunction is used in a simple relaxation oscillator operating at around 1KHz; the approximately saw-toothed waveform at its emitter is passed through a high impedance low-pass filter to produce an approximately sine-waved signal at the base of the amplifier stage employing the NPN transistor.

Two output signals are taken from the amplifier, one from the collector, and the other from the emitter. The former is fed to the emitter of the RF buffer/modulator to produce modulation of the carrier, while the signal from the emitter is taken to the AF output socket for external test purposes. Note that the supply voltage for the AF stages is controlled by the "CW-MOD" switch, hence this section is operative only when the switch is in the "MOD" position.

The pitch of the audio tone delivered by the unijunction AF oscillator is determined by the resistor in its emitter circuit. The value of 8.2K shown on the circuit diagram gives approximately 1KHz, a useful and convenient pitch for many applications. However if another pitch is preferred the resistor value may be changed accordingly; for example, a value of 15K will give the often-used 400Hz. Note, however, that the resistor value should not be reduced below 3.3K to ensure reliable operation.

Power requirements for the instrument are 9V at 1.5-3.5mA, which could easily be provided by a small transistor radio battery if the instrument were required to be used in situations remote from the power mains. However, as mains operation is likely to be more convenient in the majority of situations, we have provided the instrument with a small internal power supply. This may be omitted if not required; the circuit is marked to show where a battery may be connected in its place.

The instrument is slightly frequency dependent upon DC supply voltage–the dependence amounting to approximately 0.1% per volt at 9V, and accordingly the mains power supply is regulated using a silicon zener diode. The diode is a nominal 9V unit, type OAZ207 or OAZ212 or a similar low-power unit. It is fed from a simple half-wave rectifier circuit using a silicon power diode type 1N3193, OA605 or similar. The power transformer is a miniature 12.6V unit.

As a result of the regulated power supply the instrument is virtually insensitive to mains voltage variations.

However if the instrument is to be operated from a battery the frequency/supply voltage dependence quoted above will become relevant and should be borne in mind. In many testing and alignment situations it may not be important (unless the battery is nearing complete exhaustion) but, if accurate work is involved, it may be necessary to either recalibrate the instrument or fit a new battery.

Perhaps it should be pointed out in passing that the instrument is by no means unusually or unduly sensitive to supply voltage variation. In fact the figure of 0.1%/V is considerably less than the sensitivity typically exhibited by many valve and transistor L-C oscillators.

Note that whereas the power switch is optional if the mains power supply is used, it will be virtually mandatory if a battery is used instead. The instrument drain is not high, but continuous operation would still result in somewhat limited battery life.

The instrument is constructed in our compact 7½in x 5in x 4in standard case, as may be seen. The front panel is undrowded and has but four controls and two output connectors. The tuning dial occupies a major portion of the panel, 4½in x 3in Perspex type with matching knob and planetary reduction drive having an approximate 6:1 ratio. For the prototype we made up a matching hair line cursor from a scrap of 1/16in plastic sheet.

At the right of the tuning dial are the band switch and output level controls, with the former uppermost. The remaining "CW-MOD" switch control is in the lower centre of the panel, with the AF output connector to the left and the RF output connector to the right.

Inside the case, a majority of the components are supported by a vertical bracket attached to the rear of the front panel via the band-switch and output level control nuts and two of the tuning dial mounting screws. The tuning capacitor and coils are mounted on one side of the bracket, while tagstrips on the other side support most of the minor components. The trimmer capacitors are mounted on the real turn-up of the bracket, such that they may be adjusted through small holes in the rear of the case.

For those who care to construct their own coils "from scratch" the details are given in the table. The bands A-B coil is wound on a 5/16in former with an iron-dust core and compensated shield can (which served additionally as a shield partition between the other two coils); the band C and band D coils are wound on ½in diameter formers fitted with ¼in ferrite cores. All output coupling windings are spaced 1/8in from the tuned windings.

The RF oscillator circuitry is supported by the bandswitch and the upper miniature tagstrip, while the AF circuitry is supported by the lower tagstrip. The small tagstrip in the centre supports the buffer/modulator components. The bandswitch section nearest the front panel is used for oscillator feedback and buffer drive switching, while the other section is used for oscillator coil and trimmer switching.

All "hot" RF oscillator wiring should be made in stoug (e.g. 16G) tinned copper wire, preferably wire which has been stiffened by stretching; the wire may be insulated where necessary with 1mm varnished cambric sleeving. The rigidity produced by this means will ensure stable internal capacitance conditions and contribute to stability and long-term reliability

Standard nylex-insulated multi- or single-strand hook-up wire may be used for the relatively "cold" AF and power supply wiring, as this is less important. However, care should be taken to ensure that components are well supported and the wiring will not be able to move about in transport. Care exercised in the construction will be found highly worthwhile in terms of stable performance of the completed instrument.

When the instrument is completed and found to be operating correctly it should be calibrated. Ideally this should be performed by comparing it with a standard signal generator, or by measuring the output frequency with a digital or heterodyne frequency meter. However, as it is probably unlikely that such instruments will be available to the everage constructor, the most convenient method will probably be detection using a calibrated short-wave receiver.

To calibrate the oscillator in this fashion, connect its RF output to the aerial terminal of the receiver (with the aerial disconnected) using a small series capacitor (i.e., 22pF). Then, with the oscillator switched to deliver a modulated output signal ("MOD"), it should be possible to detect the signal by tuning the receiver over the appropriate band. When it is desired to check that one has in fact found the oscillator signal, the easiest way is to listen whether the modulation tone disappears when the oscillator is switched for "CW" operation.

Note that because the instrument delivers a significant amount of second harmonic along with the fundamental output, it will be possible to detect weak signals at twice the fundamental output frequency. In cases where the oscillator frequency seems wildly in error, one should always check that it is not in fact the second harmonic which is being detected. Actually it is a good idea to check this point wherever possible, as it is easy to be misled.

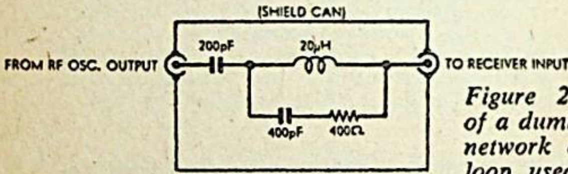
It will usually be necessary to make use of the second harmonic in calibrating the IF band of the instrument, as few receivers will have facilities for reception of signals between 350 and 500KHz. Thus the receiver broadcast band will be used to detect the second harmonics, which will fall between 700KHz and 1000KHz (1.0-MHz).

Another point which should be noted is that if the receiver used for calibration is a superheterodyne–and this is most likely–confusion and/or error may be introduced by the response of the receiver to signals at frequencies which are "IF images" of those corresponding to the dial markings. It may be remembered that a superhet by it very nature tends to respond to signals separated both above and below its local oscillator frequency by its IF, the selection of the "desired" frequency (usually the lower of the two) and the rejection of the "undesired" or "image" frequency being performed by the aerial and RF (if present) tuning circuits.

On the broadcast band the response of the receiver to images will generally be almost negligible due to the relatively large signal-image separation compared with the receiver front-end bandwidth. However, in many receivers the sensitivity to image signals becomes appreciable on the short-wave bands, particularly at the upper frequency end, as a result of the increase in the absolute bandwidth of the simple front-end tuning circuitry employed.

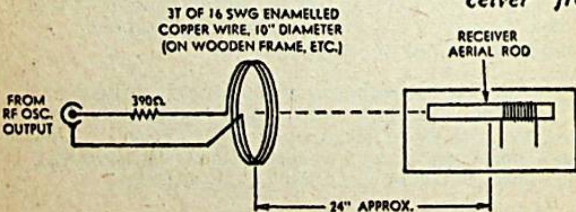
This being the case it will generally be advisable to commence calibration of the oscillator on the lower bands—where it will usually be easy to distinguish between the intended and image responses—and progress upward in relatively small increments which will permit one to "keep track" of the correct receiver response frequency. In addition it is worth bearing in mind that the two receiver responses are always separated by twice the receiver IF (i.e., 910KHz), and that the "true" response is usually that corresponding to the lower test oscillator frequency or the higher of two receiver dial settings.

If it is intended to produce an original scale for the instrument, the main concern will be to ensure that the four bands are contiguous—i.e., that they provide continuous coverage of the spectrum from 360KHz to 18MHz. However, if the instrument is to be calibrated to



(a) STANDARD DUMMY AERIAL NETWORK

Figure 2: Details of a dummy aerial network and field loop used for injecting oscillator signals into receiver front-ends.



(b) FIELD INJECTION LOOP

a copy of the prototype dial scale, continuous coverage will result automatically. In both cases adjustment will be made using the coil cores and trimmers, with the former used for the low frequency ends of bands, and the latter for the high frequency ends—in that order.

Because bands A and B share a common coil, the coil core cannot be adjusted independently for each. However, if the core is adjusted for the low end of band B, the low end of band A should automatically be correct. The recommended procedure is therefore to calibrate band B first—low end, then high end—and then to adjust the high end of band A using only its trimmer. The short-wave bands are of course independent and may be calibrated whenever desired. Remember to calibrate the low ends first, using the cores, and then the high ends using the trimmers.

When the instrument is calibrated and ready for use, probably in many cases its major role will be in the alignment and testing of radio receivers. It may therefore be appropriate to conclude this article with a brief discussion of these operations for the possible assistance of the inexperienced.

The general procedure when performing either testing or alignment of radio receivers is to commence at the "rear" or "output" end of the instrument, and work towards the "front" or "input" end, checking and/or adjusting each section or stage progressively. In this way the completed stages may be used to assist in the checking or adjustment of the remainder.

As the output section of a radio receiver normally deals only with AF or "audio" signals, it is checked by feeding an AF signal to its input. With the instrument just described this may be performed simply by connecting the AF output signal to the input grid or base, with the oscillator switch set to "MOD" in order that the AF section is operative. If the receiver audio section is operating correctly and the volume control is set to permit signal amplification, this should produce a 1KHz tone from the speaker.

Further tests concerning sensitivity, frequency response, distortion and signal-to-noise ratio require more elaborate instruments such as an audio signal generator, an electronic voltmeter, an oscilloscope and a distortion meter. However, the simple test just described can give at least a rough indication as to whether or not the audio section is operating normally.

With most modern superheterodyne receivers the next section to be checked is that comprising the last intermediate frequency ("IF") stage and the second detector. This is most easily checked by adjusting the oscillator to produce a modulated (RF) signal at the designated intermediate frequency (usually 455KHz or thereabouts). When this signal is fed into the grid or base of the last IF amplifier stage—via a 0.1uF blocking capacitor to prevent disturbing bias conditions—a tone should again be heard from the speaker. If the set is to be aligned, the core(s) in the last IF transformer should then be adjusted for a maximum or "peak" in the tone amplitude. If the transformer has two cores, these should be adjusted to produce peaks in positions where the cores are apart rather than together.

If there is more than one IF stage, the preceding stage(s) should be checked and aligned in a similar fashion by moving the point of signal injection to the earlier grids or bases progressively. It will probably be necessary to reduce the oscillator output each time this is done to allow for the gain of each stage and prevent overload of the later stages.

The mixer/converter stage and IF input transformer are checked by injecting the 455KHz modulated signal at the grid or base of the mixer—again using a 0.1uF blocking capacitor in series. The oscillator output may have to be increased at this point, as the impedance to ground is usually quite low as far as IF signals are concerned. At this stage the IF alignment may be completed by adjusting the IF input transformer cores for a peak as before.

Checking and alignment of the tunable "front end" portion of a receiver is performed in a manner almost identical with that used for the IF alignment, although here the peaking is performed at two different frequencies—at least; some receivers require adjustment at three frequencies, although nowadays such receivers are rather rare. Incidentally the method used is also applicable for receivers using the now obsolete "TRF" or tuned-radio-frequency system.

Briefly the technique is to peak the local oscillator, aerial and RF (if used) tuned circuits, in that order. Firstly at a frequency near the low end of the receiver band using the receiver coil cores and then at a frequency near the upper end of the band using the corresponding trimmer capacitors. In both cases the receiver and oscillator tuning dials should be set to the same frequencies, and the adjustments made to produce maximum tone output from the speaker. The two frequencies elected for broadcast band alignment are usually 600KHz (0.6MHz) and 1.3MHz (1300KHz).

Note that the oscillator tuning should always be performed before the aerial and RF tuning. Unless this is done it will not prove possible to obtain the aerial and RF peaks at all, at the dial settings concerned.

For receivers intended to operate with an external "capacitive" aerial-earth system, this final check and alignment should be performed with the signal from the oscillator fed to the aerial and earth terminals via a network known as a "standard dummy aerial". The circuit of the most commonly used network of this type is shown in (a) of figure 2. The idea of the network is to present to the receiver input circuit an approximate duplication of a typical aerial-earth system; if this is not done the RF alignment made with the oscillator would not necessarily give correct operation when the receiver was connected to an aerial and earth.

Receivers incorporating an internal loop or ferrite rod aerial cannot readily be aligned or checked by the method just described. However, in such cases it is possible to feed signals to the set by connecting the oscillator to a loop of wire which is placed about 24 inches away from the receiver and "broadside on" to the axis of the internal aerial. The details of a suitable injection loop are shown in figure 2(b).

Note that with receivers having ferrite-rod aerials, the ferrite rod itself is the "core" which determines the low-end tuning of the aerial input circuit. Thus the aerial peaking at the low end of the band is adjusted by sliding the coil along the rod until the signal peaks. Adjustment of the upper end of the band is by means of a trimmer capacitor as before.

The older type of portable receiver having an open loop aerial usually does not provide for adjustment of low-end aerial tuning. In such receivers it is only possible to peak up the aerial circuit at the upper end of the band.

And with those brief comments we must draw the present article to a close. It is hoped that our new RF oscillator will prove a worthy addition to our popular range of compact test instruments, and that readers will find it as useful and reliable as the prototype performance would suggest.

SPECIFICATION

An RF test oscillator producing either CW or modulated signals which may be tuned continuously from 360KHz to 18MHz in four switched bands. Fully solid-state design using four silicon transistors. RF output approximately 100mV RMS, controlled by a simple low resistance potentiometer. A more elaborate 70 ohm ladder attenuator system could be used if required.

Modulation fixed at approximately 25%. AF output fixed at approximately 200mV RMS. Tone fixed at nominal 1KHz or 400Hz or as desired. Frequency drift approximately .07% maximum. Frequency shift due to loading variations approximately .001%; shift due to 10 per cent mains voltage variation less than .02%; shift from CW to modulated modes less than .02%. FM extremely small. Operates from either 240V AC mains or 9V battery. AC power consumption less than 4 watts; battery current drain 3.5mA maximum.

COIL DATA

All coils wound with 32 S.W.G. Enamel Wire on ½" diameter formers.

Band A & B	Primary 210 T tapped at 15T Secondary 80T wound 1/8" from cold end,
Band C	Primary 45 T feedback winding 7T, output winding 3T 1/8" from hot end.
Band D	Primary 15T tapped at 3¾T. Secondary 1½T 1/8" from hot end.

OSCILLATOR PARTS LIST

1 case, 7½in x 5in x 4in, with wrap-around front panel
1 Dial assembly, 4½in x 3in, with 6:1 reduction unit, knob, cursor
1 10-415pF single gang variable capacitor
1 2-section, 2-pole, 4-position, small rotary switch
1 500 ohm switch potentiometer
1 Set of coils (see text)
1 Power transformer, 240V-12.6V at 150mA
1 On-off slider switch
2 Co-axial connectors
2 small instrument knobs
4 3-30pF miniature compression-type trimmers

SEMI-CONDUCTORS
2 BF115 or similar NPN silicon planar
1 2N2646, 2N2160 or similar unijunction
1 BC108, 2N3565 or similar NPN silicon
1 1N3193, OA605 or similar
1 OAZ207, OAZ212 or similar 9V zener

RESISTORS (Half-watt, 5 p.c.)
1 x 2.7 ohm, 1 x 68 ohm, 1 x 100 ohm, 2 x 220 ohm, 2 x 470 ohm, 1 x 820 ohm
2 x 1K, 1 x 3.9K, 1 x 6.8K, 1 x 8.2K, 1 x 10K, 1 x 15K, 1 x 33K, 1 x 47K, 2 x 100K

CAPACITORS (LV plastic type)
1 x 390pF, 1 x 470pF, 1 x .0015uF, 3 x .0047uF, 1 x .0068uF, 1 x .01uF, 2 x .047uF, 4 x 0.1uF
Electrolytics: 1 x 100uF 6VW, 1 x 100uF 100VW, 1 x 1000uF 25VW

MISCELLANEOUS
Handle, rubber feet and case assembly screws; scrap aluminium sheet for mounting bracket; miniature tagstrips–2 x 10-lug, 1 x 4-lug, 1 x 3-lug. Mains cord and plug, screws and nuts, washer, connecting wire, solder, etc.

A NEW SOLID-STATE AF SIGNAL GENERATOR

An up-to-date laboratory instrument design employing silicon transistors and a microcircuit, capable of delivering either sine or square waves over a wide frequency range at low distortion up to 10V RMS amplitude.

SPECIFICATION
An AF signal generator which delivers high-quality sine or square wave signals tunable continuously over the range 3Hz–300KHz in five switched bands. The instrument is fully solid-state and employs nine silicon transistors, five diodes and an integrated RTL microcircuit.

Maximum output level is greater than 10V RMS, with an output impedance of less than 600 ohms. Both coarse and fine attenuators are provided for adjustment of output level, the former providing a total of 70dB attenuation in 10dB steps while the latter control provides more than 20dB of additional stepless attenuation. The output level applied to the final (coarse) attenuator is monitored by an output level meter which is calibrated in volts, millivolts and decibels.

Output level is flat within 0.5dB over the full frequency range. Total harmonic distortion, hum and noise at 10V RMS sinewave output is less than .06% between 100Hz and 10KHz, rising to 0.1% at 50Hz and 30KHZ and to approximately 0.25% at 10HZ and 300KHz. Squarewave rise and fall times are less than 100nS; overshoot and droop at 5Hz are both less than 10%. Above 10Hz droop negligible.

Output level meter response is flat within 0.5dB from 3Hz–300KHz. The instrument operates from 240V AC, having a power consumption of approximately 6 watts.

The AF signal generator to be described has been designed as a modern solid-state counterport of previous valve designs. It offers the full functional and performance features expected of a laboratory signal generator, yet may be constructed at an outlay considerably less than the cost of comparable commercial instruments.

The frequency range covered is from 2Hz to 300KHz, the instrument having five switched decade bands which are continuously tunable. Alternative output waveforms are available over the full frequency range, either low-distortion sine waves or square waves. Total harmonic distortion, hum and noise components at maximum output (10V RMS) for sine wave output is less than .06 per cent between 100Hz and 10 KHz (see specification panel); while the squarewave rise and fall times are less than 100nS, with overshoot, and droop at 5Hz both less than 10 per cent. At frequencies above 10Hz the droop on squarewave is negligible.

The output level is maintained flat within 0.5dB over the full frequency range, and is monitored by a 3-inch rectangular panel meter. Both coarse and fine output attenuators are provided, the former giving a total of eight 10dB steps, while the latter provides more than 20dB of additional stepless attenuation. By means of the output meter the output level may be set conveniently and accurately to any level between 10V RMS and a few hundred microvolts, at a source resistance of less than 600 ohms on all settings.

The instrument operates from 240V AC, and has a power consumption of approximately 6 watts. It is housed comfortably in a case measuring a compact 12in x 6in x 4½in.

During the development of the present design a considerable number of AF oscillator basic circuits and configurations were critically examined and compared in our laboratory. The configurations tested included many based on integrated micro-circuit amplifiers, and using field-effect transistors both for impedance matching and automatic level control.

Although it was found possible to obtain quite good performance from some of these configurations, to date we have been unable to find a circuit or configuration which is capable of performing as well as a discrete-components circuit using bipolar silicon transistors. Hence the latter approach would appear to be capable of providing close to the best performance available at the current state-of-the-art from circuitry of moderate complexity, and it is for this reason that we have adopted it for both the recent and present projects.

The oscillator circuit is of the "Wien bridge" type, and for those unfamiliar with this configuration it is shown in basic form in figure 1. As may be seen it consists of a high-gain differential amplifier fitted with two separate feedback circuits. One circuit, consisting of resistors R1 and R2 and capacitors C1 and C2, connects from the amplifier output back to the "+" input and therefore provides positive feedback; the other circuit consists of resistor R3 and the thermistor (negative temperature-coefficient resistor), and is connected between the output and the "-" input to provide negative feedback.

The configuration in which R1, R2, C1 and C2 are connected is knowns as the "Wien network", and is capable of performing in a manner roughly corresponding to an L-C tuned circuit. At a particular frequency determined by the values of the four elements (and possibly modified by source and output loading impedances), the transmission loss of the network falls to a minimum while the phase shift also passes through zero.

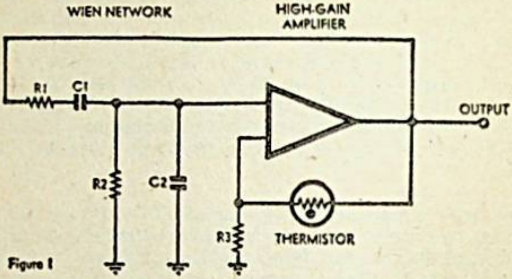


Figure 1

This "pseudo-resonance" occurs at frequency Fo, where Fo may be found from the following equation providing the driving source impedance is negligibly low compared with the series elements (R1, C1) and the output loading impedance is negligibly high compared with the shunt elements (R2, C2):

$$Fo = \frac{1}{2\pi\sqrt{R1.R2.C1.C2}} \quad \ldots (1)$$

It may be noted that the right-hand side of this equation bears a formal resemblance to the expression for the resonant frequency of an L-C tuned circuit.

If R1 and R2 are made equal in value, and C1 and C2 similarly given equal values, equation (1) reduces to

$$Fo = \frac{1}{2\pi\ RC} \quad \ldots (2)$$

In this case the transmission loss falls at Fo to a minimum of 3.0 —or in other words there is a maximum transmission "gain" of 0.33. The phase shift is zero as before.

For an active system such as a feedback amplifier to produce standard oscillations at a particular frequency, it may be recalled, the overall gain around the feedback loop must be at least unity and the phase shift either zero or a multiple of 360 degrees. In simple terms this means that the amplification must have a gain at least equal to the loss of the feedback network, and have a phase shift such that the output produced in response to a feedback signal is synchronous with the output signal from which the feedback signal was derived.

Hence for the circuit of figure 1 to produce sustained oscillations at the "pseudo-resonant" frequency, the gain of the amplifier must be at least 3 times, to compensate for the loss in the Wien network. And the phase shift must be either zero or a multiple of 360 degrees, as the phase shift of the Wien network at Fo is zero.

While the criterion for sustained oscillation of such a system is simply that the overall loop gain be at least unity for zero phase shift, a loop gain of higher than unity is undesirable because this corresponds to oscillations which grow in amplitude. What in fact happens with a loop gain exceeding unity is that the oscillation amplitude grows until saturation, cutoff or other non-linear limiting effects within the amplifier act to reduce the instantaneous loop gain on signal peaks. An equilibrium is then reached, with the oscillation amplitude remaining constant–but with significant distortion of the waveform on one or both peaks.

In order to produce sustained oscillations at a constant amplitude and with low distortion, therefore, it is necessary to maintain the overall loop gain accurately at unity. There are a variety of methods whereby this may be achieved, but probably the most generally gives quite good results is that shown in figure 1. As may be seen it involves a negative feedback circuit consisting of resistor R3 and a thermistor.

In theory, providing the amplifier has an open-loop gain of greater than the required 3 times, simple "linear"negative feedback could be used to adjust its gain accurately to that figure. However, the feedback would have to be critically adjusted, both initially and thereafter on an almost continuous basis, to compensate for amplifier gain and Wien network drifts.

By using a thermistor in the feedback circuit, in the position shown, the circuit is made to perform in its own continuous and automatic gain adjustment. This occurs in the following manner:

The thermistor has a negative temperature coefficient of resistance–in other words, its resistance falls as its temperature rises. Hence when the circuit of figure 1 is first switched on the thermistor will have a high resistance, there will be little negative feedback, and the high gain around the positive feedback loop will speedily initiate oscillations of rising amplitude.

As the oscillations grow, the temperature of the thermistor will rise also, as the latter and resistor R3 are effectively connected in series across the amplifier output and will accordingly draw signal current. Hence the resistance of the thermistor will fall, the negative feedback will increase and the effective amplifier gain will drop.

It should be apparent that an equilibrium will be reached, as the output amplitude can only rise to the point where the thermistor has increased the negative feedback on the amplifier to correspond to an effective gain of 3 times–giving unity loop gain. If the oscillation amplitude tends to rise above this level, the thermistor will reduce the loop gain slightly below unity and the oscillations will begin to die away; conversely if the amplitude tends to fall, the thermistor will increase the loop gain slightly above the unity to correct it.

In short, the "non-linear" negative feedback action produced by the thermistor acts to continuously and automatically maintain the loop gain accurately at unity, and the oscillation amplitude constant. By employing a thermistor with a suitable temperature/resistance characteristic, the output amplitude may be maintained at a level well below amplifier limiting, and the output waveform may thus be arranged to have low distortion.

The actual degree of distortion present in the output will naturally depend largely upon the linearity of the amplifier itself, as one might expect. In order to produce an oscillator with very low distortion, it is therefore necessary to base the design in the first instance upon an amplifier having low inherent distortion at signal levels below limiting.

The amplifier configuration employed in the present design may be seen by reference to the main circuit diagram. It comprises four silicon NPN transistors T1, T2, T3 and T4, together with associated components.

The first stage employs T1 in a conventional common-emitter configuration, with the base connected directly into the Wien network and the negative control feedback signal applied to the emitter via the unbypassed 470 ohm resistor (equivalent to R3 in figure 1). The base bias divider for T1 forms part of the resistive shunt Wien element (R2), the remainder being formed by a 50K wire-wound potentiometer. The latter is ganged to a similar pot in the series arm, the two forming the "fine tuning" control of the instrument. A fixed 4.7K resistor in series with the second pot forms the remainder of the series resistive element (R1), balancing the equivalent resistance of the base bias divider.

The capacitive Wien elements (C1 and C2) are switched to provide the five decade tuning ranges. It will be noted that the values of the pairs of capacitors are related by a factor of 10 for all but the highest frequency range, where the values are lower than might be expected. The reason for this is that stray wiring capacitance in fact provides the remaining capacitance for this range.

Further stray capacitance across the bias divider also tends to restrict the extent of the highest range. This is compensated by the 4.7pF capacitor across the 4.7K series element, which thus ensures that the highest band maintains the correct decade ratio.

Transistor T2 forms the second stage of the amplifier, and is again a conventional common-emitter stage which is direct coupled to the collector of T1. A series R-C step circuit connected from the collector of T2 to earth is used to modify the loop gain/phase characteristic of the amplifier at high frequencies, ensuring that the oscillator remains free from parasitic oscillations on the highest range.

Transistor T3 is used as an output emitter-follower stage for the amplifier, contributing a slight voltage loss but appreciable current gain. This gives the amplifier a low output impedance and thus ensures correct operation of the feedback circuits.

In place of the usual resistive load for T3, a fourth transistor T4 is used, biased to draw a collector current equal to the optimum emitter current of T3. As a result of the high effective AC collector-emitter resistance of a bipolar silicon transistor when biased in the "pentode" or "constant current" region, T4 thus provides T3 with a load which represents at one and the same time a low DC resistance combined with a high AC resistance.

Because of this, the peak-to-peak emitter current excursions of T3 for a given output signal amplitude are a minimum. Accordingly, the current gain of T3 varies far less than is usually the case during the signal cycle, and harmonic distortion is reduced considerably.

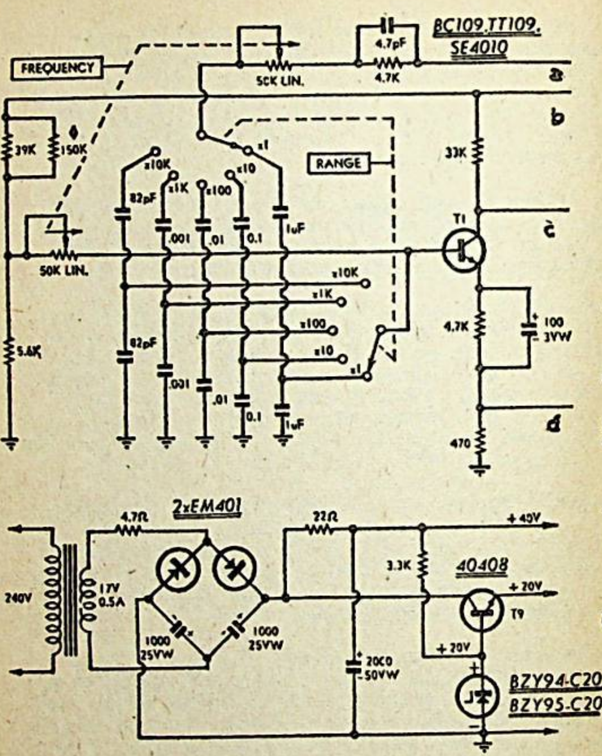
The actual distortion level produced by the oscillator depends in a rather complex fashion upon the DC operating conditions in the various stages of the amplifier. However, in most cases there will be a rather well-defined minimum in output distortion, at a particular overall operating point. Hence in order to set the circuit for minimum distortion, the most convenient method is to adjust the bias on T1 by varying the high-value shunt across the upper bias divider resistor (identified on the circuit with a diamond symbol).

The minimum distortion point will generally correspond approximately to the "half supply voltage" condition at the emitter of T3; i.e. the emitter voltage of T3 will tend to be somewhere near +10V. However, some idea of the possible deviation from this situation may be gained from the fact that the prototype unit shown produces minimum distortion with the emitter of T3 at +12V.

The thermistor used in this circuit is an STC type R24, whose resistance/temperature characteristic is such that the oscillator output level will fall between about 1.2 and 1.5V RMS. At this level the dynamic control response of the R24 thermistor provides quite good oscillator amplitude stability, both with respect to disturbance recovery and also to ambient variations.

In order to provide the instrument with alternative squarewave output, switching is provided which, when required, directs the output from the oscillator through a squaring circuit. The latter amploys an integrated digital microcircuit–a Motorola device, type MC792P, which is a triple three-input gate based on RTL circuitry. Two of the MC792P gates are wired in a Schmitt-trigger type configuration, which provides initial squaring of the oscillator sinewave, while the third gate provides follow-up squaring and buffering.

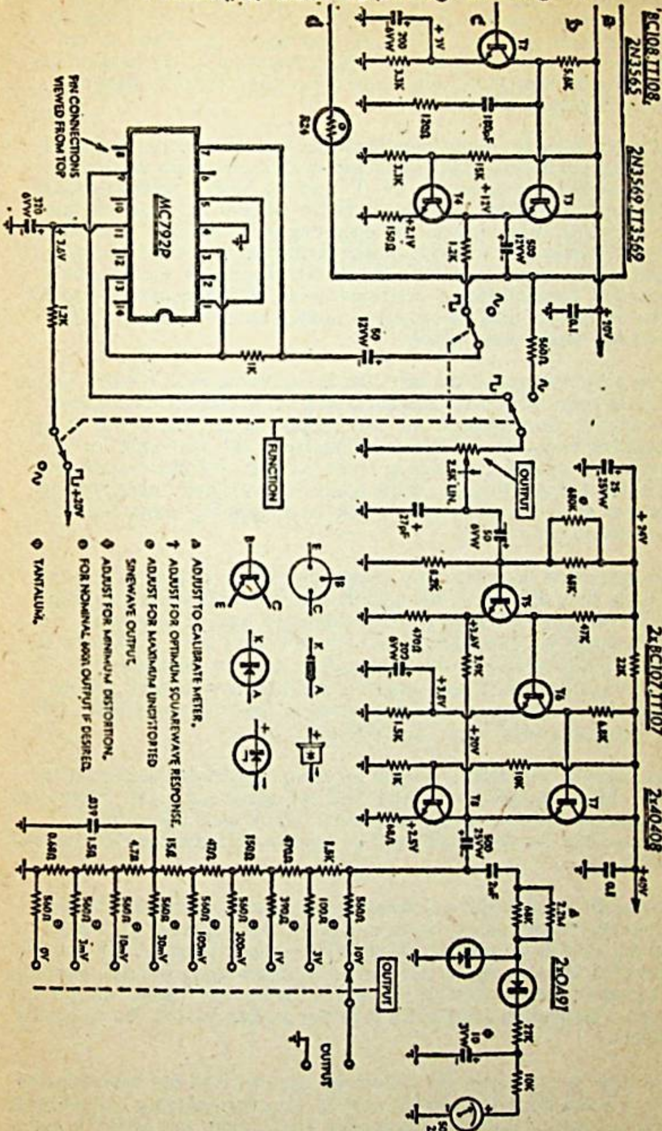
The output amplitude from the squaring circuit is approximately 3.2V P-P, which is close to that of the oscillator itself.



A.F. SIGNAL GENERATOR

Connects with drawing on Page 51

Connects with drawing on Page 50



In order to fully qualify as a laboratory-grade instrument, it is necessary that an AF signal generator be capable of delivering an output level of at least 10V RMS, at a source impedance of no higher than approximately 600 ohms. Accordingly the present instrument is provided with an output buffer amplifier designed to satisfy these criteria.

The amplifier is a three-stage feed-back circuit based on the same configuration used in the oscillator section, and again using four silicon NPN transistors T5, T6, T7 and T8. Heavy negative voltage feedback is applied, operating both for DC and AC, and this provides a high order or gain and operating point stability. The effective signal voltage gain of the amplifier is approximately 9.3 times, and within approximately 0.2dB of this figure over the frequency range of the oscillator–a performance which is adequate to provide the instrument with an output of more than 10V RMS on both sine and square-wave signals.

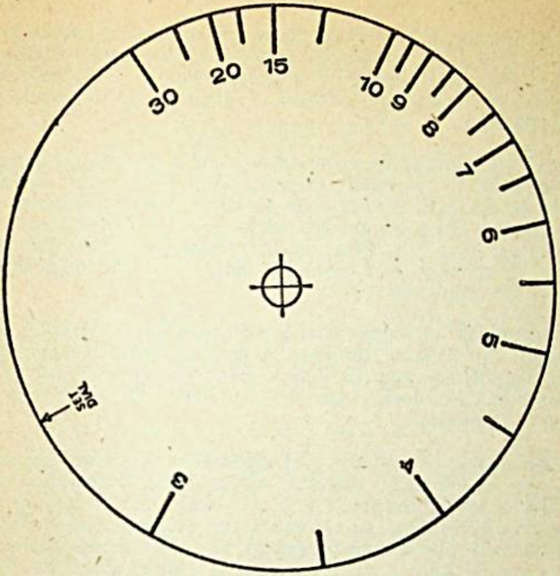
The output stages of the amplifier are operated from a +40V supply rail, in order to provide the required peak-to-peak output voltage capability. As a result of the high supply voltage, the use of heavy negative feedback, and the use of transistor 27, the amplifier is typically capable of delivering more than 10.5V RMS output over the full frequency range of the oscillator, with hum, noise and distortion at an almost unmeasureable level (typically around –90dB relative to full output).

In order that the instrument should provide an output impedance of less than 600 ohms, the amplifier output stage must be capable of developing 10V RMS across a minimum load of this value, with negligible distortion. Hence as the output stage is basically an emitter-follower operating in class A it must be operated at a quiescent current at least equal to, and ideally somewhat greater than, the peak signal current required in the latter condition. This is equal to (1.414 times 10)/600, or 23.6mA.

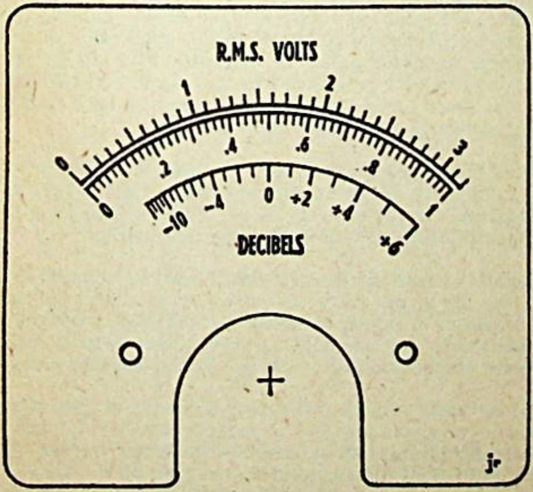
To allow a comfortable margin for component tolerance variations and similar factors, the output stage quiescent current in the present design has been set at approximately 37mA. This figure should ensure that the instrument will be capable of delivering rated output at all times.

Naturally enough this quiescent current level, combined with the moderately high supply voltage used for the output stage, results in appreciable power dissipation in the output transistor T7 and its load T8. Each in fact is required to dissipate some 740mW, dictating the use in these positions of a device having both higher voltage and higher dissipation capabilities than the devices used in the foregoing stages.

As may be seen from the circuit the device which has been selected for T7 and T8 is the 40408, an RCA type employing a metal-header TO-5 encapsulation. For the present application, this device is quite conservatively rated, with a BVceo rating of 90V and a dissipation



The frequency dial of the new generator.



An actual-size reproduction of the prototype meter face.

capability of some 860mW at 50 degrees C. ambient. However, to provide as large a "safety margin" as possible for the instrument, in keeping with its pretentions, it is recommended that the devices be fitted with small spring-clip radiators to lower the case-ambient thermal resistance.

The undistorted output capability of the buffer amplifier is naturally dependent upon the quiescent operating point, as with the oscillator amplifier. In this case, the optimum operating point corresponds closely to the situation where the emitter voltage of T7 is at half the supply voltage–i.e., 20V. As before, setting-up of the operating point is performed by means of a high-value shunt across the upper bias resistor of the input stage (T5).

"Fine" adjustment of output level is performed by a potentiometer at the input to the buffer amplifier. A small capacitor shunting the rotor of the potentiometer to earth is used to compensate for stray capacitance break-through at the highest frequencies (evident only as squarewave overshoot).

Coarse attenuation of output level is performed by a low-impedance divider chain at the buffer amplifier output. The resistor values used in the divider are multiples of 15 and 4.7, which are the preferred values giving attenuation steps closest to the required 10dB (within 1 per cent, excluding resistor tolerance). The total divider resistance is approximately 2.2K. A small capacitor is shunted across the three lowest ranges for stray capacitance compensation at the highest frequencies.

A 560 ohm resistor is fitted in series with the output on the zero attenuation (10V) range, to ensure that the buffer amplifier output stage cannot be disturbed or damaged by the application of very low resistances across the output of the instrument. Similar resistors are not required for the remaining ranges, but can be fitted where shown if it is desired that the instrument present a fairly constant output impedance of nominally 600 ohms.

Output level monitoring is performed by a passive half-wave voltage doubling rectifier and meter circuit connected to the output of the buffer amplifier. The meter used is a 3in rectangular 50uA type, that used on the prototype instrument being a Japanese unit type VP-2A.

As the metering circuit is a passive one, it must be connected as shown to the top of the coarse attenuator. This is quite standard practice. However, it should be noted that because of this the meter reading will strictly indicate only the open-circuit output level for any given combination of coarse and fine attenuator settings.

This must be borne in mind particularly when the output of the instrument is connected into low impedance circuitry. In such cases, the meter reading will indicate not the actual output, but rather the signal level being applied to a voltage divider formed by the load impedance, together with the effective output resistance of the instrument on the coarse attenuator range concerned.

The power supply of the instrument consists of a "full-wave" voltage doubler rectifier, employing a power transformer having a nominal secondary rating of 17V at 0.5A. A single section of R-C filtering is used following the rectifier for the +40V supply line, while a simple zener-reference series regulator circuit is used for the +20V oscillator supply line. The series regulator transistor is again an RCA type 40408 device, fitted with a small metal clip radiator as with T7 and T8.

As may be seen, the instrument is housed in a compact case measuring 12in x 6in x 4½in, the front panel of the case being of the "wrap-around" type.

The frequency dial is on the left of the front panel and consists of a calibrated circular plate mounted on a planetary reduction unit. A small fixed cursor fashioned from a scrap of acrylic sheet is mounted above the dial, permitting convenient and quite accurate adjustment.

In the centre of the panel (top to bottom) are the sine-square function switch, the output "fine" or "set level" control, and the frequency range switch. To the right of the last-named and underneath the output level meter are the coarse attenuator switch and the output terminals. Matching handles at either end of the panel permit convenient handling of the instrument, while also tending to protect the meter and control knobs from damage.

The meter is fitted with a face having three scales, the two uppermost scales providing the usual 10dB-related 0-1 and 0-3.16V ranges, while the lowest provides for relative dB measurements (FSD is + 6dB). The face for the prototype meter was produced photographically from original artwork and duplicate copies are available to readers via the Information Service at 50p each. We understand that in due course the meters may be made available with this face fitted by the makers.

Inside the case of the instrument a majority of the minor components are mounted on a printed wiring board, which measures 10in x 5in. The board is mounted to the rear of the front panel by two small sheet-metal brackets which are fastened to the panel via the handle mounting screws. The ganged potentiometer which performs the oscillator tuning is also mounted on the board, being coupled to the rear of the planetary reduction drive via a flexible ¼in-to-¼in shaft coupling.
The buffer amplifier output transistors are attached to the wiring board via small spring clips, which, as mentioned earlier, provide additional area for thermal dissipation. The clips used in the prototype are of the type sold in most hardware stores for use as cupboard door catches and as mounting clips for small tools and utensils.

The only minor components associated with the oscillator-amplifier circuitry proper which are not mounted on the board are the coarse attenuator resistors, which are mounted on the attenuator switch and supported by a dummy switch wafer, the attenuator compensating capacitor wired between rotor and ground of the fine attenuator.

Apart from these components there are the metering components, which mount on a small section of miniature resistor panel attached to the rear of the meter, and the power supply components.

The power supply components are supported by an "L"-shaped metal bracket which mounts in the case at the rightside rear. The bracket is designed both to support and shield the supply from the oscillator and amplifier, and to this end the power transformer is mounted inside the bracket rather than on the outside. As the 2000uF filter electrolytic has a grounded can, this is mounted outside the bracket; however, the remaining components fit comfortably inside supported as necessary by three tagstrips. A fourth 2-lug strip supports the 40408 radiator clip.

The wiring board is best wired up as the first step in assembly, after which the output attenuator may be wired and the connections made between the board, the frequency range switch and the attenuator switch. The meter circuitry may then be assembled on the rear of the meter, and finally the board mounted on its brackets and the connections made to the fine attenuator and function switch. The power supply section may be wired last, and in the interests of safety tested briefly before connection to the board.

Care should be taken when wiring the transistors and integrated microcircuit to the board, to ensure that these components are not damaged by overheating. A small, well-tinned iron is desirable, with the joints made thoroughly, but with heat applied for as short a period as possible. It is recommended that the transistor leads be cut not shorter than about ½in, unless the constructor is well experienced in the art of printed wiring assembly.

To prevent possible vibration damage of the thermistor it is a good idea to anchor it to the board with a piece of cellulose tape after soldering it into circuit.

When the board has been attached to the front panel, the planetary reduction drive and dial may be assembled. Before tightening the flexible coupling overall accuracy of within 5 per cent, or to go to the trouble of calibration using a technique such as Lissajous figures. Which of these alternatives is taken will no doubt depend upon the calibration accuracy required.

When connecting the drive to the tuning pot, ensure that the pot is turned fully anti-clockwise and that the dial is set accurately to the "set dial" arrow.

Once the instrument has been completed, the one remaining task is that of calibration and adjustment for optimum performance.

Frequency calibration of the instrument is most easily carried out with a digital frequency meter; in general this will also give the most accurate results. However, if a digital frequency meter is not available, the constructor has to accept the calibration provided by the particular range capacitors and resistors employed.

The low-end calibration of all of the ranges in common is determined by the ganged potentiometer, while the common high-end calibration is determined largely by the fixed series feedback resistor (shown on the circuit as 4.7K). The absolute calibration of each range individually is determined by the two capacitors concerned, while as mentioned earlier the small capacitor across the fixed feedback resistor largely determined the calibration at the high end of the uppermost range. These relationships should be borne in mind if it is found necessary to modify component values during calibration.

Adjustment of the oscillator operating poiht is necessary if the instrument is to give minimum distortion and this can really only be done by trial and error using a high-quality harmonic distortion meter or wave analyser. Measurements should be taken with a variety of values for the high value bias divider shunt (marked on the circuit with a diamond symbol) and from the values it should be apparent which value will give the minimum distortion.

Perhaps it should be noted in passing that the last-described adjustment will only be necessary when it is in fact made possible by the availability of the required instruments; if the instruments are not available, there should be no reason to perform the adjustment. It is likely that with no special setting-up of the oscillator bias the distortion level of a typical instrument will still be below 0.2 per cent over a major part of the frequency range.

Adjustment of the quiescent bias on the buffer amplifier may be necessary to ensure that the amplifier is capable of delivering maximum output without clipping. The easiest way of performing this adjustment is to connect an oscilloscope to the output of the instrument and to adjust the bias until the onset of clipping is symmetrical when the fine attenuator is turned to maximum (corresponding to somewhat more than 10V RMS). Bias adjustment is performed by alteration of the value of the high-value shunt in the bias divider of T5, as noted earlier.

Calibration of the output level meter circuit is best performed by connecting to the instrument output a high-impedance AC voltmeter of known high accuracy, such as a digital voltmeter. If a high impedance meter is not available, suitable allowance will have to be made for the voltage division ratio which will relate the meter reading with the output level. Note that the meter calibration should be made at full-scale deflection for greatest accuracy, with the instrument set for sinewave output and the coarse attenuator set for the 10V range (i.e., the meter should be calibrated at 10V RMS).

It should be noted that the meter circuit calibration should be performed only after the buffer amplifier bias has been adjusted to permit the latter to supply 10V RMS without clipping. If the two operations were done in reverse order, the meter calibration might well be significantly in error due to the waveform distortion.

LIST OF COMPONENTS

1 Case 12in x 6in x 4½in, with wiring board and power supply brackets
1 Rotary dial plate (see text)
1 wiring board
1 50uA 2K meter, 3in rectangular
1 Power transformer, 240V to 17V at 500mA nominal
1 3-pole 3-position rotary switch
1 2-pole 5-position rotary switch
1 single-pole 9-position rotary switch, with dummy wafer
5 Control knobs
2 Output terminals
2 Chrome handles
1 Type R24 thermistor
1 Planetary reduction drive, ¼in to ¼in
1 Flexible shaft coupling, ¼in to ¼in

SEMICONDUCTORS
2 Silicon rectifiers type IN5059, OA605 or similar
2 Germanium diodes type OA91 or similar
1 Nominal 20V zener, type BZY94-C20 or similar
1 BC 109
2 BC107
1 BC108
2 2N3569
3 40408 or similar
1 MC792P integrated circuit

RESISTORS
1 2.5K linear pot
1 50K-50K ganged wire-wound pot, high quality type
Half-watt 5 per cent type:
4.7 ohms, 22 ohms, 68 ohms, 120, 150, 2 x 470, 2 x 560, 2 x 1K, 2 x 1.2K, 1.5K, 3 x 3.3K, 3.9K, 2 x 4.7K, 2 x 5.6K, 6.8K, 8.2K, 2 x 10K, 15K, 2 x 22K, 33K, 39K, 47K, 2 x 68K, 150K, 680K, 2.2M
High stability, preferably 1 per cent type:
0.68 ohms, 1.5 ohms, 4.7 ohms, 15 ohms, 47 ohms, 150, 470, 1.5K

CAPACITORS
Low voltage, polyester or polystyrene:
4.7pF, 27pF, 180pF, .039uF, 2 x 0.1uF, 2uF
High stability, close tolerance:
2 x 82pF, 2 x .001uF, 2 x .01uF, 2 x 1.0uF
Electrolytics:
10uF 3VW (tantalum), 25uF 25VW, 50uF 6VW, 50uF 12VW, 100uF 3VW, 2 x 200uF 6VWm 320uF 6VW, 500uF 12VW, 500uF 25VW, 2 x 1,000uF. 25VW, 2,000uF 50VW

MISCELLANEOUS
Mains cord and plugs; grommets; spring clips for 40408 transistors; 8-lug length of miniature resistor panel; 2 x 8-lug tagstrips; 1 x 5-lug tagstrip; 1 x 2-lug tagstrips; scrap acrylic sheet for dial cursor; conneeting wire, solder, nuts, screws, washers, etc.

SIMPLE SHORTED-TURNS TESTER

A low-cost instrument small enough to fit into a tool kit. Easily built, it provides a reliable and rapid way of detecting shorted turns in TV line output transformers, deflection yoke windings, and other coils and transformers.

Line output transformers, deflection yoke windings, width coils and other ferrite or air cored coils in TV receivers can easily develop a fault in the form of leakage or a short-circuit between turns. Yet a fault of this type can be very difficult to isolate. Unless the component is obviously charred or otherwise showing evidence of overheating it is generally necessary to check out all of the likely alternative causes of the fault symptoms, and then if the cause is still not apparent, try substituting a new coil or transformer.

For this reason many TV service organisations supply their service engineers with a selection of replacement line output transformers and yokes.

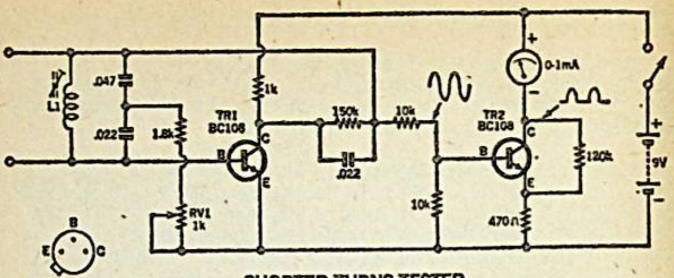
Unfortunately the time needed to try a substitute line output transformer – assuming one has one – can be quite considerable. What is really needed is a small instrument capable of indicating quickly whether or not a shorted turn is present – and that is exactly the function of the unit to be described. It is a low cost battery operated unit which gives reliable indication as to the presence or absence of a shorted turn, and it is compact enough to fit easily in the service engineer's tool kit.

The operation of the shorted turns tester depends upon the fact that a shorted turn behaves basically like a very low resistance which effectively 'loads down' all of the windings on the coil or transformer concerned. As such it tends to absorb most of any magnetic energy which might otherwise be stored in the windings and fed back into the circuit later.

The circuit of the tester consists basically of a transistor oscillator followed by an indicator stage which monitors the activity of the oscillator. Initially the oscillator is adjusted to be just oscillating reliably, as shown by the meter of the indicator stage. Then the coil or transformer to be tested is connected across the tuned circuit of the oscillator.

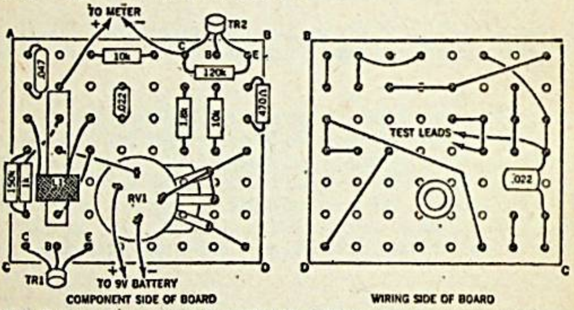
If the unknown coil or transformer is good, the effect of connecting it across the tuned circuit is merely to change the frequency of oscillation. But if it has a shorted turn, the loading reflected is sufficient to cause the oscillator to drop out, or at least fall significantly in activity.

As may be seen from the circuit, the oscillator is a straightforward Colpitts type using a BC108, 2N3565 or similar NPN silicon transistor. The coil is a standard horizontal ringing or 'sine-wave' inductor, and is readily available. A variable resistor between the junction of the two tuned circuit capacitors and the earthed emitter is used to adjust the oscillator activity.



SHORTED TURNS TESTER

The circuit diagram of the shorted-turns tester.



The wiring diagram of the tester. Note that there is one component, a polyester capacitor, underneath the matrix board.

The indicator stage is a simple meter amplifier using a second BC108 or similar. A voltage divider formed by the two 10K resistors and connected between the top of L1 and negative holds TR2 base potential at approximately 0.4 volts. Thus TR2 only conducts while there is oscillation, and then only on the most positive peaks. With a suitable meter in the collector circuit, we therefore have an indication of oscillator activity.

The 470 ohm resistor in the emitter circuit of TR2 was chosen to limit the meter reading to less than 1mA when the oscillator has maximum output (due to minimum RV1). If this emitter resistor is decreased, the maximum meter reading will rise, and vice versa.

The 120k across TR2 allows a small current to flow through the meter (approximately 0.1mA) even when the oscillator is not operating. This is to remind the user to switch the instrument off.

All of the parts should be available from most electronic component dealers, including the line stabilising coil. If you obtain a coil with reasonably high Q and inductance approximately 10mH this would suffice.

The meter we used was a small Japanese type, but any 1mA FSD meter could be used, so long as the dimensions suit your case. We assembled our instrument in a moulded plastic case. The matrix board on which the circuit is wired was secured in the case by passing the switchpot shaft through holes in both the matrix board and the case, and securing the nut on the outside.

One should ensure that the matrix board and the case material together are not too thick, to allow the threaded portion of the switchpot bush to extend through far enough to accommodate the nut. The solder joints under the matrix board should be kept tidy, because large lumps of solder between the board and the case will cause the same problem.

The wiring diagram indicates the layout and interconnection of the components on the matrix board, so the reader should have no difficulty wiring this section. There is only one component under the board, a 0.22uF low voltage polyester capacitor laid flat. Small matrix board studs were used to secure the switchpot tags to the board. The transistors and the sinewave coil were also soldered to studs. With the board assembled, the battery clip, test leads and meter leads can next be connected. Leave the alligator clips off the test leads at this time until the leads are threaded through the case.

The case dimensions and cutouts are shown. Quite often the meter comes seated in a cardboard template, which can be used to trace out the necessary holes. Be careful while drilling the plastic case, as this material is often very brittle and may crack. With the case prepared one can secure the meter and place the grommet in the test lead hold, before securing the matrix board and connecting the meter leads.

Some meters are not marked for polarity, so you must determine the correct polarity before you connect the leads. The easiest way to do this is by connecting a resistance of about 2.2-4.7k in series with the meter, and then connect a 1.5V battery across the combination. The needle will move clockwise when the positive battery lead is connected to the positive meter terminal.

Before the test leads are threaded through the case, tie a knot in them, to eliminate pressure on the joints if the leads are pulled. The 9 volt battery will just fit between the case and the switchpot. We glued a piece of cardboard to the case as shown, to isolate the battery case from TR1. A piece of foam rubber under the battery holds it in place when the bottom is secured. The bottom is secured via one screw into a threaded pillar mounted on the matrix board. The last part to be added is a suitable knob.

When the instrument is complete, it may be tested. Switch the unit on, but do not advance the pot. The meter should read approximately 0.1mA, indicating that the oscillator is not operating. Turn the pot until the meter jumps to approximately 0.4mA. The oscillator is now operating and if you short the test leads, the meter reading should drop, indicating that the oscillator has stopped.

Now unshort the test leads and set the meter reading at 0.6mA. At this level the oscillator will not be affected if a high impedance shunts the coil, but will drop out if the shunt is relatively low. If the meter reading is originally set too high, some high impedance windings may appear to be okay when they are not.

If you have a good line output transformer about, connect the test leads across each winding in turn. The meter reading should remain fairly constant, or increase slightly. Leave the test leads across any of the windings and thread a short piece of wire or solder around the ferrite core. Then short the ends of the wire, whereupon the meter reading should drop to minimum, indicating a shorted turn. All of the windings should give the same reading so it does not matter which you test.

A final word of warning. If a line output transformer is tested in circuit, it would be advisable to disconnect any circuitry which might show up as a shorted turn, such as the EHT rectifier diode filament lead in valve receivers. Removing the rectifier diode from its socket will effectively eliminate this example, but there are other possibilities, so be careful and disconnect all the leads if the tester indicates a shorted turn.

A NEW SOLID-STATE AF MILLIVOLTMETER

A compact and up-to-date design using a new integrated linear amplifier microcircuit. While simply constructed at low cost it offers a high standard of performance.

Measurement of small AF alternating voltages is frequently necessary in developmental and testing situations concerned with audio and super-audio amplifiers and preamplifiers, filters, and transducers such as microphones, gramophone pickups, tape heads and loudspeakers. The compact and up-to-date instrument described in this article may be used to perform such measurements conveniently and reliably, either in laboratory situations or in the field.

The performance of the new instrument is significantly high. It will permit measurements of signals from as low as a few hundred microvolts up to 30V RMS, with a response flat within +/- 0.5dB from below 10Hz to above 100KHz, and improved scale linearity. These improvements, together with increased input impedance and an increase in reliability and stability, are largely due to the use of a new integrated linear amplifier microcircuit as the "heart" of the instrument.

The microcircuit used in the new design is a General Electric type PA 230, which is a low-cost monolithic silicon device specifically designed for use as a low level preamplifier in AF and LF equipment. Operating from a single 12V DC supply, the PA230 will operate from -55 to +110 degrees C. It delivers up to 10V P-P output, is protected against output short-circuits, and features a minimum gain of 4000 (typically 7000) with an open loop bandwidth of 500KHz typical Distortion is typically 1% at full input at 1KHz on open loop, and input noise voltage with 600 ohm generator impedance and 30Hz-20KHz bandwidth is typically 2uV. The device is supplied in a dual-in-line package and has maximum dissipation of 800mW at 25 degrees C ambient.

Cost of the PA230 is quite modest for a linear microcircuit and makes the device quite attractive for many domestic and industrial applications. It may be obtained on order, via the usual parts suppliers.

As the heart of the new millivoltmeter the PA230 is used to perform all of the necessary voltage amplification; as may be seen from the circuit diagram, it is connected in a stabilised feedback amplifier configuration. An emitter-follower stage using a low-noise NPN transistor is employed at the PA230 input to provide current amplification and give the instrument a high input impedance.

The meter movement and associated rectifier bridge are connected in series with the negative feedback applied around the PA230, making the feedback substantially proportional to meter current. As a result the feedback not only provides gain and calibration stability but also corrects for non-linearity in the meter rectifier. Calibration adjustment is performed using a small slider potentiometer which varies the reedback ratio.

The meter movement employed is a 100uA rectangular 4in type, having an internal resistance of approximately 1K. The meter is protected from transient and overload damage by means of an R-C filter/damping circuit together with a low-current silicon diode.

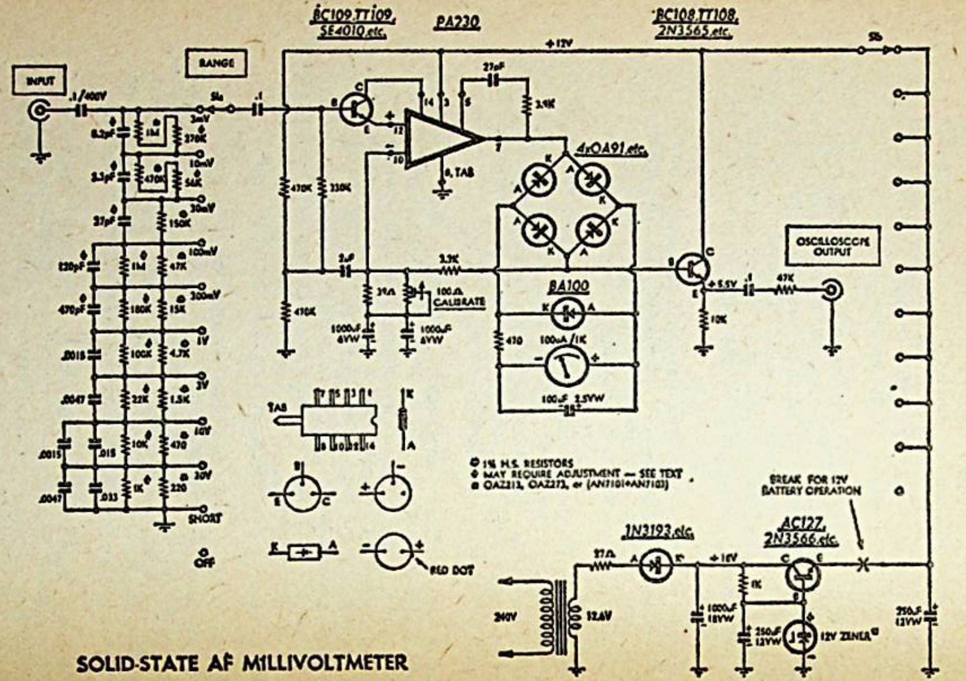
It should be noted that the feedback loop around the PA230 performs DC stabilisation as well as definition of signal gain and linearity. It is for this reason that the lower feedback resistors are taken to earth via blocking capacitors. Two capacitors are used in this position to provide an adequately low impedance to below 10Hz.

The negative feedback signal is applied directly to the "–" differential input of the PA230, while the input signal from the emitter follower is presented to the "+" input. By this means the application of the feedback does not reduce the input impedance, but rather raises it due to the action of the differential input stage in the PA230. To preserve the high input impedance of the emitter follower the bias divider is bootstrapped to the negative feedback input via a 2uF blocking capacitor. As a result of these measures the input impedance of the basic circuit is greater than 2M shunted by approximately 10pF.

A series R-C step circuit connected between the output pin 7 and pin 5 of the PA230 provides phase correction at high frequencies to compensate for stray capacitance in the meter circuit. With the values shown the circuit is quite stable, and has a basic response which is virtually flat between 20Hz and 100KHz, falling smoothly beyond these limits to readh -0.5dB at approximately 8Hz and 400KHz.

The input to the circuit is controlled by a switched voltage divider which provides the instrument with nine measurement ranges. The divider taps are arranged to give inter-range ratios as close as possible to a constant 0.316, corresponding to 10dB.

Ideally the use of multiples of the preferred resistor values 15 and 4.7 would give the closest approximation to exact 10dB ratios; however in practice the divider is complicated due to the loading effect



SOLID-STATE AF MILLIVOLTMETER

of the circuitry following the divider, viz, the meter amplifier with its 2M/10pF input impedance. The operation at high frequencies is also complicated by stray capacitance in the switch, together with a small variation in the input capacitance of the meter amplifier due to gain variations within the feedback loop.

These effects necessitate correction of the divider resistance ratios using series and shunt "paddler resistors", together with high-frequency compensation using small shunt capacitors. Using these techniques the divider is made to produce range ratios which remain within o.5dB of the required 10dB, from zero frequency to better than 100KHz. The overall response of the instrument for all ranges is thus within +/- o.5dB from below 10Hz to above 100KHz, with an overall input impedance of 1M minimum shunted by approximately 25pF.

A tenth position of the input switch connects the metering amplifier input to earth, providing a "shorted" mode in which the meter registers only the residual instrument noise. This permits changes to be made in external circuitry, repositioning of test leads and other operations which might otherwise produce violent meter deflection.

To permit inspection and analysis of signals such as noise, ripple and crosstalk, the instrument is provided with an oscilloscope output. The output circuit is isolated from the metering circuit by means of an emitter follower and series buffer resistor, to ensure that loading effects do not effect instrument accuracy. The signal fed to the output circuit is that available in relatively undistorted form at the top of the resistive portion of the feedback divider.

The signal level available at the oscilloscope output socket is approximately 1V p-p for full scale deflection, at an impedance level of approximately 50K. With the capacitive loading provided by a typical shielded cable and oscilloscope input, the effective output bandwidth will be 10Hz–30KHz +/- 3dB.

Power supply to the instrument is controlled by a second pole of the metering range switch (S1b). The eleventh and final position of the switch thus becomes an "off". Power requirements of the basic instrument are 12V DC at approximately 5mA, which may be provided by either the external mains supply or an external battery.

The internal power supply is a simple regulated circuit designed to deliver modest current at a low effective source impedance and with a very low ripple level. It employs a half-wave rectifier circuit with a 1N3193 or similar silicon diode fed from a miniature 12.6V power transformer.

The remainder of the power supply is a simple series regulator/electronic filter using an NPN transistor and a zener diode. The transistor may be either a germanium type AC127, or similar, or a silicon planar type 2N3566 or similar. The 12V zener diode used as the voltage reference may be either a type OAZ273, an OAZ213 or a series combination of two lower voltage planar types.

Physically the new instrument is housed in the small flanged front case used for most of our current instrument designs, measuring 7½in x 5in x 4in. The meter occupies a major portion of the panel area and is to the left, with the range switch in the upper right-hand corner and the input socket beneath it. The mains cord entry and oscilloscope output socket are at the rear of the case.

The shield is a vertical plate attached perpendicularly to the rear of the front panel via two of the meter mounting studs.

A second, smaller shield plate is used between the range switch and the metering amplifier panel, to prevent spurious coupling at high frequencies due to stray capacitance. In the prototype this sheild is cut from thin sheet brass, being supported by earthed lugs on both the amplifier panel and the range switch.

The attenuator components are supported around the range switch by the appropriate switch lugs, with the section nearest the front panel used for the input switching and the rearmost section for component anchoring. The centre switch section is that used for power switching. The metering amplifier is supported by a small length of miniature resistor panel bolted to the vertical shield plate, while the metering circuitry, oscilloscope output stage and power supply regulator are mounted on a second length of miniature resistor panel attached to the rear of the meter and supported by the meter terminal bolts. The power transformer and rectifier are mounted in the rear of the case.

Most of the wiring should be visible from the photographs, and constructors should find little difficulty in duplicating the original. However the circuit is not unduly critical, and providing the usual precautions are taken concerning signal lead dress, isolation of input and output leads, and avoidance of earth loops, little trouble should occur despite quite considerable differences in layout. The connections to the various semiconductor devices are shown on the circuit; note that the PA230 device is shown viewed from the "top", with the plane of the connection lugs and tab towards the rear.

When the instrument is connected to the power upon completion, it will be found that the meter pointer swings rapidly to full scale; however, this should not be viewed with alarm nor taken as a symptom of a wiring or component fault. The effect is due to the unbalanced situation prevailing before the 1000uF capacitors in the feedback circuit have charged to their operating voltage, and is quite normal upon switch-on. The meter movement is shunted quite heavily during this time by the BA100 diode and 100uF shunt capacitor, and thereby protected from damage.

The current flowing through the meter normally falls below full scale after about 15 seconds. However no attempt to calibrate or to use the instrument for serious measurements should be made for at least 2 minutes, as this time is required before the feedback circuit is fully stabilised. It will be seen that during this period the residual meter reading will fall asymptotically to less than 5 per cent of full scale, corresponding to 200uV equivalent input.

The initial calibration of the instrument should ideally be carried out by comparison with a reference instrument, using an audio generator set to approximately 1KHz at a level of 3.16mV corresponding to FSD using the 3mV nominal sensitivity of the basic instrument. The calibrate potentiometer is then adjusted to produce a reading of "10" on the 3mV range.

With the basic sensitivity set, the attenuator divider may then be checked and if necessary adjusted to ensure that the range ratios are accurately set to 10dB and within +/- 0.5dB from 10Hz–100KHz. Adjustment of the basic (LF) ratios may be performed using the padder resistors identified on the circuit diagram with a diamond symbol; high frequency compensation may be adjusted, if required, using the capacitors similarly identified.

If a reference instrument of known accurate calibration and frequency response is not available, neither of the foregoing adjustments may be carried out. The constructor will have to rely upon the accuracy provided by the attenuator divider component values shown, while "basic" calibration will have to be performed using any available signal source having an amplitude know as accurately as possible.

It should be noted that, in common with the majority of electronic meters, the instrument is fundamentally an average-reading one which is calibrated in RMS values assuming a sinoidal waveform. This fact should be taken into account particularly when performing initial calibration, and also borne in mind when subsequently using the instrument for measurement upon non-sinoidal signals.

SPECIFICATION

A fully solid-state AC millivoltmeter for the AF spectrum, using three transistors and an integrated linear amplifier microcircuit. A mains power supply is included, but the instrument may be operated from a 12V battery if required.

Nine input measuring ranges cover the FSD range 3mV–30V RMS, with 10dB range ratios. The indicating meter employs three scales: a full length 0-10 scale, a 0-3 scale with appropriately shortened length, and a dB scale.

Accuracy of the instrument will be determined largely by the meter movement employed (typically +/- 2 per cent of FSD) and the range attenuator resistors. Linearity is similarly dependent upon the meter, that of circuit being better than +/- 2 per cent from 30-100 per cent of FSD.

Frequency response of basic (3mV FSD) instrument better than +/- 0.5dB from 10Hz–400KHz. Minimum bandwidth on all ranges 10Hz –100KHz, again +/- 0.5dB.

Input impedance 1M minimum shunted by approx. 25pF. Residual hum and noise equivalent to less than 200uV RMS input, with input open circuited but shielded.

An oscilloscope output is provided, giving approx. 1V P-P at FSD. Output impedance 50K, bandwidth approx. 10Hz–30KHz +/- 3dB.

Power requirements 12V at approx. 5mA, supplied either from internal mains supply or external battery.

MILLIVOLTMETER PARTS LIST

1 Small instrument case, 7½in x 5in x 4in, with flanged front panel. Handle, rubber feet, scrap aluminium for vertical shield bracket, scrap brass for switch shield.
1 100uA 1K meter movement, 4in rectangular type
1 3-section 1-pole 11-position rotary switch
1 Miniature power transformer, 12.6V secondary
1 Medium bar-type control-knob
2 Co-axial sockets

SEMICONDUCTORS
1 PA230 linear amplifier microcircuit
1 BC109 or similar low-noise NPN silicon transistor
1 BC108 or similar NPN silicon transistor
1 AC127 or similar NPN transistor
4 OA91 or similar germanium diodes
1 BA100 or similar silicon diode
1 1N3193 or similar power diode
1 12V zener diode (see text)

RESISTORS
1% High stability types: 220ohm, 470ohm, 1.5K, 15K, 47K, 150K, 470K, 1M
5% half-watt types: 27ohm, 39ohm, 470ohm, 2 x 1K, 3.3K, 3.9K, 2 x 10K, 22K, 47K, 56K, 100K, 180K, 270K, 330K, 2 x 470K, 1M
1 100-ohm slider pot

CAPACITORS
LV Ceramic or plastic: 3.3pF, 8.2pF, 2 x 27pF, 120pF, 470pF, 2 x .0015uF, 2 x .0047uF, .015uF, .033uF, 2 x 0.1uF, 2uF
1 0.1uF 400V plastic or paper
1 100uF 2.5VW electrolytic
2 250uF 12VW electrolytic
2 1000uF 6VW electrolytic
1 1000uF 18VW electrolytic

MISCELLANEOUS
1 x 6-lug section and 1 x 10-lug section of miniature resistor panel: 1 x 8-lug tagstrip; grommet for mains cord entry; mains cord and plug; connecting wire, solder, bolts, nuts, washers, etc.

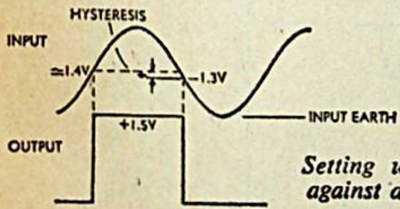
BUILD YOUR OWN FREQUENCY METER, TRIGGER DIVIDER, TACHOMETER WITH 3 IC's

DESCRIPTION: The circuit uses three inexpensive integrated circuits, two type FuL914 and one FuL923. The first 914 acts as a Schmitt trigger. The input is held just above the triggering voltage by the 47K tab pot and is coupled to the input by a 5uF capacitor and a 100K linear potentiometer, which acts as a sensitivity control. Using this method, less than 150mV (typically 110mV) is required to trigger the Schmitt trigger.

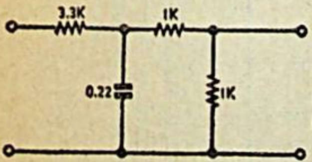
Although not necessary, the second stage, an FuL923, provides frequency division and extends the range of the instrument from 1 to 2MHz. It also makes reading more convenient and accurate in some parts of the scale, e.g. 10-30 ranges. This facility can be switched in or out as shown. If desired, this section may be left out by breaking the circuit at the dashed lines drawn and connecting the output of the Schmitt trigger into the input of the second FuL 914.

This is connected as a monostable. It has five "on" times selected by selecting various capacitor sizes as shown. The 5K tab pot allows fine control of the "on" time.

Pulses produced by the monostable are fed via a 1K resistor to a meter, bridged by a 10uF capacitor and a silicon diode. These pulses charge the capacitor which discharges through the meter. Since the RC time constant of the meter-capacitor is fixed, the voltage will depend on the pulse rate (pulse height and width are constant) and thus on the frequency. Should an excessive pulse rate (and hence voltage) appear at the meter, the BA100 will conduct and shunt the meter protecting it from damage. Meter zeroing is accomplished by the 5K tab pot and the 180-ohm resistor.



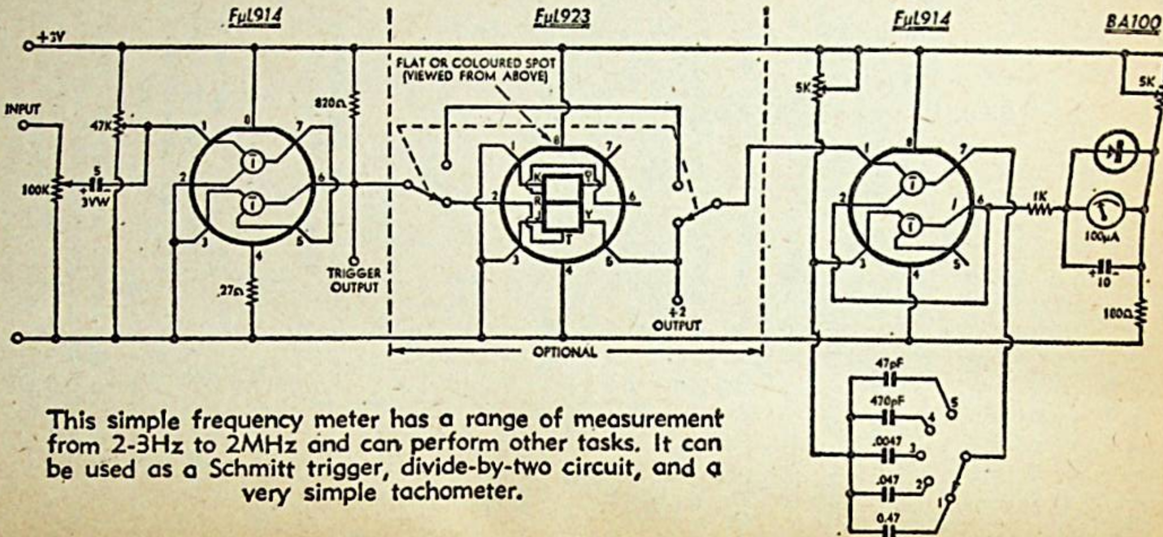
Setting up the Schmitt trigger against an AC input waveform.



Use a circuit similar to this to minimise voltage spikes, when using the instrument as a tachometer.

CONSTRUCTION: The unit was constructed on 4in x 3in printed circuit board. The components were assembled, the board was drilled and the copper laminate cut to provide insulation where required, and the components were soldered in. A 7in x 4in x 3in box was used for the case and a pair of batteries were used for the power supply. These should be the large torch type as "penlight" cells do not give sufficient life with the current drawn, about 20 to 35mA.

FREQUENCY METER, TRIGGER, DIVIDER, TACHOMETER



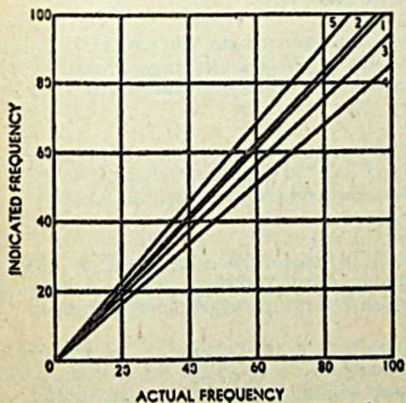
This simple frequency meter has a range of measurement from 2-3Hz to 2MHz and can perform other tasks. It can be used as a Schmitt trigger, divide-by-two circuit, and a very simple tachometer.

CALIBRATION: The instrument requires careful setting up. First, using an AC input and monitoring the Schmitt trigger at the output (end of the 820-ohm resistor) with a ORO, set tieh 47K resistor to give equal mark/space ratio in the square wave resulting from the sine wave input.

An alternative method is to monitor the DC voltage from negative to the 820-ohm resistor with a multimeter and observe the setting when the meter indication changes (from about 0.2 volts to about 1.5 volts). Move the tab pot back very slightly. Moving back too far will result in the trigger reverting to its original state. Careful setting can get the trigger to sit exactly in the centre of its hysteresis point. (See diagram.) This process ensures maximum sensitivity.

Having set up the Schmitt trigger, it is necessary to calibrate the monostable section. Firstly, set the meter zero using the 5K tab pot in the voltage divider. This backs off the voltage due to the saturated transistor in the output of the monostable. Now, using a 50-ohm mains source, preferably a 6.3V filament transformer, and with the 100K set on minimum and with the unit on the 100Hz range, increase the sensitivity until the meter reads positively.

Adjust the 5K tab pot fine timing control until the meter indicates 50. Note that the meter zero moves with this adjustment. Re-set the meter to zero, and recalibrate to 50. Repeat this successive approximation until the meter indicates 50 with signal and zero with no signal.



Range-to-range accuracy depends primarily on the tolerance of the timing capacitors.

If an accurate signal generator is available, it will be possible to check the other ranges. The accuracy of the instrument depends on the linearity of the meter used and on the accuracy of the timing capacitors. Using ordinary capacitors, the accuracy will be plus or minus 10 per cent or so. Any range can be calibrated using the above procedure and a signal generator. However, the other ranges may then be wrong.

This problem can be overcome by using expensive precision capacitors, by the use of a calibration table (the method which I chose), or by having a second bank on the switch to select from five 2K tab pots in place of the 1K fixed resistor (each being calibrated individually). As can be seen from the graph, the linearity within a particular range is controlled by the linearity of the meter, and is typically between plus or minus 1 and 2 per cent. However, between ranges it is only plus or minus 10 to 15 per cent. This large error can be eliminated by any of the methods mentioned above.

USE. It is advisable to leave the sensitivity on minimum to reduce the possibility of damage to the input of the trigger by excessive voltage. The meter is protected so overloading will not damage it, but this should be avoided where possible.

Switch the instrument on and set the range switch to the range to be measured. These are:

1	100Hz FSD
2	1KHz FSD
3	10KHz FSD
4	100KHz FSD
5	1MHz FSD

Now increase the sensitivity until the meter reads and then advance it just sufficiently to get a reliable reading. Using minimum sensitivity ensures loading the source by a minimum amount—this can be most important in tuned circuits, etc.

Using the Schmitt trigger. Only AC facilities are included. However, having an input on the 47K pot or on the trigger input will allow DC use. The output can be taken off either before or after the divide-by-two FuL923.

Using the divide-by-two range. Place the input into the Schmitt trigger in the normal way and take the output out after the divide-by-two range. Switching the input and output of the FuL923 gives a 1:1 or 1:2 ratio.

Use as a tachometer. The first consideration is voltage—200 to 400 volt spikes appear on the primary of the ignition coil and an integrating /dividing network must be used. A suggested one is shown:

Consideration must be given to the polarity of the earth (if the car source is to be used as a power supply) and the number of cylinders and 2 or 4 stroke must be used to give the necessary pulses/sec to rpm conversion. The triggering spikes can be used to advantage as it should be possible to use only the 914 monostable and one timing capacitor. This would give a very inexpensive tachometer, expecially if a zener-controlled power supply were used.

From the above description, it can be seen that the instrument is a useful and flexible piece of equipment to have. It can be made up in a variety of forms depending on the desired use(s).

TESTING TRANSISTORS USING A MULTIMETER

This article shows how transistors may be tested for failure, using only a standard multimeter and some knowledge of transistor construction. In addition to finding out whether a transistor is "good" or not, the same tests can be used to determine its polarity, NPN or PNP.

So far as is known, ordinary bipolar transistors are not subject to aging and gradual failure as is normal with valves and other thermionic devices. However, it is thought that transistors are adversely affected by various forms of radiation and that the effect is additive over a period of time.

On the surface of the Earth, where the level of cosmic radiation is normally quite low, the effects of radiation in terms of gradual degeneration of transistor action and parameter changes are not significant. Up to the present time particular devices have not been in service for a sufficiently long period for any deterioration to be observed.

For many practical purposes then, a transistor may usually be considered as a device which either functions correctly or fails more or less completely with no intermediate modes of operation. In addition, under normal circumstances the transistor may be considered as having an indefinite life expectancy with failure only due to faulty manufacture or to an adverse change in circuit conditions.

To test a transistor to see if it is simply "good" or "bad", we may consider it as consisting essentially of two semiconductor diodes connected back-to-back, as shown in figure 1. The condition of the transistor can then be determined with reasonable confidence by individuality testing the two diode junctions using an ohm meter. In some cases it will also be necessary to perform a further test of the two junctions as a whole, again using an ohm meter.

A primary characteristic of a semi-conductor diode junction is that it will pass current relatively easily in one direction, but not in the other. If an ohm meter is placed across a diode and the resistance noted, then the meter leads are reversed and the reading again noted, there will thus tend to be a low apparent resistance one way and a very high apparent resistance in the other direction. This is shown pictorially in figure 2.

Although the resistance of a junction in the reverse direction is very high it is not infinite. This is because there is a certain amount of current leakage in the reverse direction. With transistors, the junction leakage currents can vary widely, depending upon the physical construction and type of semiconductor used, silicon or germanium. In general, there is considerably less leakage in silicon device junctions than there is in germanium devices.

Before showing how to test a transistor, there are a few points which we should mention about the use of multimeter ohms ranges. In particular, it should be noted that there are precautions which should be taken both to ensure the validity of the testing and—equally important—

to avoid damaging a good transistor.

Basically there are two things about the multimeter ohms ranges which should be known: the voltage which appears at the meter test terminals, and the maximum current which can flow through the circuit under test. It is necessary to know the voltage used, because there may be instances in which the base-emitter junction of the transistor under test will have a reverse breakdown voltage sufficiently low to cause it to enter breakdown during testing. Unless this is borne in mind an otherwise "good" transistor can be unjustly discarded as "shorted"

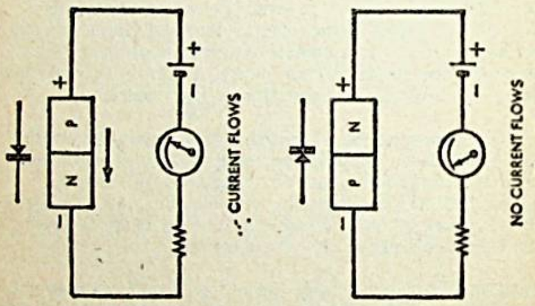


Figure 2

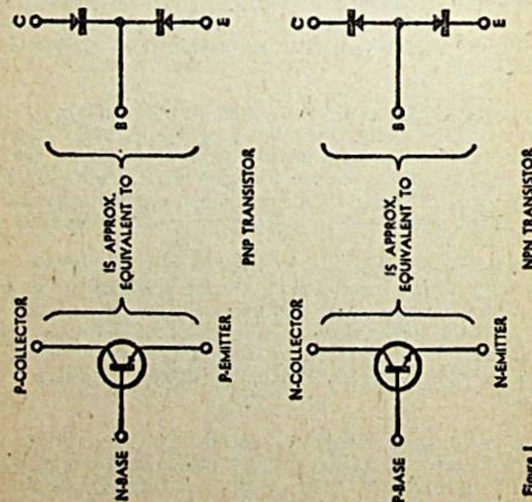


Figure 1

The base-emitter junction of a transistor is similar to a zener diode in that it will not pass significant current in the reverse direction until a sufficiently large voltage is applied. However if the ohm meter test voltage is sufficiently large then the base emitter junction will show a low resistance in both directions, because in the reverse direction it will pass appreciable "breakdown" current. This situation would normally indicate a faulty transistor.

Usually multimeters use a 1.5 volt cell on the lower ohms ranges and fortunately the base-emitter breakdown voltage of most transistors is in excess of this. However, there are a few exceptions. There are usually germanium drift field types, together with a relatively few high frequency junction types. Some can have breakdown voltages as low as 0.5V, but this is uncommon.

In practice, these transistors will not show a complete short in the breakdown direction. But, rather, they will give a reading less than maximum resistance, but noticeably higher than the reading obtained in the normally low resistance direction. So long as there is a reasonable ratio between the two the junction can therefore be assumed to be functional. Nevertheless, under no circumstances should you use a test voltage of more than 1.5V.

The actual voltage at the test leads of your multimeter can be determined in several ways. Possibly the easiest way of all is to consult a circuit diagram, but this is not always available.

If another meter, preferably a high resistance unit, is available, the voltage at the test terminals may be measured with the range switch in the ohms times one position. The voltage will probably measure a little less than 1.5V because of the loading effect of the second meter. Failing this, you could open up the meter and trace out the circuit and note the battery voltage used, or otherwise contact the manufacturer.

The next thing to find out about the multimeter is the maximum test current that can flow when a transistor or diode is connected in the low resistance direction. If the test current is too large, one or both of the transistor junctions could be damaged.

There are two ways in which a semi-conductor junction can be destroyed by excessive current. The actual semi-conductor can be made to dissipate too much power, resulting usually in a short circuit, or the fine wires connecting the external leads and the semi-conductor can be caused to fuse, producing an open circuit and again, in effect destroying the junction.

Most multimeters use the ohms measuring circuit shown in elementary form in figure 3. In essence the circuit consists of a cell or battery in series with a meter and a resistor Rs which may be adjusted for zero ohms reading. Another method of adjusting for zero ohms is to have a variable shunt across the meter as shown in dotted form. The unknown resistance Rx completes the circuit.

Figure 3 shows the basic ohms measuring circuit often used in multimeters. The diode test procedure is shown in figure 4, while figure 5 shows typical base connections.

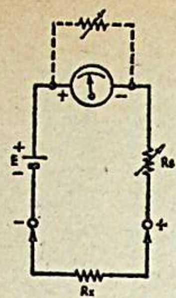


Figure 3

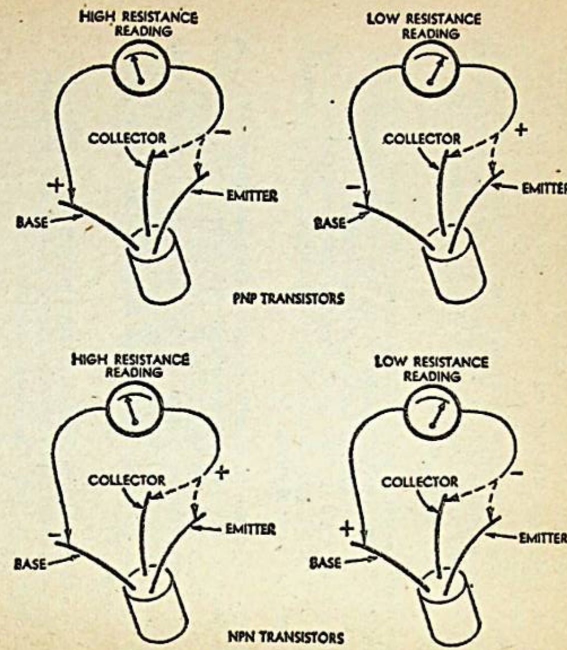


Figure 4

WARNING:- TERMINAL VOLTAGE OF OHM-METER SHOULD NOT EXCEED 1.5V WITH MAXIMUM TEST CURRENT OF 1mA

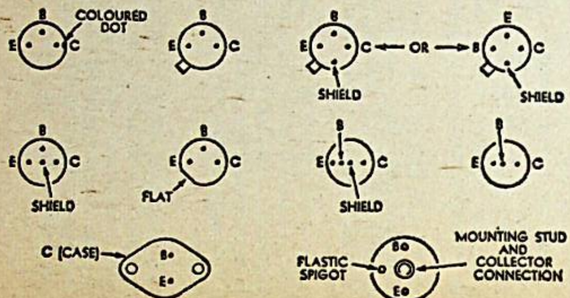


Figure 5 COMMON TRANSISTOR BASE CONFIGURATIONS

When Rx is zero, as is the case if the test leads are shorted, the series resistor Rs may be adjusted to give a full scale deflection on the meter. Now, if the unknown resistance Rx happens to be of the same value as Rs then half the short circuit current will flow in the circuit, and the cell voltage will be dropped equally across Rs and Rx. But, more importantly, the meter will now read half scale.

The half scale reading is usually used as a reference, because the ohms scale on a multimeter is not linear. Naturally zero deflection indicates an infinite resistance at the test terminals, while full scale deflection corresponds to zero ohms or, in other words, a short circuit. Hence the upper half of the scale corresponds to a finite resistance range while the lower half corresponds to an infinite range.

Various ohms ranges may be provided, usually by increasing the value of the series resistor Rs by the required multiplication factor. Hence, for a fixed battery voltage, minimum test current will occur on the highest ohms range. However, most multimeters employ more than 1.5 volts for the higher ohms ranges, depending upon the sensitivity of the actual meter movement.

Consequently for transistor testing you should use the highest ohms range which still provides a voltage of no more than 1.5V, and the short-circuit current for that range should not exceed 1mA. Again, this may be found by either consulting the circuit diagram, or measurement with a milliammeter.

Commonly, the range which satisfies these conditions is the R x 100 ohms. However, this may not always be the case, as the ohms measuring circuit will vary with different instruments. In addition, the scales may not always be designated in multiples; some meters may specify the centre scale reading of each range or, in a few instances, they may have ohms division ranges.

The general methods for testing a transistor for both polarity and condition will be given in the following paragraphs, and in general apply equally to low, medium and high power transistors.

The "initial" method for checking PNP and NPN transistors is shown pictorially in figure 4, together with instructions. The upper two diagrams show the procedure and results with a PNP transistor while the lower two show the same procedure and the results obtained with an NPN device.

In the first instance it is not necessary to know the polarity of the transistor under test. From the results obtained by the procedures shown you will be able to tell not only that the device is a good one but also determine if it is PNP or NPN.

Firstly, connect either test lead to the base of the unknown transistor and use the other as a "wander" lead connecting it to the emitter and collector, one after the other. It is not necessary to test the emitter and collector in that order. You may, if desired, check between base and collector and then between base and emitter.

With both measurements the meter should read either high or low resistance. If one junction reads high and the other low, then the transistor is faulty. However, similar readings for both junctions (i.e., both high or both low resistance) does not necessarily mean that the transistor is good.

It could be that both junctions are short circuited, or that both are open circuited. To examine this possibility a further test is required. This involves nothing more than repeating the test with the "base" and "wander" leads transposed, to reverse the battery polarities applied to the transistor junctions.

If the transistor is in fact a good one, then both junctions will give a resistance reading opposite to that obtained before you changed the lead connected to the base. For example, say you measured a high resistance for both junctions in the first case then after reversing the base and "wander" leads, you measured a low junction resistance. This would indicate a good transistor.

But if the junctions do not measure "both high" then "both low" after reversing the leads, or vice versa, then the transistor is faulty.

From the tests just described, you can also determine the transistor type. For this, however, the polarity of the voltage appearing at the multimeter test leads must be known.

Referring to figure 3, again, it can be seen that the voltage across Rx, due to the cell E, will have the polarity indicated, negative on the left terminal and positive on the right. But, if we wish to use the meter to measure a voltage instead of a resistance, then the polarity of the applied voltage will have to agree with the polarity shown across the meter.

In other words, the lead polarity for an input voltage is opposite to that of an output voltage. And the input terminals of a multimeter are marked for input voltage polarity. Hence, the meter output voltage on the ohms ranges will generally be positive at the lead marked negative and coloured black, and negative at the lead marked positive and red.

So, we suggest that you put the red test lead in the socket marked negative and the black lead in the socket marked positive, so as to minimise confusion when you test a transistor. Then the red lead will actually have a positive output voltage and the black lead will be negative.

Now, assuming that the transistor is a good one, a PNP type will give a low resistance reading for both base-emitter and base-collector junctions if the negative test lead is applied to the base. While an NPN device will give low resistance readings when the positive lead is connected to the base.

An easy way to remember the polarities is to associate the polarity of the test lead connected to the base with the N in PNP and the P in NPN. However, be careful not to be confused.

Naturally enough the tests thus far described depend upon one having knowledge of the lead connections of the transistor to be tested. At the very least, it is necessary to know which is the base lead! To assist in determining the connections for unmarked and unknown transistors, we have shown the more popular connection conventions in figure 5.

It should be noted, however, that even more so than with other conventions, transistor basing tends to be noteworthy for its exceptions. Thus the conventions shown in figure 5 should be regarded as "likely" connections, yet gives sensible results using another configuration, then the chances are that it is a good unit whose only fault is "unconventional" lead connections.

As noted earlier there is a "second order" test which may be necessary in some cases, more frequently with power and high voltage transistors than with the more common types. The test is for voltage punch-through between emitter and collector.

If a sufficiently large voltage is applied between the emitter and collector of a transistor, then voltage "punch-through" can occur. The punch-through phenomenon can result in the transistor passing the "diode" tests just described, but, at the same time being effectively shorted between collector and emitter.

Figure 6, at right, shows the test procedure for collector to emitter voltage punch through. It is simply a measurement of the resistance, which should be relatively high for a good transistor, with both meter polarity connections.

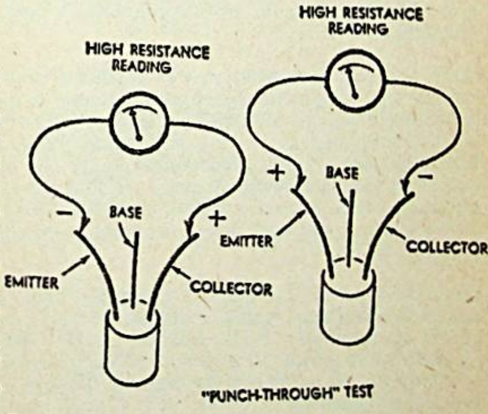


Figure 6

What actually happens when a transistor is punched through is that the depletion region at the collector-base junction increases, as the collector-emitter supply voltage is increased, until the region extends into the base-emitter depletion region. At this point, if excessive collector current flows, a narrow channel is fused between the collector and emitter. The channel may typically be converted to the same semiconductor "type" as the emitter and collector regions (i.e., N-type for an NPN device), without effectively disturbing the performance of the original junctions as shown by the simple diode tests.

The test for the punch-through condition is simply an extension of the diode test, using the ohm meter to measure the resistance between the emitter and collector. A transistor which has suffered a voltage punch-through will give a very low resistance reading between collector and emitter, both with one meter polarity and with the other.

Conversely, a good transistor will give high resistance readings between collector and emitter, again with both meter polarity connections. However, the resistance reading will generally be slightly less in one direction than in the other. This effect will be more noticeable with power transistors than with others.

Figure 6 illustrates, in a similar manner as that for the diode test, the procedure for punch-through testing.

TRANSISTOR AND DIODE TESTING: BASIC PRINCIPLES

In many modern applications the performance of common semiconductor devices is sufficiently defined by a small number of basic tests. Explained in this article are the basic tests appropriate for diodes, bipolar transistors, field-effect transistors and unijunctions.

Modern semiconductor diodes and transistors are intrinsically very reliable devices, but they are subject to rather wide production variations in performance parameters and may also be speedily damaged by abnormal or inappropriate circuit conditions. Device testing is therefore a frequent necessity.

There are many tests which may be applied to a semiconductor device, ranging from simple inter-electrode continuity or differential resistance tests using a bench multimeter to complex and refined measurements of parameters such as gain, rise-time, admittances and loss factors using high-frequency or pulse measurement equipment. Specialised transistor and diode testing instruments cover a wide range, from simple and inexpensive bench or pocket testers, measuring leakage current and DC current gain, to elaborate and costly computer-controlled parameter plotters.

While there exist a large number of testable characteristics and parameters for each semiconductor device, it is by no means necessary that all or even a majority of device parameters be tested to determine the suitability of a device for a given circuit application. Rather, it is probably true to say that in the majority of applications—particularly those in the servicing or development of domestic equipment—the performance of common semiconductor devices may be sufficiently defined or predicted by a relatively small number of basic tests.

This means that a large majority of semiconductor device tests can be performed quite satisfactorily using a relatively simple and unpretentious testing instrument. In fact such an instrument often has a distinct advantage over more complex and refined testing equipment: it is usually simpler in operation, and can generally be used to perform tests at a significantly higher rate, even when operated by unskilled personnel. In addition it will generally involve a far lower initial outlay.

It is true that, even in the most routine applications, there will be occasions when measurements will be required of parameters beyond the range of a simple test instrument. However such occasions will generally be either sufficiently infrequent to justify individual test set-ups on each occasion, or otherwise concerned with sufficiently few or specialised parameters to justify construction of a custom-designed test instrument.

Knowledge of the basic tests appropriate to the various commonly encountered semiconductor devices is essential not only for the design and construction of suitable test instruments of the former type, but also for their efficient and effective use. The following discussion of the basic tests used for evaluating diodes, bipolar transistors and unijunctions may help the reader to acquire this useful knowledge.

DIODES: Three of the most basic yet useful tests employed in evaluating semiconductor diodes are illustrated in the diagrams of figure 1. These consist of tests measuring (a) the voltage drop Vf of the diode during forward-bias operation, (b) the voltage drop BV of the diode during reverse-bias operation in the "breakdown" or avalanche-mode condition, and (c) the leakage current Ir during reverse-bias operation at voltages substantially below BV.

The forward voltage drop Vf is a characteristic which can be of considerable importance when a diode is to be used as a rectifier, non-linear element or level detector at low signal levels. It gives an indication of the order of forward bias voltage which must be applied to the device before appreciable current flows (for an "ideal" diode, Vf would presumably be infinitely small).

The diagram of figure 2 shows typical voltage-current characteristic curves for silicon and germanium diodes. As may be seen the two types are basically similar, the main difference being that forward conduction of a silicon diode occurs somewhat more abruptly and at a slightly higher voltage than for a germanium diode.

Fairly obviously the forward voltage Vf varies as a function of the forward current, varying appreciably at low current levels but less at higher current levels. In order to measure the forward voltage on a comparative basis it is therefore necessary to perform the measurement with a specified forward current passing through the diode, as in figure 1 (a). (The symbol consisting of a circle enclosing an inverted "v" represents an ideal "constant current generator", which may in practice be approximated by a high voltage in series with an appropriately high resistance.)

The choice of the test current level used for this measurement is influenced by two rather opposing considerations. One is that higher current levels tend to give a truer picture of the "conducting" state of all devices, by ensuring that each device is operated on the high-slope (low resistance) portion of its forward characteristic. The other consideration is that device power dissipation during the test will rise fairly rapidly with test current, so that a high test current level risks damage of low-power devices.

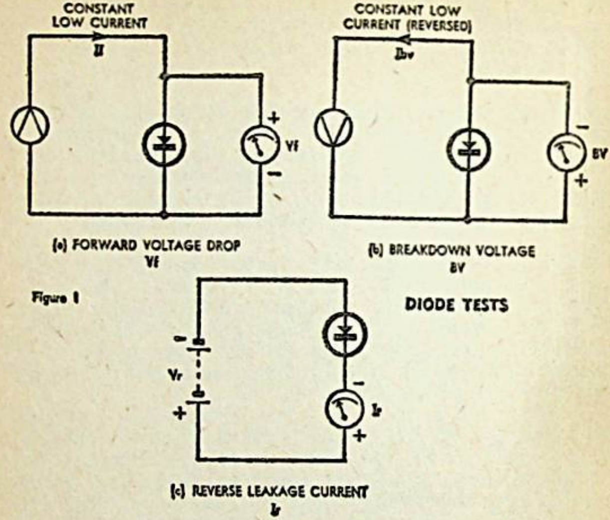


Figure 1

DIODE TESTS

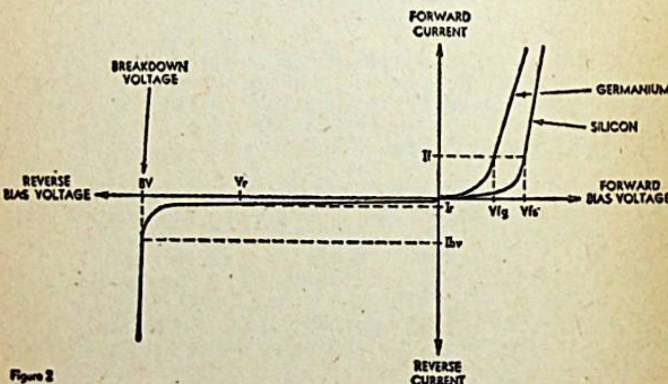


Figure 2

In practice a test current level If of the order of 2 milliamps strikes a suitable compromise between these factors, permitting useful and meaningful comparison of devices while ensuring that device power dissipation during the test is very low. At this current the forward voltage drop Vfg for typical germanium diodes lies between approximately 0.35V and 0.55V, while for typical silicon diodes Vfs lies between approximately 0.5V and 0.9V.

The significance of readings obtained from a pracical test circuit similar to that shown in figure 1(a) will depend mainly upon the approximation used for the constant current generator, and the current taken by the voltmeter relative to the diode current. For most practical purposes a 10V supply in series with a 4.7K resistor gives an adequate approximation for the current source, while a 0-2V voltmeter based upon a 50uA or 100uA meter movement will give neglibible loading.

In passing, it may be noted that a voltmeter reading of zero during the Vf test does not indicate a "perfect" diode, but rather that the device is short-circuited. Conversely a full-scale reading would indicate an open-circuit. (To allow for the latter possibility it will normally be necessary to protect the voltmeter from overload.) Hence the Vf test is well suited to be that first applied to an unknown device.

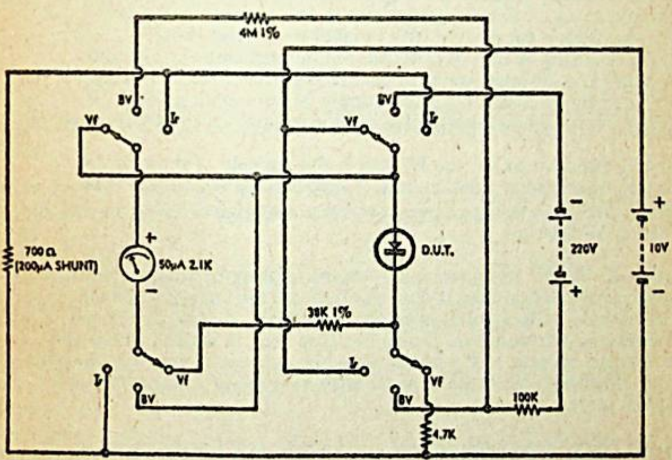


Figure 3

DIODE TEST CIRCUIT

Another important and useful measure of diode performance is the voltage drop BV of a diode when operated in the reverse-bias "breakdown" or avalanche region. This is often called the "breakdown voltage" Diode dissipation in the breakdown region tends to be relatively high, as appreciable current may be drawn by the device while it is sustaining a high-voltage drop.

For diodes to be operated in the breakdown region as "zener" or voltage reference diodes, BV is important because it represents the working voltage of the device. Conversely for diodes which are not to be permitted to enter the breakdown region, BV is again important because it indicates the limit below which must be kept the peak inverse voltage (P.I.V.) applied to the device in operation.

As shown by the diagram of figure 1(b), the measurement of BV is performed in a fashion very similar to that used for Vf. A constant current—in this case in the reverse direction—is forced through the device, and a voltmeter used to measure the voltage drop.

As before, the current level Ibv used for the test is dictated by the opposing considerations of reading validity and device protection. Higher current levels ensure that all devices are tested in the breakdown mode yet, even more so than before, also cause increased power dissipation and risk damage of small devices due to overheating.

With currently available low-voltage devices, a test current of the order of 2 milliamps again strikes a suitable compromise. At this current most devices are operated well into the breakdown region, yet device dissipation during the test is limited to approximately 200 milliwatts or less.

By intentionally degrading the approximation used for the constant-current supply device dissipation can in fact be kept below about 120 milli-watts even for devices with breakdown voltage drops of 100V or higher. Reading validity is reduced, but in most cases this is quite acceptable. A suitable practical circuit to perform the test of figure 1(b) may thus consist of a 220V supply in series with a 100K 1W resistor, and a 0-200V voltmeter again using a 50uA or 100uA movement.

As with the test for Vf, the BV test is also capable of showing up shorted or open-circuited devices. A zero reading indicates a short, while a full-scale reading suggests either a very high breakdown voltage or an open circuit.

Although the BV test gives an important measure of diode performance in the reverse-bias mode, it does not indicate the behaviour of the diode at reverse bias voltages below breakdown. A third useful test of diodes is therefore that shown in figure 1(c), in which is measured the leakage current Ir at a low reverse bias voltage. As may be seen the diode is simply connected in series with a source of voltage Vr and a current meter.

As one might expect the reverse bias voltage used for the Ir test is again something of a compromise. The lower the voltage used, the lower the leakage current levels involved; hence in order to be able to use simple metering circuitry it is desirable to use a fairly high bias voltage. On the other hand increasing the bias voltage generally means that devices with low breakdown voltage will be taken into the breakdown region and therefore invalidate the measurement.

It is true that diodes with very low breakdown voltage will generally be those intended for use in the breakdown mode, as "zeners" or voltage reference diodes; hence for these devices the reverse leakage current before breakdown will in most cases be of little or no concern. This allows us to set Vr at a convenient level for the purposes of measurement —say 10V.

At this reverse bias typical modern diodes exhibit quite a low order of leakage current. Silicon diodes will generally draw less than 10-15uA, while germanium diodes may range somewhat higher.

Figure 3 shows the circuit of a practical diode tester which would perform the three tests just described. As may be seen a 4-pole 3-position switch is used to determine the test performed; if the reader cares to trace through the circuit in each case he will find that the test circuits correspond closely to those shown in figure 1.

Note that the meter movement is marked as having an internal resistance of 2.1K. This does not signify the use of a non-standard meter, but merely the use of a standard 50uA/2K meter together with an overload protection circuit consisting of a shunt silicon diode and a series 100 ohm resistor. For details of this method of meter protection readers are referred to the article by Philip Watson in the December 1959 issue of "Radio Television and Hobbies".

The meter ranges provided for the three tests performed by this circuit are as follows:

Vf: 0-2V
BV: 0-200V
Ir: 0-200uA

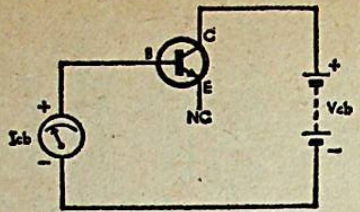
Incidentally the letters "D.U.T." adjacent to the diode symbol in figure 3 are commonly used to indicate that the symbol represents the "device under test".

BIPOLAR TRANSISTORS: Five of the most basic and useful tests available for evaluating bipolar transistors are illustrated in figure 4. (The circuit symbols and polarities shown are those for NPN devices; for PNP devices the circuits would be unchanged but the transistor symbols and polarities changed appropriately.) There are two tests concerned with leakage currents, and two concerned with breakdown voltage drops; the final test deals with DC current gain.

Although a bipolar transistor has two P-N junctions, in normal operation only one is operated under reverse bias conditions: the collector-base junction. The leakage current of this junction is therefore of considerably greater interest than that of the base-emitter junction. This is particularly so, as the collector-base junction leakage can significantly influence the transistor operating point and its thermal behaviour in circuits involving high resistance base bias.

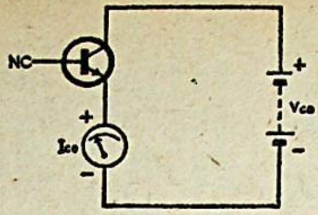
The usual method of measuring the reverse-bias leakage current of the collector-base junction is shown in figure 4(a); as may be seen the circuit involved is similar to that used in figure 1(c) for a diode. As the test is made with the emitter disconnected the leakage current measured is designated Icho, where the "O" signifies that the third electrode of the device is left open circuited.

In this case the choice of reverse-bias voltage Vcb is simpler than in the diode test, as few transistors have collector-base avalanche voltages below about 25-30V. A test voltage of 10V is therefore in order, and this figure usually permits useful Icbo readings to be made using a 50uA movement.



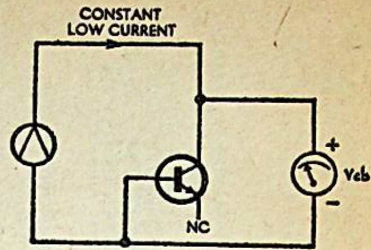
(a) COLLECTOR-BASE LEAKAGE WITH EMITTER OPEN-CIRCUIT

I_{cbo}



(b) COLLECTOR-EMITTER LEAKAGE WITH BASE OPEN-CIRCUIT

I_{ceo}



(c) COLLECTOR-BASE BREAKDOWN VOLTAGE WITH EMITTER OPEN-CIRCUIT

BV_{cbo}

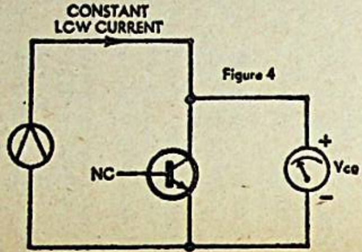
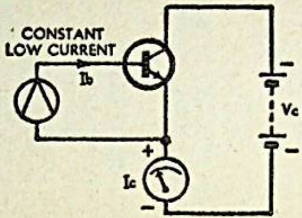


Figure 4

(d) COLLECTOR-EMITTER BREAKDOWN VOLTAGE WITH BASE OPEN-CIRCUIT

BV_{ceo}



(e) DC CURRENT GAIN β OR h_{FE}

$I_c = \beta.(I_{cbo} + I_b)$

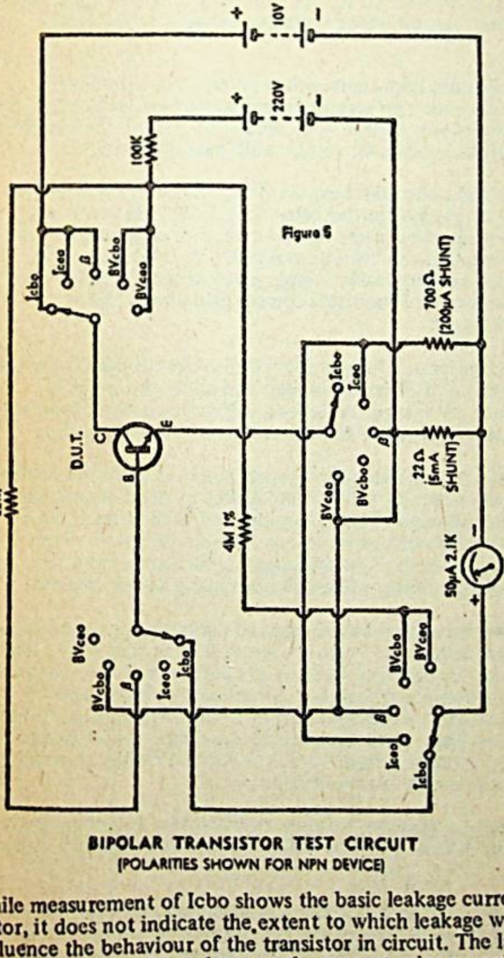
i.e. if $I_b \gg I_{cbo}$

$\beta = \frac{I_c}{I_b} = K.I_c$

where $K = \frac{1}{I_b}$ = constant

BIPOLAR TRANSISTOR TESTS
(POLARITIES SHOWN FOR NPN DEVICES)

Low and medium power silicon transistors typically have Icbo leakage of less than 5uA; germanium transistors of the same type may range to seven or eight times this figure. As with diodes, a full-scale reading on this test suggests either excessively high leakage or an internal collector-base short. With silicon transistors an effective reading of zero is often obtained, many devices having Icbo figures of the order of a few tens of nanoamps; however, a germanium transistor giving a zero reading for Icbo will usually prove to be open-circuit.



BIPOLAR TRANSISTOR TEST CIRCUIT
(POLARITIES SHOWN FOR NPN DEVICE)

While measurement of Icbo shows the basic leakage current of a transistor, it does not indicate the extent to which leakage will tend to influence the behaviour of the transistor in circuit. The latter behaviour will tend to depend largely upon the current gain; accordingly it is usual to make a second leakage measurement, using the circuit shown in figure 4(b). The quantity measured in this case is Iceo, the collector-emitter (amplified) leakage current with the base left open circuit.

As with the Icbo test there is little worry about avalanche breakdown and the testing voltage Vce may be a convenient 10V. A current meter range of 200uA is usually appropriate for most modern transistors, as high-gain silicon devices have a typical Iceo of 30uA or less while germanium types may range up to 5 or 6 times this figure. A hard full-scale reading would suggest either excessive leakage or a collector-emitter short.

Breakdown voltages are important with transistors, as with diodes, and in this case there are two measurements usually made: BVcbo, the collector-base breakdown voltage with emitter open circuit; and BVceo, the collector-emitter breakdown voltage with base open circuit.

Although at first sight one might expect BVcbo to be less than BVceo for a particular device—the emitter being further from the collector than the base—the reverse is generally the case. This is because BVcbo involves what one might call "simple avalanche" breakdown, while BVceo describes behaviour under "amplified avalanche" conditions as a result of the influence of transistor current gain upon collector-emitter leakage current.

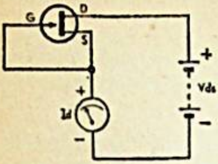
It may be seen from figure 7(c) and (d) that the circuits used for measurement of BVcbo and BVceo are similar, and are both similar to that used for diode BV testing. As before, a constant-current source is used, with a sensitive voltmeter to measure the device voltage drop.

In determining the appropriate test current, there are again the opposing considerations of reading validity and device protection, and the designers of test equipment diverge considerably in the current level chosen. Some advocate currents as low as 100uA; others are of the opinion that currents up to a few milliamps are desirable from the viewpoint of reading validity, while still not risking device damage.

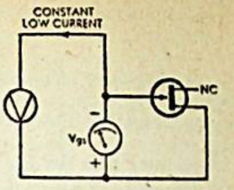
The present writer is inclined to lean toward the latter school of thought, particularly if precautions are taken to ensure that devices with a high breakdown voltage are arranged to draw less current and thus dissipate less heat. Thus a BVcbo or BVceo test circuit employing a (220V + 100K) approximation to a constant current source will give quite useful and meaningful readings, while limiting the maximum device dissipation to approximately 120mW, a figure which should be within the ratings of most devices commonly encountered.

As before the use of a sensitive voltmeter is necessary if loading errors are to be avoided. A basic 50uA or 100uA movement is usually found adequate.

A third breakdown voltage is occasionally measured: BVebo, the reverse-bias breakdown voltage of the base-emitter junction. This is not as important as the other two, but can be important in applications

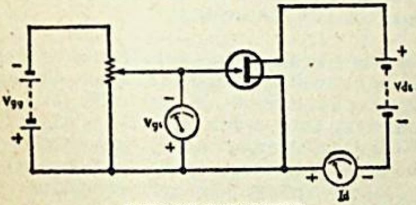


(a) ZERO-BIAS CHANNEL CURRENT
Idss



(b) GATE-SOURCE BREAKDOWN VOLTAGE
WITH DRAIN OPEN-CIRCUIT
BVgso

Figure 6



(c) TRANSFER CHARACTERISTIC

$$gm = \frac{\Delta Id}{\Delta Vgs}$$

(also $Vp = Vgs$ for $Id \to 0$)

JFET TESTS
(POLARITIES SHOWN FOR N-CHANNEL DEVICES)

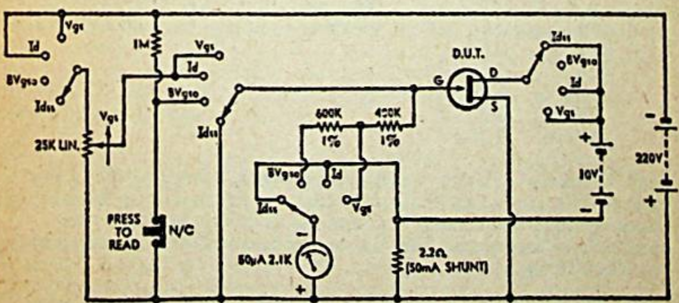


Figure 7

JFET TEST CIRCUIT
(POLARITIES SHOWN FOR N-CHANNEL DEVICES)

where high reverse base-emitter voltages can occur. As BVebo may be measured in a similar way to that used for BVcbo, this test can usually be made quite easily when required simply by temporarily substituting the emitter for the collector lead in a BVcbo test circuit.

The fifth transistor test illustrated in figure 4 is that for DC common-emitter current gain, shown in diagram (e). This transistor characteristic is commonly known as beta or hFE.

It can be seen from the diagram that beta is measured by driving the transistor with a constant low base current, and measuring the resultant collector current with a known collector-emitter voltage Vc applied. The collector current will be equal to beta (Icbo + Ib).

If the external base current Ib is large compared with the internal collector-base leakage current Icbo–and with modern transistors this is fairly easy to arrange–the Ic will be a close approximation to beta-Ib. Thus since Ib is held constant at a known value Ic will be directly proportional to beta, and the collector current meter may be calibrated directly in terms of current gain.

With most modern low- and medium-power transistors Ib may be set at 10uA, delivered by a circuit consisting of a suitably high voltage in series with an appropriate resistor (220V and 22M, for example). Then if Ic is measured during the test with a meter of say 5mA full-scale deflection (F.S.D.), the meter may be calibrated directly with a 0-500 beta scale. Such a scale is appropriate for the majority of devices in current use, although a few very high gain devices have gain ranges extending to 600-700.

The choice of collector-emitter voltage Vc applied during the beta test is not a critical one. The main considerations are that Vc be high enough to ensure that no device can saturate, yet low enough to prevent excessive dissipation. The range between these limits is a wide one, and a convenient figure of 10V is often selected.

Figure 5 shows a practical bipolar transistor test circuit based upon the five tests shown in figure 4. Again a protected 50uA meter movement is used, with shunts and multipliers arranged to give the following ranges:

Icbo:	0-50uA
Iceo:	0-200uA
beta:	0-500
BVcbo:	0-200V
BVceo:	0-200V

If the BVcbo of a device were to be measured, this test could be performed with the circuit of figure 5 using the method noted earlier, which involves temporarily inserting the device with collector and emitter leads transposed. The reading required will then be obtained on the BVcbo range.

Before leaving bipolar transistor testing it should be noted that the DC current gain (beta) measurement just described is only one of a number of current gain measurements which may be made on a bipolar transistor. Another important parameter is hfe, the "AC" or incremental current gain, which gives a more accurate description of the behaviour of a transistor in a common-emitter amplifier circuit.

Although hfe and other related gain parameters are not particularly difficult to measure, they generally involve somewhat more elaborate test circuits than that shown in figure 4(e). For this reason provision for measurement of these parameters is not usually made on simple or "general purpose" testers, being provided mainly by more specialised instruments.

In general, measurement of beta as the sole gain test applied to a device by a general-purpose tester may be justified on the grounds that it is simply provided, speedily performed and gives at least a convenient indication of the order of current gain provided by the device in a "typical" circuit.

FIELD-EFFECT TRANSISTORS (FETS): Although these devices fall into a number of different types, all consist basically of a "channel" element whose effective conductivity is a function of the potential applied to a closely coupled but isolated "gate" control element. In junction-type FETS (JFETs) the gate-channel isolation is provided by a reverse-biased P-N junction, while in metal-oxide-semiconductor (MOSFETs) and other insulated-gate devices (IGFETs) the isolation is provided by a thin layer of metal oxide. In both main types of FET the "common" and "output" ends of the channel are termed the "source" and "drain" respectively; MOSFET devices usually have an additional electrode, the substrate, which is in most cases connected to the source.

An important parameter describing FET performance is Idss, the channel current which flows when zero external bias voltage is applied to the gate. Idss is a measure both of the channel conductivity of a device and of its internal stabilising behaviour, and thus gives a good indication of the suitability of a device for practical circuit applications.

The measurement set-up required for Idss is quite simple, and is illustrated in figure 6(a). A known drain-source voltage Vds is supplied to the device with the gate electrode tied to the source, and a meter the resultant current. Typical devices exhibit an Idss between about 1 and 20mA; in general an Idss of greater than 20mA predicts difficulty in arranging for the device concerned to give useful gain at acceptable channel current levels.

The choice of testing voltage Vds is not critical, but too high a value will risk over-dissipation of devices exhibiting high Idss. A suitable value for most purposes is 10V.

An important characteristic of JFETs is BVgso, the breakdown voltage of the gate-channel junction with the drain open-circuited. BVgso shows the absolute limit of gate reverse bias which may be applied to the device for normal operation.

The basic test circuit used for measuring BVgos is shown in figure 6(b), and it may be seen that this circuit is similar to that used to measure breakdown voltages in diodes and bipolar transistors. The main difference is that in this case the test current level must be kept down to prevent excessive dissipation in the small and relatively fragile gate-channel junction; a convenient current level is 200uA, obtained using a 200V supply and a 1M series resistor.

It is most important to note that the BVgso test must never be applied to MOSFET or other IGFET devices, as breakdown of the gate insulation of these devices causes permanent damage.

A further aspect of FET performance which is generally quite important is the transconductance, which determines the "gain" of a device. As this parameter varies with channel current and is not constant a variety of characteristics are used to describe it, including the complete Id/-Vgs transfer curve.

For JFET devices the transfer curve closely approximates a parabola which is tangential to the Bgs axis at the so-called "pinch-off" voltage Vp and intersects the Id axis at Idss. At the latter point the curve has maximum slope, so that the maximum transconductance of a device occurs at zero gate bias and is commonly symbolised as Gmo.

If Idss and the pinch-off voltage Vp are known, Gmo can be found quite easily from the following expression:

$$Gmo = -2.Idss/Vp.$$

Figure 6(c) shows a basic circuit which permits both plotting of the Id/–Vgs tranfer characteristic of a device and determination of Vp. (The symbols and polarities shown are for N-channel devices, as with figure 6(a) and (b); for P-channel devices they would be changed appropriately.) As may be seen the circuit consists simply of an arrangement whereby the channel current Id may be measured at various values of applied reverse gate bias voltage –Vgs. The pinch-off voltage Vp is found simply by noting the value of –Vgs at which Id effectively falls to zero.

A practical circuit which would perform the FET tests described in figure 6 is shown in figure 7. The circuit has been arranged to employ a single meter movement which is switched to measure –Vgs and Id alternatively for transconductance measurements. It again uses a 50uA protected meter movement, with 10V and 220V supplies to provice the operating voltages and currents. In order to prevent damage to IGFETs a normally closed pushbutton is connected across the BVgso constant-current source so that the test cannot be made unless the button is pressed.

The meter ranges for the FET test circuit of figure 7 are as follows:

Idss:	0-50mA
BVgso:	0-50V
Id:	0-50mA
–Vgs:	0-20V

UNIJUNCTIONS: Probably the most common requirement of these devices is that they be capable of producing oscillations in a simple relaxation circuit. A useful "first test" for a UJT is therefore one in which the device is connected into a suitable relaxation circuit and its ability to oscillate determined. In general if a UJT will oscillate in this fashion it will be at least potentially suitable for most other applications.

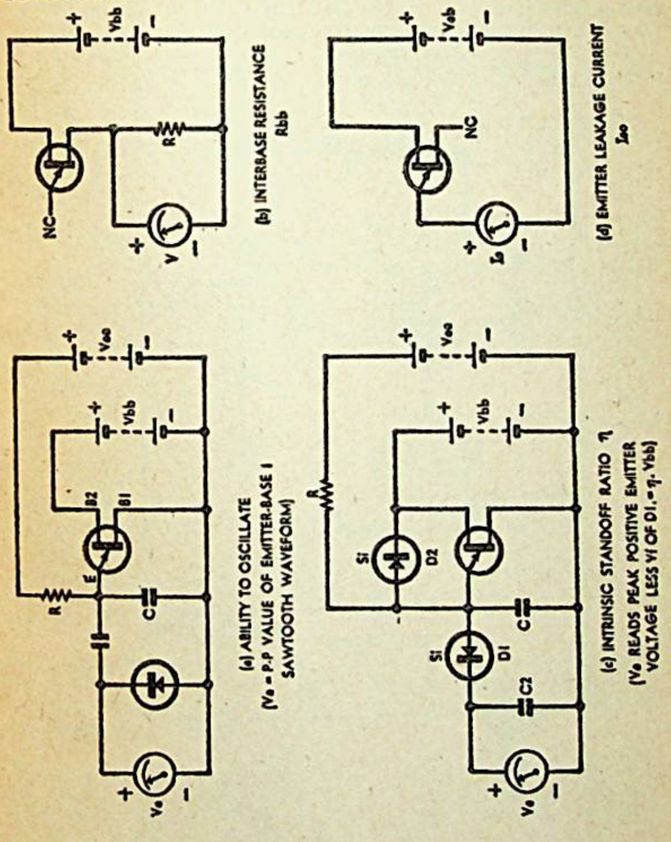
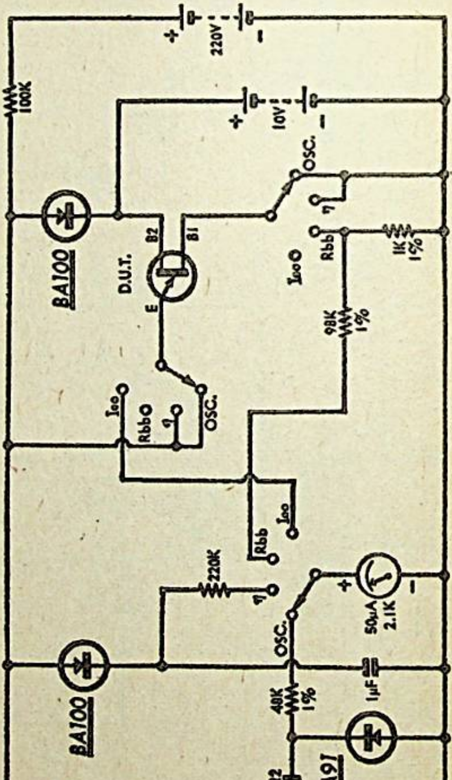


Figure 8 UNIJUNCTION TESTS

Figure 8(a) shows a basic circuit which performs this test. The base electrodes B1 and B2 of the UJT are connected to an interbase supply Vbb, while the emitter is connected to a simple R-C charging circuit connected to a second supply Vee. A simple half-wave meter rectifier circuit is used to indicate the presence of the expected sawtooth waveform at the emitter, the circuit performing in a similar fashion to a simple neon-tube relaxation oscillator.

A zero reading on this test generally indicates that the device concerned is faulty.

A further UJT characteristic used in circuit design work is Rbb, the internal "interbase" resistance. A convenient test circuit for measuring Rbb is shown in figure 8(b). As may be seen, it simply connects the interbase resistance of the device in series with a resistor R across the supply Vbb, to form a voltage divider. A voltmeter connected across R may be calibrated directly in terms of interbase resistance.



UJT TEST CIRCUIT

Figure 9

A supply voltage Vbb of 10V is usually quite satisfactory for this test, providing sufficient voltage for convenient measurements yet ensuring that the test device is not subjected to excessive dissipation. As typical devices have an Rbb falling in the range 2-5K a convenient value for R is 1K, in which case a 0-5V meter will give a useful Rbb scale having F.S.D. corresponding to 5K.

For critical design of UJT circuits an important characteristic of a device is the so-called "intrinsic standoff ratio", which is nothing more than the proportion of the device interbase resistance effectively lying between the emitter junction and the B1 electrode. The intrinsic standoff ratio is symbolised by the lower-case Greek letter Eta, as shown in the test circuit of figure 8(c).

In this circuit the UJT is again connected in a simple relaxation oscillator configuration. However this time use is made of the fact that in such a circuit the peak voltage reached by the emitter before conduction is equal to the forward-bias voltage drop of the emitter-base junction plus a proportion of Vbb equal to the intrinsic standoff ratio. Hence by coupling a large capacitor C2 to the oscillator charging capacitor C by a silicon diode D1, the voltage developed across C2 during oscillation will automatically equal that fraction of Vbb given by the intrinsic standoff ratio.

Diode D2 permits calibration of the meter Ve directly in terms of the intrinsic standoff ratio. In the absence of a UJT, D2 clamps the junction of R and C at a voltage which exceeds Vbb only by the forward voltage drop of a silicon diode junction. As this voltage across C corresponds to a hypothetical intrinsic standoff ratio of unity, the meter may thus be arranged simply to give a full-scale reading. The fraction F.S.D. indicated by the meter when a UJT is in circuit will then be equal to the intrinsic standoff ratio of the device concerned.

A full-scale reading on this test indicates that the device concerned has an open-circuited emitter. Conversely a zero reading indicates that there is an internal emitter-base short.

A fourth UJT characteristic which is useful in both design and servicing work is Ieo, the emitter junction leakage current. A device with excessive Ieo is generally unsuitable for level detection and timing applications, although it may be satisfactory for pulse generation. Typical devices have Ieo figures of less than 1uA, although some devices may range as high as 15uA.

It may be seen from figure 8(d) that the test circuit used for Ieo is similar to that used for leakage testing of diodes and bipolar transistors. A convenient value for the testing voltage Veb is 10V, while the meter may conveniently be a 50uA unit.

Figure 9 shows a practical circuit which would perform the four UJT tests described in figure 8. As with the previous circuits it employs a single protected 50uA meter movement and 10V and 220V supplies. The meter ranges for the four tests are as follows:

OSC:	0-10V p-p
Eta:	0-1
Rbb:	Inf – 1K
Ieo:	0-50uA

Although the practical circuit of figure 9 and those given in figures 3, 5 and 7 could be constructed as individual device test units, a more economical approach is to combine them into a single comprehensive test set.

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