

BERNARDS

BUILD YOUR OWN ELECTRONIC EXPERIMENTERS LABORATORY USING IC's

BY

B. B. BABANI

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ELECTRONIC
EXPERIMENTERS
LABORATORY
USING IC's**

**BY
B. B. BABANI**

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CONTENTS

SIMPLE AUDIO FREQUENCY OSCILLATOR USES 2 LOW COST IC'S	5
TTL PULSE DETECTOR USES INEXPENSIVE IC, LED HIGH IMPEDANCE VOLTMETER BASED ON LOW COST OP AMPLIFIER	15
SQUARE WAVE OSCILLATOR/PULSE GENERATOR USES 555 IC	16
DIGI DESIGNER	28
5-STAGE LOGIC PROBE HAS 7-SEGMENT READOUT	31
REPAIRING CRACKED PC BOARDS	35
LOW COST SCALER EXTENDS DIGITAL COUNTER TO 200MHz	40
BUILD OUR DIGI-METER : A DVM WITHOUT TEARS	41
BREAKDOWN VOLTAGE TESTER FOR SEMICONDUCTORS	49
LOW RANGE OHMMETER	56
EXPERIMENTAL BRIDGE TYPE WIDE RANGE OHMMETER	65
SYNCHRONOUS GATING UNIT FOR AUDIO TESTING	66
SIGNAL TRACER USES IC	74
SIMPLIFIED VERSION OF SQUARE WAVE GENERATOR	89
LIST OF EQUIVALENTS FOR CIRCUITS USED IN THIS BOOK	90
	92

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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SIMPLE AF OSCILLATOR USES TWO LOW COST ICs

Here is the design for an easy to build audio oscillator which uses only two low cost ICs. It generates both sine and square waves at low distortion, covering the frequency range from 3Hz to 30kHz. You can power it from batteries, or from a simple mains supply.

Looking through a recent edition of "Radio Communication", my interest was aroused by a very simple Wien bridge oscillator circuit by D.S. Jones. The circuit uses the ever popular and very low priced type 741 IC op amp. The supply suggested is 12 volts and the current just a few milliamps. Tuning is by a ganged 10k potentiometer and the frequency coverage is from 3Hz to 40kHz in four ranges. This looked like a good starting point for a low cost, simple, yet very worthwhile, economy audio oscillator.

By rearranging some of the time constants in the Wien bridge, the frequency coverage could easily be changed to go from 3Hz to 30kHz in four ranges. Instead of a 12V DC supply, this has been increased to 15V and could even be increased further to 18V. Hopefully, the increase in supply voltage would reduce the harmonic distortion to some degree.

As this audio oscillator set out to be an economy unit, a fairly obvious saving would be to do without a level meter. However, by taking some other quite inexpensive steps, the output voltage may be determined with quite good accuracy without the use of a meter.

The sine wave output is over one volt and by adding a preset potentiometer in series with the output load, the voltage across the load could be adjusted to one volt against a suitable meter. As the output voltage is almost constant over the entire range, it is then only necessary to provide a means of voltage division and the output voltage can be attenuated by a known amount.

By suitable switching, the sine wave output can be put through a type 555 IC used as a Schmitt trigger to give a square wave output. As the output of this stage is also in excess of one volt, the same procedure has been adopted to reduce the output to one volt and this is attenuated in the same way as for the sine wave output.

The oscillator circuit is the Wien bridge and for those unfamiliar with this arrangement it is shown in principle in Figure 1. As may be seen it consists of a high-gain differential amplifier 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 so 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 R1, R2, C1 and C2 is known as the “Wien network”. At a particular frequency determined largely by the values of the four elements, the transmission loss of the network falls to a minimum while the phase shift also passes through zero.

If R1 and R2 are equal in value, and C1 and C2 also equal in value, the frequency of oscillation is equal to the reciprocal of $2\pi RC$. Under these conditions, the transmission loss falls to a minimum of 3. In other words, there is a maximum transmission gain of 0.33.

To produce sustained oscillations, the overall loop gain must be at least unity and the phase shift either zero or a multiple of 360 degrees. For the circuit of Figure 1 to produce continuous oscillations, the amplifier gain must be at least 3 to compensate for the loss in the Wien network.

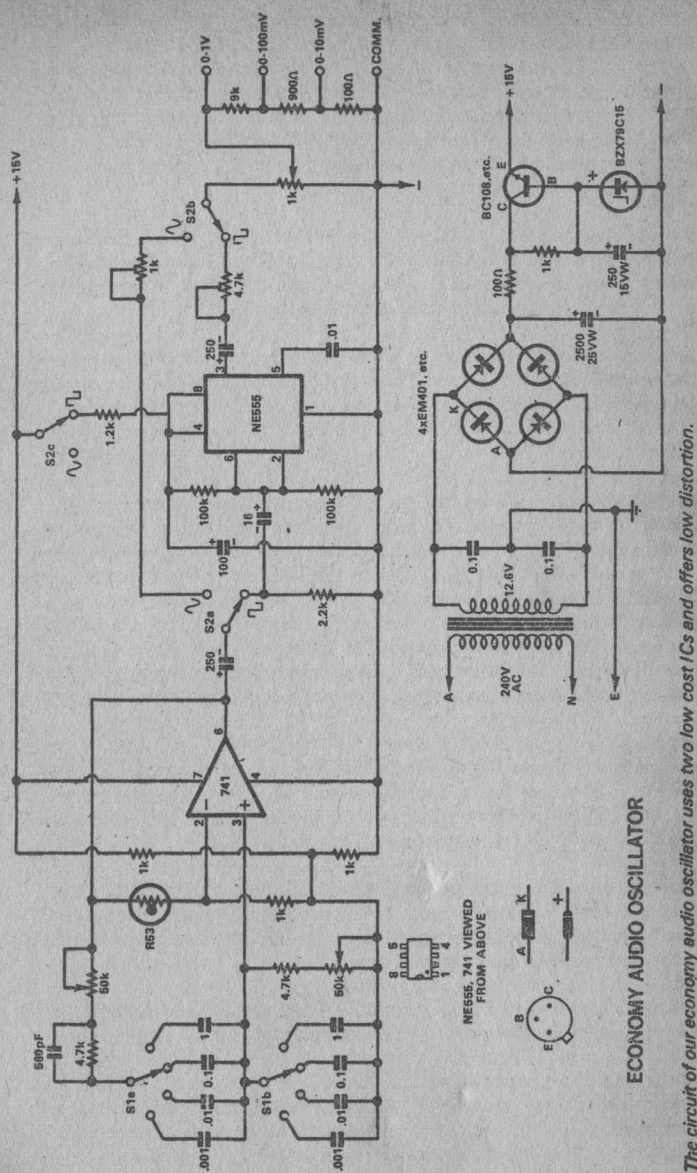
It is necessary for the amplifier to have an overall loop gain of at least unity for sustained oscillations. But if the gain is more than one, the oscillations will continue to grow until the amplifier is driven into clipping. To keep the gain at unity and oscillation at constant amplitude, a thermistor is used in the negative feedback network.

Since the thermistor has a negative temperature coefficient of resistance, its resistance falls as its temperature rises. When power is first applied to the circuit the thermistor will have a relatively high resistance and so there will be little negative feedback. The resulting high gain around the positive feedback loop will rapidly build up oscillation.

As the oscillations grow, the temperature of the thermistor rises, as this and resistor R3 are effectively connected in series across the amplifier output and so draw signal current. Hence the resistance of the thermistor falls, negative feedback increases and the effective amplifier gain drops.

Consequently an equilibrium is reached, as the output amplitude can only rise to the point where the thermistor has increased the negative feedback to correspond to an effective gain of three – giving unity loop gain. If the oscillations tend to rise above this level, the thermistor will reduce the loop gain and the oscillations will begin to die away. Conversely, if the amplitude tends to fall, the thermistor will increase the loop gain slightly above 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 at unity, and the oscillation amplitude constant. By using a suitable thermistor the output amplitude may be maintained at a level well below limiting, giving an output waveform with low distortion.



ECONOMY AUDIO OSCILLATOR

The circuit of our economy audio oscillator uses two low cost ICs and offers low distortion.

Looking now at the main circuit, the relative parts may be clearly seen with reference to the foregoing description of the theoretical aspects of this type of circuit. The positive and negative feedback circuits are easy to see, along with the two resistors and capacitors for each frequency range. Before leaving the oscillator section of the circuit, it may be seen that one of the 4.7k resistors is shunted with a 560pF capacitor. This is to compensate in some measure for stray capacitance in the circuit, which becomes significant at the higher frequencies.

The oscillator circuit is followed by a Schmitt trigger which can be used to convert the sine wave output from the oscillator to a corresponding square wave. The Schmitt trigger uses a type 555 timer IC as suggested by Walter G. Jung and the circuit performs this task adequately.

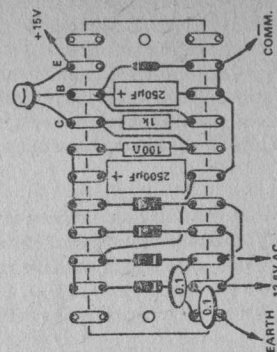
In order to select either sine or square wave output, a three-pole two-position toggle switch is used. This diverts the sine wave either through or around the Schmitt trigger. At the same time, the positive supply rail is switched on or off in accordance with the requirements of the Schmitt trigger.

Whichever output is selected, it is fed into a 1k linear potentiometer. Before being fed into this potentiometer, it is passed through a variable resistor in the form of a preset pot, which is set to give a predetermined output level. More will be said about this later on. To give a wider range of control over the output level, the output of the 1k pot is fed into a voltage divider. The divider provides for divisions of 1, 10 and 100, corresponding to attenuations of 0dB, 20dB and 40dB, respectively. In addition, the 1k pot is fitted with a knob calibrated 0 to 10, which means that attenuations may be achieved fairly accurately down to 60dB.

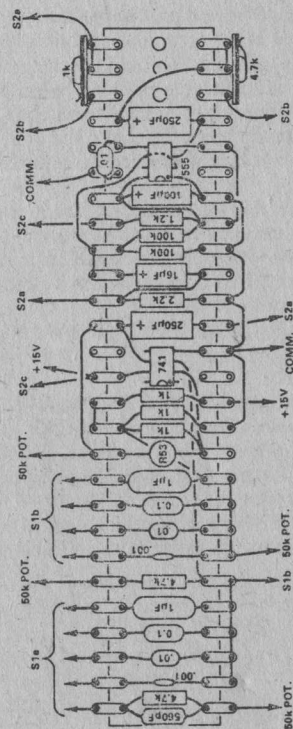
With the preset pots just referred to adjusted so that we have a level of one volt across the 1k pot, we are able to make settings between 0 to 1V, 0 to 100mV and 0 to 10mV. We may therefore set the output level fairly accurately anywhere between 1mW and 1V.

We elected to incorporate a power supply within the unit, although it would be quite possible to run it from batteries. The power supply is regulated at 15V and although it is a simple circuit, steps have been taken to keep hum and noise to a minimum.

The transformer has a 12.6V secondary winding rated at 150mA. Four diodes in a bridge give about 18V DC across the 2500µF filter capacitor. A BC108 transistor is used in a simple series regulator circuit which is referred to a 15V zener diode in its base. The two 0.1µF capacitors, one on each side of the secondary of the transformer help to reduce noise from the mains. The negative supply line is left floating as is normal practice with this type of instrument. Metal parts of the case are connected to the mains earthing system.



the simple regulated power supply is shown at right. Make sure that the thermistor is firmly secured in position.



These two wiring diagrams should make construction a straightforward process. The oscillator tagboard is shown at left, whilst

Some comments on the more important components may be helpful. Perhaps the item of most interest is the 50k linear ganged potentiometer. In the interest of economy, we used an imported deposited carbon unit under the name of Soanar. This has proved to be quite satisfactory; we have tried a number of them in the circuit and each one follows the calibrations well enough for all practical purposes. For readers who decide to use this potentiometer, we are printing a full size copy of the scale which may be used in any way that is considered best.

Before leaving the subject of carbon potentiometers, we can see no reason why some other brands of potentiometer should not be tried. However, it may be that the scale which we have calibrated against the Soanar unit may not suit another brand. In this, and other circumstances yet to be touched on, it would be necessary to calibrate your own scale.

When considering what potentiometer to use initially, we invested the idea of using a moulded carbon unit but as near as we can ascertain, these are not available in the type and resistance value required. In past projects we have used a wire wound potentiometer, made by Naunton. These are still available and may be used if any reader so desires. However, there are two points worth considering. It would be necessary to calibrate your own scale, and the price could be rather daunting. The choice of course is up to the individual.

The output level control is a 1k moulded carbon potentiometer in the prototype. While an ordinary carbon pot may be quite satisfactory, we considered that the moulded element would be more reliable from a linearity point of view. This is important as we have fitted a knob to this control which is calibrated 0 to 10. This facility, in conjunction with the output divider, makes it possible to set the output level with reasonably good accuracy.

The output divider just mentioned is made up of a number of resistors to give the correct ratios. The bottom element is 100 ohms and this presents no problem as it is a preferred value. The next section calls for 900 ohms and this is made up with 220 ohms and 680 ohms in series. The top section has a value of 9k and this is made up with 2.2k and 6.8k in series. We used Philips ½ watt types here and although we did not specify close tolerance, we have found them to be quite satisfactory.

The set of capacitors for the frequency ranges should be of good quality and preferably of 1 percent tolerance. The tolerance applies to all except the 1μF. Although it would be ideal to have this in the same tolerance, it would be quite expensive and we understand that availability is open to question. We found that with 10 percent tolerance units in circuit the amount of error was not significant for normal purposes.

The thermistor is an STC type R53 and this does an excellent job in this circuit. While we have not tried other types, we feel that it would be wise not to substitute for this item.

Both ICs are readily available. The type 741 comes in a number of brands but the 555 IC timer is only available from Signetics. Both ICs are 8-pin DIL are used in the prototype but other packages are also available if required. For our purpose, the smaller package is more convenient.

The prototype was built into a plastic case with an aluminium front panel and one which we have used a number of times before.

Most of the components are mounted on either one of two miniature tag boards and these are shown in diagram form. Care should be taken with the thermistor to stabilise it mechanically and this may be done by simply tying or gluing it to the adjacent 1μF capacitor. Sockets for the two ICs are fixed to the board by drilling either holes on the board carefully with a fine drill. Lugs of the sockets are pushed through the board and leads taken as shown in the diagram. The photographs show the location of all major components. The boards are stood off the case by 5mm or so by means of spacers, or by using a couple of extra nuts on each of the mounting screws.

The power transformer which we used had a centre tap on the secondary winding, but this is not required. We simply cut it off to about 10mm long and located it so that it was not likely to come into contact with any other parts of the unit when it was all assembled. A lead which should not be forgotten is from the mains earth termination to a lug on the metal front panel.

Once the unit is assembled it is ready for adjustment and calibration. Before going ahead with these operations, a simple check that it is working would be in order. Feed the output into a CRO or listen on a pair of headphones. If you choose the latter method, make sure that the oscillator is operating at a frequency which is covered by the phones in use. Assuming that you get sine waves and square waves on the CRO, or sounds resembling a flute with the switch in the sine wave position and sounds like a clarinet when the switch is in the square wave position, you are in a position to carry out adjustments.

Set the dial knob assembly on the potentiometer spindle such that the pointer corresponds with the marks on the scale for extreme limits of travel. Set the output knob on the output pot spindle so that 0 and 10 correspond with minimum and maximum output, respectively.

Set the range switch and frequency dial to give an output of about 100Hz. Set the output level control to 10 and take the output from sockets "E" and "0-1V" and connect to an AC voltmeter of good accuracy and high impedance. Set the toggle switch to sine wave output and adjust the 1k preset pot to give a reading of 1 volt. Set the toggle switch to square wave output and adjust the 4.7k preset pot to give a reading of 1 volt.

These Lissajous figures show (left to right) vertical to horizontal frequency ratios of 1:1, 2:1, and 3:1.

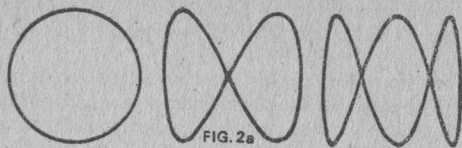


FIG. 2a

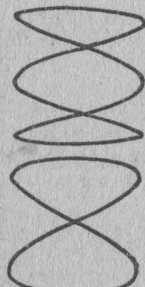
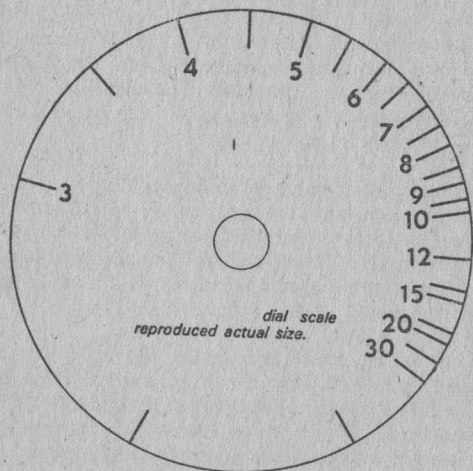


FIG. 2b

These Lissajous figures show vertical to horizontal frequency ratios of 1:3 (top) and 1:2 (bottom).

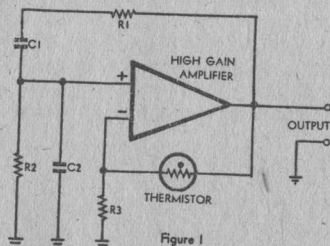


Figure 1

Above is the schematic of a Wien bridge oscillator showing the high gain amplifier and its associated positive and negative feedback networks.

If you have access to a calibrated CRO, the two preset pots just mentioned may be set against the CRO graticule. For sine wave output, the level should be set to give a reading of 2.8V peak-to-peak and for square waves, a reading of 2V peak-to-peak should be obtained.

If you have used the same type of 2-gang potentiometer as for the prototype, together with a copy of our scale, then you will not have to carry out any further calibrations. For readers who have used the wire wound, or some type of pot different from that on the prototype, then it will be necessary to calibrate for one range. This could conveniently be from 30 to 300Hz, with the range switch set to "X10".

If you are fortunate enough to have access to a digital frequency meter, then the job may be done quite easily. Another method is to generate Lissajous figures with the aid of an oscilloscope and another known source of audio frequency. The second source could be an accurately calibrated oscillator, a test record, an electronic organ or the mains frequency of 50Hz.

For those not familiar with the Lissajous figure method of frequency comparison, the following briefly outlines the procedure. The output of the known frequency source is fed to the horizontal amplifier of the CRO and the output of the oscillator to be calibrated is fed to the vertical amplifier. When the gains of the two deflection amplifiers are suitably adjusted, a series of patterns known as Lissajous figures are produced. Typical frequency patterns are shown. These enable the unknown frequency to be calibrated easily.

These figures differ from the normal pattern produced by a sine wave oscillator, since the horizontal deflection signal is a sine wave and not the normal sawtooth produced by the oscilloscope timebase generator.

Some examples of Lissajous figures are shown in Figure 2a. Each represents a different frequency ratio - 1:1, 2:1 and 3:1. Each ratio is that of the vertical frequency to the horizontal frequency. If the signal applied to the horizontal amplifier is 50Hz and the resulting ratio is 2:1, then the vertical frequency will be 100Hz.

The method of determining the ratio is to count the loops along the top of the pattern and the number of loops along one side. The ratio of the two numbers is the frequency ratio.

This basic rule makes it possible to identify a variety of ratios. For example, five loops along the top and two at the side would indicate a ratio of 5:2 and, related to 50Hz, this would indicate a frequency of 125Hz. Incidentally, submultiples can be identified in the same way and Figure 2b indicates the pattern shape for $\frac{1}{2}$ and $\frac{1}{3}$ of 50Hz, corresponding to 25Hz and 16.7Hz.

If you attempt to simulate these patterns on a CRO you will find that the patterns tend to drift. This is because the phase relationship of the two frequencies is changing because of small frequency drift of the two sources. Unless the two sources are very stable, the circle pattern will slowly change to an ellipse to a straight line as the phase relationship of the two signals changes. A circle represents a phase difference of 90 or 270 degrees while a straight line represents a phase difference of zero or 180 degrees. This drift in the pattern is only a portion of a cycle and so it has negligible effect on calibration accuracy.

The foregoing information on Lissajous patterns may be used to calibrate your oscillator and this may be applied most easily to the range 30 to 300Hz, as mentioned earlier. Once you get the easy points, such as 50Hz, 100Hz, 150Hz, etc., other intermediate and less obvious patterns will become obvious by virtue of interpolation of the already determined points. It is not a difficult operation and indeed, can be quite fascinating.

List of Component Parts

- 1 Case 184 x 115 x 118mm, with aluminium front panel
- 1 Transformer, 240V primary, 12.6V secondary at 150mA.
- 1 50k linear dual-ganged potentiometer (see text)
- 1 1k linear moulded carbon potentiometer
- 1 2-pole 4-position rotary switch
- 1 3-pole 2-position toggle switch
- 4 Banana sockets, 1-black, 3-red
- 1 3-way terminal strip
- 1 Jabel handspan knob for frequency scale
- 1 Scale (see text)
- 1 Knob with indicator dot for range switch
- 1 Knob calibrated 0 to 10 for 1k pot
- 1 Tagboard with 27 prs tags
- 1 Tagboard with 11 prs tags
- 4 Diodes, EM401 or RS276-1139 or BY127 or 1N4007
- 1 Zener diode, BZX79/C15 or RS276-624/564
- 1 Transistor, BC108 or RS276-2009
- 1 IC, 8-pin DIL 741 or RS276-010/1741
- 1 IC, 8-pin DIL 555 or RS276-1723
- 2 8-pin DIL IC sockets
- 1 1k linear trimpot
- 1 4.7k linear trimpot
- 1 Thermistor, STC type R53

RESISTORS (½ or ¼ watt)

- 2 100 ohms
- 1 220 ohms
- 1 680 ohms

- 4 1k
- 1 1.2k
- 2 2.2k
- 2 4.7k
- 1 6.8k
- 2 100k

CAPACITORS

- 2 0.001µF 630V polystyrene, low tolerance
- 2 0.01µF 630V polystyrene, low tolerance
- 2 0.1µF 100V polycarbonate, low tolerance
- 2 1µF 200V polycarbonate (may be low tolerance if available)
- 1 0.01µF 25V redcap
- 2 0.1µF 100V polycarbonate
- 1 16µF 40VW electro
- 1 100µF 12VW electro
- 2 250µF 12VW electros
- 1 250µF 15VW electro
- 1 2500µF 25VW electro

SUNDRIES

3-core flex with 3-pin plug and flex clamp, screws, nuts, spacers, hookup wire, solder.

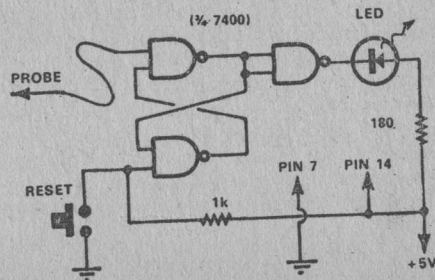
TTL PULSE DETECTOR USES INEXPENSIVE IC, LED
When one is testing an experimental logic circuit, or troubleshooting in a completed digital instrument or system, it is often necessary to detect single, fast pulses. At times this can be surprisingly difficult even using a wideband scope with extensive triggering facilities. Some logic probes have provision for "stretching" short pulses, but even this can be of limited help.

I have found that this simple circuit is very useful in this sort of situation. It will only detect negative-going pulses, but in practice this has not proved a problem as it is usually possible to look for a pulse where it is negative going.

As you can see, it is a simple R-S flip-flop formed from two gates of a 7400 IC. A third gate is used as an inverting driver for the indicating LED. The circuit is reset before each "search" by pressing the pushbutton.

If the probe is taken down to the low logic level (0.8V or less) for even as short a time as about 10 nanoseconds, the circuit will quickly flip over to the set state and latch there, indicating that a pulse occurred.

Almost any low cost LED can be used.



HIGH IMPEDANCE VOLTMETER BASED ON LOW COST OP AMP

Here is an easily built high impedance electronic voltmeter using a single 741 op amp IC. While the basic instrument is designed for DC measurements, probes are also described to allow its use for measuring both RF and low frequency AC signals.

Specifications

Ranges: DC – 0-1, 0-10, 0-100 volts. AC and RF – 0-1, 0-10 volts RMS.

Input impedance: On DC – 22 megohms, all ranges.

On AC and RF – 5 megohms approx., both ranges.

Frequency: 20Hz to 50MHz, using two probes.

Power supply: 2 x 9V batteries. Drain approx. 2mA.

When working on solid-state circuits there are many occasions when a DC voltmeter of really high resistance is very useful, because it can give substantially true readings in high-resistance circuits without upsetting their working conditions to any serious extent.

Needing such an instrument quickly and having no calibrated vacuum-tube voltmeter handy, the writer considered what might be done by using a 1-milliamp FSD meter which was on hand, in conjunction with an inexpensive linear integrated circuit. Such an arrangement has the advantage, apart from the possibility of high input resistance, that calibration is unnecessary if accurate resistors are available, because a reading of one milliamp full scale can easily be made to mean one volt, with other ranges in proportion merely by switching one resistor. Accuracy and stability can be very good because of the large amount on negative feedback employed with the integrated circuit. Furthermore, a 1-milliamp meter is cheaper and much more robust than those of ten or twenty times greater sensitivity which are usually found in normal high-resistance voltmeters – which, in any case, will have a resistance seldom exceeding half a megohm on the lower ranges.

Having worked out the details of such a DC instrument, the idea of extending its usefulness to radio-frequencies and, consequently, to AC voltage measurements present itself. The need here was for an RF voltmeter of low losses, low capacity, and an input impedance of several megohms, which could be applied to a tuned circuit or oscillator while leaving its "Q" almost unaffected. Therefore it was arranged that an AC or RF probe on a long flexible lead could be plugged into the DC instrument, the probe containing a full-wave peak-to-peak rectifier circuit using low-loss materials. It was then only necessary to change

one resistor in series with the meter movement by a DC-AC switch to have the meter show RMS values of AC voltage, with the original linear scale remaining accurate throughout except for the first one-third of the 1-volt scale where the curvature of the rectifier characteristic produces increasing non-linearity. A calibration curve for this part of the scale is shown hereafter.

At first sight it seemed that one AC probe would suffice for both audio and radio frequencies. However, it was desired to extend the audio range with full accuracy down to 20Hz and this required capacitors of 1 μ F in the rectifier circuit. Flat lacquered polyester-foil capacitors were used here and were quite satisfactory at the lower frequencies, but it was feared that their losses would be too high at radio frequencies, where much smaller capacity would in any case be sufficient.

Since low RF losses were an essential requirement of the project, a second probe was made for this work, using small polystyrene-foil capacitors and glass-fibre insulation, designed for frequencies from 100kHz upwards; while the other probe covers the range 20Hz to 200kHz.

Figure 1 shows the circuit of the DC instrument and Figure 2 the two probe circuits. Note that the low-frequency probe has a DC-AC switch so that it can be used for both purposes. AC and RF measurements are confined to the 1-volt and 10-volt ranges only, as anything exceeding 40 volts RMS could break down the rectifier diodes in the probes.

The Integrated Circuit

For readers who may not be very familiar with these devices, a very brief explanation may be in order. The type of IC used here is called an operational amplifier. Originally devised for carrying out certain mathematical operations in computer work, it is now used increasingly for all sorts of linear amplification purposes in both DC and AC circuits, and for these purposes it has potent advantages in space, cost, simplicity, stability and economy of working current.

An operational amplifier of this type is, in essentials, a direct-coupled multi-stage DC transistor amplifier of very high overall gain, intended specifically to be used with very heavy negative feedback. The initial gain, called open-loop gain, may be anything from around 2000 times to 200,000 times or more, depending on the type chosen, but the gain after feedback is applied, called the closed-loop gain, is seldom required to be more than 1000 times and is commonly much less.

This massive negative feedback not only extends the frequency response and reduces distortion and noise in the usual manner of NFB; it also makes the whole circuit largely independent of manufacturing tolerances in the IC itself and in its supply voltages. As a result, its performance in a given circuit can be predicted and determined almost complete-

by the choice of the external feedback components. Feedback can be either DC as in the present design, or AC as in the case of pre-amplifiers for audio frequency work, or both. There are, however, frequency limitations depending upon the type of IC chosen and very few are useful beyond a megahertz or so irrespective of the amount of feedback used. They are, as previously said, primarily DC amplifiers.

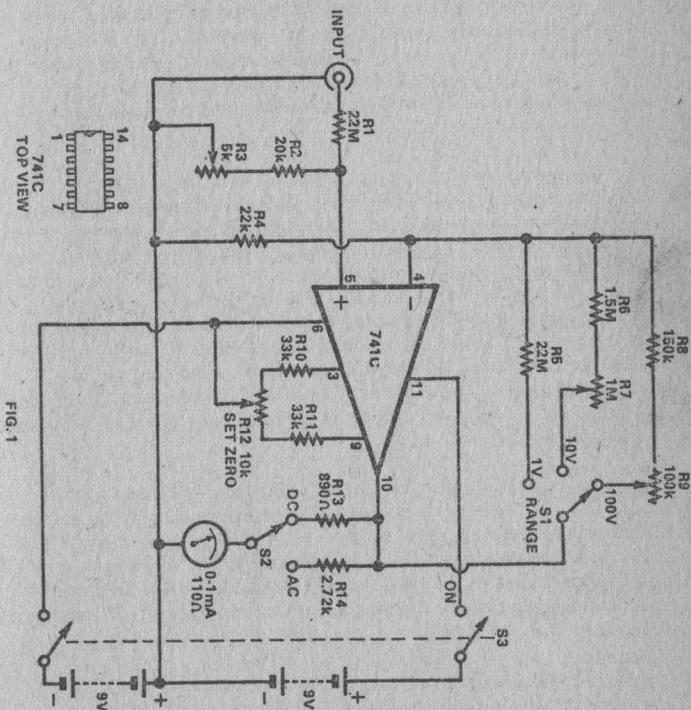
Usually an operational amplifier, or "op-amp", has two inputs and one output, plus a few other connections for supply voltages and other purposes. The input connections are called "inverting" and "non-inverting" inputs respectively, meaning simply that an input signal applied to either of them will appear at the output terminal either with the same or with the opposite polarity, according to choice of input terminal. To avoid drawing the whole complex internal circuit of the IC, the symbol shown in Figure 3 is used to represent it, namely, a triangle placed sideways with inputs on the left and output at the apex on the right. The inputs are labelled "+" for the non-inverting one and "-" for the inverting one. A signal applied to the "+" input appears at the output in phase with the input; if applied to the "-" input the output is in phase opposition; i.e. there is phase reversal.

Figure 3 shows the two basic circuits for an "op-amp". In 3a showing the "inverted" circuit, the input goes through R1 to the inverting terminal and negative feedback is applied through R2 to the same terminal. If the source of the signal has low impedance, the closed-loop gain of this system is, practically, $R2/R1$ and by varying R2 it can be made almost anything from unity up to the full open-loop gain. The input impedance of this arrangement is relatively low - virtually equp to R1 - and its output impedance fairly high; these properties are useful in many circuits but not for the present application where we need very high input and low output impedances, the latter to suit the low resistance of the milliammeter.

We therefore turn to Figure 3b, showing the non-inverting configuration. Here we feed the signal into the "+" terminal, but the feedback remains connected to the inverting terminal. The closed-loop gain will now be equal to $R3/R2$ and the input impedance will be high - several megohms at least - and output impedance 75 ohms or less. These characteristics evidently suit our present purpose and this circuit was chosen.

Several types of linear op-amps are available at moderate prices having widely-different characteristics. For this instrument the popular type 741 seemed most suitable. This is quite a complex affair containing twenty transistors, many resistors and a capacitor for frequency compensation and having a typical open-loop gain of 200,000 times. It is available either in a TO99 round can or as 14-pin or 8-pin dual-in-line packages; the 14-pin version was chosen. A second version, type 741C, is often available and sometimes at a rather lower price; this appears to be a so-called "Commercial" version with somewhat looser tolerances,

Complete circuit of the Op Amp voltmeter. The required range is selected by varying the feedback network. AC and RF voltages may be measured by means of external probes. (See Fig. 2)



...found quite satisfactory and used in this instrument. A constructor may use whichever type suits him or is procurable, and the package version desired.

The practical circuit evolved into that shown in Figure 1. Starting with the meter, we connect in series with it a resistor of a value which, together with the meter's internal resistance, will total 1000 ohms and thus give a full-scale deflection (1mA) with 1 volt applied across them. This will be the output required from the op-amp on all ranges and the meter scale divisions can be read as volts or fractions of a volt, multiplied by the range switch setting. The meter used here is an imported one of SEW make, type MR-85P with square face measuring $4\frac{5}{8} \times 4\frac{3}{16}$ inches, and its measured resistance was 110 ohms, requiring a series resistor of 890 ohms. This can be made with sufficient accuracy from 820 ohms and 68 ohms in series, both of 1pc tolerance, these being preferred values. (R13 Figure 1.)

While a large meter like this one is nice to work with, easily read and robust, any 1mA meter can be used. If its internal resistance is not known or cannot be measured, it can be calibrated in terms of whatever 1 volt standard meter may be available, using a variable pre-set trimmer resistor of around 2000 ohms as the series resistor; the accuracy will then be no better than that of the standard voltmeter used.

It was desired to have a DC input resistance of at least 20 megohms, this being determined by R1. The nearest preferred value is 22 megohms, and the required tolerance is 1pc; this being difficult to procure, it was made up from two 10 megohm and one 2 megohm units of 1pc in series. Both input circuits of the op-amp have to be returned to the earth line through resistors of only moderate value and of the same resistance, otherwise the slightly differing input currents can set up standing voltages large enough to make zero-balancing difficult. Thus R2 and R3, taken together, form a voltage-divider with R1 which reduces the input signal applied to the op-amp; a value of 22k was considered the highest desirable. Thus an input of 1 volt will be reduced to 1 millivolt at the input of the 741.

Since this must produce an output of 1 volt for application to the meter, the 741 must give a gain of 1000 for this range and the feedback circuit on the inverting input must be proportioned accordingly. This input, as above-mentioned, must be returned to the earth line through a resistance similar to R2 + R3, namely 22k. Thus R4 is determined and the feedback resistor R5 then automatically becomes 22 megohms, the ratio of R4 to R5 producing the required gain of 1000.

For the 10-volt and 100-volt ranges it is only necessary to change the feedback resistor to increase feedback and reduce gain in the required ratio; thus R6 + R7 becomes 2.2 megohms for the 10-volt range and

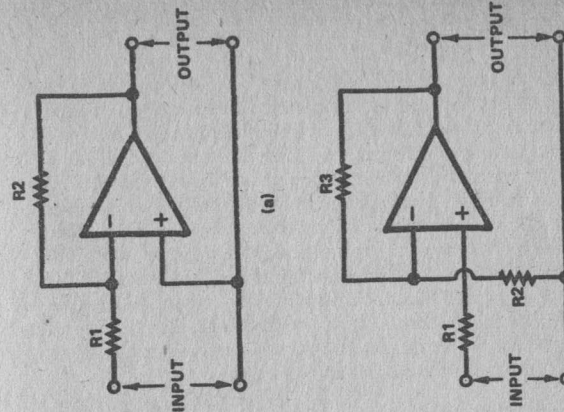


FIG. 3

The final circuit is based on (b). This gives a high input and low output impedance, ideal for this requirement.

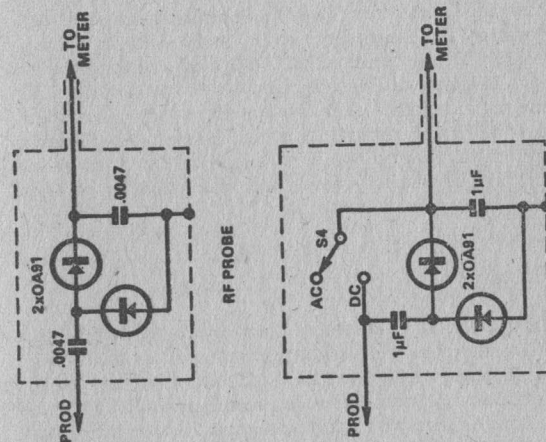


FIG. 2

Circuit for the RF and AC-DC probes to suit the voltmeter. Each probe is constructed in a small IF transformer can.

R8 + R9 becomes 220k to give the 100-volt range. These three conditions are shown in Figure 4 with the respective input and output voltages and gains indicated.

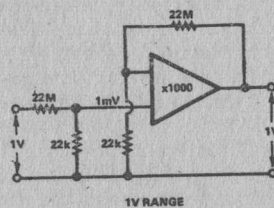
All resistors so far mentioned should ideally be of 1pc tolerance, and if these are available the accuracy of the instrument will be dependent only on that of the meter movement without further calibration. However, close tolerance resistors may sometimes be hard to get, so variable pre-set resistors have been added at the signal input (R3) and the 10-volt and 100-volt feedback resistors (R7 and R9) to permit exact adjustment, either to the required tolerance on a bridge or by direct comparison with another voltmeter. These can be miniature carbon trim-pots if low cost is important but they are awkward for precise adjustment and of doubtful stability of setting. A much better solution is the miniature Cermet trim-pot of the 10-turn type as made by Spectrol, Beckman, Morganite and others; these are probably worth the extra cost in terms of accuracy and stability.

It is necessary, in order to bring the output voltage to zero in the absence of an input signal, to balance the input currents of the two input terminals. Provision is made for this in the internal circuit by connecting a potentiometer (R12) across terminals 3 and 9, with its slider going to the negative supply rail (terminal 6). Additional resistors R10 and R11 were found desirable to make the adjustment less critical. R12 is a panel control with a suitable knob, used to set the meter reading to zero on the 1-volt DC scale only; this setting holds good on all other ranges; It is quite stable and shows very little drift after a few minutes' warmup, but it may be checked occasionally when great accuracy is called for.

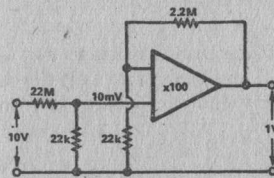
AC and RF Ranges

As stated earlier, the rectifier system in the probes actually produces a DC output proportional to the peak-to-peak value of the measured AC voltage, which is roughly 2.83 times the RMS voltage. If applied to the DC voltmeter circuit as it stands, the scale readings would have to be divided by this factor, a very awkward procedure, or else the meter would have to be re-calibrated with an additional peak-to-peak AC scale; both expedients to be avoided. The alternative is to reduce the milliammeter sensitivity until it reads the RMS value of the peak-to-peak voltage, thus preserving the original single scale for all purposes. This can be easily done by changing the multiplying resistor in series with the meter, and switch S2 does this by bringing in R14 to replace R13. Its value can easily be calculated.

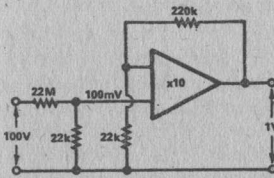
The combined resistance of the meter and R14 must now be 1000×2.83 , or 2830 ohms; subtracting the meter's internal resistance of 110 ohms (in this case) give 2720 ohms for R14. If desired, this could be made a variable resistor to permit fullscale deflection to be adjusted on



1V RANGE



10V RANGE



100V RANGE

FIG. 4

Showing how the meter ranges are provided by changing the feedback resistor values. Values for other ranges can be calculated.

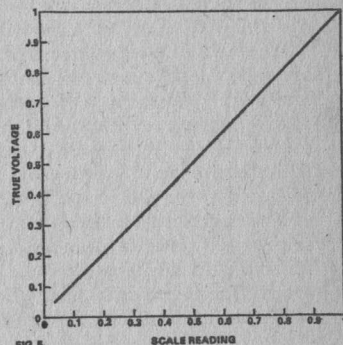


FIG. 5

Correction curve for the 1V AC range. Although the error is small this curve will correct for it where it may be important.

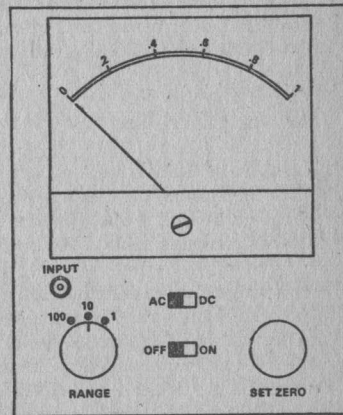


FIG. 6

The physical form of the finished meter is not critical, but the panel arrangement shown should serve as a guide.

the 10-volt scale by comparison with another AC voltmeter, but most such voltmeters (of the rectifier type) are not accurate to better than 4 or 5pc and the use of the calculated resistor value is likely to give greater accuracy.

As mentioned earlier, the lower one-third of the 1V AC scale becomes non-linear, but the curve of Figure 5 shows that this is not serious. Nevertheless, the curve gives quite accurate correction down to 0.1V RMS at least.

It is worth remembering that by leaving S2 in the DC position while measuring AC, one gets, effectively, the peak-to-peak reading instead of RMS. In other words the sensitivity is increased by the factor 2.83 and although the scale is not calibrated for it, this increase is often quite useful for comparative readings of RF voltages too small to be read comfortably on the RMS scale.

The effective input impedance is lower when using the RF and AF probes. For the peak-to-peak rectifier arrangement used the effective load presented to the circuit under measurement is usually taken to be one-quarter of the DC load resistance, in this case 22 megohms (R1). Thus the AC and RF probes load the circuit under measurement with about 5 megohms, with a few picofarads of capacity in parallel; this is negligible for most tuned circuits.

Two small 9-volt transistor radio batteries are adequate for the supply and as the total drain is only around 2 milliamps their life will be long. Lower voltages down to perhaps 3 volts could be used without much effect on performance but the 9-volt supplies are recommended. Note that both positive and negative supplies are needed.

RF and AC Probes

These are of similar construction, the only difference being in the size of the capacitors, the insulation used and the provision of an AC-DC switch on the AC probe, allowing it to be used for DC measurements without changing cables. Figure 2 shows the circuits.

Each probe is assembled in a square IF transformer can about 1½in square by 2½in high. The open end carries the insulating panel supporting the whole circuit assembly and a sharp 1¼in spike to serve as a prod, while the shielded coaxial cable carrying the DC output to the meter is brought out through a hole in the solid end of the can. The shielded coax cable is essential here to avoid pick-up of spurious signals, hum, etc, because of the very high input resistance of the meter circuit. It is about two feet long for convenience and is terminated at the meter end with any small coaxial connector of good insulation resistance, to suit a matching socket on the front panel.

The diodes in both probes are type OA91, of the germanium point-contact type, with a peak voltage rating of about 115 volts. They are chosen in preference to silicon types, firstly because point-contact types are usable to higher radio-frequencies than other constructions; secondly, because their conductivity starts at lower voltage than silicon types, thus contributing to better sensitivity and scale linearity; and thirdly, because their slightly higher leakage gives faster response and quicker recovery after a momentary overload. Silicon diodes of good quality have such low leakage that if the probe capacitors accumulate an overload charge it takes quite a time to leak away, and in the meantime the meter is paralysed.

Capacitors in the RF probe are small tubular polystyrene types (not polyester) of excellent "Q" and insulation. Uncased silvered-mica types would be equally good but probably more bulky, and most available mica capacitors seem to come in "moulded mud" cases whose RF losses are not to be trusted. Ceramic capacitors are quite unsuitable because of high RF losses and should not be used.

The AC-DC probe uses flat lacquer-coated polyester foil capacitors of 1μF which fit the can fairly snugly; the writer used "ELNA" brand. This probe has a miniature slide-switch of the smallest available size mounted on one side wall, to change the probe from AC to DC operation simply by shorting input to output. The prod consists of a 1½in length of 6BA threaded brass rod, sharpened to a point at one end and mounted at the other by nuts in the centre of a square panel of insulating material. The latter is mounted over the open end of the can by whatever mounting means are provided – lugs, spade bolts, or whatever.

The AC-DC probe used ¼in bakelite sheet for this panel, but in the RF probe a piece of polyester-glass-fibre sheet salvaged from a high-quality printed circuit was used. Sheet polystyrene would have lower losses, but is hard to find and ceramic materials, also good, are too difficult to work.

Inside the can the components are supported on a vertical panel of the same material, cut to fit the internal dimensions of the can and mounted to the top panel by a generous fillet of Araldite adhesive or else by small lugs and bolts. No specific constructional details are given as so much depends on the components actually available.

The AC-DC probe will need careful arrangement to leave room for the switch without fouling the rather large capacitors. For the screened leads the writer used PVC insulated, screened and PVC sheathed coaxial cable of 3.5mm overall diameter – a pleasantly flexible size which imposes minimum strain on the meter. A short length of tinned-copper braid equipped with an alligator clip is attached to the probe can to serve as earth connection to the circuit being measured.

A useful accessory is a further alligator clip provided with a 6BA nut mounted coaxially on its shank; this can be screwed onto the prod

when a temporary connection is needed to a circuit, so leaving both hands free.

Construction

The prototype is assembled on a sloping panel measuring $7\frac{1}{4} \times 6$ in, set at a 30-degree angle on the front of a pine-board box of sufficient depth to house the batteries. Figure 6 gives the panel layout.

The 741 IC is best used in a 14-pin DIL socket to avoid soldering directly to the IC pins, which can be a tricky job. These sockets exactly match the holes in 0.1in matrixboard or Veroboard, and their pins can easily be soldered to wires or strips on the reverse side. The writer favours Veroboard for this, and a suitable connection scheme for the circuit of Figure 1 is easily arrived at. If suitably dimensioned, this board, carrying most of the components, can be mounted directly on the meter terminals by loops of 18-gauge tinned copper wire.

Note carefully the small notch at one end of the 741 to identify it with the pin numbering diagram (see Figure 1). The socket should have a similar small notch at one end and the 741 must be inserted with the notches coinciding. There is nothing else to prevent the 741 being inserted back to front, which may do it no good at all. And when inserting or removing it, be very careful to do so with an absolutely straight pull or push, otherwise the fragile pins are almost certain to be crumpled or otherwise damaged, perhaps beyond repair. Once safely in its socket, the less it is moved the better, but the ability to remove it during servicing or circuit changes is a great advantage over permanent soldering into the circuit.

The Range Switch S1 should preferably be a miniature rotary wafer type to conserve space. The "ON-OFF" switch S3 must be a double-pole type, to switch both batteries. For this and also for S2 the writer used small slide switches, again to save panel space.

Because the instrument is virtually self-calibrating if 1pc resistors have been used throughout, no special instructions are needed in that case.

If the suggested variable resistors have been included, two courses are open. If a Wheatstone Bridge is available or can be borrowed, each combination of fixed and variable resistor can be set up to the exact value needed before wiring into the circuit and further calibration is then unnecessary. The values required are

$$\begin{aligned} R2 + R3 &= 22k \\ R6 + R7 &= 2.2 \text{ megohms} \\ R8 + R9 &= 220k. \end{aligned}$$

R4 and R5 cannot well be made adjustable without badly complicating the scale-setting procedure; they must be of 1pc tolerance.

Failing the bridge, each DC range can be set at full-scale deflection by comparison with another voltmeter, starting with the 1-volt range. A supply of finely-adjustable voltage, or a potentiometer across a fixed supply can be used. The 1-volt range is best set up by feeding in an accurate one volt DC and adjusting R3 for full-scale reading, having first zeroed the pointer using the Zero Set knob R12. The 10-volt and 100-volt ranges can then be set up in succession by adjusting the appropriate resistor in the feedback chain (R7 and R9 respectively.)

With the DC ranges correct, the AC ranges should not need further calibration. However, if an AC voltmeter of known accuracy at 10 volts is available for comparison the 10-volt AC range can, if desired, be set correctly at 10 volts full scale by using, say, a 5k variable resistor for R14. The 1-volt scale will then also be correct and both will be accurate when using the RF probe over its designed frequency range.

Parts List

RESISTORS

R1 22 megohms, 1pc (see text)

R2 20k 5pc

R3 5k linear trim-pot

R4 22k 1pc

R5 22 megohms, 1pc (see text)

R6 1.5 megohm, 5pc

R7 1 megohm linear trim-pot

R8 150k 5pc

R9 100k linear trim-pot

R10 } 33k 10pc

R11 }

R12 10k linear potentiometer, panel mounting

R13 890 ohms, 1pc (see text)

R14 2720 ohms, 1pc (see text)

All fixed resistors are $\frac{1}{2}$ watt carbon film or metal oxide film

CAPACITORS

2 5000pF, tubular polystyrene pigtail type

2 1 μ F flat lacquered polyester foil, 200VW

SEMICONDUCTORS

1 Integrated circuit operational amplifier, uA741C or

RS276-007/010/1741

4 Diodes, OA91 or RS276-1102/1136.AA117-118 1N618

MISCELLANEOUS

Meter: 0-1 milliamp, SEW Type MR-85P or similar, to choice

- S3 } Miniature slide switches
 - S4 }
 - S1 } Any small one-pole three-way switch, panel mounting
 - S2 } Miniature slide switch
 - 1 } Small coaxial socket, panel mounting
 - 2 } Coaxial plugs, to suit above
 - 1 } 14-pin DIL socket
 - 2 } 1-inch knobs, to choice
 - 2 } 1F transformer cans, approx. $1\frac{1}{8}$ inch square x $2\frac{1}{4}$ inch high
 - 4 } feet of 3.5mm screened and PVC sheathed coaxial cable
 - 2 } 9-volt transistor radio batteries, P1 size
 - 2 } battery connectors for above
- Matrix board or Veroboard, 0.1in hole centres
 Front panel, cabinet, wire, screws, nuts, screwed brass rod 6BA,
 three alligator clips.

SQUARE WAVE OSCILLATOR/PULSE GENERATOR USES 555 IC

With a little care, the popular 555 or Tandy Archer RS276-1723 monolithic timer IC can be used as a square wave oscillator or variable duty cycle pulse generator. The circuits and design details are given in this short note.

The popular 555 timer IC can be connected as an oscillator, as shown in Figure 1. The capacitor C is charged by current flowing through R1 and R2 in series, and the time the output is high will be given by:

$$T1 = 0.685(R1 + R2)C \text{ seconds}$$

where the resistance is in megohms and the capacitance in microfarads.

At the end of this time, a ground is applied to pin 7 from within the device, and the capacitor is discharged through R2 in a time:

$$T2 = 0.685R2C \text{ seconds.}$$

The pulse repetition time will therefore be:

$$T3 = 0.685(R1 + 2R2)C \text{ seconds.}$$

With this simple circuit, the duty cycle, or the ratio of the pulse "ON" time to repetition time will be:

$$(R1 + R2)/(R1 + 2R2).$$

The duty cycle can be set from slightly more than 50pc to nearly 100pc by a suitable choice of the values of R1 and R2. The lowest value of R2 will be determined by the maximum current that may flow into pin 7 without exceeding the dissipation rating, when it is held low during the discharge cycle.

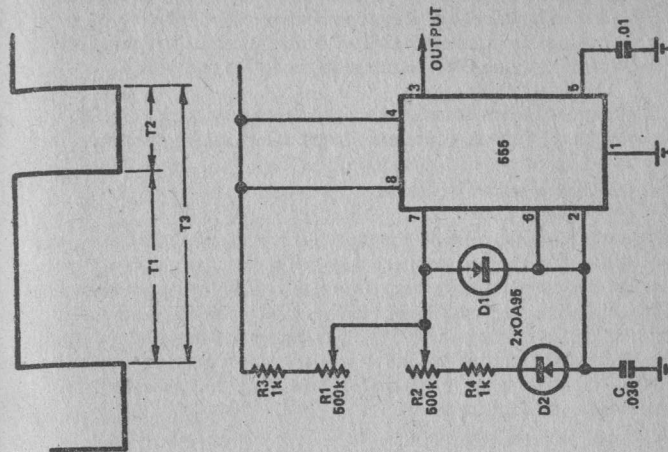


FIG. 3

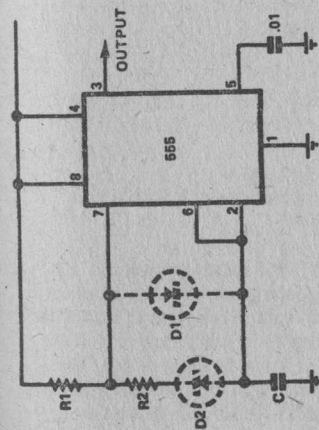


FIG. 1

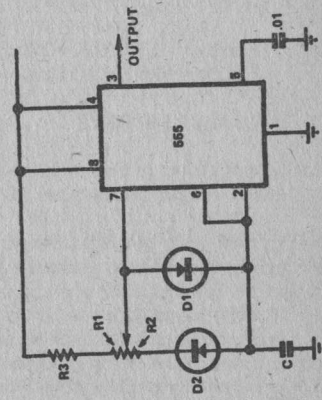


FIG. 2

The addition of the diode D1 (shown dotted) between pins 6 and 7, allows the capacitor to charge through R1 only, and discharge through R2, so the duty cycle is now $R1/(R1 + R2)$. It can therefore be varied over almost the full range from 0 to 100pc. The charge time will not be completely independent of R2, as some current will flow through it due to the voltage drop across D1. This problem can be overcome by inserting a second diode D2 in series with the lower end of R2.

Two applications of this circuit are a fixed frequency, variable pulse width oscillator, Figure 2, and a wide range square wave generator, Figure 3.

Variable Pulse Width Oscillator

As the repetition time is set by the sum of $R1 + R2$, and the duty cycle by $R1/(R1 + R2)$, the two resistors may be replaced by a potentiometer so that the duty cycle may be varied from 0 to 100pc without changing the repetition time (or frequency). R3 must be included to limit the maximum current into pin 7, and this will determine the minimum pulse width. In practice there is a small variation in the repetition rate, but this is of little importance in applications such as a pulse width control to vary the speed of a DC motor.

Wide Range Wave Generator

If R1 and R2 in Figure 1 are both varied, but remain equal in value, the repetition rate will be varied, while the output remains a square wave. A suitable circuit is shown in Figure 3. Resistor R3 is again added to limit the maximum current into pin 7, and R4 must then be included to balance its effect so that $(R1 + R3) = (R2 + R4)$.

The limits of the frequency range are given by

$$F_{\min} = 0.73/(R1 + R3) \text{ Hz.}$$

$$F_{\max} = 0.73/R3C \text{ Hz.}$$

and their ratio is $(R1 + R3)/R3$.

For the values shown in Figure 3, the lowest frequency is 40Hz, and the highest 20kHz, with a range of 500:1.

The output will only be a square wave if R1 and R2 track exactly. The maximum error is likely to occur if a log. potentiometer is used, to avoid crowding the frequencies at one end of the scale. This type of potentiometer is usually made with two linear segments so that it approximates a logarithmic curve and the tracking error is likely to be biggest at the crossover between the two segments. In practice, the ratio of off to on time may vary to as much as 40:60 instead of 50:50 over this part of the range. If a reasonably accurate square wave is needed over the full range, linear potentiometers will provide closer tracking.

THE DIGI DESIGNER

If you're involved with logic circuit design using digital ICs, you'll find this little unit irresistible. It provides all the basic requirements for fast and easy lash-up of new circuits, yet also doubles as a logic trainer.

A breadboarding aid of one sort or another is almost essential when one is attempting to try out even a simple circuit using the usual dual in-line ICs. Without such an aid, progress tends to be painfully slow, and the ICs and other components are likely to be damaged due to repeated lead bending and soldering operations. If you've tried it, you'll know what I mean.

In its most basic form, a digital bread-boarding aid may be little more than a scheme which allows the ICs to be plugged into sockets, with their connections brought out to convenient solder pads or terminals. But before it can be put to much use, this basic setup must be supplemented with a power supply, a source of test signals, provision for monitoring logic levels and a means of setting control signals.

With the Digi Designer, we have provided not just a basic breadboard aid, but all of these supplementary facilities as well. Although it measures only 19 x 21.6 x 10.2 cm. the Digi Designer combines a multi-device breadboard socket, a power supply, four logic level monitors with lamps, a dual-polarity clock generator with six switched frequencies, four logic level switches and two dual polarity pushbutton pulsers with bounce suppression.

It thus forms a complete miniature logic laboratory, with which you can easily and quickly lash up – and test – any number of possible circuits and logic configurations. All you need is the ICs and some lengths of hookup wire, and away you go!

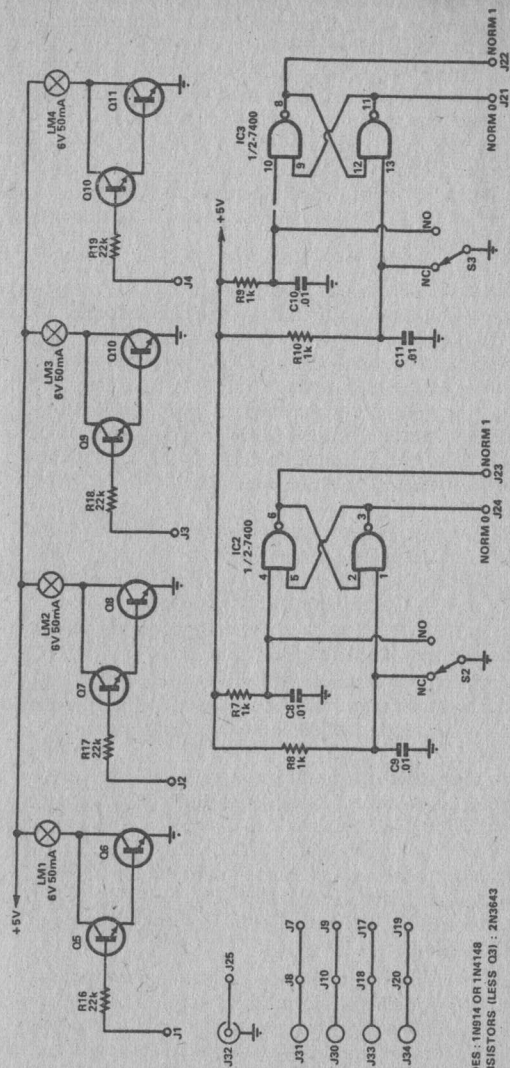
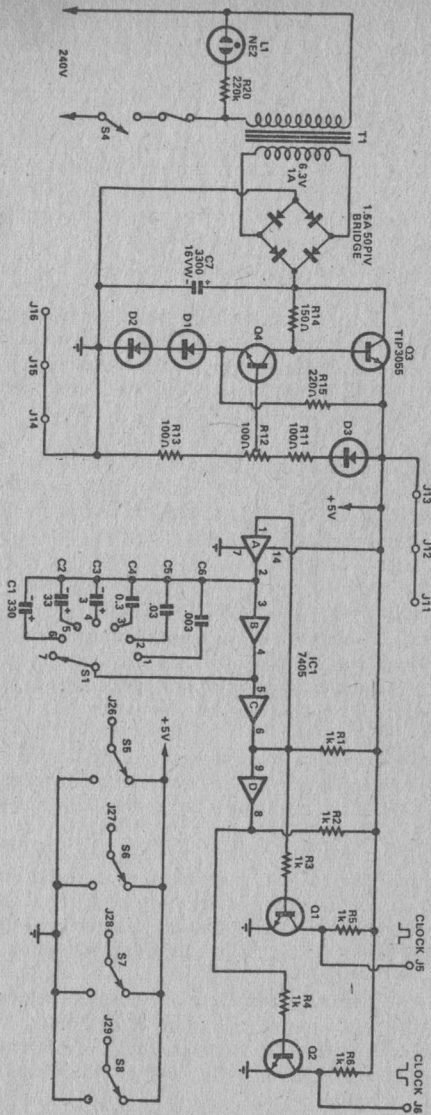
Needless to say, it would also be very suitable for use as a logic trainer in schools and colleges. Provided with a few standard gate and flip-flop ICs, it can readily be used to demonstrate virtually all of the basic laws of digital logic.

Heart of the Digi Designer is a multiway breadboard socket, Based on a precision moulding of high-impact plastic, the socket provides some 328 nickel-silver spring contact clips, grouped to allow multiple connections to IC pins and other devices.

The clips are designed to provide low resistance contact with IC pins, transistor leads, component pigtails and solid-core hookup wire from 22 to 26 gauge. At the same time they require only a relatively low insertion force, which should avoid strain to multi-pin ICs and similar devices.

The fact that ordinary solid-core hookup wire may be used for the interconnections makes the unit very convenient in use. There are

DIGI DESIGNER



DIODES: 1N614 OR 1N1486
TRANSISTORS (LESS Q3): 2N3643

The circuit of the Digi Designer is shown above. While simple and straightforward, it provides all of the required features with no sacrifice in performance. Note the dual polarity clock generator.

no special patchcords to become lost or broken, and leads of any length may be fabricated with a minimum of effort right on the spot.

The socket is mounted centrally on the aluminium front panel of the Digi Designer, with the lamp monitors, clock generator and mains switch, pilot and fuse above it. On the lower part of the panel are the logic switches, the pushbutton pulsers and a coax socket/pin jack combination which may be used either to feed external signals into the unit, or extract signals from it.

Immediately adjacent to the breadboard socket are pin jacks supplying plus 5V and ground, also four pairs of pin jacks connected to screw terminals. The latter again provide a convenient means of interfacing the circuitry with external instruments or other Digi Designers.

The circuit of the unit reveals a very practical engineering approach to its design. Considering the functions performed, there are surprisingly few parts and the wiring is quite straightforward. In fact the Digi Designer is a good example of the all-too-often forgotten art of "getting the most from the least".

The power supply is a fairly conventional series regulator using a plastic version of the well-known 2N3055. Two normal silicon diodes are used as the reference, with a third in the feedback divider presumably for temperature compensation.

The clock generator uses three of the inverters from a 7405 IC in cascade, with a capacitor across the centre element. The capacitor value is switched to provide the six frequencies: 1, 10 and 100Hz, and 1, 10 and 100kHz. Two discrete transistors are used as output buffer stages, with a fourth inverter from the 7405 used to provide the complementary output.

The lamp indicator circuits are very simple, each comprising only a 6V/50mA lamp, an input resistor and two general purpose NPN transistors connected in the Darlington configuration.

In contrast with these, the pushbutton pulser circuits seem to reveal a "BBbelt and braces" design philosophy. Each uses two gates from a 7400 IC as a simple R-S flip flop, which is normally quite capable of giving full suppression of button contact bounce. But just to be sure, the designer has also fitted the inputs with R-C integrator circuits. This may not be a case of gliding the lily, though, as the SPDT buttons used are of a rather simple and rudimentary type.

Inside the case, all the components are mounted on the rear of the front panel - including the power transformer. The power supply and clock generator wiring are on a fairly large printed wiring board, while the lamp drivers and pushbutton circuits are mounted on two much smaller boards. These are supported solely by the wiring, but this should cause no problems in view of their modest size and weight.

5-STAGE LOGIC PROBE HAS 7-SEGMENT READOUT

For anyone working with logic circuits, it should be invaluable. Detecting no less than five distinct logic conditions, it indicates them simply and elegantly using a 7-segment readout.

The logic probe described is the latest in a line of probes I have constructed. Working with industrial control equipment using TTL devices I have tried most available logic probes and noted the following.

Probes including pulse stretching circuitry can be a problem as it often proves necessary to move the probe supply leads to the IC being tested, to prevent false triggering. Where stretching is included, provision to disable should also be provided.

Secondly the probe must show a large number of logic states or it will be put aside for an oscilloscope.

Finally the probe must obtain power from the circuit under test, be hand held, and have protection on the input and output against the most common forms of abuse.

The probe developed is unusual in that it will detect 5 logic states and display these on a 7-segment LED display. For details of the display and logic states covered, see Figure 1.

Pulse stretching circuitry was omitted from this device, as I use a separate probe for this form of measurement.

Input Detection

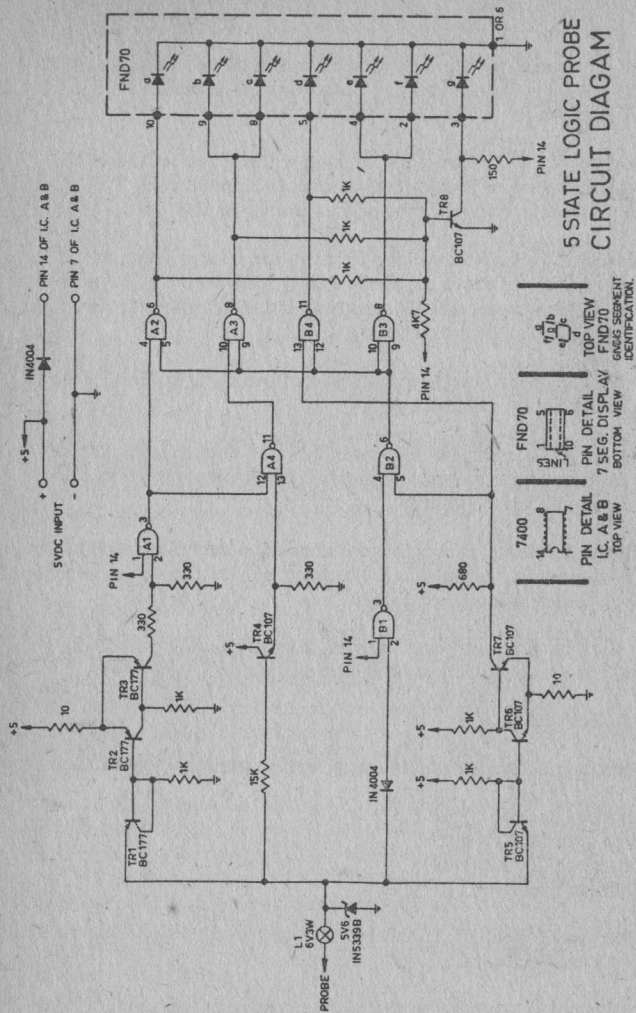
Individual detectors for the following 4 logic states are provided.

1. Short to positive (TR2 & 3)
2. True logic '1' (TR4)
3. True logic '0' (Gate B1)
4. Short to negative (TR6 & 7)

The 5th state (open circuit or incorrect logic level) is displayed if all the above states are absent.

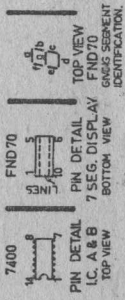
Detecting a short to positive or a short to negative relies on a similar arrangement, and only the short to negative circuit will be detailed.

Transistors TR6 & 7 form a Schmitt-Trigger with TR6 normally in conduction. Placing the probe on a voltage less than the 50mV dropped across the 10 ohm emitter resistor, diverts base current from TR6, taking it out of saturation, bring TR7 into conduction.



The complete probe circuit. There are four detectors, two for the "short to rail" states. Circuit is fully protected.

5 STATE LOGIC PROBE CIRCUIT DIAGRAM



Transistor TR5 has a dual role, normally acting as a blocking diode, but when forward biased cancelling the base emitter drop of TR6.

To activate the Schmitt-Trigger the probe must contact a potential below 50mV with a shunt resistance less than 10 ohms.

To detect logic '1' the emitter follower TR4 drives a standard TTL input. Whereas a gate will respond to logic '1' and open circuit in a similar manner, TR4 will only follow a true logic '1'. The base emitter drop of TR4 also ensures a full logic level must occur to initiate a change of state.

Logic level '0' is taken to a standard 74 series gate input for detection. Including a diode in series with the gate reduce the '0' level accepted to the worst case level.

Overvoltage Protection

The input probe is by-passed by a zener diode, thereby limiting the maximum voltage to reach the input detectors to 6 volt positive and 0.5 volt negative. Lamp L1 is the load for the zener, having resistance below 2 ohms in normal use, but increasing to 25 ohms with 20 volts on the probe. Fusing of the lamp takes place as the voltage increases. These components are for protection only and voltages above 5 volts should be avoided in normal operation.

Protection from reversal of the supply takes the form of a diode in series with the IC positive pin. Provided the reverse voltage does not exceed 5 volts it is not necessary to similarly protect the input detectors - indeed the short to positive detector would cease to function with diode protection.

Output Decoding

With detection of the input states completed the information enters a gate network for decoding to a form suitable for the 7-segment display. The gates determine which segment (or segments) should be illuminated for a given input. In addition, only one level must be displayed at a time.

The short to negative or positive rail is only a specialized instance of a logic '0' or '1' and hence more than one input detector will be activated by a short condition. However the gating is arranged so the short detectors override the normal logic detectors.

Four gates decode segments 'A' through 'F' while segment 'G' (open circuit indication) is controlled by TR8. The latter transistor by-passes all current from segment 'G' while any other segment is illuminated.

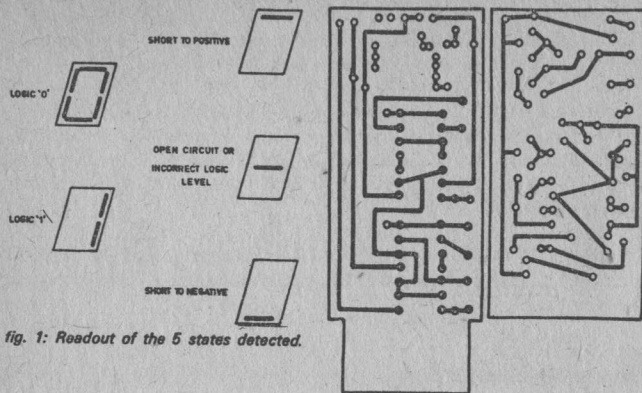


Fig. 1: Readout of the 5 states detected.

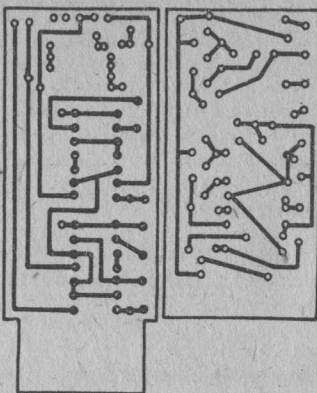
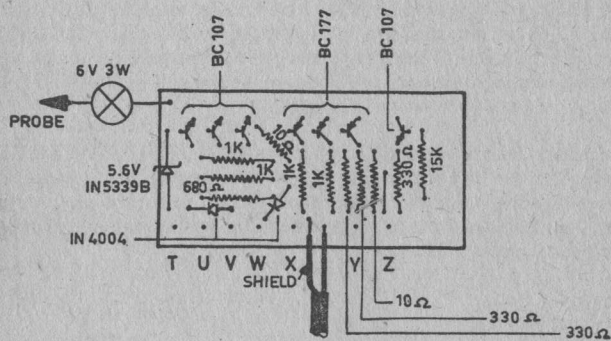


Fig. 2 (above): The board patterns, shown actual size. Below is a close-up of the wired-up board assembly.



LINK WITH OTHER P.C. BOARD

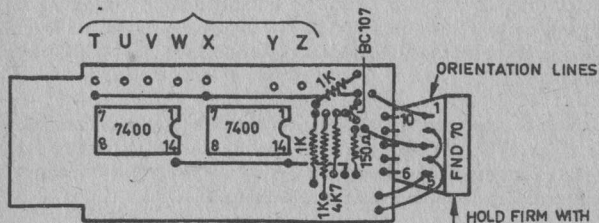


Fig. 3 (above): Wiring diagrams for the two boards, also showing interconnections.

The display segments are driven directly from the TTL outputs, current limiting being effected by the internal level '1' gate resistance. The circuit has been arranged so that the power dissipation is divided equally between ICs A and B.

The brightness difference that should in theory result from operating certain segments in parallel while others are driven singly, is not discernible in operation.

General Information

A few points may not be clear from the foregoing description.

A short to rail value of 50mV/10 ohms was taken rather than a true short circuit. This overcomes any voltage drop on the supply lines and enables input protection to be incorporated. In practice the '0' level will always exceed 50mV and a '1' level – even in open collector applications – will be many times this value below the rail voltage.

The input loading of this device is 1UL for all conditions excepting a short to positive or negative rail. It is assumed any short to rail will pass the extra 5mA required for these detectors.

Supply current for this probe can reach 100mA, and while in most instances this will be available from the unit under test, it can be provided by a separate battery or supply. However if this is done, ensure that the negative of the extra supply is interconnected with the negative of the unit under test.

The 7-segment display legend, while being easily interpreted is also capable of being read upside down. To prevent ambiguity the top and bottom should be identified by a red dot placed adjacent to segment 'a' and a green dot placed adjacent to segment 'd'.

Construction Details

The accompanying photographs give an idea of construction used for the prototype. The components are mounted on two PC boards measuring 28mm wide by 50mm long. A miniature plastic torch case is used as a housing, requiring only the addition of a probe at the reflector end, and a hole for the 7-segment display to be viewed at the opposite end.

Approximately a metre of shielded cable is provided for the power lead, this terminating with E.Z. mini hooks or similar.

Detail of the PC boards and their assembly is given in Figures 2 and 3. Care should be taken to keep all components as close to the PC boards as possible or trouble may occur when fitting the boards into the housing. Metal can transistors will require an insulating sleeve to prevent adjacent cans shorting.

- 1 6 volt 5 watt lamp
- 2 Mini Hooks
- 2 10 ohm ¼ watt resistors
- 1 150 ohm ¼ watt resistor
- 3 330 ohm ¼ watt resistors
- 1 680 ohm ¼ watt resistor
- 7 1K ohm ¼ watt resistor
- 1 4.7k ohm ¼ watt resistor
- 1 15K ohm ¼ watt resistor
- 1 Penlite torch 20 x 35 x 95mm
- 1 Shielded mic. Cable (1 metre long)

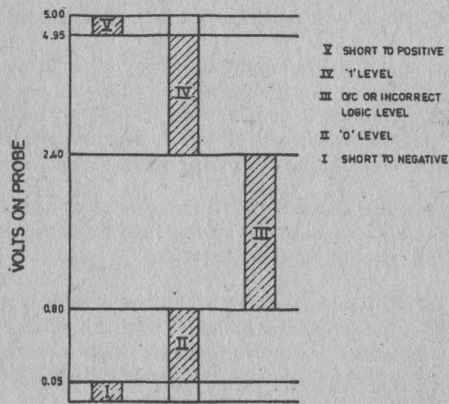


Fig. 4 The voltage levels which define the five states detected by the probe.

REPAIRING A CRACKED BOARD

A cracked printed wiring board may usually be repaired with a little care. If both the base and the copper are cracked, drill small holes through each run of copper, on opposite sides of the crack.

From thin tinned copper wire, make up a staple for each pair of holes. The staples should be just wide enough to bridge the two holes when the crack is closed. Make the sides of the staple fairly long.

Push the two halves of the board together, push one staple through the holes from the copper side, then gently twist the ends of the staple together on the component side, with a pair of pliers. Solder the twist, and clip off the surplus. Solder the tinned copper wire to the copper pattern. Where more than one staple is required, insert alternate staples from opposite sides.

DIGITAL COUNTERS TO 200MHZ

Here is a new design which should be of special interest to lab workers, mobile two-way radio technicians and VHF radio amateurs. It is easily built at low cost, and uses a state-of-the-art ECL integrated circuit to extend the range of your present DFM or counter to beyond 200MHz.

You may remember, I described our first frequency scaler to extend the range of existing digital counters. It used some of the first emitter-coupled logic (ECL) integrated circuits, which had then just been released. By using some seven of these devices I was able to achieve a maximum frequency of slightly more than 120MHz, with a division ratio of either 10:1 or 20:1 to cope with counters with maximum counting rates as low as 6MHz.

The performance of this design was considered quite good at the time, particularly in view of its modest cost, and many of the units were built up. They have been used in many research, development and teaching laboratories as well as by mobile radio technicians and VHF radio amateurs. However, integrated circuit technology is progressing at a very high rate, and it is not surprising that since the original unit was developed, higher performance devices have been produced.

This fact made itself evident some weeks ago when I was looking through literature describing some of the recently released devices. Despite the relatively short time that has elapsed since it was described, the 1969 scaler is now well and truly obsolete. Hence the reason for developing the new design described in this article.

The new scaler is completely different from the old one in terms of circuit configuration. It uses only three ICs, two transistors and five diodes – three of which are in the power supply. In terms of cost and complexity, it should be both cheaper and easier to build than its predecessor.

Yet at the same time its performance is significantly improved. The upper frequency limit is now slightly above 200MHz, thus extending well past the 144-148MHz amateur band and the upper VHF mobile radio band. The frequency division ratio is now 100:1, so that the scaler will extend to 200MHz any counter capable of counting up to 2MHz at present.

Input sensitivity is high, less than 800mV peak-to-peak input being required for reliable division up to 200MHz. Input impedance is low, approximately 50 ohms, but at VHF this is generally no problem. In most cases a small "sniffer loop" or length of wire for capacitive pickup at the end of the input cable will be quite sufficient. The input is protected from overload damage.

Output from the scaler is approximately 4.5V peak-to-peak, with a low output impedance. This should be more than adequate for all but the most insensitive counters.

Basically the scaler consists of two cascaded decade divider stages, a very high speed input stage followed by a lower speed stage. If the unit is to be used with a counter having a basic frequency capability of 20MHz or better, the second decade may be omitted, reducing the cost and simplifying the unit still further.

In this "shortened" form the scaler may easily be built into existing 70MHz frequency meters to convert them at low cost into full 200MHz instruments. This is a simple procedure, and is described later in the article.

Heart of the scaler in both its "long" and "short" versions is a new ultra-high speed ECL decade divider IC, which performs the input divider function. This device is the Fairchild Semiconductor type 95H90, for which Philips also have an equivalent, type GHJ121.

The 95H90 is a medium-scale integration (MSI) device which contains four very high speed J-K flip-flops. These use ECL circuitry and are arranged for either BCD decade counting, 10:1 frequency division, modulo-11 counting, or 11:1 frequency division. It has a guaranteed minimum input counting rate of 220MHz, and typical devices can exceed 300MHz in a suitable circuit configuration.

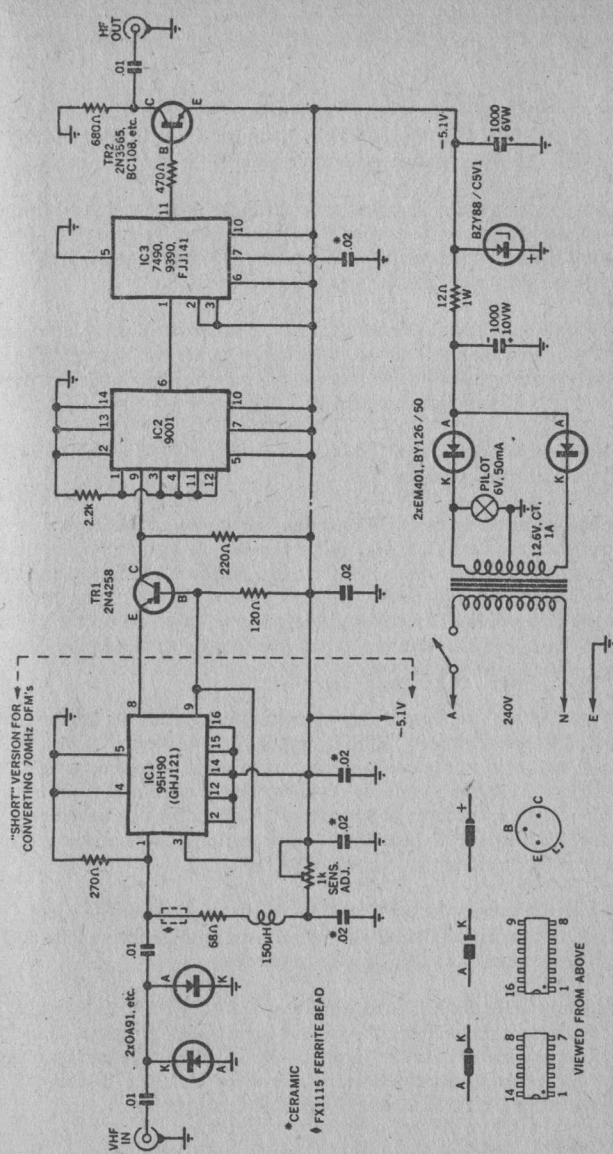
Here the device is used for straight 10:1 frequency division, and the simple circuit configuration tends to limit operation to about 210MHz.

The 95H90 is not a cheap device, costing around \$15 plus tax. However as it needs only the addition of a few minor components to provide a full 200MHz-plus counting or divider decade, it is still significantly cheaper than any other alternative approach.

The input of the scaler connects to the input of the 95H90 via two coupling capacitors in series, with the junction of the capacitors connected to ground via inverse-parallel diodes. These protect the 95H90 against input overload damage.

A DC bias voltage is also fed to the input of the device to set the conditions for maximum input sensitivity. The bias circuit is decoupled by a ferrite bead and a small RF choke. Without these components, spurious coupling back through the supply line tends to degrade the 95H90 performance and reduce the maximum frequency of operation.

The second decade of the scaler uses TTL (transistor-transistor logic) devices: a 9001 and a 7490 or 9390 or FJJ141. The 7490 / 9390 / FJJ141 is actually a full MSI decade divider itself, but with a guaranteed maximum counting rate of 10MHz. Used alone for the second decade, it could in the worst case limit the upper frequency of the scaler to 100MHz. For this reason the input flip-flop of the device is left unused, and a type 9001 flip-flop IC used instead. With a typical maximum counting rate of 50MHz, the 9001 ensures that the upper frequency limit of the scaler is set solely by the 95H90 input device.



The circuit of the scaler, which uses only three ICs and two transformers.

Because the second decade uses TTL devices whose logic levels are different from those of ECL, level translation is required between the two stages despite the operation of both from a common positive-ground 5V supply. The translator used is a simple inverter using a 2N4258 very high speed PNP switching transistor.

To prevent disturbance of the second divider decade due to output loading, a buffer stage is used between the 7490 and the output connector. The buffer uses a 2N3565, BC108 or similar general-purpose NPN silicon transistor.

Power for the scaler circuits is provided by a simple supply using a full-wave rectifier connected to a centre-tapped 12.6V transformer winding. A 400mW zener diode is used to stabilise the supply to the ICs to a nominal 5.1V. The current drain of the scaler circuits is approximately 210mA.

With the exception of the power transformer, mains switch, pilot lamp and input / output connectors, all of the scaler components and wiring are mounted on a small printed wiring for ease of assembly. The board measures $4\frac{1}{2} \times 2\frac{3}{4}$ in (114 x 70mm) and is coded 72 / s10.

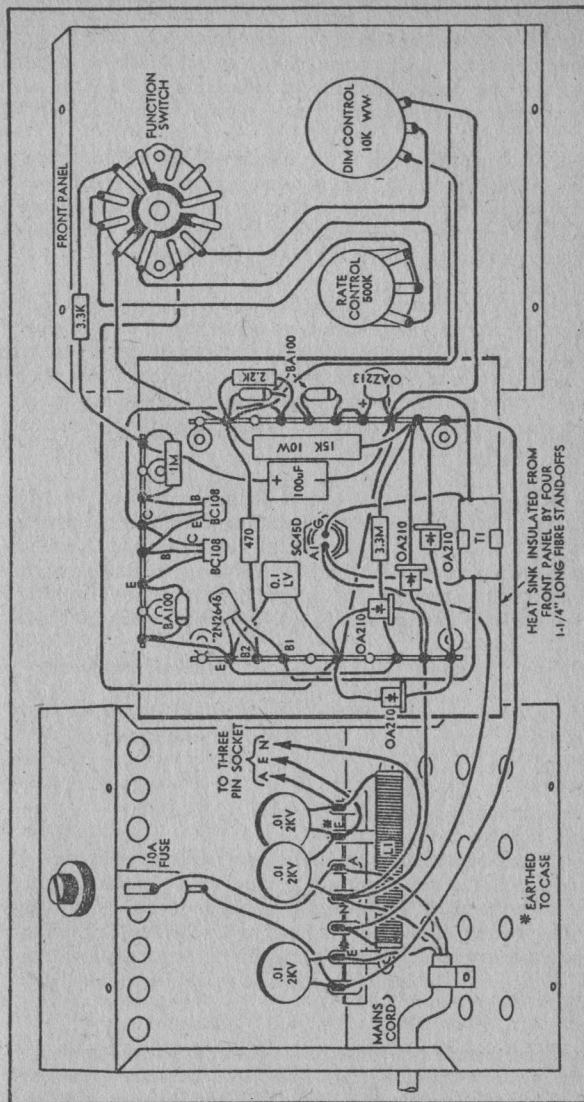
The prototype of the "long" version of the unit was built in a small plastic-and-aluminium instrument case,

Inside the case the printed board is attached to the rear of the front aluminium panel, via four $1\frac{1}{4} \times \frac{3}{8}$ in Whitworth screws with multiple nuts used for spacing. The lower two screws are also each used to fasten one side of the input and output connectors, to reduce the number of screw heads visible on the front panel. The power transformer is mounted on the bottom of the plastic case, as far to the rear as possible to provide maximum clearance between it and the wiring board.

The prototype "short" version of the scaler has been built into the original 70MHz digital frequency meter. Here it simply consists of the "front half" of the board, mounted in the DFM in place of the small board originally used to perform 2:1 input frequency division. The scaler board is simply sawn in half along a line drawn between two "notches" which have been provided for this purpose in the etched copper pattern. The notches are approximately midway along the long edges of the board.

The scaler board has been designed with its width the same as the small divider board in the original DFM, and with its mounting hole centres also the same. It is therefore a simple drop-in replacement.

Because the input gate and first counting decade of the DFM already use ECL devices, there is no need for the scaler's logic translator stage. Thus in this short form the scaler consists simply of the 95H90 device and its power supply and input biasing components. It is powered from the -5V supply already provided in the DFM for the other ECL devices.



The lower input frequency limit for reliable operation of the 95H90 device with sinewave inputs is approximately 1MHz, so that it is really not suitable for use with all input signals. However, on the other hand, the existing input shaping circuitry of the DFM is not capable of working at VHF, so that it also cannot be used with all input signals. A switch must therefore be used to connect either into circuit alternatively between the DFM input connector and the gating circuitry. The switch may logically be marked "HF-VHF".

Note that the correct position of this switch will depend purely on the frequency of the input signal, and not on the range selected. Even when one switches down to lower ranges to look at the less significant digits of a VHF signal, it is still necessary to have the switch in the "VHF" position, as the input circuitry must still handle the fill signal frequency.

On the prototype instrument, rather than add a further switch to the front panel, I merely changed the connections to the switch used originally to select either AC or DC input coupling, so that this switch has become the "HF-VHF" switch. The DC input facility had never been used since the instrument had been built, so that it is no great loss. The original coupling capacitor is simply wired permanently in series with the HF input circuitry, as shown in the small diagram.

Because the scaler involves a 10:1 input division instead of the 2:1 division used originally, the appropriate timebase frequencies for correct readings will now be 100kHz and its sub-multiples, not 50kHz and its sub-multiples. Happily these frequencies are already available on the timebase board, so that it is simply a matter of changing over the connections from the range switch wafer S1c to the timebase board. The new frequencies used are as follows:

Range (HF)	Timebase Frequency
0-1999Hz	1Hz
0-19.99kHz	10Hz
0-199.9kHz	100Hz
0-1.999MHz	1kHz
0-19.99MHz	10kHz
0-35MHz or more	100kHz

Strictly it would be necessary to divide these frequencies by ten when the input switch is in the VHF position, to compensate for the 10:1 division introduced by the scaler. However, this would involve either a further wafer on the range switch, or another 10:1 frequency divider IC. Even then it would be necessary to add further provision to reduce the lower sampling rate, as for the lowest rating the gating time would be no less than 10 seconds.

This additional complexity seems to me unwarranted, when one can easily perform a mental x10 multiplication whenever the scaler is switched into circuit. All one has to remember is that readings obtained with the input switch in the VHF position are one-tenth of their correct value; thus the lowest range really reads in units of 10Hz instead of 1Hz, for example, while the highest range reads directly in HMz and not in units of 100kHz.

Note that the original 2:1 input divider board also included a flip-flop used to divide the lower sampling rate to 0.125Hz. This rate was necessary to allow correct operation with the 0.5Hz timebase frequency used for the lowest frequency range. As the lowest timebase frequency will now be 1Hz, this low sampling rate is no longer required. The 0.25Hz output from the input board may now be taken directly to the sampling switch, giving a lower sampling rate of 1 every four seconds.

After the scaler is completed, there is only one adjustment required to ensure that it is operating correctly. This is exactly the same for either the "long" or "short" version: adjustment of bias for maximum input sensitivity.

The best way to make the adjustment is to feed a VHF signal into the scaler input, with its output connected either to a digital counter or to an oscilloscope. Then the bias preset pot on the scaler board is simply adjusted for reliable operation with the smallest possible input voltage. As viewed on a counter, this will be evident as the bias setting at which one can reduce the amplitude of the VHF input furthest before the reading starts to fall from its correct value and being to vary randomly. On an oscilloscope, it will be the setting where the input can be reduced furthest before the output wave form begins to exhibit "drop-out". Its as simple as that!

PARTS LIST

- 1 Instrument case, as required.
- 1 Printed wiring board, 4½ x 2¾in, 72 / s10.
- 1 Stepdown transformer, secondary 12.6V CT at 1A.
- 2 Coaxial connectors, panel mounting.
- 1 Miniature SPST toggle switch, 240V rating.
- 1 Ferrite bead, type FX1115 or similar.
- 1 6V 50mA pilot lamp bezel.
- 1 150uH peaking inductor.

SEMICONDUCTORS

- 1 95H90 or GHJ121 decade divider IC.
- 1 9001 high speed flip flop IC.
- 1 7490 or 9390 or FJJ141, RS276-1808 decade divider IC.
- 1 2N4258 very high speed PNP silicon transistor.
- 1 2N3565, BC108, RS276-2009 or similar.
- 2 EM401, RS276-1139, BY126 / 50 silicon diodes.
- 2 OA91, RS276-1102/1136 or similar germanium diodes.
- 1 BZY88 / C5V1 or similar 5.1V 400mW zener diode.

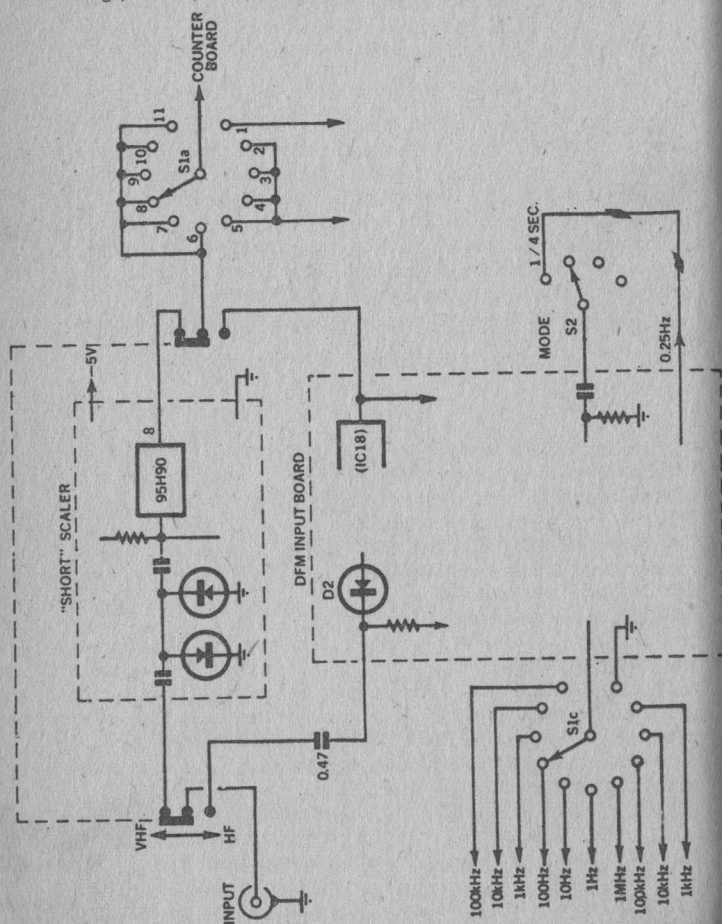
RESISTORS

- 5% half watt: 1 x 68 ohm, 1 x 120 ohm, 1 x 220 ohm, 1 x 270 ohm,
1 x 470 ohm, 1 x 2.2k.
- 1 12 ohm 1 watt.
- 1 1k miniature tab pot.

CAPACITORS
 3.01 μ F disc ceramic
 5.022 μ F disc ceramic.
 2 1000 μ F 10 VW electrolytic

MISCELLANEOUS

Mains cord and plug, clamp for cord, connecting wire, solder lugs, screws, nuts, etc.



How to convert the DFM to count up to 200MHz. Few parts are needed.

BUILD OUR DIGI-METER: A DVM WITHOUT TEARS

Here's a basic DVM which almost anyone should be able to put together with complete success. Despite this its performance is very impressive.

Not very long ago, before integrated circuits came along, digital volt-meters or "DVMs" were large and costly instruments. Because of their cost and complexity they tended to be found only in the rarified atmosphere of the research laboratory. The idea that a DVM would ever be used for everyday measuring jobs seemed a futuristic pipe-dream; as for building a DVM into a piece of equipment as a direct replacement for a moving coil meter movement, that was virtually unthinkable!

Yet, thanks to modern semiconductor technology, and the increasingly complex ICs which it has produced at ever-lower costs, both these dreams have now become a reality. Low cost multi-range DVMs are now available for little more than the cost of a good analog multimeter, and are already being used even in servicing and hobbyist situations. Not only this but more and more original equipment manufacturers ("OEMs") are fitting single-range DVM "panel meters" into equipment in places where once they would have used moving coil movements.

The latest generation of these digital panel meters are truly impressive little devices, a tribute to modern semiconductor devices and advanced circuit design. A particularly outstanding example is the Model AD2010/E recently released by Analog Devices, Inc., of Norwood, Massachusetts.

The size of the AD2010/E is a compact 45 x 76 x 20mm, less than many moving coil movements. It weighs only 113 grams (4 ounces). Yet crammed into this compact package is a complete 3½-digit DVM with a full-scale sensitivity of 199.9mV and an accuracy of 0.05pc plus/minus 1 digit in the least significant position. It also offers such features as bright seven-segment LED readouts with 7mm-high digits, display storage, automatic polarity selection, automatic zero correction, internal overload protection, blanking of leading zeros, BCD outputs for data logging, a DC input resistance of 100 megohms, a normal-mode 50Hz hum rejection of 40dB, and a common-mode rejection of 60dB.

In short, it is a true state-of-the-art single range DVM, compressed into a little black box half the size of many conventional meter movements.

As soon as we saw the AD2010/E a few weeks ago, we wondered whether it might be possible to use it as the heart of a basic and easy to build multi-range DVM for the home constructor. When we

approached Parameters with the idea, they were confident that it could be done. And so, with their help and co-operation, the Digi-Meter was born.

As you can see from the photographs and the circuit, it is essentially little more than the AD2010/E fitted into a case, with a power supply and a simple input attenuator circuit to provide four higher voltage ranges. All the real credit must go to Analog Devices! To make the instrument just that more attractive, however, we have added a simple constant-current source circuit, to provide four additional resistance ranges.

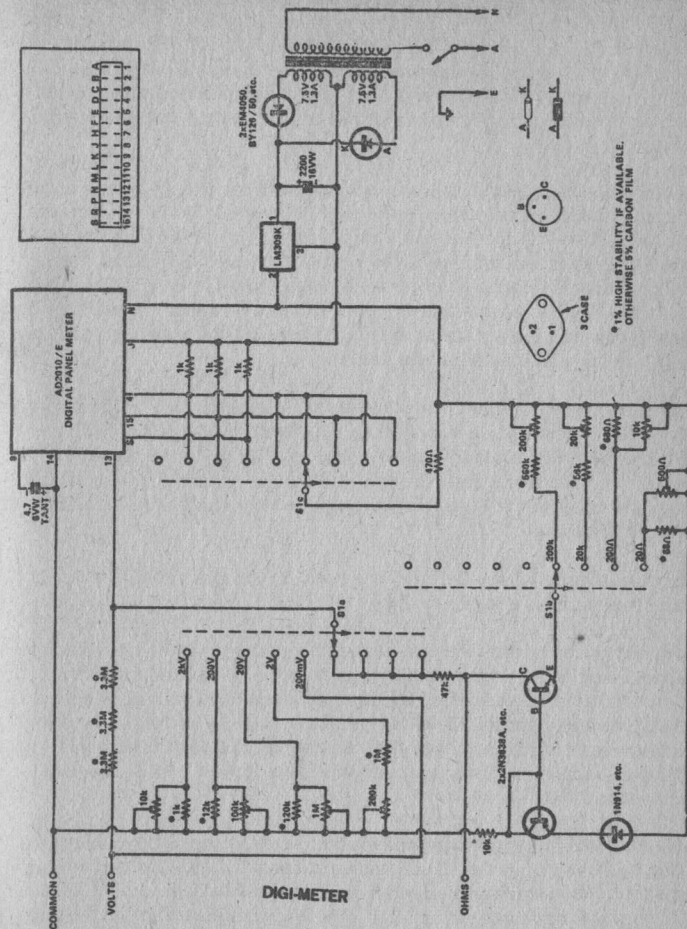
The circuit itself must surely be about as simple and as straightforward as one could get with a practical multi-range meter.

As mentioned earlier, the basic DC input resistance of the AD2010/E is 100 megohms, and its sensitivity 199.9mV. To provide the higher voltage ranges we have used an input voltage divider system similar to that in the more complex DVM. Three series-connected 3.3M resistors are switched into circuit in series with the AD2010/E input, with different shunt resistors switched directly across the input for each range.

It would not be sufficient simply to use standard close tolerance fixed resistors for the range divider shunt resistors, because the closest tolerance normally available is 1pc; this would quite seriously degrade the basic 0.05pc accuracy of the AD2010/E. Hence it is necessary to use preset adjustable pots for each range, to allow more accurate calibration. It is still desirable to use high stability resistors, if they are available, but more in the interests of stability than for close tolerance.

The advantage of the particular input attenuator system we have used is that each of the high voltage ranges may be calibrated independently; there is no adjustment interaction (apart from the internal calibration of the AD2010/E, which may be regarded as a reference). The fixed 9.9M resistance in series with the input on all the higher voltage ranges also gives additional overload protection. Note that the input resistance on the higher voltage ranges varies between about 10 and 11 megohms, which is quite suitable for most measurements.

The constant current source used for the resistance range is a simplified version of that used in the more elaborate DVM, and is also designed to operate from the same 5V DC supply used for the AD2010/E. It uses a single forward-biased 1N914 or similar silicon diode as the basic voltage reference, with a second diode-connected transistor to balance the Vbe drop of the actual current source device. Both transistors are high-gain PNP silicon devices such as the 2N3638A. Preset resistors are used again in the emitter feedback circuit, to allow accurate calibration of the four ranges. The currents for the four ranges are 1μA, 10μA, 1mA and 10mA.



The circuit is simplicity itself, thanks to the Analog Devices digital panel meter. Only a handful of other components are required for the power supply and regulator, input attenuator and current source circuit.

Provision of the decimal point selection circuit proved to be a little tricky, because the AD2010/E is designed so that all decimal point LEDs are normally "on". For its normal application as a single-range panel meter, this is no problem as the unwanted points are simply blanked by shorting the appropriate pins to the negative rail. But in the Digi-Meter we had basically only one single-pole 9-position switch section available for decimal point selection — hardly the most appropriate for the job.

After a little head-scratching we came up with the system shown on the circuit, and it seems to work quite well. Essentially we have taken each of the AD2010/E decimal point pins permanently to the negative rail via 1k resistors, and then used the switch to re-apply 5V DC to the pin of the decimal point needed for each range, via a 470-ohm protective resistor. The 1k resistors appear to be quite low enough to reliably blank the undesired LEDs, while at the same time being large enough to allow any one to be unblanked when desired.

Our prototype Digi-Meter was built up in a nominal 7 x 5 x 4 inch vinyl-covered metal instrument case. This produces a neat and professional little unit, yet involves a minimum of work. We mounted the AD2010/E directly on the front panel (it clips into a plain rectangular hole 80 x 45 mm), along with the range switch, the input terminals and a small mains on-off switch.

Most of the rest of the wiring is mounted on the rear panel, which like the front panel is also removable.

The mains cord enters the case at the lower far end of the rear panel, its entry being made via the usual grommetted hole. It is clamped in the approved manner with a P-clamp, and the active and neutral wires terminated in a section of "B-B" connector strip. The earth wire is soldered to a lug screwed firmly to the metal panel, so that even when the panel and transformer are removed from the rest of the case, both remain earthed.

The only component not actually mounted on either the front or rear panels of the case is the IC voltage regulator, which is mounted on a simple heatsink bracket bent up from a scrap of 16G aluminium sheet. The bracket measures 65 x 60 x 10mm, and is mounted in the bottom of the case by two of the screws used to attach the rubber feet.

If the usual four mounting holes for a TO-3 device are drilled in the heatsink, this will suit either the LM309K device or the TO-3 to measure a voltage in a low resistance circuit, (say less than about 10k), the effect will be very small, and can generally be neglected.

For measurements in higher resistance circuits, the only way of correcting for the error produced is to first make the measurement, then

simulate the resistance of the circuit by connecting a resistor of similar value across the Digi-Meter input terminals. The residual reading produced is then subtracted from the original reading, algebraically as before, to obtain the corrected reading.

When completed and power is to be applied for the first time, it would be a wise precaution to temporarily detach the 30-pin connector from the AD2010/E until you make sure that the power supply is correctly delivering the nominal 5V. This will prevent any possible damage should a wiring error have been made.

When you have satisfied yourself that the voltage socket between pins J and N on the socket is between the acceptable limits of 4.8V and 5.2V, turn off the power temporarily and refit the 30-pin connector to the AD2010/E. Reapplying the power should then result in normal operation, with digits visible on the readout display.

The only remaining operation is that of calibration. This is a little easier than with the more complex DVM, at least in one respect: the 200mV range is already calibrated. If the full accuracy of the AD2010/E is to be retained for the other ranges, however, it will be necessary to perform a full-scale calibration using a reference instrument and a stable source of adjustable DC.

Probably few readers will be in a position to do this, so that in many cases it will be necessary to do the next best thing: calibrate the 2V and 20V ranges using a source of reasonably known voltage, and then use a few suitable voltages present in an amplifier or other piece of equipment to bring the 200V and 2kV ranges into as close agreement as possible with the lower ranges.

The most convenient source of a reasonably known voltage for most readers will probably be a new mercury cell of the type used in cameras having inbuilt cadmium sulphide exposure meters.

When new, these cells have an open-circuit terminal voltage of 1.3566V, which drops to around 1.3524V after 12 months at an average temperature of about 21 degrees C. It is therefore reasonable to assume that virtually any such cell will have an unloaded voltage of very close to 1.35V, providing it is in reasonably good condition.

Even better, a cell bought as new from a large city store, where it is likely to be fairly fresh, could be reasonably assumed to have an unloaded terminal voltage of 1.355V. Such a cell could therefore be used to calibrate the 2V and 20V ranges to better than 1 per cent.

Calibration of the resistance ranges is best done using a precision decade resistance box, if access to such a device is possible. Some readers may be fortunate enough to gain access to such a decade box at their local technical college. For those unable to do this, probably the only practical alternative is to obtain a few representative close-tolerance resistors, and use these as references.

A final note: when using the Digi-Meter, you should notice that there will be a very small but non-zero residual reading on the 2V range for open-circuit input. Similarly there will be a more significant residual reading on the 200mV range under the same conditions. Both are normal, and are due to the small bias current drawn by the input circuit of the AD2010/E.

On the 2V range the residual reading should be so small that for most measurements it may be ignored. However, for readings where the maximum possible accuracy is desired, it is only necessary to subtract the residual reading algebraically from the reading produced for the voltage being measured, to obtain the best estimate of the true voltage.

With the 200mV range the situation is not quite so simple, because the effect of the bias current will depend upon the resistance of the circuit in which the measurement is being made. If the Digi-Meter is being used to measure a voltage in a low resistance circuit, (say less than about 10k), the effect will be very small, and can generally be neglected.

For measurements in high resistance circuits, the only way of correcting for the error produced is to first make the measurement, then simulate the resistance of the circuit by connecting a resistor of similar value across the Digi-Meter input terminals. The residual reading produced is then subtracted from the original reading, algebraically as before, to obtain the corrected reading.

The Parts Needed

- 1 Case
- 1 Digital panel meter, Analog Devices type AD2010/E
- 1 Power transformer, 15V CT at 1.3A, low profile
- 1 Miniature rotary switch, 2-sections each 2-poles 9-positions.
- 1 Miniature toggle switch, SPST 240V rating
- 1 instrument knob, large
- 3 Screw terminals, 2 red and 1 black.

SEMICONDUCTORS

- 2 BY126/50, EM4050, RS276-1101/1105 or similar 50V silicon diodes
- 1 1N914 or RS276-1139 silicon diode
- 2 2N3638A, MPS3638, TIS50, 2N5447, SK3025, RS276-2021, or similar silicon PNP transistors
- 1 LM309K. μ A7805 or similar 5V-1A regulator

RESISTORS

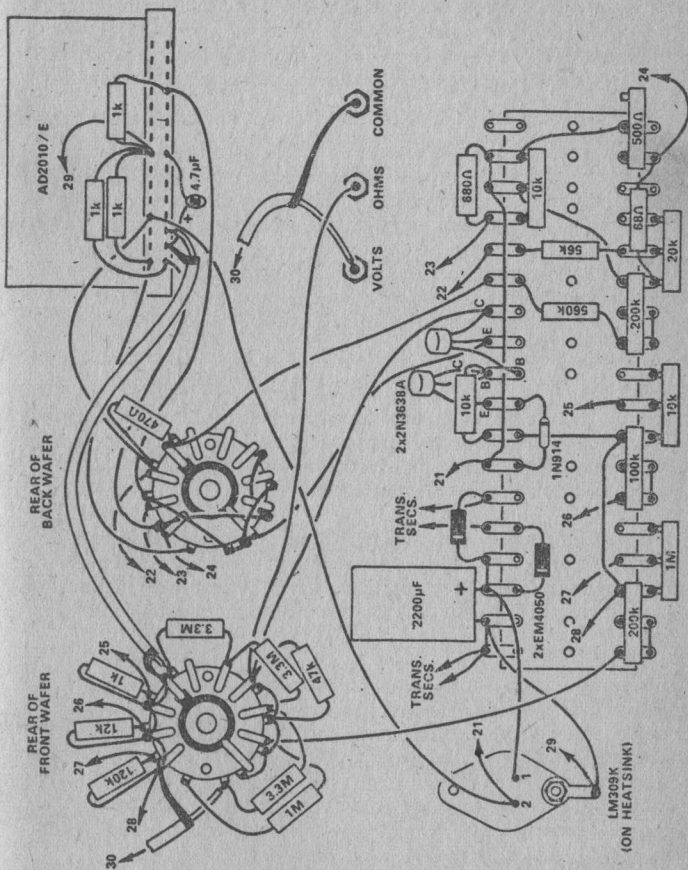
- Half watt, 5pc: 470 ohms, 3 x 1k, 10k, 47k
- High stability for preference: 680ohms, 680ohms, 1k, 12k, 56k, 120k, 560k, 1M, 3 x 3.3M
- Miniature preset pots: 500-ohms, 2 x 10k, 20k, 100k, 2 x 200k, 1M (all linear).

CAPACITORS

- 1 4.7 μ F 6VW tantalum
- 1 2,200 μ F 16VW electrolytic.

MISCELLANEOUS

Mains cord and plug, handle and rubber feet for case, 30-pin connector for AD2010/E, 18-lug pairs section of miniature resistor strip, 3-bay section of "B-B" connector strip. P-clamp for mains cord, scrap of 16G aluminium for regulator heatsink, connecting wire, solder, etc.



The wiring diagram of the Digi-Meter,

How often have you wanted to select a transistor with especially high breakdown voltage, for a special circuit? Or picked up a zener diode, only to find its identification has rubbed off? If the answer is often, you need this easily built little tester.

Here is a handy, low cost device which should be useful in the laboratory, to servicemen, amateurs and enthusiasts alike. It is designed to make breakdown voltage tests of a non-destructive nature, on bipolar transistors, FETs, diodes (including zener diodes), and other semiconductor devices.

With this tester, such measurements as BV_{ces} , BV_{cer} and BV_{ceo} may be made directly. If the type of transistor is unknown, PNP or NPN, this may also be determined directly. Unknown diodes may also be checked as to whether they are silicon or germanium, by virtue of the forward voltage drop, together with the reverse breakdown voltage for each device. The zener voltage may be also readily determined for zener diodes. By a little exercise of the imagination, other tests may also be made on specific components. More will be said about testing procedure later on.

Basically, the tester consists of a high voltage supply which is directed across the item to be tested, in series with a limiting resistor. Facilities are included to introduce the item to be tested into the circuit, via suitable switching. Finally, a high resistance voltmeter is provided, which is connected across the device being tested.

Contrary to what was once thought, it is possible to break down the junctions of solid state devices without causing damage, providing the actual power dissipation within the device is kept to a low value. A typical example in practice is the now common zener diode voltage regulator.

In designing a tester of this sort, a number of arbitrary and other decisions have to be made in order to arrive at a practical unit. The extent or value of the high tension voltage to be used must be considered, in the light of the actual maximum voltage breakdown of the more common devices to be tested, against the economics of generating a useful voltage. In this unit, we settled for just a little over 400V. This is easy and economical to generate, and most devices which one could reasonably consider for testing would come within this figure.

While it is essential to have a voltmeter of very high resistance to measure the breakdown voltage of the device, the accuracy of the volt-

age reading is not all that important, within reasonable limits. This means that the design of the voltmeter part of the circuit may contain some short cuts in order to keep costs down. More will be said about this and similar items as we go along. Meanwhile, let us have a look at the circuit of the complete instrument.

The power supply, including the 400V high tension and the +12V and -12V supplies for the IC are all obtained from one source. A small transformer with a secondary AC voltage of 150, rated at 30mA are fed into a voltage doubler, giving a total output of about 420V DC. A little over 400V of this is applied via 440k to the device to be tested. Under pre-test conditions, this resistor is shorted to ground, thus dissipating within itself just a little under 400mW.

The maximum power which can be dissipated within a device being tested, is when its breakdown voltage is half the supply voltage, and amounts to a bit less than 100mW.

In order to protect the operator against an electric shock, the 440k resistor is normally shorted to ground via a push-button switch. A socket is provided for small transistors and the three connections are duplicated on three banana sockets for large transistors and other devices. A toggle switch reverses the supply voltage to accommodate NPN and PNP devices, while a second three-position toggle switch permits switching for three different test conditions. These are - collector to emitter with base shorted, with 10k in the base and with the base open circuit.

The voltmeter is designed around the ubiquitous 741 IC. This device is itself quite cheap, and it also allows the use of a relatively cheap 0-1mA meter movement.

In order to keep the input resistance to the device high and to restrict the voltage applied to the non-inverting input, a voltage divider consisting of 22M and 22k is used. By varying the negative feedback from the output to the inverting input, the gain, and so the effective voltage read by the meter, is controlled. The meter has 900 ohms in series with its internal 100 ohms, resulting in a 0-1 volt movement. Combined with the negative feedback just mentioned, this gives the requisite voltage ranges of 10V, 100V, and 400V.

Although the amount of current delivered by the 741 IC is restricted, we considered it desirable to add some extra protection to the meter. An ordinary silicon diode, such as an EM401 or BY127 or RS276-1139 has been connected across the movement and this should prevent any overload from exceeding about six times, giving adequate protection. A 10k trimpot is connected between pins 1 and 5, to adjust the zero meter reading.

To provide the positive and negative supply voltages for the 741 IC, we have connected two 12V zener diodes in series. The junction of the diodes becomes the reference point for all measurements involving the IC. The two zener diodes are fed from the 400V supply via 60k.

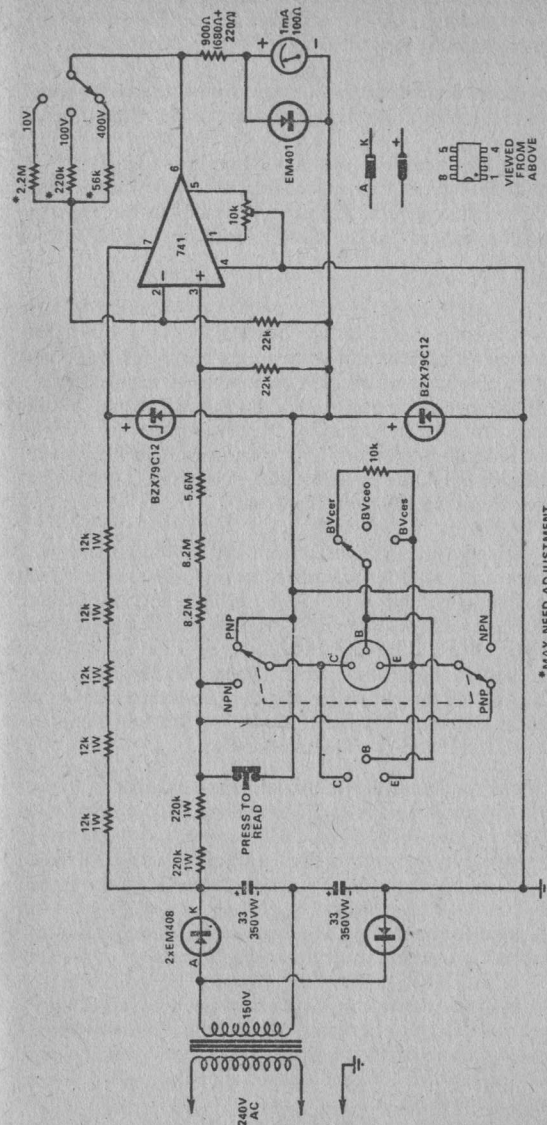
So much for the basic circuit details. A number of components need clarification and so some comments may be helpful. The power transformer which we used is a standard stock item. This transformer has a multitapped secondary but we have ignored the taps and used the full available 150V AC. There is also a 6.3V secondary winding, which we have not used. More will be said about these points under constructional details. If you wish, and another suitable type of transformer is available to you, there is no reason why it should not be used.

The 440k series resistor is made up of two 220k 1W resistors. Although one 470k 1W resistor may do the job satisfactorily, it is not desirable to put too high a voltage across any resistor, notwithstanding its power dissipation capabilities. A somewhat similar situation exists with the 60k resistance in series with the two zener diodes. Not only do we have a high voltage to cope with but the power dissipation of nearly 3 watts must also be considered. Although it is all right to apply high voltages across wire wound resistors, providing the power dissipation is adequate, such resistors are expensive and we settled for five 12k 1W units as an economy measure.

The 22k resistor at the input to the IC should be of low tolerance and perhaps the best way to achieve this end is to make up the value with two 8.2M units and a 5.6M unit, all of low tolerance, in series. Both 22k resistors should also be of low tolerance.

The 900 ohm resistance in series with the meter is made up with a 680 ohm and a 220 ohm resistor in series. These should also be low tolerance types. The 56k resistor should, according to calculations, be 55k. The three resistors involved in the voltage range selection should ideally be tailored to size during calibration of the system against a meter of known accuracy. In point of fact, we fitted a 2.7M resistor in place of the 2.2M and shunted it down until we got the required accuracy. Similarly, we used a 270k resistor, shunted down, instead of the 220k. Also, the 56k must be shunted to give a full scale reading of 400V.

Before leaving the subject of resistors, we have used the term "low tolerance". In our unit, we used 5 percent tolerance resistors where this was required and then did the final calibration with the three resistors in the feedback network, as described in the last paragraph. If you have no means of calibration, then it would be wise to use 2, or even 1 percent tolerance resistors where required and hope that this will lead to a satisfactory level of voltmeter accuracy. Although accuracy is not vital, as suggested earlier, we feel that it should be around the 5 percent mark.



BREAKDOWN VOLTAGE TESTER

The complete circuit of the breakdown voltage tester. The required range is selected by varying the feedback network.

Although the push-button switch only calls for a single pole, we used a double pole unit as this was available at the time. Also, the single-pole three-position rotary switch, was not available at the time we needed it and we substituted a double pole unit.

The 741 IC is available in a variety of packages but we chose the 8-pin DIL version as this is compact and is easy to fit into the wiring board.

We built the prototype Breakdown Voltage Tester into a plastic moulded case with aluminium front panel, and the result is a neat and compact unit. The actual physical arrangement of the tester is not critical and it may be built into some other type of box if you choose to do so.

Construction is a relatively simple matter but like any other project, it is wise to tackle it in some logical order. The majority of the components are mounted on a length of miniature wiring board and this could be done first. One operation which calls for more than usual care, is the mounting of the socket for the 741 IC. Eight small clearance holes must be accurately drilled in the board so that the socket fits neatly into these holes. The holes should be just large enough to take the pins of the socket. With this done, the pins should then be carefully wired to their terminations as indicated on the diagram.

The rest of the wiring is quite straightforward and should present no difficulties. However, as mentioned earlier, the resistors associated with the calibration of the meter should be given consideration at this stage, according to your approach to this part of the circuit. If you intend to follow our method, then a 2.7M resistors should be fitted instead of the 2.2M and a 270k resistor should replace the 220k. A 56k resistor will be fitted anyway, and all three resistors will then be shunted down after the unit is complete and ready for calibration. For the time being, omit the lead from the 220k resistor to the 8.2M resistor.

Before leaving the wiring board assembly, perhaps we should point out that the two resistors connected in series to make up 900 ohms, share one pair of tags at one end of the board. This means that the resistors are stood off the board at one end, to form a triangle, with a soldered joint at the top. Also, the centre hole in the board at this point is used for screwing the board to the case later on. The equivalent hole at the other end is also used for mounting, which means that the 33 μ F electrolytic must be moved to one side while mounting is done.

The transformer is then mounted on the back of the box and at an angle. The angle mounting was necessary to fit the transformer into the available space. Before fixing the board which has already been completed, leads for external terminations should be provided. These involve switching, meter leads, etc.

Mounting of the components on the metal panel should present no problems. However, if you have a transistor socket the same as ours,

the question arises how best to fix it. We drilled a hole so that the socket was just a neat fit, set it so that about half of it protruded through the panel and then put a couple of blobs of Araldite between the socket and the back of the panel.

When wiring the components on the panel, some care should be taken with the two toggle switches as these can be a little tricky. The three banana sockets are connected in parallel with the transistor socket. The protective diode is wired directly across the meter terminals. We also ran a lead from the mains earth screw on the terminal strip, to the earth point on the wiring board and continuing onto a solder lug under one of the meter fixing screws.

Having completed your Breakdown Voltage Tester, before proceeding to adjust and calibrate the unit, it is always wise to make a thorough check of the wiring, polarity of components, etc. Satisfied that all is well, set the rotor of the 10k trimpot to its mid-position. Switch on and check for obvious signs of distress.

Measure the voltage from the voltage doubler output, between it and earth. This should be about 420V. Check the voltage between earth and the other end of the two 220k resistors. With the push-button switch operating correctly, this voltage should be about 12V, which is the voltage across one of the zener diodes. A check across both zener diodes in series should give 24V.

At this point, there may be some error in the meter zero reading. This error will be a maximum with the voltage selector set to "10V". Adjust the 10k trimpot for zero reading on the meter, with the selector set to the 10V range.

To calibrate the meter, it will be necessary to have a source of variable voltages around 10V, 100V and 400V. A good multimeter is also needed to calibrate against. It does not matter in which order the voltage ranges are calibrated. Connect the negative lead of the multimeter and the negative lead of the voltage source to the junction of the two zener diodes. With the 8.2M voltmeter input resistor still isolated from the rest of the circuit, this is connected to the positive pole of the voltage source and the positive lead of the multimeter.

Set the appropriate range on the voltage selector switch, adjust the input voltage from the source, for the correct reading on the multimeter. Now proceed to shunt the appropriate resistor in the feedback network until the meter on the tester reads correctly. Solder this new resistor permanently in place. Repeat this procedure for the other two ranges and calibration is complete. Having removed the multimeter and voltage source leads from the unit, connect the 8.2M resistor to the push-button switch side of the 220k resistor. This done and with the voltage switch set to 400V, pressing the push-button should give a meter reading just exceeding full scale.

Although the use of the polarity Tester is more or less self-explanatory, some comments and suggestions may be helpful. The simplest devices to test are diodes in their various forms. Bipolar transistors, whether small or large, NPN or PNP, silicon or germanium, can also be tested simply and easily. Junction FETs may also be tested for BV_{gso} .

Before making a test, always start with the voltmeter set to 400V and then if the subsequent reading is low, the range may be changed to 100V or 10V as required. After taking a reading, always reset the voltmeter to 400V.

To test any diode for its reverse breakdown voltage, or to find the operating voltage of a zener diode, connect the cathode of the diode to the collector point on the transistor socket, or the equivalent banana socket point via an alligator clip and banana plug. The anode of the diode is connected to the emitter point. Set the switch to "NPN" and press the button for the reverse breakdown voltage reading. Setting the switch to "PNP" will give the forward voltage drop, which will be about 0.6V for a silicon device and about 0.2V for a germanium device.

To test small bipolar transistors in TO18, TO5 packages, etc., they are plugged directly into the small socket provided. However, in some cases, the connections are such that they may not be plugged in directly, without bending the leads so that the right connections are made between the socket and the device. The popular BF115 is a case in point. The shield lead will be ignored, but the base and emitter leads must be reversed. Whenever there is a doubt, the appropriate manufacturers reference should be consulted.

With the device connected, the polarity switch should be set to NPN or PNP as required. Pressing the button will give the breakdown voltage readings for BV_{ces} , BV_{ceo} and BV_{cer} , according to the position selected on the switch. In most cases, there will be little difference in readings between BV_{cer} and BV_{ceo} . BV_{ces} will generally be somewhat higher.

If the bipolar transistor is of the larger variety and cannot be plugged into the small socket, then it may be literally hung on the three alligator clips, clipped to the appropriate connections on the transistor. Testing is then carried out as previously described.

If you wish to test any bipolar transistor for BV_{cbo} , then this may be done by using the alligator clips, or by plugging the appropriate leads into the socket. In this case, the collector will be connected as before but the base will be connected to the emitter point and the emitter will be left open. The reading will then be taken by pressing the button and with the switch set for polarity. The three-position switch has no effect on this reading. It may be of interest to mention that this test gives virtually the same reading as BV_{ces} .

There are many occasions when the type and particularly the polarity of a transistor are unknown. If the transistor is a good one, with the polarity switch the wrong way only a very low reading will be observed. With the polarity switch correctly set, the normal voltage reading will be obtained. If the transistor is short circuited, no reading will be obtained in either polarity. If the transistor is open circuited, the full voltage reading will be obtained in both directions. By thoughtful use of the polarity and the three-position condition switches, the state of a transistor can be determined with reasonable certainty.

To test a junction FET for BV_{gso} , connect the gate to the collector point and the source to the emitter point, with the source left open. Set the polarity switch to PNP for N-channel devices and to NPN for P-channel devices. Pressing the button gives the breakdown voltage reading.

The foregoing is just a brief run down on testing solid state devices for breakdown voltage effects. For a more in-depth treatment, readers may wish to refer to Transistor and Diode Testing in our Bernard's Book No.206 "Practical Transistor Circuits for Modern Test Equipment" on pages 80 to 96.

Parts List

- 1 Case 184 x 115 x 118mm, with aluminium front panel.
- 1 Power transformer, 240V primary, 150V, 30mA secondary.
- 1 Meter, 1mA FSD, 100 ohms, 75 x 65mm
- 1 IC, 741 8-pin DIL. or RS276-007/010/1741
- 1 IC socket, 8-pin DIL.
- 1 Miniature tag board, 25prs tags
- 1 3-way terminal strip
- 2 33 μ F 350VW electrolytics
- 2 Diodes, EM408 or similar or BY127-1N4006
- 1 Diode, EM401 or similar or BY127-1N4002, RS276-1129
- 2 Zener diodes, BZX79C12, RS276-563/623
- 1 Miniature toggle switch, double-pole, 3-position
- 1 Miniature toggle switch, DPDT
- 1 Miniature push-button switch, single-pole, normally closed
- 1 Rotary switch, single-pole, 3-position
- 1 Transistor socket
- 3 Banana sockets
- 1 Knob

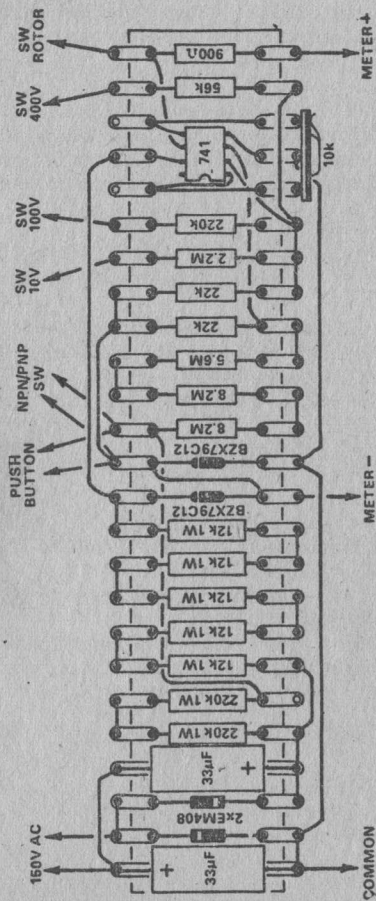
RESISTORS ($\frac{1}{2}$ watt unless stated otherwise)

- 1 220 ohms (low tolerance)
- 1 680 ohms (low tolerance)
- 1 10k

- 1 10k linear trimpot
- 5 12k 1W
- 2 22k (low tolerance)
- 1 56k (see text)
- 2 220k 1W
- 1 220k (low tolerance)
- 1 2.2M (low tolerance)
- 1 5.6M (low tolerance)
- 2 8.2M (low tolerance)

SUNDRIES

3-core flex and plug, flex clamp, screws, nuts, solder lugs, solder, hook-up wire.



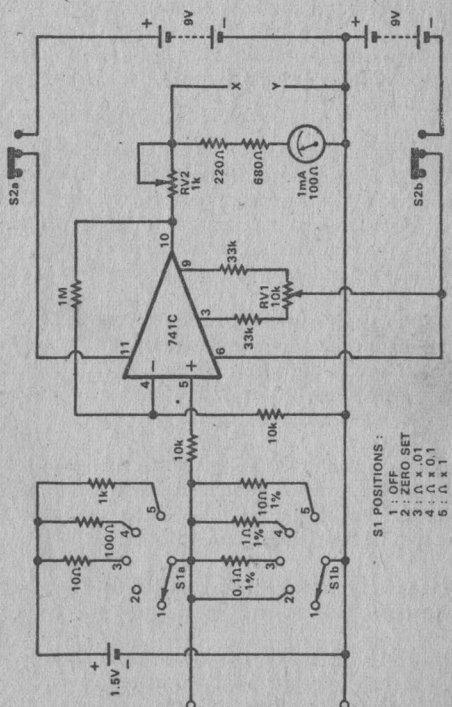
This wiring diagram of the miniature tagstrip should make construction a straightforward process. Follow the diagram in conjunction with the circuit

LOW RANGE OHMMETER

This ohmmeter fulfils a need for measurement of resistors from 0.01 ohm to over 100 ohms by comparing a resistor of unknown value with the selected 1% tolerance reference resistor. In operation the meter is zeroed by means of RV1 with the range switch set at position 2. In positions 3, 4 and 5 a voltage of almost 15mV across the reference resistor, coupled with an IC gain of 101 produces an output of about 1.5V and by adjusting RV2, the meter is set to FSD.

When connected, an unknown resistor forms a parallel circuit across the reference resistor lowering the voltage across it, thereby reducing the meter reading. Accuracy is better than 1% + or - the reference resistor tolerance. A meter having a mid-scale reading of 10 is necessary. Any VOM having this type of ohms scale could be used as the read-out by connecting the lower end of the 680 ohm resistor to 0V and connecting the VOM set at the 1V range between points X and Y.

A low range ohmmeter



- S1 POSITIONS:
- 1: ZERO SET
 - 2: 20Ω SET
 - 3: Ω x 0.1
 - 4: Ω x 0.1
 - 5: Ω x 1

Here is an interesting experimental project which should appeal to beginners and advanced workers alike; a simple bridge type ohmmeter using a novel form of null indicator. The circuit would also lend itself to a wide range of other applications.

Following the description of our most recent RC Bridge there was some discussion in our laboratory concerning bridge circuits in general and the manner in which they can be modified to perform specific functions. The discussion considered in particular the various forms of null indicator available and their suitability for unskilled operators who, ideally, need an unambiguous "high" or "low" indication or, in some cases a simple "go" or "no go" indication.

Arising out of this was the circuit we are about to describe; a null indicator based on an operational amplifier IC ($\mu A741$) and a pair of light emitting diodes. Not only is it a novel circuit but, as far as we know, also an original one. More importantly, it would seem to have considerable potential in many applications, particularly those involving unskilled operators, as already mentioned.

Fig. 1 is a simplified explanatory circuit, in which the operational amplifier is being used as a comparator. If the voltage at the non-inverting input is higher than that at the inverting input, the op amp will saturate, and the output voltage will rise to a value close to that of the positive supply rail, V_{cc} . This will forward bias one of the light emitting diodes, which will give a visible indication, and reverse bias the other diode, which will not emit.

When this situation is reversed, and the voltage at the inverting input is higher than that at the non-inverting input, the op amp will saturate in the reverse direction, and the output voltage will fall to a value close to that of the negative supply rail, $-V_{cc}$. This will forward bias the other light emitting diode, and reverse bias the first one.

Since the op amp has a very high gain, the change-over between the two conditions will occur very rapidly and over a very small range of differential input voltages. Thus the null of the bridge will be indicated by the point at which the two LED's change from off to on and vice-versa.

To prove the circuit, we made up a simple bridge network similar to that shown on the left hand side of Fig. 1.

It consists of a 1k linear pot, a set of three standard (close tolerance) resistors, a battery, and provision to connect the unknown resistor (R_x). The 1k pot forms a pair of resistors whose ratio is variable over a wide range although, in practice, we limit this range so as to provide a practical scale. From the mid position of the moving arm, which represents a 1 to 1 ratio, we provide for an extremity of 100 to 1 in either direction.

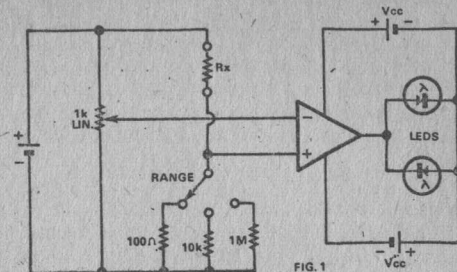


FIG. 1

Fig. 1. Simplified diagram showing the basic bridge configuration. Note that the 1k linear pot provides two of the four resistors making up the bridge. The other two are the unknown resistor and one of the standard resistors selected by the range switch.

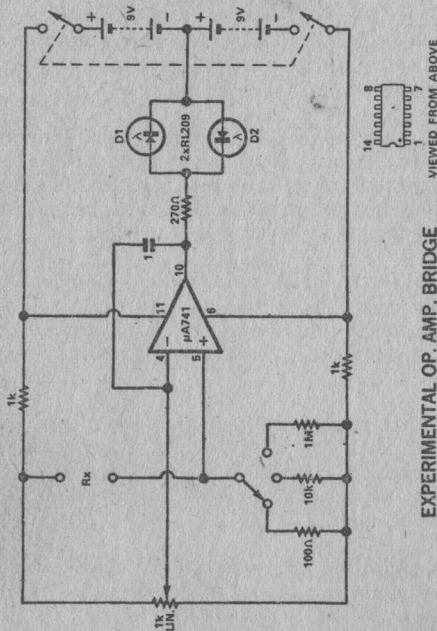


FIG. 2

EXPERIMENTAL OP. AMP. BRIDGE

Fig. 2. The circuit as finally developed. The bridge circuit remains essentially the same, except for the addition of two 1k current limiting resistors. The one battery pack, supplying 9V each side of a centre tap, serves both the bridge and the IC.

The three standard resistors are 100 ohms, 10k and 1M. In conjunction with the pot movement just discussed, this gives a range of measurement from at least 10 ohms to 10M, beyond this with some limitations, and with useful overlap between ranges.

After ironing out a few minor problems we came up with a circuit which we feel is worth passing on to readers. As presented it is a simple battery operated bridge, using relatively few low cost components, but capable of quite useful accuracy. It would seem to be an ideal project for YRCS groups or similar organisations.

Alternatively, by using closer tolerance components, it can be upgraded to provide a much higher order of accuracy, but at some increase in cost. And for the really keen experimental types there is an almost unlimited scope for its exploitation in automatic control circuits.

The practical circuit is shown in Fig. 2, and is worth discussing in some detail. The 741 op amp requires nominal voltages of plus and minus 15, but it will work quite satisfactorily for our purpose from voltages as low as plus and minus six. Since 9V batteries are readily available in convenient sizes, we elected to use two of these.

This gives us a total of 18V, which also becomes the power supply for the bridge. The latter should be as high as possible when high value resistors are being measured, otherwise current due to thermal noise in the resistors may be comparable with the current due to the supply.

On the other hand, a high supply voltage causes problems when low values of resistance are being measured, as the battery drain increases enormously. To avoid this, limiting resistors are connected in series with the bridge. To keep the input voltages to the op amp centered about zero, equal value resistors are fitted at the top and bottom of the bridge.

These resistors also limit the current to the op amp inputs, so that it is not necessary to provide separate input resistors. A protective resistor is required for the LEDs, to keep the forward current within the maximum allowable. Because the two diodes are back to back it is not possible to exceed their reverse voltage rating.

When the prototype was constructed, some trouble was experienced with oscillation at the null point. This took the form of cycling between the LEDs, which made both LEDs appear to be on at the same time. This was prevented by using a 1uF capacitor to provide negative feedback from the output to the inverting input.

In order to prevent oscillation if the wiper arm of the pot should become dirty and go open circuit, it is necessary that the wiper arm be connected to the inverting input, and not the non-inverting input.

The circuit does have one minor disadvantage. Because the null indicator operates from the same supply as the bridge, and because there will be some leakage between the op amp input terminals and the supply rails, there will be some error introduced into the null indication at higher resistance values. In practice this is insignificant up to about 10M, which should be regarded as the upper limit for accurate measurement.

The only close tolerance components are the three standard resistors. We used 1% types but, if these are not available, it may be possible to trim 5 or even 10% types to the correct values. To ensure stability with respect to temperature, only good quality components should be used.

We constructed our unit on a panel measuring 111mm x 161mm, which happened to suit a box we had on hand. Since the size of the panel and the type of box is not critical, we suggest that the reader make his own choice in this regard, possibly using something which is on hand.

A good idea of the panel layout and general assembly can be obtained from the photographs. The pot is in the top centre of the panel and the two switches and the unknown resistor terminals at the bottom. The LEDs we used fit neatly into a 1/8in hole and we mounted them at the top of the panel, about 20mm apart, where they are easily seen.

The only other piece of metal work is a simple battery clamp made from a scrap of aluminium. It is roughly "U" shape, but with wings extending outwards from the uprights of the "U". One 9V cell is held under each of these wings, but they also support a piece of veroboard on which is mounted the IC and other minor components. The bracket and the veroboard are clearly visible in the photograph. The bracket is fastened to the front panel by the pot, avoiding the need for unsightly screws.

We mounted the three standard resistors on the back of the single pole three position range switch. It was necessary to add a dummy terminal to the switch, to provide the common point for the resistors, by attaching a solder lug to a spare hole fortunately provided. Alternatively, there is ample room to mount the resistors on the veroboard with the remainder of the components, although this will entail running more wires to the switch.

The 741IC we used was in a 14 pin dual-in-line package. These devices are also available in an 8 pin dual-in-line package and in an 8 lead metal can. All are suitable for use in the bridge. We used veroboard with the correct hole spacing to suit the dual-in-line package.

In order to reduce the height needed for the op amp, we mounted it directly on the veroboard. As only pins 4, 5, 6, 10 and 11 are used, we did not solder the remaining pins to the veroboard. This makes it easier to remove the op amp if necessary. Take care not to damage the IC while it is being soldered, due to excess heat. It is a good idea to leave it to the last to solder in.

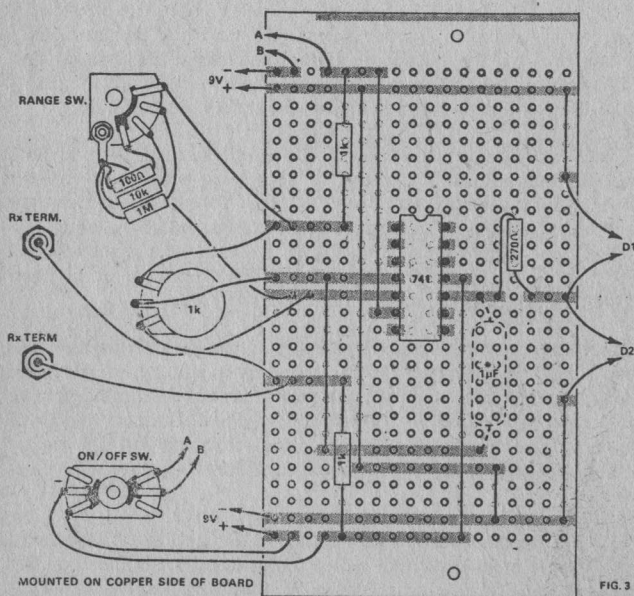


Fig. 3. Wiring diagram showing the general layout of components and the pattern of the Veroboard. Layout is not critical and almost any other physical arrangement would be acceptable. All components mount on the panel, simplifying the choice of a box.

As shown in the photographs, the IC and the remaining three resistors are mounted on top of the veroboard. The LED's and the 1uF capacitor are mounted underneath. This saves space, in the case of the capacitor, and enables the LED's to fit into their mounting holes in the front panel. The veroboard is attached to the battery mounting clamp using two self-tapping screws and some washers for spacers. Leave sufficient clearance under the board to avoid shorts to the battery case.

We chose to polarise the unknown resistor terminals by selecting one red and one black. While resistors are not normally polarised, the bridge may be used in other applications, such as checking the forward and reverse leakage of a diode. The red terminal should indicate positive.

We used pointer knobs for the two rotary switches and a larger round knob for the potentiometer. To this we added a clear plastic pointer, fastened to the bottom of the knob with a strong epoxy glue. Make sure that the line on the pointer is truly radial and in line with any lines or arrows already on the knob.

Finally, there is the matter of a scale. In the case of previous bridge projects we have been able to publish a calibrated scale and supply details to any firms wishing to supply etched labels. We are deliberately avoiding this approach on this occasion for a number of reasons.

Such a scale is valid only if the pot used by the builder is identical with that used in the prototype. This means, in practice, nothing less than a wire wound type of the same make and model. Unfortunately, wire wound pots are quite expensive and, to keep the cost down in what is, after all, mainly an experimental project, we used a linear carbon pot, which is a good deal cheaper.

Unfortunately, carbon pots, as a class, introduce two unwanted variables. They are not necessarily as linear as the wire wound type, even from unit to unit of the same model, and the angle of rotation varies from make to make. Thus, a scale prepared for brand "A" will not necessarily suit brand "B". But the linearity is another matter.

Because of these variables, the best approach is to make one's own scale; an interesting exercise in its own right. There are two approaches. One is direct calibration against a large number of close tolerance resistors, ideally one for each calibration mark.

The other is direct calibration of only two points, one at each extremity of the scale, and all other points by calculation on the assumption that the pot is linear.

Obviously the first would be the more accurate, but is seldom practical. The second is more practical, but leaves the accuracy at the mercy of the pot linearity. We used the second method.

For this method we require three close tolerance resistors, 1k, 10k and 100k, a protractor, and ordinary drawing instruments. The first job is to determine the angle between two selected values near the extremities of the pot rotation. Place a scrap of paper on the panel where the scale will finally fit and secure it with sticky tape. Fit the pointer so that it swings approximately evenly each side of centre. This setting is not critical at this stage.

Set the range switch to 10k and note the positions at which the 1k and 100k resistors are nulled. The 1k should null near the furthest anti-clockwise position, and the 100k near the furthest clockwise position. If these positions are reversed, reverse the outside connections to the pot.

Remove the pointer and the piece of paper and, with the protractor, measure the angle between the two marked points. This will be easier if the centre of the pot can be marked, using lines at right angles, before the paper is removed from under the pointer. The angle should be about 200 degrees.

Let us call this angle W . Then, if s is the fraction or multiple of 1 by which the standard has to be multiplied, and p is the angle in degrees that actually corresponds to this value of s , we can say $p = 0.611 \times W \times (s-1) / (s+1)$. This formula enables the values of p to be calculated from the values of s . The angle p is measured from the vertical, which corresponds to the position $s = 1$, $p = 0$. The angles are measured positive clockwise and negative anti-clockwise.

As an example, we have calculated the angle at which the scale marking "0.5" would be placed. We have taken the measured value W as 200.

$$p = 0.611 \times 200 \times (0.5-1) / (0.5 + 1)$$

$$p = 122.2 \times (0.5-1) / (0.5 + 1)$$

$$p = 122.2 \times (-0.5) / (1.5)$$

$$p = 122.2 \times (-0.333)$$

$$p = -40.74 \text{ degrees}$$

This means that the scale point marked "0.5" should be positioned 40.74 degrees anti-clockwise from the centre of the scale.

These calculations will be much easier if a slide-rule or a calculator can be used, as well as being much less prone to error.

Once the values for the scale have been obtained, the scale can be drawn on a suitable piece of white card. We made the scale with a diameter of 80mm, the major graduations 6mm long, and the minor graduations 3mm long, using India ink. Black ball point pen would provide an acceptable substitute for ink. We used rub-on lettering to mark the major points on the scale and also to label the functions of the rotary switches.

When the scale is finished, suitable cutouts must be made for the pot and switch shafts, the LED's and the terminals. With the scale fitted, the pointer can be attached to the pot shaft, and correctly aligned with the scale. This is done using the 10k close tolerance resistor, with the range switch on the 10k range. The pointer should be attached to the shaft so that when the bridge is balanced, the pointer reads 1.0. Provided there are no serious linearity errors in the pot, the scale should then be accurately calibrated for all ranges.

Care should be exercised in regard to battery life. The current consumption is highest on the lowest resistance range, and unlike an ohm-meter, current is consumed even when the unknown terminals are open circuit. Fortunately, the LED's give a visible indication that the power has been left on. Otherwise, due to the intermittent nature of the current drain, battery life should be quite long.

PARTS LIST

1 $\mu A741$ op amp, RS276-007/010/1741 or equivalent

1 red LED's, CEMA RL209 or similar

2 1k $\frac{1}{4}$ watt resistors

1 100 ohm 1% $\frac{1}{4}$ watt resistor

1 270 ohm, $\frac{1}{4}$ watt resistor

1 10k 1% $\frac{1}{4}$ watt resistor

1 1M 1% $\frac{1}{4}$ watt resistor

1 1k linear pot

1 1 μ F capacitor

2 9V batteries with clips

2 terminals, 1 red, 1 black

1 2 pole 2 pos rotary switch

1 1 pole 3 pos rotary switch

3 knobs to suit (see text)

Vero board, scrap aluminium, hookup wire, solder, self-tapping screws, white card, 5mm spacers, solder lugs, screws and nuts, small piece of clear plastic.

Note: resistor wattage ratings and capacitor voltage ratings are those used in our prototype. Components with higher ratings may generally be used provided they are physically compatible. Components with lower ratings may also be used in some cases, providing ratings are not exceeded.

A fully solid-state instrument which produces synchronously graded AF tone bursts suitable for a wide variety of audio measurements. The use of integrated microcircuits gives the instrument a standard of performance comparable with quite costly commercial units, despite the fact that it is easily constructed at low cost.

Although at first sight the instrument described in this article might appear to be suitable only for a few specialised and somewhat esoteric applications, this is not the case. In fact it has many uses, of which probably the most practical and straightforward is the measurement of short-term or "dynamic" power output capability of audio power amplifiers.

Other direct uses include testing of amplifier stability, measurement of loudspeaker system response and damping, measurement of room and auditorium acoustics, measurement of the performance of filters and automatic gain and level control circuits, sonar equipment testing, and psychometric research and testing. The instrument may also be used to produce sound effects for live theatre and motion pictures, and in the production of electronic music.

In addition it may be used directly or indirectly as a signal shaper-squarer, a wide-range binary scaler-frequency divider, or a synchronous digital gate controller suitable for counting, frequency and time measurement, or clock word generation.

In view of the likely interest in the use of tone-bursts to measure the short-term power output capability of audio amplifiers, a brief discussion of this application will now be presented by way of a preamble to the description of the instrument itself. It is hoped that the discussion will be found of value not only to those who may contemplate construction of the instrument but also to those who wish solely to gain a theoretical knowledge of this relatively recently evolved technique.

It is commonly found that the maximum power output which may be obtained from an audio power amplifier, before discernible waveform clipping or other distortion is incurred, depends significantly upon the period of time over which the power is to be delivered. Generally, the shorter the period involved the higher the power which may be delivered—although this simple relationship will naturally be limited by transformer losses and other effects influencing the power frequency response.

Thus the maximum power output delivered by a power amplifier when tested under continuous sinewave conditions may often be significantly less than that which it is capable of delivering during the short and relatively infrequent amplitude peaks present in typical musical programme material. Hence for the many audio amplifiers

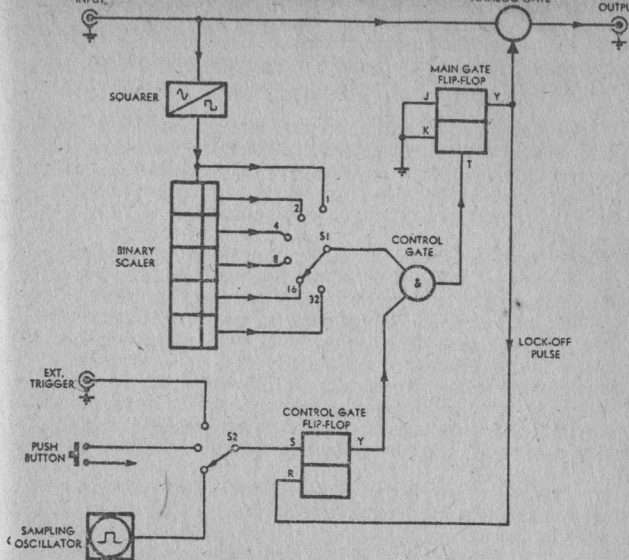


Fig. 1 BASIC OPERATION OF THE TONE-BURST GATE

Specification

A synchronous analog gating unit which may be used to derive pedestal-free bursts of input signals between 10Hz and 60KHz. The integral number of cycles in each burst and the burst repetition rate may be adjusted independently, or locked if desired. The instrument is fully solid-state and employs low-cost RTL digital microcircuits. It may also be used as a signal shaper-squarer, a scaler frequency-divider or a synchronous digital gate controller.

ANALOG INPUT:

- Frequency Response: 10Hz—60KHz.
- Minimum Input: Less than 150mV RMS.
- Maximum Input: Greater than 600mV RMS.
- Switching Ratio: Better than 40dB.
- Insertion Loss (gate open): Less than 6dB minimum.
- Switching Transients: Less than 30mV P-P.
- Input Impedance: Approximately 33K.
- Output Impedance: Maximum 6K, minimum 500 ohms.
- T.H. Distortion (gate open): Less than 0.2% at 250mV RMS input.
- Gate offset or pedestal: Negligible.

BURST LENGTH: Adjustable to 1, 2, 4, 8, 16 or 32 cycles; also continuous.

BURST REPETITION RATE: Internal trigger generator provides a range of approx. 3Hz—500Hz. Single bursts triggered manually by panel push-button. External triggering by +3V pulse, duration less than 20uS. Inter-burst interval may also be locked to equal the burst length.

SYNC OUTPUT: Rectangular waveform with level 0V when gate is closed and +3V when gate is open.

intended to measure musical signals rather than sinewaves, a conventional power output rating based upon continuous sinewave measurement may be both unduly conservative and of only incidental relevance, despite the relative ease and high order of objectivity with which the measurement may be made.

Aware of the limitations of a continuous sinewave power rating when used alone as a performance criterion applied to their products, manufacturers of high fidelity amplifiers some time ago adopted other rating methods which have been employed either as supplements to the basic continuous sinewave rating, or as substitutes for it.

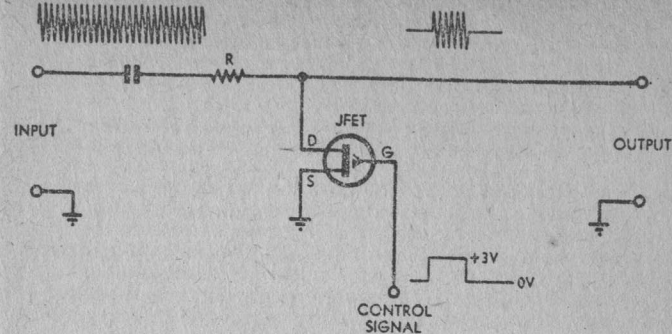
It is undoubtedly true that academic purism was by no means the sole motivation behind the adoption of these other—usually more flattering—rating methods. It is also true that some of the rating methods adopted were and still are, of dubious technical value. However, it is neither appropriate nor possible in the present article either to discuss in detail the various rating systems or to examine the motivation behind their introduction. Sufficient to say that one of these ratings, the so-called “music power” rating, was introduced in an effort to provide a measure of the actual power output capability of an amplifier when reproducing a musical programme.

Unfortunately while such a rating is easily proposed and validly justified, the measurement ideally required is not at all simple or straightforward in execution. The basically statistical nature of musical signals itself produces both theoretical and practical problems when one attempts to measure the power level involved, quite apart from the perhaps more obvious difficulty in selecting the “typical” musical programme to be used as a basis for inter-amplifier comparison.

The method originally adopted for “music power” measurements bypassed these problems altogether, by making a convenient and usually fairly valid assumption regarding the cause of the diminished power output capability of an amplifier when tested under sinewave conditions. The assumption was that the diminished power output was due solely to imperfect power supply voltage regulation under the steady drain of continuous operation at maximum power output.

The converse corollary of the assumption was that under the transient conditions of a musical programme the power supply was sufficient to cope with the relatively infrequent demands for full power, and was therefore able to maintain supply voltages at their full amplitude. Hence in order to measure the “music power” output capability of an amplifier, it was reasoned, it was not necessary to attempt to take measurements under actual musical conditions but simply necessary to take a “normal” continuous sinewave measurement with the amplifier supply voltages maintained at their full amplitude to duplicate the situation prevailing during brief musical peaks.

Consequently an established technique for the measurement of amplifier “music power” is to substitute for the internal amplifier power supply an external heavy-duty regulated supply—delivering the same voltages—and to measure the maximum continuous sinewave output under these artificial conditions.



(a) BASIC JFET SHUNT ANALOG GATE

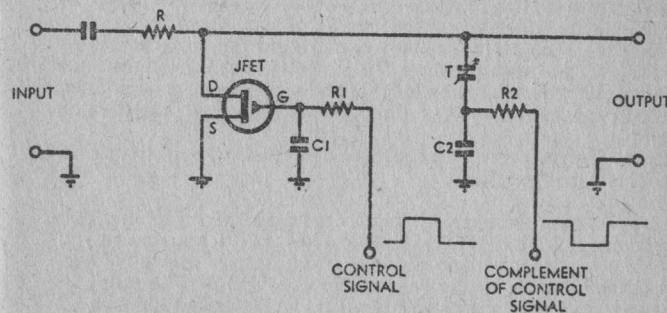


Fig. 2 (b) MODIFIED ANALOG GATE WITH TRANSIENT SUPPRESSION

Although this technique is widely used it is not without its critics. Many are not happy about the assumptions involved, and are dubious about the validity of simply equating sinewave conditions with a substitute power supply to transient conditions with an amplifier's actual power supply. Quite apart from such criticisms there is the practical inconvenience involved in matching a substitute regulated supply to an amplifier and connecting it in circuit for the testing.

Fortunately an alternative approach exists, and one which is not only more valid theoretically but also more convenient in practice. The method involves testing using short bursts of sinewaves delivered to the amplifier from an instrument such as that to be described, with the amplifier itself unaltered and powered from its own internal supply. The “tone bursts” used are quite easily arranged to have a duration and repetition rate which effectively duplicates infrequent musical programme peaks, while their relatively simple and basic nature makes them both easily reproducible and also more readily measured than a musical signal.

It should be noted that a tone burst power output measurement does not really give the information originally signified by the term "music power"—nor, for that matter, does a measurement made using the "substitute power supply" method. Both methods give essentially the absolute maximum power output capability of an amplifier, under short-term conditions; and this is not strictly equivalent to the somewhat vague notion of "music power output".

Probably the most accurate way of visualising the situation is to regard the "music power" capability of an amplifier as a dynamic or variable parameter which depends at any instant upon the immediately preceding amplifier conditions, and which accordingly varies over a range defined at its lower limit by the normal continuous sinewave output rating and at its upper limit by the normal continuous sinewave output rating and at its upper limit by the rating provided by a tone-burst or "substitute supply" measurement.

Late in 1965 the American Institute of High Fidelity published a revised version of their earlier "IHF Amplifier Standard" rating system, in which they advocated the use of the term "dynamic power rating" to signify the short-term output power capability of an amplifier, replacing the term "music power rating". The method suggested for the measurement of dynamic power rating was the tone-burst method, using bursts of 10ms duration at a low reception rate.

Let us now turn to the instrument to be described, considering first its principle of operation.

The input signal to be gated is fed to both an analog gate and to a shaper-squarer. The latter derives from the signal a squarewave equivalent having extremely fast rise- and fall-times corresponding to the zero crossings of the original signal waveform. The output of the squarer is fed to a group of flip-flops which are arranged in "toggling" mode and connected in cascade to form a binary scaler or frequency divider.

A six-position switch (S1) is arranged to select either the direct output from the squarer or the scaled-down signal from a chosen scaler stage. The selected signal from the switch is then passed to an AND logic gate whose second input is controlled by a simple set-reset state and thus holds the AND gate closed; thus the signal selected by the switch cannot progress further.

The "set" (S) input of the control flip-flop is connected to a second switch (S2) which may be used to select the desired source of burst triggering signals. There are two triggering sources internal to the instrument, one a free-running "sampling oscillator" of adjustable frequency—for repetitive operation—and the other a manual on-off circuit operated by a front-panel push-button. A third position of the switch allows the use of external triggering sources.

On the arrival of a triggering pulse at its S input, the control flip-flop switches to its set state and accordingly opens the control AND gate. This excursion from the signal selected by S1 to pass through the gate and reach the "toggle" (T) input of a second flip-flop, which is a J-K type whose output controls the analog gate.

Under quiescent conditions this "main gate" flip-flop is in the reset state, and accordingly the analog gate is closed. The analog (tone) input signal is therefore prevented from reaching the output terminal. However, the arrival of the first toggling edge from the control gate switches the main gate flip-flop into the set state, and the gate is immediately opened to pass the tone signal.

The period of time for which the gate remains open depends upon the setting of switch S1, because the second toggling edge to reach the main gate flip-flop via the control gate will cause the former to be reset once more—closing the analog gate with it. And depending upon the setting of switch S1, this turn-off edge will correspond to the end of a single tone cycle, or two cycles, or four, eight, sixteen, or thirty-two. Hence the analog gate will have been open for a period which accurately corresponds to an integral number of tone cycles, and a "burst" of the desired number of tone cycles will have been passed to the output terminal.

In order to prevent further opening and closing of the analog gate due to the arrival of further toggling edges at the main flip-flop (from S1), a pulse derived from the main gate output is used to reset the control gate flip-flop immediately following the resetting of the main gate flip-flop. This closes the control gate and prevents further signals from reaching the main gate flip-flop from S1. Both flip-flops thus remain reset, and the analog gate remains closed, until the arrival of another triggering pulse from S2.

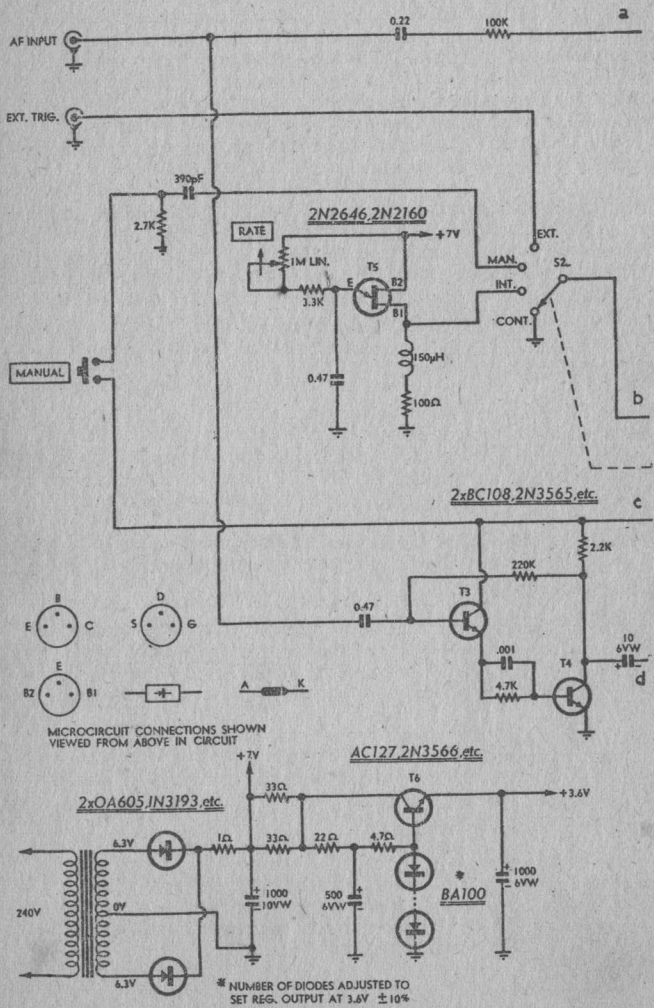
From the foregoing it should be apparent that it is the triggering pulses, supplied to the control gate flip-flop from S2, which initiate each tone burst; while the duration of each burst is governed by the tone-derived signal selected by S1. The burst repetition rate and the burst duration may therefore be adjusted independently.

Before turning to consider the full circuit diagram of the new instrument it may be worthwhile to look briefly at the operation of the analog gate circuit employed.

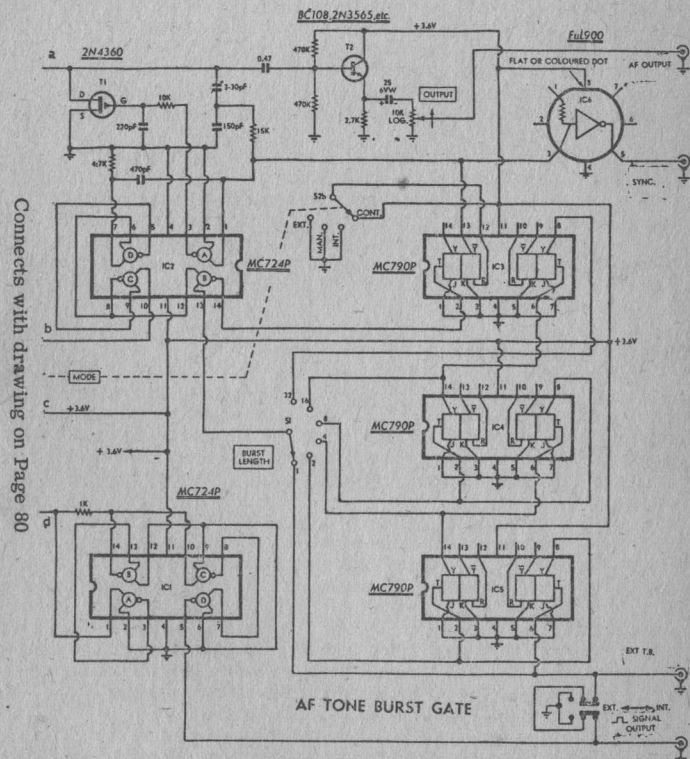
A number of circuits have been developed for this purpose; however, the circuit which has been selected in the present instance employs a p-channel junction field-effect transistor (JFET) as a voltage-controlled shunt switch. Its basic operation is shown in the diagram of figure 2(a).

As may be seen the gate circuit proper consists simply of a series input isolating resistor R and the JFET, whose drain-source channel is connected between the output terminal and ground. The effective drain-source resistance of the JFET is controlled by means of a control signal applied to the gate electrode.

When the control voltage is at zero, there is no applied transverse electric field in the JFET channel. Accordingly the drain-source is low—a few hundred ohms—and the input signal suffers heavy attenuation as a result of the divider action produced by resistor R. Under these conditions the gate is effectively "closed".



Connects with drawing on Page 81



When the gate is taken positive, however, the reverse bias applied to the gate-channel junctions of the JFET produces a transverse field which "pinches off" the channel. The drain-source resistance rises to a very high value, and the input signal therefore appears at the output terminal suffering only slight attenuation. The gate under these conditions is effectively "open".

The switching action of this type of analog gate is free from many of the problems associated with other circuits. Because the drain-source channel requires no DC bias, and furthermore acts over a fairly wide signal amplitude range as an almost perfectly linear resistor (of controllable value), there is virtually no DC offset component or "pedestal" associated with resultant tone bursts.

The available channel resistance range is quite wide, even with moderate control voltages, giving the gate a high "on" to "off" ratio combined with a low insertion loss in the "on" state. Hence the gate circuit used in the final instrument exhibits an on-off switching ratio of more than 40dB, with an insertion loss of only about 5dB.

The shunt JFET gate is also capable of switching at a high speed although the operation at high speeds is complicated by inter-electrode capacitances. Drain-source capacitance tends to increase the insertion loss and reduce the switching ratio at high signal frequencies, while gate-source capacitance tends to slow down the rise and fall of the gate control voltage. However, more troublesome than both of these is gate-drain capacitance, which tends to feed control signal energy into the signal circuit.

The effect of the latter is that sharply differentiated switching "spikes" tend to appear across the analog output terminals. A large positive spike appears as the gate opens, and a smaller and narrower negative spike as the gate closes. The differences in amplitude and duration arise as a result of the differing dynamic R and C situations present in the JFET for turn-off and turn-on. As the spikes represent spurious gate output or "noise", it is important that they be suppressed.

A number of suppression techniques may be used. However, the method which has been adopted in the present instrument is shown in the diagram of figure 2(b). A two-step process, it works partly by slowing down the attack and decay of the JFET gate control signal by means of the low-pass filter formed by resistor R1 and capacitor C1. This reduces the charge stored in the JFET gate-drain capacitance, by reducing the rate of change of voltage. (The reduced switching speed is still high relative to the analogue signals being switched, so that the gate operation is not impaired.)

The second step involves cancellation or nulling the remaining spikes, in effect by injecting into the analog circuit spikes of opposite polarity. This is performed by an adjustable trimmer capacitor T which is connected to a source of signal complementary to the JFET control signal. A second low-pass filter comprising R2 and C2 is used to adjust the attack and decay of the nulling signal to correspond with the JFET gate signal.

At a critical adjustment of the trimmer the net charge transfer to or from the analog circuit during gate opening and closure is reduced to a very low level. The tone bursts produced by the gate are then quite "clean".

Reference to the main circuit diagram of the complete instrument will show that many of the logical functions are performed by integrated digital microcircuits. Six such devices are used, five of which are in the "dual in-line" encapsulation while the remaining device is of the TO-5 eight-lead type. All six devices are of the economy plastic RTL type, the dual in-line devices being from the Motorola Semiconductors' MC700P range and the TO-5 device from the Fairchild Semiconductor FuL900 range.

The cost of these devices is quite modest when one considers the number of discrete components which would be required to perform the same functions—if one includes the functionally identifiable devices within the microcircuits, for example, the total transistor count for the complete tone-burst gate is greater than 90! From this figure it may also be deduced that, quite apart from the cost advantage, the use of microcircuits simplifies quite drastically the construction of the instrument.

The shaping-squaring function of the instrument is performed by transistors T3 and T4 together with IC1, which is an MC724P quadruple 2-input gate. The transistors are employed in a simple input preamplifier circuit having a moderately high input impedance. Output from T4 is coupled to gate A of IC1, and the output of this gate in turn feeds to gate B. Both gates are used simply as squarer/inverters, with a small amount of positive feedback from B's output back to A's input via a 1K resistor to improve input sensitivity. The two remaining gates of IC1 are connected in cascade from the output of B, to give additional squaring; the output from gate D accordingly has quite short rise—and fall—times.

Although in the simplified instrument block diagram of figure 1 the output of the shaper-squarer block is taken directly to the scaler and selector switch, in the instrument itself this connection is made by means of a small slider switch mounted on the rear of the case. In addition both the shaper output and the scaler-selector input are made available at adjacent coaxial connectors. This allows the internal connection between the two to be broken at will, and external arrangements made either for use of the squared signal and/or for use of substitute gating signals.

As a result of this simple and inexpensive refinement not only is the instrument made more flexible in terms of its main function, but it is also rendered capable of performing additional functions. Thus the squarer section becomes available for external use, as does the scaler section. Use of the latter section as an external scaler/frequency divider will be discussed further shortly.

The scaler section itself consists of five binary J-K flip-flops, each of which is half an MC790P microcircuit. Devices IC4 and IC5 provide four of the flip-flops, while the fifth is the element in IC3, remote

gate A or IC2 for amplification and incidental inversion. The direct flip-flop output is also differentiated to produce the control gate flip-flop "lock-off" signal. Additionally it is used to derive the analog gate transient suppression signal, and also to provide an oscilloscope synchronising signal via IC6. The latter device is employed as an isolating buffer-inverter, and accordingly delivers a +3V rectangular signal corresponding to the "open" state of the analog gate.

It is the IC6 "sync" output connector which may be used as the output of the scale section of the instrument when the latter is required for external use. With the signal to be scaled or divided applied to the "Ext T.B." input connector, the slider switch in the "Ext" position, and +3V DC applied to the external triggering input, the selector switch S1 may be used to adjust the scaling ratio over the binary range from 1:2 to 1:64. Note that when the instrument is used in this way the main gate control flip-flop is used as a sixth scaler stage.

As mentioned earlier there is a fourth position of the "mode" switch S2, in which the analog gate is kept "open" continuously to permit setting-up of auxiliary equipment. To provide this mode of operation a second pole of S2 is used to apply +3.6V to the "reset" input of the main gate flip-flop, holding the latter locked in the reset state. As output is taken from the Y-complement output of the flip-flop rather than from the Y output, this results in the analog gate being locked in the "open" condition.

The analog gate of the instrument is of identical form to that already discussed and illustrated in figure 2. It employs a P-channel economy silicon JFET type 2N4360 or similar, and as a result of the relatively modest demands made by this mode of operation, most such devices should work satisfactorily despite the wide spread in parameters. A small proportion of devices may prove unsuitable by virtue of unduly low gain or channel resistance, but these should be sufficiently rare to make replacement an economic proposition.

In order to achieve maximum switching ratio combined with minimum insertion loss, an emitter follower stage is employed following the analog gate as an impedance transformer and buffer. As the dynamic range of the gate itself is but modest, the emitter follower functions quite satisfactorily from the +3.6V microcircuit supply line.

For optimum performance the analog gate and emitter follower should be operated at a signal input level of between 250 and 500mV RMS, as this provides the best noise/distortion compromise. In view of this, and because in most applications a burst signal of adjustable amplitude will be required, the output of the emitter follower is controlled by a simple potentiometer attenuator.

The power supply of the instrument is a relatively simple circuit supplying approximately 7V unregulated for the unijunction trigger oscillator and a regulated 3.6V + or - 10% for the remaining circuits. The unregulated supply is a simple full-wave rectifier employing two silicon diodes type OA605, 1N3193 or similar, and a miniature centre-tapped 12.6V step-down transformer.

The regulator is a simple series type employing either a germanium type AC127 or a silicon type 2N3566 medium-power NPN transistor. The base reference voltage is well filtered and regulated, the latter being performed by a string of forward-biased silicon diodes in view of the relative scarcity of low-voltage zener diodes. The number of diodes is arranged so that the regulated output voltage at the transistor emitter is within + or - 10% of its nominal 3.6V value.

As may be seen from the photographs the instrument is housed in the small flanged front panel instrument case used for most of our recent instrument designs, measuring 7½in x 5in x 4in. There are only four main front-panel controls, comprising the "burst length" and "mode" switches together with the "burst rate" and "output" potentiometers. The remaining fittings consist of the push-button for manual operation and the four main coaxial connectors.

At the rear of the case are the squared-signal output and external timebase input connectors, the "Int-Ext" slider switch and the mains cord entry.

Inside the case, most of the components are mounted on a "Veroboard" perforated printed-strip conductor panel measuring 7in x 3¾in; the panel is a section of stock Veroboard strip. The position of components and interconnection of copper conductors should be fairly clear from the board diagram provided, and also from the internal photographs.

The Veroboard wiring panel is supported by two small sheetmetal brackets which are attached to the rear of the front panel by the input connector mounting screws and the "burst length" switch mounting bush respectively. The brackets mount the panel at a distance of 2in behind the front panel, and close to the top of the case.

Wiring of the components on the board is a fairly straightforward operation, and should present few problems. However, care should be taken to ensure that components are not damaged either by mechanical strain during the lead bending and insertion, or by overheating during the soldering operation. The soldering itself should be done quickly and carefully with a low to medium power iron having a small bit; a steady and sure hand is required to ensure that the copper conductors are not overheated nor accidentally shorted by tiny solder links.

The microcircuit connection leads should not be cut, but simply passed through the board holes, bent over carefully and soldered. It will be found that the "dual in-line" devices have small moulded projections which automatically space the device from the board. The TO-5 device should be spaced approximately 3/8in from the board, which spacing may also be used for the transistors.

To prevent strain on the Veroboard panel the power transformer is mounted in the lower rear of the case, with the mains cord, primary and secondary leads terminated on a 5-lug tagstrip alongside. The mains cord is clamped inside the case to prevent strain on the tagstrip.

Three minor components are supported by convenient switch and potentiometer lugs behind the front panel rather than on the Veroboard panel. These components are the 2.7K resistor and 390pF capacitor associated with the manual triggering button, and the 1000uF 10VW power supply input reservoir electrolytic capacitor. The first of these connects between the push-button and the input connector earth bus, while the second connects between the appropriate lug on S2 and an unused lug on S1. The 1000uF electrolytic connects between the +7V lug on the "burst rate" pot and an earth lug on the "mode switch".

When wiring the unit it is wise to connect only four BA100 diodes into the power supply reference circuit. This will ensure that when the unit is first switched on the regulated output will not be excessively high. Upon switch-on the voltage should be checked, and an additional diode or diodes added as necessary to bring the supply within the required range.

The only other adjustment to be made to the completed instrument is that concerning the switching transient suppressor trimmer. This adjustment is most easily made with a sensitive wide-band oscilloscope connected to the AF output, and in the absence of AF signal input. To permit gate operation under these conditions an external square-wave gating signal must be connected to the "Ext T.B." socket, with the slider switch set appropriately.

It will be found that at a critical setting of the trimmer, switching transients are reduced to a minimum. The trimmer should be locked in position at this setting with a small dab of nail lacquer or similar cement, to prevent vibration disturbance. If difficulty is experienced in obtaining a null balance, or if the null is insufficiently accurate to reduce the transients below 30mV P-P, it may be necessary to modify the R-C filter values in either the JFET gate circuit, or the trimmer feed circuit, or both.

LIST OF PARTS:

- 1 Small instrument case, 7½in x 5in x 4in, with flanged front panel.
- Handle, rubber feet, scrap aluminium for wiring board brackets.
- 1 7in length of Veroboard panel
- 1 Miniature power transformer, 12.6V CT
- 1 Rotary switch, 1 pole, 6 positions
- 1 Rotary switch, 2 poles, 4 positions
- 1 Slider switch, 2 poles, 2 positions
- 1 Miniature push-button
- 6 Small coaxial sockets
- 1 Medium fluted control knob
- 2 small fluted control knobs
- 1 Small skirted knob, numbered

SEMI-CONDUCTORS AND I.C.s

- 3 MC790P dual flip-flops
- 2 MC724P quadruple gates
- 1 FuL900 inverting buffer
- 1 2N4360 junction FET
- 1 2N2646, 2N2160 or similar UJT
- 3 BC108, 2N3565 or similar
- 1 AC127, 2N3566 or similar
- 2 OA605, 1N3193 or similar
- 4-5 BA100 or similar (see text)

- RESISTORS (5 p.c. tolerance, ½W for fixed values)
- 1 x 1 ohm, 1 x 4.7 ohm, 1 x 22 ohm, 2 x 33 ohm, 1 x 100 ohm,
 - 1 x 1K, 1 x 2.2K, 2 x 2.7K, 1 x 3.3K, 2 x 4.7K, 1 x 10K, 1 x 15K,
 - 1 x 100K, 1 x 220K, 2 x 470K
 - 1 x 10K log pot
 - 1 x 1M linear pot

CAPACITORS

(Fixed values, low voltage plastic, except electrolytics)

- 1 x 150pF, 1 x 220pF, 1 x 390pF, 1 x 470pF, 1 x .001uF, 1 x .22uF,
- 3 x .47uF
- 1 3-30pF ceramic air trimmer
- 1 10uF 6VW electro
- 1 25uF 6VW electro
- 1 500uF 6VW electro
- 1 1000uF 6VW electro
- 1 1000uF 10VW electro

MISCELLANEOUS

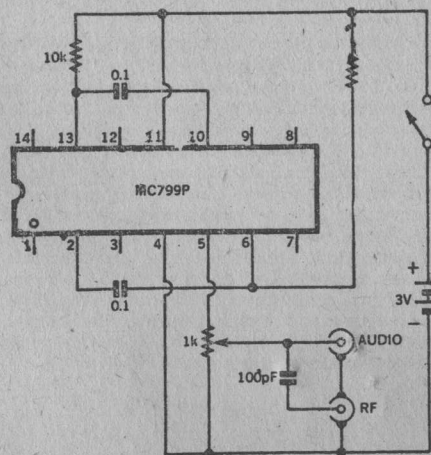
- 1 x 150uH RF inductor, power cord and plug, grommet for cord entry,
- nuts, bolts, washers, solder lugs, connecting wire, solder, etc.

SIGNAL INJECTOR USES IC

I decided to make up an injector using a Motorola IC type MC799/P and the circuit shows what I came up with. It is basically a battery powered 1000Hz multivibrator that generates square waves. The amplitude of the square wave output is fully variable and sufficient to test a loudspeaker.

The injector also provides a wide bandwidth RF signal suitable for testing RF and IF stages. Current drain is of the order of 80mA which allows a reasonably long battery life. The resistors used were ¼W types and the 1K potentiometer is a carbon type with switch. The two 0.1uF capacitors are 10V disc type, the 100pF capacitor is also a disc type.

Construction may be left to the individual ideas and requirements of the builder. Many ideas will no doubt suggest themselves, possibly dictated by what materials and facilities happen to be available at the time.



SIMPLIFIED VERSION OF SQUARE WAVE GENERATOR

After experimenting and consulting the Fairchild data sheets, which gave the output of the FuL914 as 16 units and the input of the FuL900 as 6 units, I decided that the FuL914 would not be overloaded by the FuL900. Tests showed that the circuit operated satisfactorily in this mode, so I built up the complete unit as illustrated. In addition, I left the No. 2 lead of the FuL914 floating in order to use it as a sync. input, should this be necessary. AC and DC outputs were also made separate.

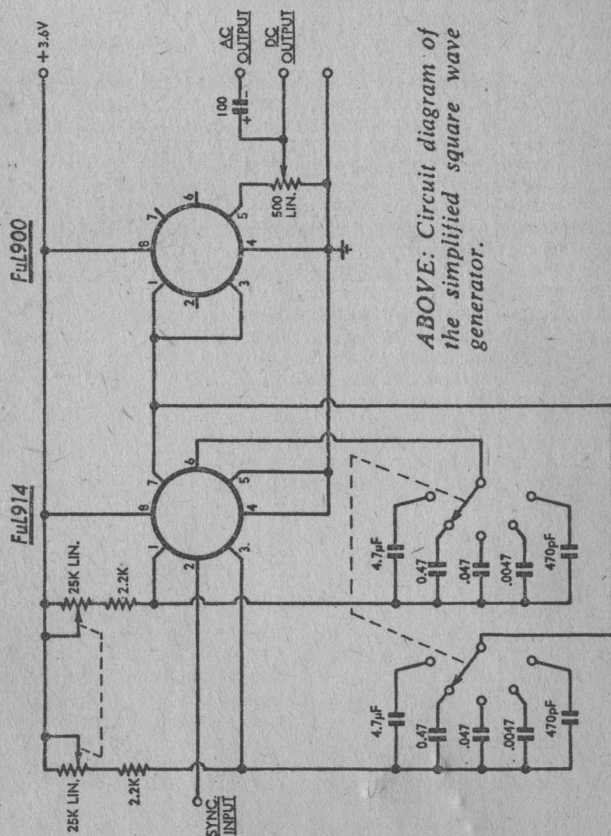
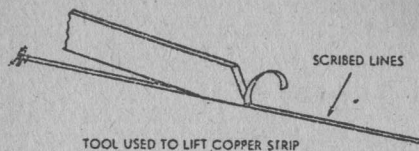
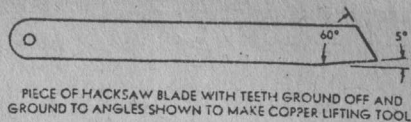
The performance seems to be unchanged by the modifications, since my generator has a frequency range of 4Hz to 2MHz. You may be interested in my experience with the modified circuit.

In addition, I would draw attention to the method of construction adopted, which is much less messy and time consuming than etching your own boards. Basically, it involves the removal of thin strips of copper from the board with a suitable scribing tool, to provide the necessary pattern. The method, incidentally, leaves intact a maximum amount of copper and, with it, the ability to make the circuit changes which are often necessary in '1 off' designs.

The approach is first to collect all the components required for the project, since physical placement is most important. Then draw the circuit on the copper side of the board in pencil, which can be rubbed out, if necessary, with an ordinary eraser. Leave room for the actual size of the components involved and, for preference, arrange them in a neat and orderly fashion. After a certain amount of planning and possible rearrangement, it will become apparent where lines have to be scribed to separate the components.

Fine parallel lines are scribed to cut through the copper and a tool, made from a specially ground section of hacksaw blade, can be used to roll the copper out from between the scribed lines. A little practice is required to get the knack; too steep an angle will cause the tool to dig in; too shallow an angle will cause the blade to skip out of the groove.

Having drilled the holes for the component pigtails, the board can be viewed against the light from both sides to pick out any copper which has been missed or copper feathers which might short across the grooves. The components can then be soldered in, taking care to use a heat shunt when soldering leads to delicate components. If the planning has been done well, the need for jumpers will have been minimised but, if desired, these can often be hidden under components. We use a standard knife devised for scribing Laminex and similar sheeting. During development of the original square-wave generator, we passed through the simpler form suggested but decided on the additional stages: (a) to render the device less dependent for its operation on a degree of unbalance in the basic multivibrator sections and (b) to preserve a near-unity mark-space ratio irrespective of output load conditions. However, if "1-off" units constructed to the simpler design work satisfactorily, a worthwhile economy will have been effected. If particular units fail to operate, it may be necessary to try another FuL914



EQUIVALENTS TABLE FOR SEMICONDUCTORS & IC'S

1N914=RS276-1139. AA117-118. 1N618. OA91/5.
1N4004=RS276-1103/1138. SK3031. BY137/400
1N4148=RS276-1139. AA117-118. 1N618. OA91/5.
1N5339B=Zener diode
2N2646=2N2160. RS276-2029. TIS43
2N3565=BC108-209. RS276-2009. 2N4967
2N3638A=MPS3638. TIS50. 2N5447. RS276-2021. SK3025.
2N3643=RS276-2033. BSW54. 2N2219A
2N4258=RS276-2023. BS W25. 2N3546.
95H90. GHJ121. Divide by 10/11 prescaler
741/C=RS276-010/1741. CA3056-3741. MC1539-1741. TBA221/2 OP Amp
7400=RS276-1801. FJH131, FLH101-105. MC408-458-508-558. Quad 2
Input NAND Gate
7490=9390. FJJ141. RS276-1808 Decade counter
9001=7470. 8470. FJJ101. FLJ101-105.
AC127=RS276-2001. 2N3566-647-1304-1308-2430. 2SD96. SK3010. AC141-
176-179-181.
BA100=RS276-1136. 1N659-914/A-4863. BA108-127-215. BAV18. BAW63.
BC107=RS276-2031. BC147-167-207-317. 2N2921-3568. 2SC458.
BC108=RS276-2009. BC148-168-208. SK3020. MPS6520. 2N3564-3711-4131.
BC177=RS276-2024. BC204-212-261-307-512-557. 2SA565.
BZX79C12=RS276-563/623. 400mW 12Volt Zener Diode
BZX79C15=RS276-624/564. 400mW 15Volt Zener Diode
BZY88C5V1=400mW 5.1 Volt Zener Diode
EM401=RS276-1139-1101-1105. BY127. 1N4007. BY126/50. BAY219. -
EM408=BY127. 1N4006. RS276-1114. 1S107.
EM4050=RS276-1139-1101-1105. BY127. 1N4007. BY126/50. BAY219.
FuL900=U8B900 Buffer
FuL914=U8A914. RS276-015. Dual 2 input gate
LM309K=UA7805. RS276-1830. Voltage regulator
MC724P=MC824-924. Quad 2 input NOR gate
MC790P=MC890-990. Dual JK Flip-Flop
MC799P=MC899-999=Dual Buffer
NE555=RS276-1723/1728. Timer
OA91=RS276-1102/1136. AA117-118. 1N618. OA95.
OA95=RS276-1102/1136. AA117-118. 1N618. OA91
OA605=1N3193. RS276-1139. 1N4003. 1S101. BY127

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