$3.00

TEST GEAR

- metering and power supply projects

PROJECTS INCLUDE:

- audio millivoltmeter
- digital display
- digital voltmeter
- simple frequency counter
- phase meter
- audio signal generator
- signal generators
- logic testers
- IC testers
- five power supplies
- decade resistance box
- scope calibrator
- beam adaptor
- silent A-B switch
- audio attenuator
- universal timer

an ELECTRONICS TODAY publication

TEST GEAR

- metering and power supply projects

EDITORIAL
Project Designs: Barry Wilkinson
Editor: Jan Vernon
Publisher: Collyn Rivers
Managing Director: Arnold Quick

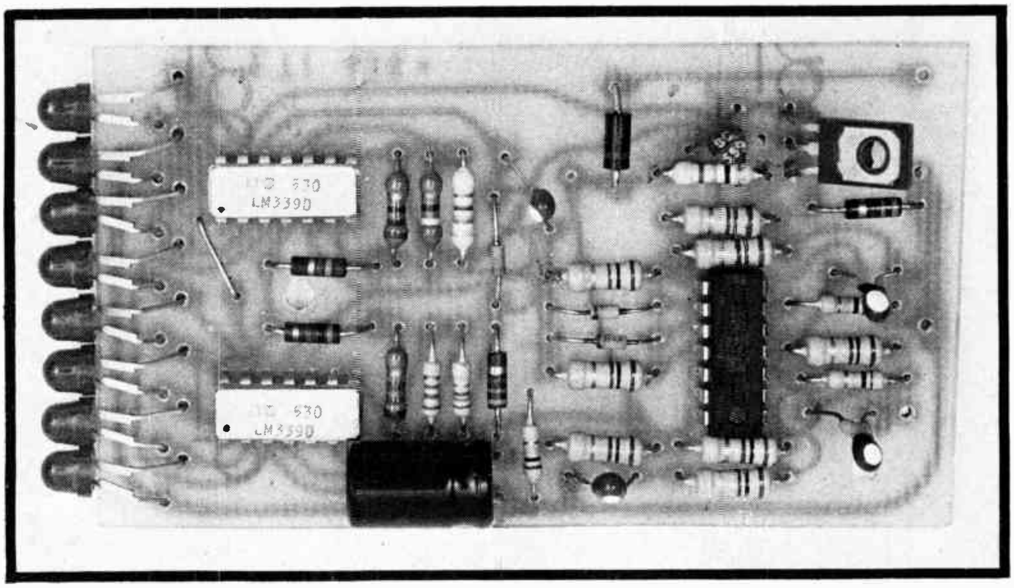
ADVERTISING
Sydney: Tel 33-4282
Melbourne: Tel 51-9836

Contents

Test Gear was printed in 1977 by Wilke and Company Limited, Melbourne.
Distributed by Gordon and Gotch.
(*Recommended and maximum price only).

an ELECTRONICS TODAY publication

AUDIO LEVEL METER



PROJECT 438

Peak and average audio levels are indicated by a bar of light.

HIGH-POWER amplifiers usually incorporate meters to indicate the output-power levels in each channel. These meters are often called VU meters but in most cases they resemble proper VU meters only in the way they are scaled.

A professional VU meter is the industry standard for measuring the levels of complex music waveforms. It has a scale marked from –20 to +3 VU (on a steady state signal VU correspond to dB) where '0' VU corresponds to a level of one milliwatt into 600 ohms. The meter has a carefully controlled time constant such that if a reference tone level is applied the pointer of the meter will take 0.3 seconds to reach 99% of the reference level, and will then overshoot by not more than 1.5% and not less than 1.0%.

The professional VU meter is thus an instrument that has been designed to give a reasonable compromise between indicating the fast peaks and the average levels of a complex music waveform.

In contrast the meters fitted to some amplifiers have scales calibrated in VU but usually relying on the inertia of the meter movement to provide meter averaging. Apart from this the 0 VU point corresponds to the rated power output of the amplifier – not to 1 mW into 600 ohms (equivalent to 75 mW in 8 ohms). Strictly speaking therefore such meters should be called level or power meters, not VU meters.

Even the best of such meters are not fast enough to indicate accurately the peak levels which occur in music and hence are useless for detecting the onset of amplifier clipping. This is vital as at clipping amplifier distortion rises rapidly.

One alternative is to use in addition to the level meter a clipping indicator that detects fast peaks which exceed a preset level. The ETI 417 OVER-LED project was such an instrument – it flashed an LED when a music transient exceeded clipping level.

The circuit described in this project is best described as a 'level meter'. It uses an array of LED diodes set to illuminate at successively higher increments in music level. With this type of display an estimate can quite easily be made of channel balance, and all transients, no matter how fast, are detected and indicated.

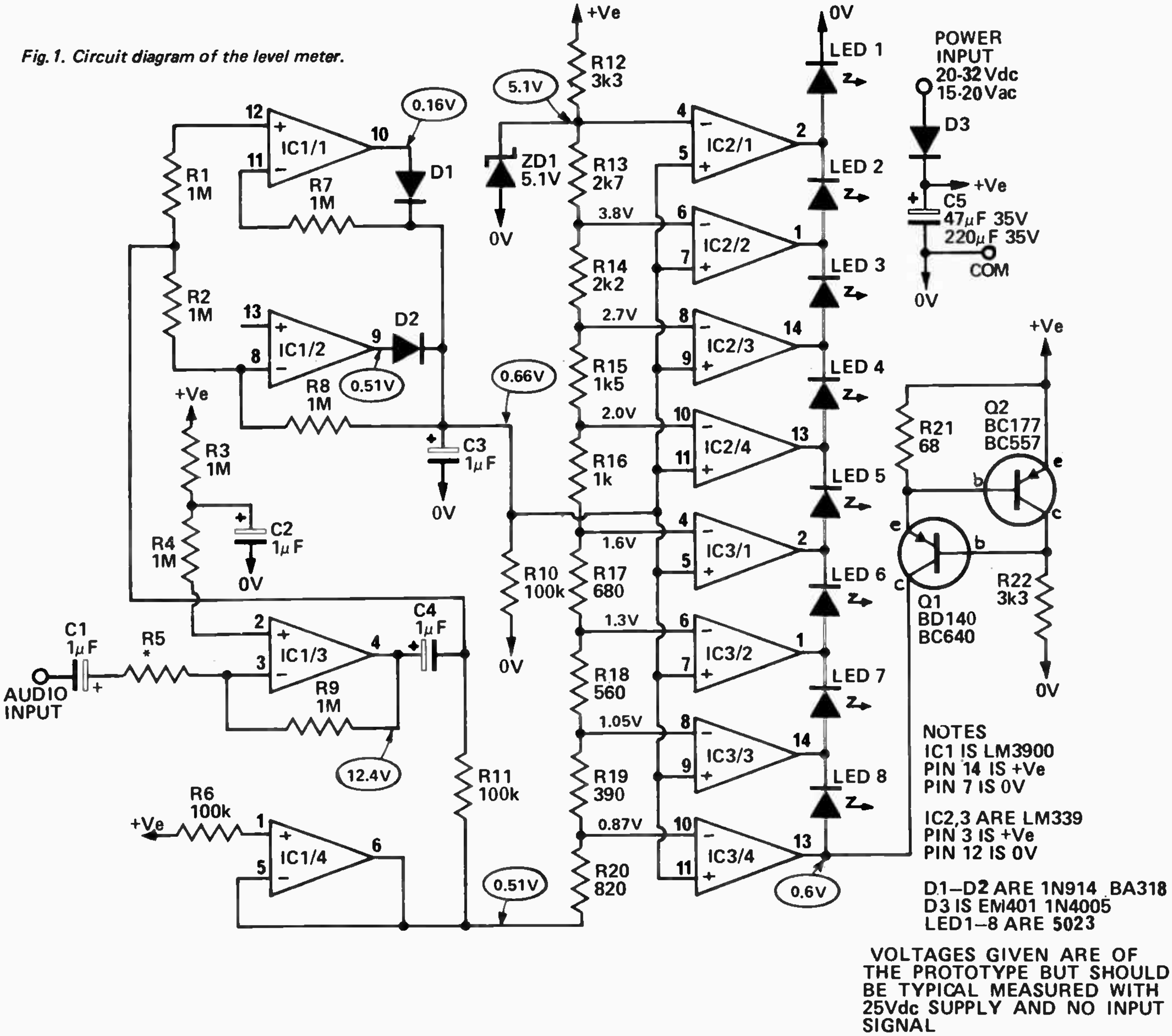
SPECIFICATION

Supply voltage	20 to 32 volts dc 15 to 20 volts dc
Supply current	16 mA dc approx
Input sensitivity (VU meter)	500 k/v
Indication	8 LEDs 3 dB apart
Attack time	1 ms
Release time	0.5 sec.

DESIGN FEATURES

The ETI 438 Level Meter can be arranged to indicate levels either in 'VU meter' format or in output power format. In the 'VU-meter' format the eight diodes light at 3 dB intervals from –18 to +3 VU where 0 VU corresponds to the nominal voltage required. Alternately as a power meter (remember that an amplifier cannot be driven beyond the clipping point) the top LED indicates maximum power and each lower LED indicates half the power of the one above it. The LEDs of the meter could thus be labelled,

Fig. 1. Circuit diagram of the level meter.



HOW IT WORKS – ETI 438

Although the circuitry of the level meter looks complicated the complete instrument only uses three ICs. These are an LM3900 which is a quad amplifier and two LM339s which are quad voltage comparators.

The input signal is amplified and buffered by IC1/3 to provide about 2.5 volts out at 0 VU input. The value of R5 is selected to give the sensitivity required for amplifiers of different power outputs. The gain of this amplifier is equal to the ratio of R9/R5.

A positive peak detector, IC1/1, and an inverting negative peak detector, IC1/2, give an output which represents the absolute peak level. Capacitor C3 and resistor R10 provide the peak hold and decay time. IC1/4 provides compensation for the 0.6 volt offsets of the LM3900 inputs.

The eight comparators are connected to a resistor divider chain the top of which is fed from a 5.1 volt supply which is stabilized by a zener. The resistor values are calculated to provide reference voltage steps at 3 dB intervals. The output of the detector is applied to all the non-inverting inputs of the comparators.

The LEDs are all connected in series and supplied with a constant current of 10 mA by the source consisting of Q1 and Q2. The outputs of the comparators are via open collector transistors which are "ON" if the input is lower than the reference voltage at the particular comparator input. With no input signal at all the comparators are all on thus shorting out all the LEDs so that none is on. As the input voltage rises the comparators turn off in sequence allowing the 10 mA to flow through the LEDs. Thus as the voltage increases a bar of light of increasing height is formed by the LEDs.

The current drawn from the power supply is about 16 mA and is independent of the number of LEDs which are on. Supply voltage is not critical and may be anywhere between 20 and 32 volts. Providing the supply is between these limits the unit will also be insensitive to supply ripple. When working from a dc supply a 47 microfarad filter capacitor is required but if an ac supply is used then the capacitor should be increased to 220 microfarad to minimize ripple. A single diode is used to both rectify the ac input and to prevent damage due to accidental reversed polarity if a dc supply is used.

PARTS LIST – ETI 438

R21 Resistor 68 ohm ½W 5%
R19 " 390 ohm ½W 5%
R18 " 560 ohm ½W 5%
R17 " 680 ohm ½W 5%
R20 " 820 ohm ½W 5%

R16 " 1k ½W 5%
R15 " 1k5 ½W 5%
R14 " 2k2 ½W 5%
R13 " 2k7 ½W 5%
R12,22 " 3k3 ½W 5%

R6,10,11 Resistor 100k ½W 5%
R1,2,7,8 " 1M ½W 5%
R3,4,9 " See Table 1 ½W 5%
R5 " See Table 1 ½W 5%

C1,2,3,4 Capacitor 1 µF 35V electro
*C5A " 47 µF 35V electro
*C5B " 220µF 35V electro

* use 47 µF for dc operation 220 µF for ac operation

IC1 Integrated Circuit LM 3900
IC2,3 Integrated Circuit LM 339

D1,2 Diode IN914, BA318 or similar
D3 " EM401, IN4005 or similar
ZD1 Zener diode 5.1 V 400 mW

Q1 Transistor BD 140, BC640
Q2 " BC177, BC 557

LED 1-8 L.E.D. 5023 or similar

PC board ETI 438

NOTE: Electronics Today is adopting the European standard method of showing component values — i.e. 1k5 = 1.5k, 2k7 = 2.7k, etc.

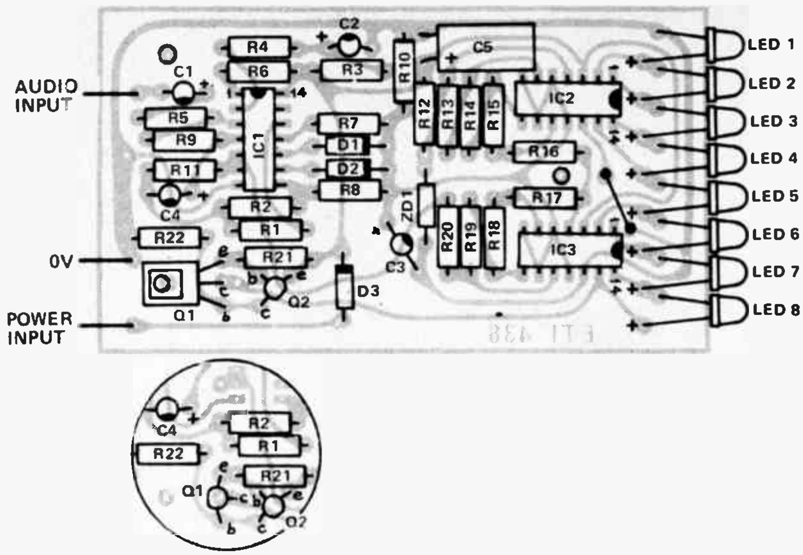


Fig.2. Component overlay using BD140 for Q1. Circled diagram shows use of alternative BC640

for example (for a 100 watt amplifier) 100, 50, 25, 12.5 watts etc.

The fast attack time of the meter (less than one millisecond) ensures that even very short transients are detected, whilst the relatively slow release time (0.5 seconds) provides a reasonably-accurate, average — level indication.

In most previous designs for such meters, discrete transistors were used to build level detectors. Temperature effects and variations in gain led to inaccuracies and to calibration difficulties. These problems have largely been overcome in the ETI 438 meter by using the LM339 IC which contains four accurate level detectors in one package. Additionally the LM339 also has an open-collector output stage which enables a constant current supply for the LEDs to be used. Thus the current and LED brightness are the same no matter how many LEDs are alight.

If required the interval between LEDs may be altered by changing the values of R13 to R20. Thus for example, a 6 dB interval could be used. Additionally the display could be extended to 12 or even 16 diodes by adding comparators and LEDs and by substituting another divider chain for R20 .(values would have to be calculated for the levels required). The positive inputs of the comparators would also be fed from C3 and R10.

A separate current source would be required as there is insufficient supply voltage available to light 16 LEDs in series. If the bottom LED in such a system indicates a level more than 30 dB down it may also be necessary to use a trimpot as the bottom resistor of the second divider chain to adjust for offsets etc.

The LM3900 is a quad differential amplifier which uses a current balancing technique at the input rather than the voltage balancing that is used with conventional operational amplifiers. Both the inputs "look" like the base-emitter junctions of normal transistors and both are at 0.6 volts with respect to ground. The currents into the two inputs must be equal if the output of the amplifier is to be in the linear region. In the case of IC1/3 the current into the positive input is set at about 12 microamps by R3 and R4. Current into the negative input is provided from the output by R9. If the current into the negative input is too low the output voltage will rise thus increasing the current into the negative input until balance is achieved. This self balancing ensures correct static biasing.

Gain is obtained by feeding a signal into R5 which adds or subtracts current into the negative input. For

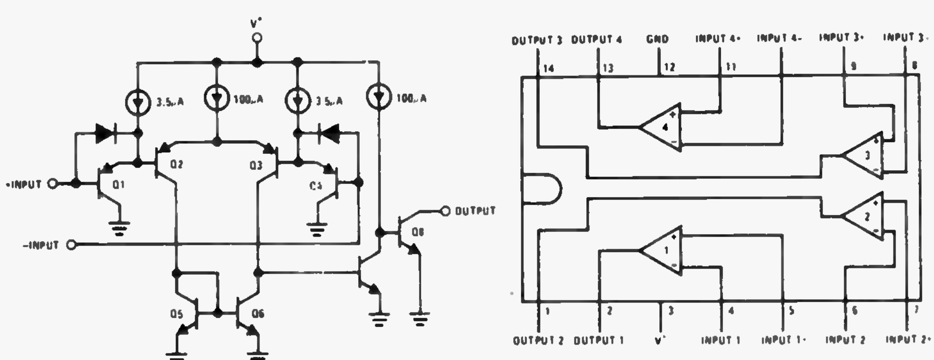


Fig.3. Internal circuitry and pin corrections of tha LM339 IC.

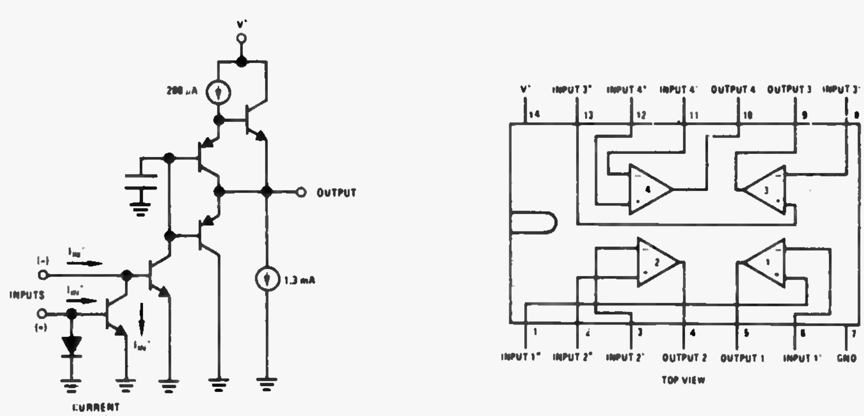


Fig.4. Internal circuitry and pin connections of the LM3900 IC.

AUDIO LEVEL METER

the amplifier to remain balanced there must be a corresponding shift in output voltage. The voltage gain is the ratio of R9 to R5.

SPECIFICATION LM3900

Maximum supply voltage	32 V
Supply current	6 mA typical
Voltage gain	2800 V/V typical
Input current range	1 µA – 1 mA
Current balance	0.9 – 1.1 at 200 µA
Bias current	30 nA typical
Output current capability	18 mA source typical. 1.3 mA sink typical

The LM339 is a quad voltage comparator where the output of each is an NPN transistor which has ar unterminated collector and its emitter connected to ground.

SPECIFICATION LM339

Maximum supply voltage	36 V
Supply current	0.8 mA typical
Voltage gain	200 000 V/V typical
Offsett voltage	2 mV typical
Bias current	25 nA typical
Response time	1.3 µS typical
Output sink current	16 mA typical
Input common-mode voltage range	0 to (V^+ – 2 volts)

CONSTRUCTION

The meter will most likely be mounted in an existing amplifier or piece of equipment and for this reason the board construction only is given.

Layout of components is non-critical but, as with any multiple IC device, construction is greatly simplified by using the printed-circuit board specified. The usual precautions with polarities of components, such as capacitors, diodes, ICs and transistors should be observed. Some care must be taken when mounting the LEDs in order to obtain even spacing and good alignment. The long lead of the LED should be inserted in the hole furthest from the edge of the board. Put a slight curvature in the leads so that the LEDs can be aligned against the edge of the board (see photo). Take care not to bend the leads too often or too close to the body of the LED as the leads break very easily.

CALIBRATION

Resistor R5 is selected from Table 1 and this will ensure a result within 10 percent of that required. Greater accuracy may be obtained by using a variable potentiometer in series with R5. To adjust this potentiometer inject a signal (around 1 kHz) equal to 0 VU (VU meter) or maximum power ($E = \sqrt{RP}$, e.g. 4 ohms and 100 watts, E = 20 volts) and adjust such that the second top LED (VU meter) or the top LED (power meter) just lights. ●

TABLE 1A – VU METER
FSD = +3 dB
R3, 4 and 9 are 1 megohm

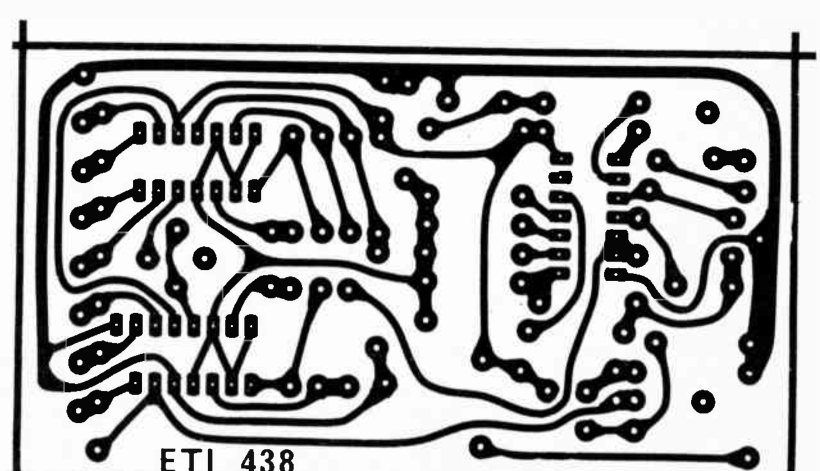
SENSITIVITY	VALUE OF R5*
50 mV	22 k
100 mV	47 k
250 mV	120 k
500 mV	220 k
1 V	470 k

**R5 = Sensitivity x 500 000 ohms.*

TABLE 1B – POWER METER
FSD = 0 dB
R3, 4 and 9 are 100 k

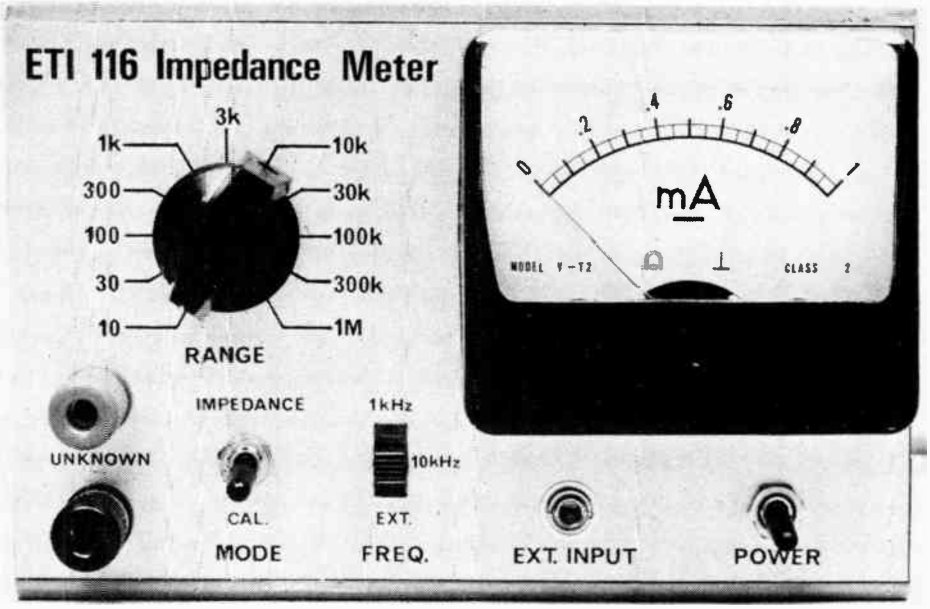
POWER OUTPUT IN WATTS	VALUE OF R5 4 Ohms	8 Ohms	16 Ohms
5	150 k	200 k	270 k
10	200 k	270 k	390 k
15	240 k	330 k	470 k
20	270 k	390 k	560 k
25	330 k	430 k	620 k
30	360 k	470 k	680 k
40	390 k	560 k	820 k
50	430 k	620 k	910 k
75	560 k	750 k	1.1 M
100	620 k	910 k	1.2 M
150	750 k	1.1 M	1.5 M
200	910 k	1.2 M	1.8 M
250	1 M	1.5 M	2 M

$R5\ k = 32\sqrt{PR}$ Where P = power in watts
R = speaker impedance in Ohms.



IMPEDANCE METER

Measure impedance directly with ETI's new impedance meter — checks capacitance and inductance too!



THIS IS an unusual project — in that we started out designing one thing and finished up developing another!

We had intended to design an RLC bridge which is a very useful instrument and perhaps the next most commonly used after the multimeter, signal generator and scope.

But whilst it is useful to be able to measure the value of an individual component, on many occasions we are more concerned with the magnitude of the impedance than we are with the actual value of C or L.

For example assume that we require to know how the impedance of a speaker varies with frequency. Due to the effects of the crossover network it will not be known whether the speaker is inductive or capacitive in the crossover region. Additionally a speaker goes capacitive below its natural resonant frequency. Hence the use of an RLC bridge to plot impedance would be very tedious indeed. We would have to determine whether the speaker was capacitive or inductive, measure the actual value and then calculate the impedance for each point to be plotted.

With the ETI impedance meter impedance can be read *directly* as a function of frequency as shown in Fig. 7.

This is just one example of the many possible applications. In addition the meter may be used to measure component values by simply referring to a reactance chart or doing a simple calculation as detailed below.

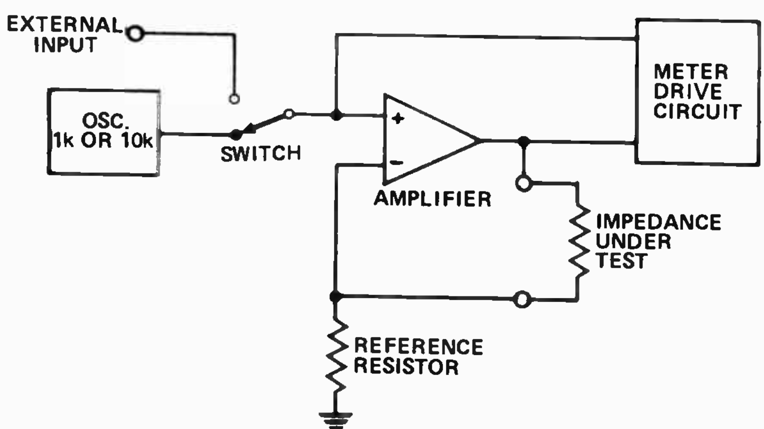
Other applications include measuring the impedances of microphones, filters, transformers and amplifier inputs etc. All can be measured as easily as one would measure a resistor using an ohmmeter. Simply by connecting the device to the input terminals of the meter and making the measurement as detailed in the "How To Use" section.

In most practical applications we require to know the magnitude of the impedance — we do not care whether the device is predominantly inductive or capacitive.

On the rare occasions that we do require to know reactance we can

Fig. 1. Block diagram of the impedance meter shows that it consists of an oscillator an amplifier and a meter circuit.

eti PROJECT 116



SPECIFICATION

Impedance measuring range	1Ω – 1 Meg Ω	
Frequency of test	20 Hz – 20 kHz 1 kHz or 10 kHz	external internal
Range of inductance	10μH – 1000 H 20μH – 100 H	external internal
Range of capacitance	100 pF – 1000μF 100 pF – 100μF	external internal
Accuracy ± 5%		

Voltage applied to unknown, max 1 V rms

When measuring items which are connected to the mains earth either the item, or the meter, must have the earth removed.

IMPEDANCE METER

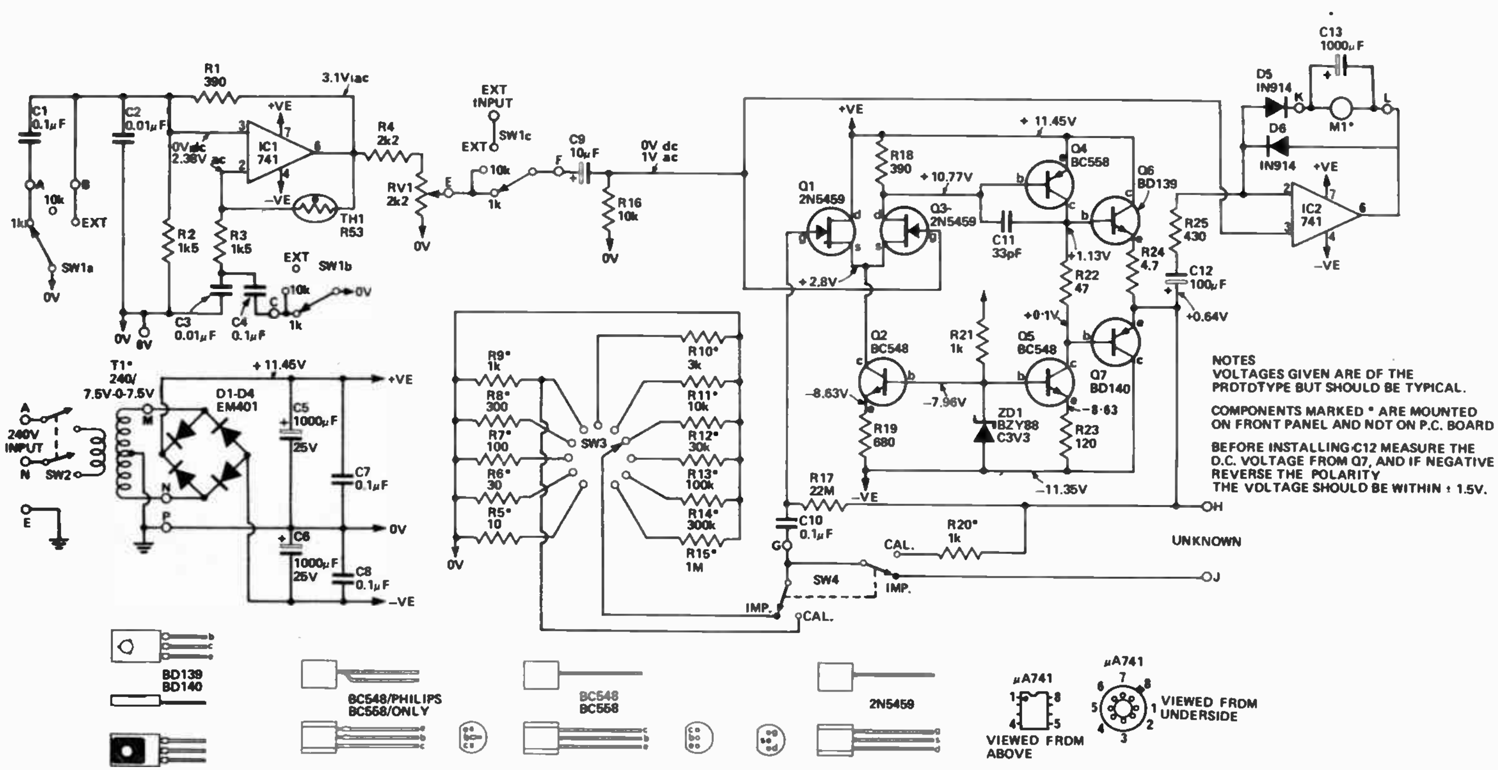


Fig. 2. Circuit diagram of the complete impedance meter.

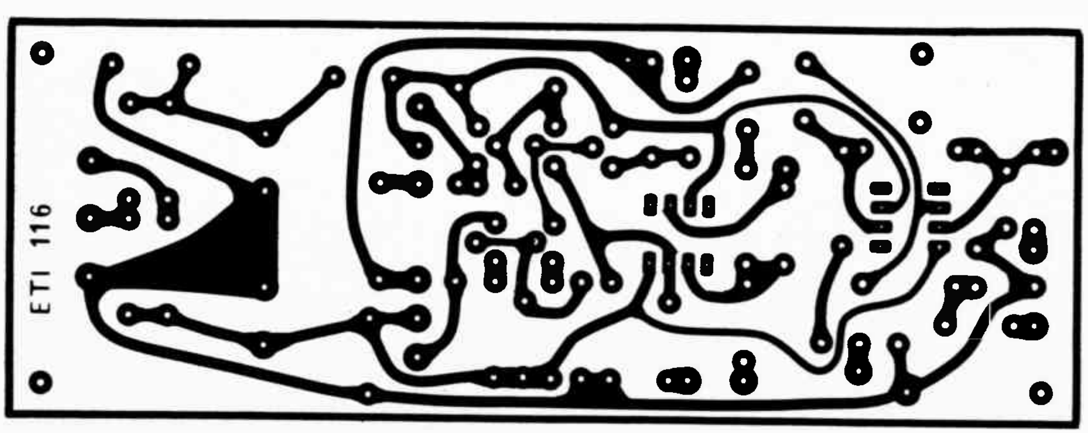


Fig. 3. Printed circuit board layout. Full size. 140 x 62 mm.

measure the dc resistance as well as the impedance and calculate from the formula

$$X = \sqrt{Z^2 - R^2}$$

where X = reactance inductive or capacitive at the frequency used

Z = magnitude of impedance (as measured on impedance meter)

R = dc resistance (as measured by an ohmmeter).

MEASURING CAPACITANCE

The value of an unknown capacitor can easily be determined by measuring the impedance and then using the reactance chart. Or, it may be calculated from the formula

$$C = \frac{1}{2\pi F X_c} \text{ (with capacitors } X_c = Z_c\text{)}$$

If the 10 kHz frequency is used this may be simplified to

$$C \text{ in microfarads} = \frac{16}{Z_c} \text{ (} Z_c \text{ in ohms)}$$

and if 1 kHz

$$C\mu F = \frac{160}{Z_c} \text{ (} Z_c \text{ in ohms)}$$

Since the meter can resolve the range 1 ohm to 1 megohm this implies a capacitance range of 16 pF to 160μF. But as explained elsewhere stray capacitance limits the lowest capacitance that can be resolved to about 100pF.

MEASURING INDUCTANCE

To determine the value of an unknown inductance the impedance is again measured and the value read off the reactance chart. Alternately the value may be calculated from

$$L = \frac{X_L}{2\pi F} \text{ high Q coils } X_L = Z_L$$

or $$L = \frac{\sqrt{Z_L{}^2 - R^2}}{2\pi F} \text{ (low Q coils)}$$

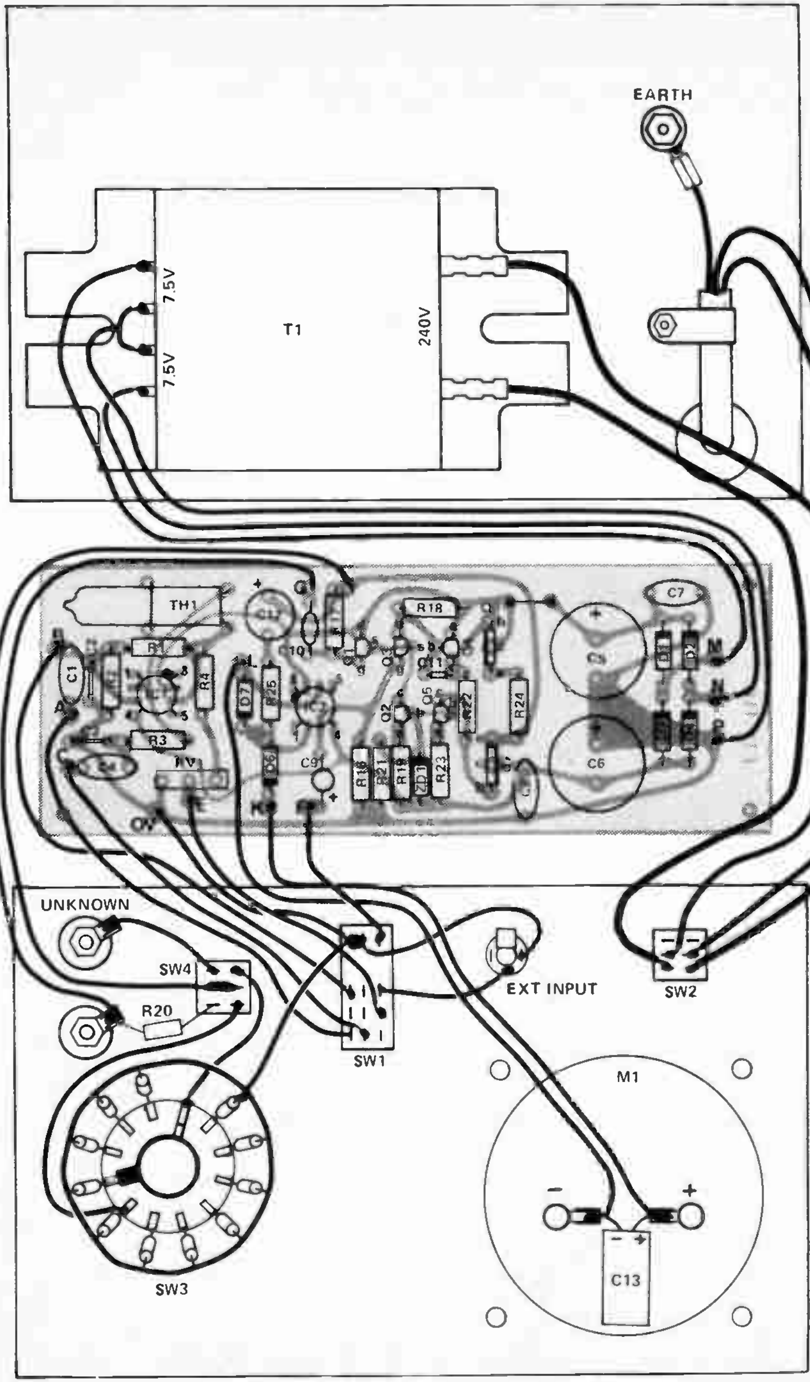


Fig. 4. Component overlay and wiring diagram for the impedance meter.

HOW IT WORKS ETI-116

The basic format of the impedance meter may be seen from the block diagram Fig.1. Firstly, we have an oscillator which may be switched to provide either 1 kHz or 10 kHz. Then we have a differential amplifier with a high input impedance, and lastly a meter drive circuit.

Either output of the oscillator, or an external frequency, as required, is passed to the non-inverting input of the amplifier. The amplifier gain is set by the ratio of the unknown impedance, Z, to the reference resistance, R. Due to feedback, the voltage across R is always equal to the input voltage and, as the amplifier requires no input current, the current through R must also flow through the unknown impedance, Z. The voltage across Z is therefore proportional to its impedance.

The meter circuit measures the output voltage by using the input voltage as a reference. Since the input voltage is equal to the voltage across R, we are effectively measuring the voltage across Z.

Refer now to the main circuit diagram Fig.2. The oscillator is of the Wein bridge type and uses a 741 IC as the amplifier and an R53 thermistor as the stabilizing element. The circuit oscillates at the frequency where the impedance of C2 and C3 is equal to the resistance of R2 and R3 respectively. Therefore, to change frequency, we simply change the values of C2 and C3. The output of the oscillator is attenuated by R4 and RV1 to approximately one volt.

The amplifier has a very high input impedance, can supply about 200 mA into a load, has an open-loop gain of 50 dB and can work into any load including a short circuit (unity gain).

An integrated circuit operational amplifier having the above characteristics (at reasonable cost) is not available, hence, a discrete seven transistor design was used. To obtain the high impedance input a pair of FETs, Q1 and Q3, used as a differential pair, operate with a constant current (4 mA) supplied by Q2. Transistor Q4 is supplied with a constant current of 22 mA by Q5, and Q4, in conjunction with the input pair, supplies the necessary overall gain. Transistors Q6 and Q7 buffer the output of Q4 and Q5 to provide the necessary current drive. The dc bias for the amplifier is provided by R17 such that an output voltage within ±1.5 volts of zero is always obtained.

The meter drive circuitry consists of a 741 IC with a meter, and half wave rectifier in series, connected in the feedback path. A second diode is used to prevent the IC being saturated on the opposite-polarity swing.

The current in the meter is half the current through R25 and, since this is proportional to the difference between input and output voltages of the amplifier, is proportional to the voltage across the unknown impedance. The meter scale is linear and the IC effectively compensates for the diode drop. Capacitor C3 provides the smoothing necessary when working at frequencies less than 40 Hz.

As previously stated the gain of the amplifier is set by the ratio of the unknown impedance 'Z' and the reference resistor 'R', and is equal to

$$\frac{Z + R}{R}$$ (where Z may be complex)

The value of R is switch selectable from 10 ohms to 1 megohm in eleven ranges. In the calibrate mode a 1 k resistor, R20, is substituted for the unknown impedance and the 1 k range selected. This provides a gain of two and thus with one volt in we have two volts out and hence 1 volt into the meter circuitry.

Thus, on calibrate, the output of the oscillator (or the external oscillator level) should be adjusted by RV1 to obtain full scale deflection on the meter. The calibrate position should also be selected before changing the unknown impedance, as an open circuit may damage the meter by driving it well beyond full scale.

IMPEDANCE METER

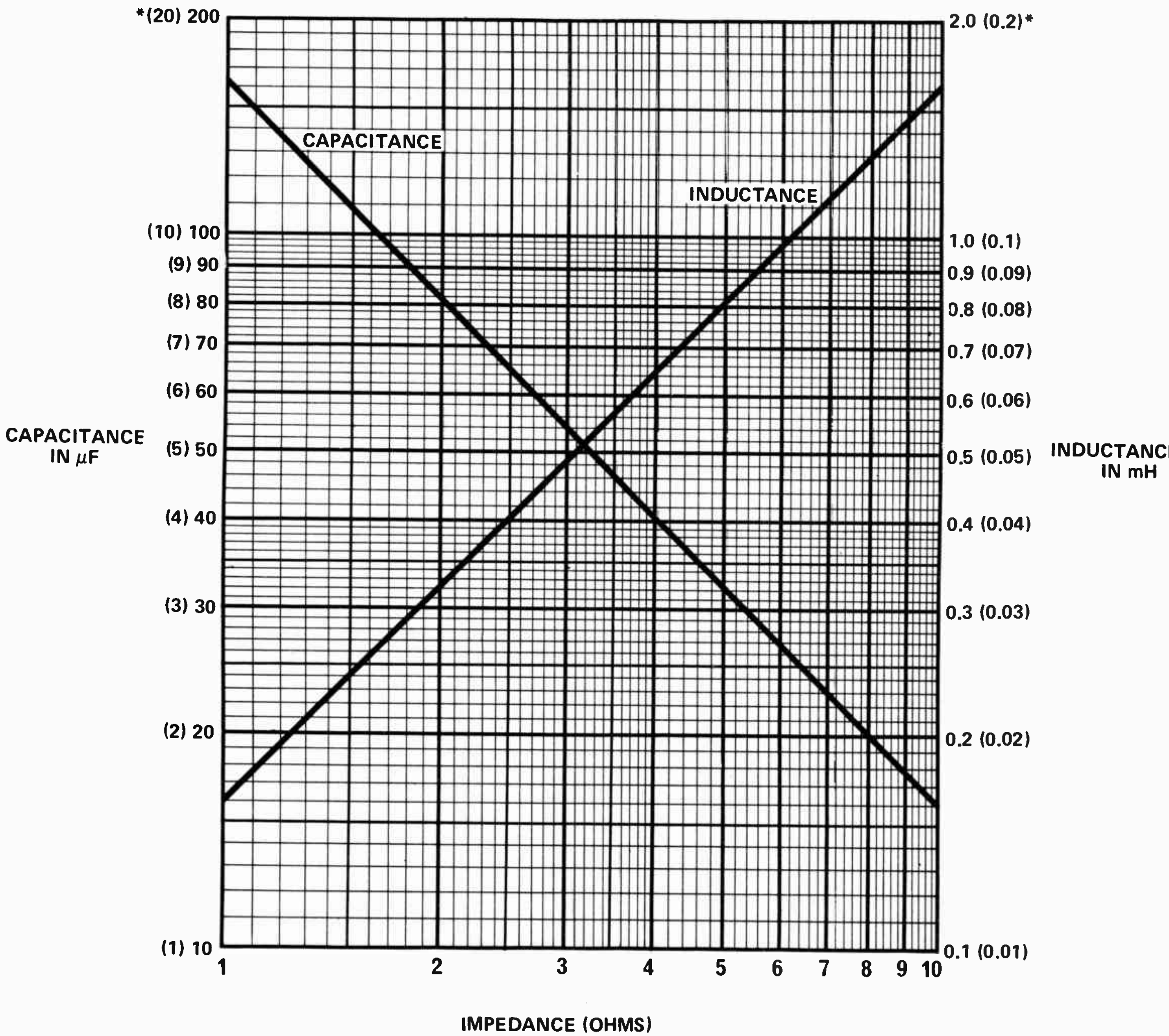
It should be borne in mind that we are determining impedances by using audio frequencies in this instrument hence components such as RF coils may well have a different impedance at RF frequencies (due to skin effect etc) than they do at audio. Additionally iron-cored coils have an inductance dependant upon the measuring frequency and upon dc current flowing. Hence such coils should be measured under conditions as close as possible to those when in circuit. Further the inductance value, as measured, will only be accurate on coils having a Q greater than 10.

If the dc resistance is greater than one tenth of the measured impedance the second formula should be used.

TURNS RATIO

To measure the turns ratio of an unknown transformer simply load the secondary with a value of resistance, R, which causes the impedance Zp (looking into the primary) to drop by 50% from the unloaded value. The turns ratio may then be calculated from

$$\frac{N_1}{N_2} = \sqrt{\frac{Z_p}{R}} \quad \text{(N = number of turns)}$$



FOR IMPEDANCES GREATER THAN 10Ω DIVIDE CAPACITANCE SCALE BY THE SCALING FACTOR AND MULTIPLY THE INDUCTANCE SCALES BY THIS FACTOR. e.g. A CAPACITOR WHOSE IMPEDANCE IS 6000 OHMS (SCALING FACTOR x 1000) AT 1 kHz VALUE IS 27/1000 = 0.027μF

*FIGURES IN BRACKETS ARE FOR 10 kHz

Fig. 5. Reactance chart for determining values of L or C from measured impedance at 1 kHz (10 kHz in brackets).

This calculation is based on the fact that an impedance in the secondary is transformed to an impedance in the primary that is proportional to the square of the turns ratio.

Many other applications can be devised for an impedance meter and the few mentioned here are indicative of the usefulness of such an instrument.

CONSTRUCTION

Any accepted construction method may be used but the use of a printed circuit board will greatly simplify the procedure.

Components should be assembled onto the printed circuit board, with the aid of the component overlay Fig 4, making sure that all polarized components are orientated correctly. Capacitor C12 should not be fitted initially as the required polarity must be determined as follows.

Temporarily connect the transformer to the otherwise completed board and switch on the power. Measure the voltage from the amplifier at point H. This should be within ± 1.5 volts of zero. If this voltage is negative reverse the polarity of C12 to that shown on the overlay. If the voltage is positive use the polarity shown. This variation of voltage at point H is due to differences in the FET transistors Q1 and Q3.

Attach wires to all output connections of the printed circuit board allowing sufficient length to terminate them in their respective positions. Instal the board in position using 12 mm long spacers and countersunk screws. Countersunk screws are necessary as they will be covered by the lid of the box. Install the power transformer and power lead, on the rear panel, together with the power-cord clamp and earth lug. Mount the slide switch to the front panel using countersunk screws.

Resistors R5 to R14 should be mounted on the rotary switch SW3 before mounting it on the front panel. If the 30, 300, 3k etc resistors are not available they may be replaced by a parallel combination; eg 30 ohms is obtained from 33 ohm and 330 ohms in parallel and 3 k from 3.3 k and 33 k in parallel.

The rest of the front panel components, except the meter, (for ease of wiring) should now be mounted together with the escutcheon. The wiring can now be completed and the meter installed and connected.

USING THE METER

The meter should be used in the following manner:—

1. Switch the cal/impedance switch to cal.

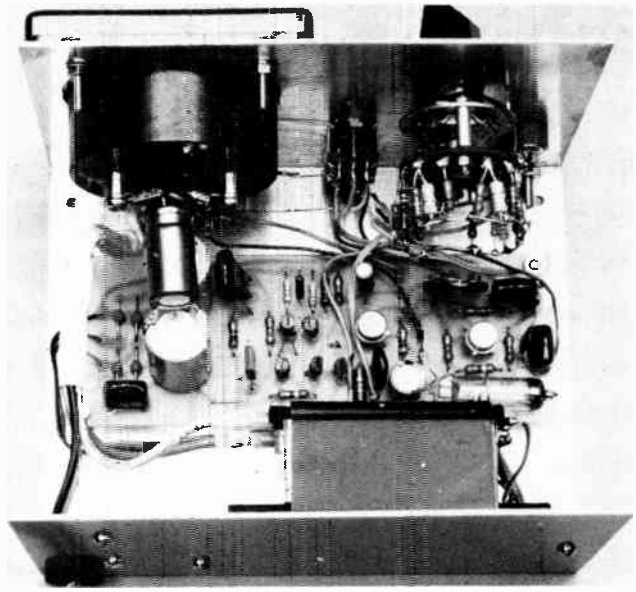


Fig. 4. Internal view of the meter shows how the board and other components are positioned.

2. Switch on power.
3. Select the required test frequency. The meter should read full scale, if not, adjust RV1.
4. If an external oscillator is used set the frequency and adjust oscillator output level to obtain full scale reading.
5. Connect the impedance to be measured.
6. Select the one megohm range.
7. Switch the cal/impedance switch to impedance.
8. Reduce the range, if necessary, to obtain a readable deflection. This reading is the required impedance; eg 0.6 on the 10 k range is an impedance of 6 k.
9. If desired the external frequency may be varied to obtain a plot of impedance versus frequency.
10. Switch back to 'Cal' before removing the impedance being measured.

TABLE 1

Error	Resistance (R2/R3)	Capacitor (C1,C4)	Capacitor (C2,C3)
1%	150k	0.001μF	100 pF
2%	68k	0.0022μF	220 pF
3%	47k	0.0033μF	330 pF
4%	39k	0.0039μF	390 pF
5%	27k	0.0056μF	560 pF
6%	22k	0.0068μF	680 pF
7%	18k	0.0082μF	820 pF
8%	18k	0.0082μF	820 pF
9%	15k	0.01μF	1000 pF
10%	13k	0.01μF	1000 pF

PARTS LIST – ETI 116

R24	Resistor	4.7 ohm ½W 5%
R5	"	10 " " "
R6	"	30 " " "
R22	"	47 " " "
R7	"	100 " " "
R23	"	120 " " "
R8	"	300 " " "
R1,18	"	390 " " "
R25	"	430 " " "
R19	"	680 " " "
R9,20,21	"	1k " "
R2,3	"	1k5 " "
R4	"	2k2 " "
R10	"	3k " "
R11,16	"	10k " "
R12	"	30k " "
R13	"	100k " "
R14	"	300k " "
R15	"	1M " "
R17	"	22M " 10%
RV1	Potentiometer	2k2 Trim type
TH1	Thermistor	type R53
C11	Capacitor	33pF ceramic
C2,3	"	0.01μF polyester
C1,4,7	"	0.1μF "
C8,10	"	0.1μF "
C9	"	10μF 16V electrolytic
C12	"	100μF 6.3V electrolytic
C13	"	1000μF 6.3V electrolytic
C5,6	"	1000μF 25V electrolytic
Q1,3	Transistor	2N5459 or similar
Q2,5	"	BC548 "
Q4	"	BC558 "
Q6	"	BD137, BD139
Q7	"	BD138, BD140
IC1, 2	Integrated Circuit	μA741C mini dip or T05
D1-D4	Diodes	EM401 or similar
D5, 6	"	IN914 "
ZD1	Zener Diode	BZY88 C3V3 or similar
T1	Transformer	240V/7.5-0-7.5V @ 1A PL 1.5-18/20VA, PL 15/20VA
M1	Meter	0-1ma FSD. 75 x 65 mm
SW1	Switch	three pole three position slide switch
SW2	"	DPDT 240V toggle switch
SW3	"	one pole eleven position rotary switch
SW4	"	DPDT toggle switch

PC board ETI-116, Metal box Dick Smith type LMB 564, Front panel, small phone socket, pointer knob, 3 core flex and plug, rubber grommet and cable clamp, four 12 mm long spacers, two terminals, nuts & bolts etc.

FREQUENCY CALIBRATION

The frequency should be within 10% of nominal if specified components are used. However, if a frequency meter is available the network can be trimmed to give the correct readings.

Measure both the 1 kHz and the 10 kHz and calculate the percentage errors. If either or both are low in frequency the resistors R2 and R3 can be paralleled with additional resistors to increase the frequency. Since this will affect both ranges choose the one with the greatest error. Table 1 gives the correct resistance to use.

Re-measure the frequencies. One frequency should now be right and the other high. The capacitors C1 and C4 or C2 and C3 can be paralleled by the appropriate capacitors as selected from Table 1.

LIMITATIONS

Due to stray capacitance, (about 15 pF) associated with the front panel terminals and the switches, the 1 megohm range is useful only up to about 4 kHz. The 300 k range is useful to about 10 kHz.

When measuring series LCR networks (where the impedance rises greatly off resonance) it is usually necessary to parallel a resistor across the network to stabilize it. Once at resonance, the resistor may be removed for the actual impedance measurement. The frequency can now be altered provided that the meter is not allowed to go off scale. The resistor used should be not more than 10 times the value of the network impedance at resonance. ●

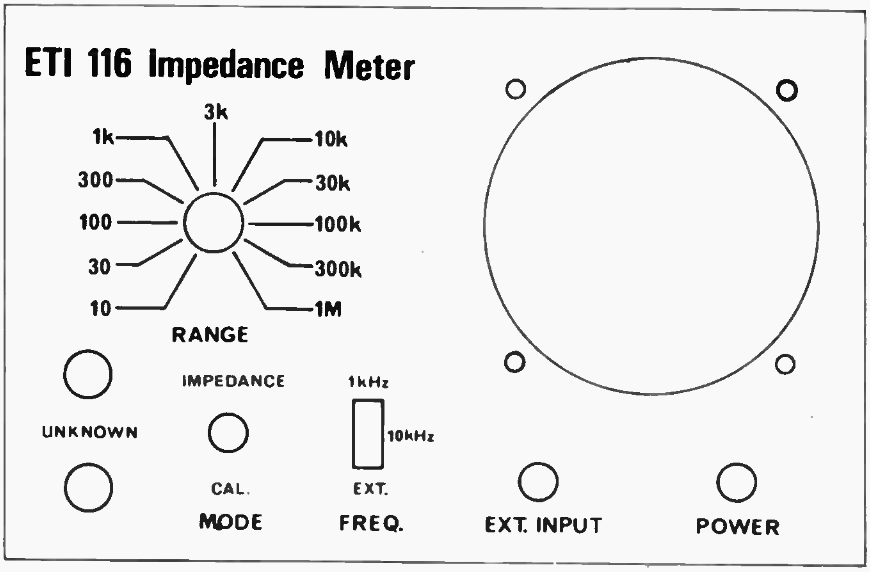


Fig. 6. Layout of front panel. Full size is 152 x 98 mm.

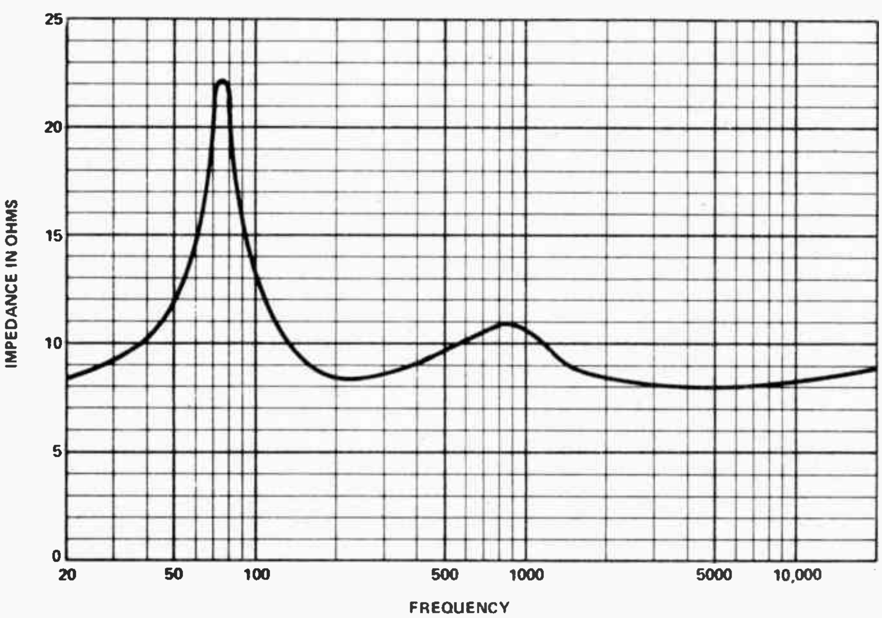


Fig. 7. Impedance-versus-frequency plot for a two-way speaker box. Note the combined speaker/box resonance is 75 Hz. The crossover frequency was 2 kHz. A plot such as this would be extremely difficult to generate using a conventional LCR bridge, but is very simply done using the ETI 116 impedance meter.

CIRCUIT DIAGRAM MARKINGS

ELECTRONICS Today International is adopting British Standard BS1852: 1967 for marking component values on circuit diagrams.

The values of components are given by figures but the decimal point is replaced by a multiplier symbol in accordance with a table of standard prefixes. This procedure greatly reduces the possibility of errors.

Examples

4 k7	equals	4.7 k ohm
47 k	"	47 k ohm
1 M 5	"	1.5M ohm
4n7	"	4.7 nF
6p8	"	6.8 pF

Where a multiplier is not needed, the symbol 'R' is inserted to signify ohms.

Example

4R7 equals 4.7 ohms

Note also that capacitors that were formerly specified as decimal fractions of microfarads (10^{-6}F) are now expressed in nanofarads (10^{-9}F).

Example

0.01μF = 10 nF

Abbreviation	Read as:	Multiplies unit by:
T	tera	10^{12}
G	giga	10^{9}
M	mega	10^{6}
k	kilo	10^{3}
h	hecto	10^{2}
da	deka	10
d	deci	10^{-1}
c	centi	10^{-2}
m	milli	10^{-3}
μ	micro	10^{-6}
n	nano	10^{-9}
p	pico	10^{-12}
f	femto	10^{-15}
a	atto	10^{-18}

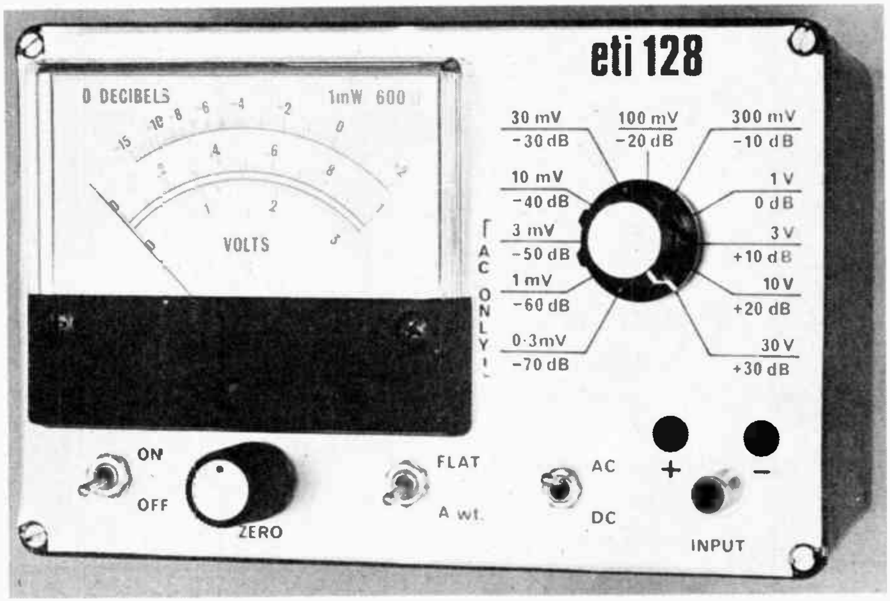
Standard prefixes. Multiplier symbols above 1000 are written with capital (upper case) letters, multipliers below 1000 do not use capitals (i.e. they are in lower case).

When spelled out in full, all multipliers start with a lower case letter (except when it is the first letter in a sentence).

Thus – 10 MW = 10 megawatts
– 10 mW = 10 milliwatts

PROJECT 128 AUDIO MILLIVOLTMETER

Sensitive instrument for 'A' weighted audio noise and signal measurements.



AN ACCURATE and sensitive ac voltmeter is needed for many audio equipment measurements.

Whilst for example, maximum power output is readily measurable with a conventional multimeter, more complex instrumentation is required for measuring noise output (a measurement required when checking signal/noise ratio).

Even signal levels as high as 100 mV, typical output of most pre-amplifiers, are not readily measured with accuracy on a conventional multimeter.

The ETI 128 Millivoltmeter is specifically designed for such measurements whilst also being useful as a general purpose ac/dc voltmeter. The lowest range, of 300 microvolts FSD, allows measurements to 80 dB below one volt, whilst other ranges allow measurements up to 30 volts ac or dc. These ranges cover most of the measurement requirements of audio work.

When measuring noise levels account must be taken of the non-linear characteristics of the ear. For this reason a network has been incorporated which tailors the meter response-versus-frequency to match the subjective response of the ear. Such a network is known as an 'A weighting network' and its use provides a measurement which is realistically related to what is heard. When measurements are made using this network the results must be quoted as being 'A weighted'. Typically this is done by quoting dBA rather than just plain dB.

CONSTRUCTION

The meter is a highly sensitive instrument and for this reason the constructional method given should be followed closely if noise and hum pickup are to be minimized.

A diecast box is used to house the meter as this provides excellent shielding against external signals. The front panel label is made from 'Scotchcal'. This is a specially prepared sheet of thin aluminium which is coated with a photo-sensitive emulsion on one side. The reverse side has a self-adhesive coating, protected by waxed paper, which is peeled off when the material is to be stuck down. As Scotchcal is only available in bulk, ETI is making available ready-to-use front panel labels made from this material. Should you require one of these labels send $4.00 and a stamped, self-addressed envelope (minimum size envelope 190 x 127 mm).

The meter used in the prototype was from Dick Smith Electronics. It measures 100 x 82 mm but requires to be rescaled. The scale as published on page 19 should be cut out and glued over the existing scale taking care not to let glue or dirt enter the meter movement. Any similar meter may be

SPECIFICATION

RANGES	
dc (FSD)	10, 30, 100, 300 mV, 1, 3, 30 V.
	auto-polarity, LED indication.
ac (FSD)	0.3, 1, 3, 10, 30, 100, 300 mV, 1, 3, 10, 30 V
	0 dB = 1 mW into 600 ohms (0.775 V)
	weighting curves, ac only, flat, 'A' weight
ACCURACY	± 3% nominal
MINIMUM READING	
Open circuit	–76 dB
Terminated 47 k	–85 dB
POWER SUPPLY	
Voltage	+6 and –6 volt (batteries)
Current	approximately 12.5 mA
Battery life	approx 100 hours (8 x 1015 cells)

AUDIO MILLIVOLTMETER

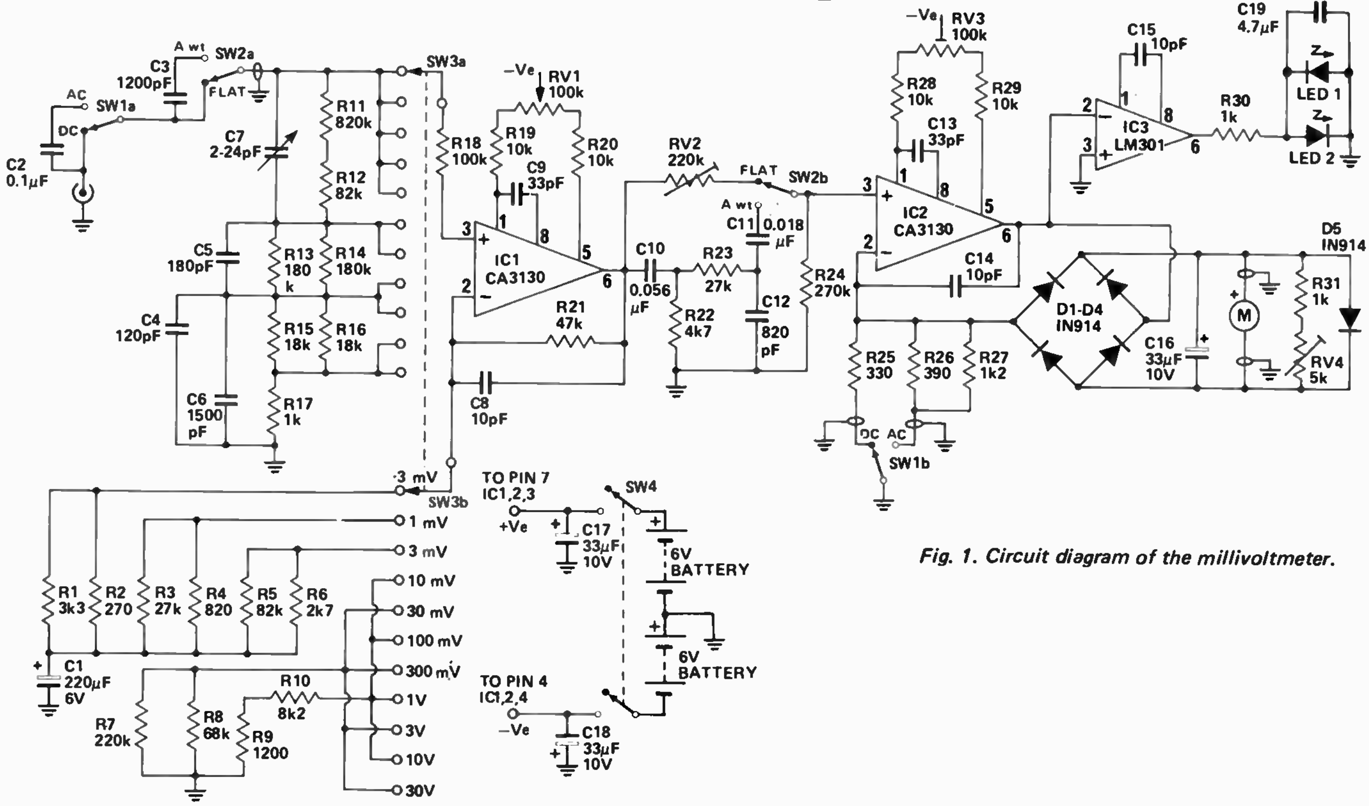


Fig. 1. Circuit diagram of the millivoltmeter.

HOW IT WORKS – ETI 128

The millivoltmeter may be separated into several sections in order to simplify the explanation of its mode of operation. These are:–

(a) Input attenuator.
(b) Input amplifier.
(c) 'A'-weight network.
(d) Meter drive circuitry.
(e) Polarity detector.

The input attenuator consists of resistors R11 to 17 and capacitors C4 to 7, and gives division ratios of 1, 10, 100 and 1000. The capacitors are required to ensure that the division remains accurate at high frequencies.

The input amplifier is a CA3130 operational amplifier where the gain is selected by SW3b. Gains of 190, 60, 19, 6 and 1.9 are available which together with the input divider ratios provide the 11 ranges required. The high gain ranges of 190, 60 and 19 are ac coupled, as the temperature stability of the CA3130 will not allow voltages of less than 10 mV dc to be used. The output of this amplifier is 60 mV when the meter is indicating full scale on any range. A potentiometer, RV1, is provided to adjust the offset voltage on the CA3130 and thus acts as a zero-set control. Since the offset voltage is affected by temperature this control is available externally.

When measuring noise in audio systems a weighting network is often used to give a measurement which is related to the non-linear response of the ear. The most commonly used weighting is known as 'A' weight and this facility is built into the meter. The 'A' weight curve is produced by a network that has a three-pole, high-pass filter and a single-pole, low-pass filter. The main section of this filter is formed by C10, C11, C12 and R22, 23, and R24 (two poles). The third pole is due to C3 and the one megohm combined resistance of R11 to R17. This later section prevents saturation of the input amplifier at low frequencies. Since this filter introduces some loss at 1 kHz, RV2 is incorporated to provide the same loss in the 'flat' mode.

The second IC acts as a meter amplifier. The input signal is rectified by the diode bridge D1 to D4 whilst the amplifier effectively compensates for the diode drops. A preset for offset adjustment, RV3, is provided for this IC. Calibration is performed by adjustment of the shunting resistance, R31 and RV4, across the meter. Due to the full-wave action of the rectifier the meter when on the dc ranges reads uni-directionally regardless of dc polarity. The output of IC2 will however either be at over one volt positive or one volt negative (voltage drops across the diodes) depending on whether the input voltage is positive or negative. This is compared by IC3 against zero volts and, depending on polarity, either LED 1 or LED 2 will be illuminated. With an ac input both LEDs will be on. These LEDs are therefore the polarity indicators. Capacitor C19 removes any high frequency components which could be coupled into the input, as the LEDs are located next to the input socket.

Due to the difference between the average and the RMS values of a sine-wave a slight change in gain is necessary in the ac mode and, this change is made by SW1b.

used as long as it has 100 microamp sensitivity.

The ac/dc and Flat/'A' weight switches are four-pole types although only the outer two poles are used. The centre two poles are earthed in order to reduce the capacitance between the two outer poles. Such precautions are necessary to prevent any possibility of instability on the most sensitive ranges. The metal bracket which supports the printed-circuit board also acts as a shield between the meter circuitry and the input stages.

Commence construction by assembling components to the printed-circuit board, making absolutely sure that all are mounted in the correct position and with the correct polarity. This should be carefully done – once the meter is

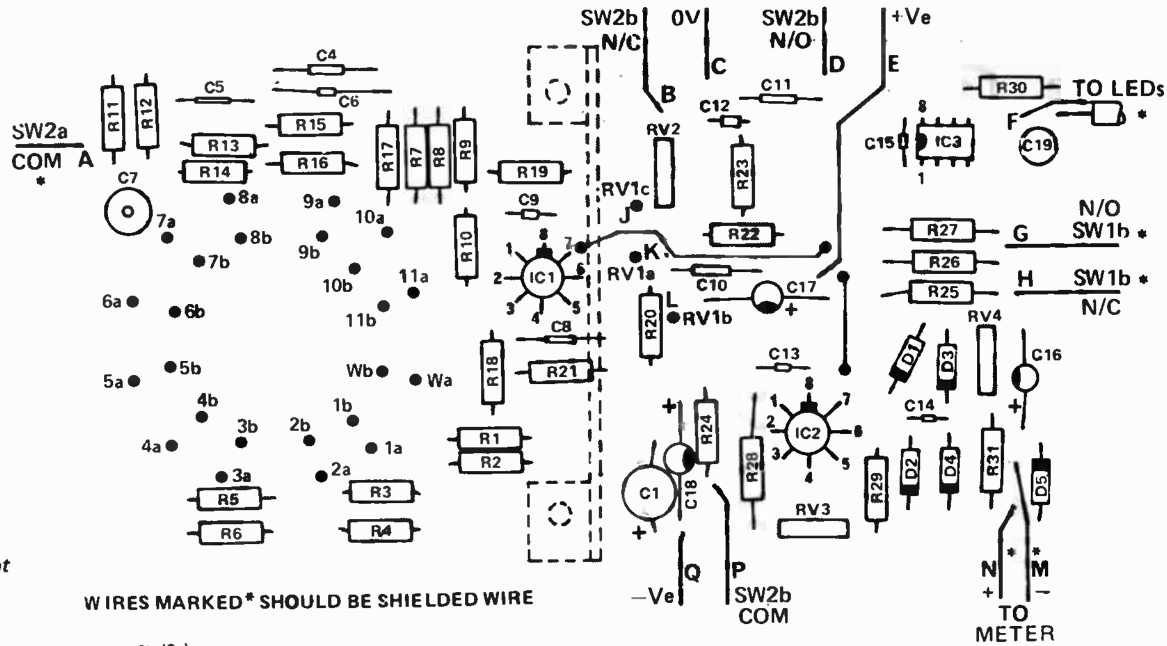


Fig. 2. Component overlay.

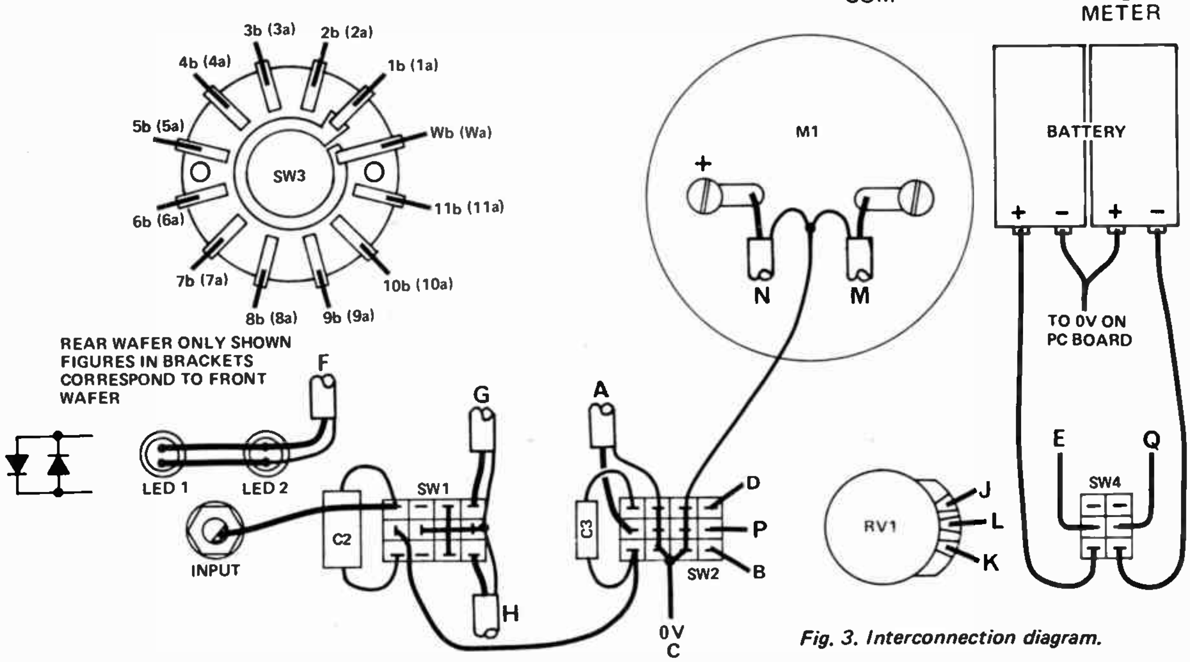


Fig. 3. Interconnection diagram.

fully assembled, it is very difficult to change components.

Assemble the front panel, fitting all switches with the exception of SW3, LEDs, potentiometer, input socket, meter, and the shield. The shield passes between the centre two contacts of the 'A'-weighted switch.

Solder a tinned copper lead to each of the 12 contacts on the rear wafer of switch SW3 (about 25 mm long). Feed these wires through the holes provided in the printed-circuit board (1b to 11b and Wb) making sure that the wiper contact on the switch goes to Wb and that the other wires are inserted in sequence. Do not solder as yet.

Assemble the printed-circuit board onto the shield and the rotary switch to the front panel. We used a 3 mm stack of washers to space the switch back from the front panel so the

PARTS LIST — ETI 128

R2	Resistor	270 ohm	2%	1/4W
R25	"	330 ohm	2%	1/4W
R26	"	390 ohm	2%	1/4W
R4	"	820 ohm	2%	1/4W
R17	"	1k	2%	1/4W
R6	"	2k7	2%	1/4W
R10	"	8k2	2%	1/4W
R15,16	"	18k	2%	1/4W
R21	"	47k	2%	1/4W
R8	"	68k	2%	1/4W
R13,R14	"	180k	2%	1/4W
R11	"	820k	2%	1/4W
R30,31	Resistor	1k	5%	1/4W
R9,27	"	1k2	5%	1/4W
R1	"	3k3	5%	1/4W
R22	"	4k7	5%	1/4W
R19,20	"	10k	5%	1/4W
R25,29	"	10k	5%	1/4W
R3,23	"	27k	5%	1/4W
R5,12	"	82k	5%	1/4W
R18	"	100k	5%	1/4W
R7	"	220k	5%	1/4W
R24	"	270k	5%	1/4W
RV1	Potentiometer	100k lin rotary		
RV2	"	220k Trim		
RV3	"	100k Trim		
RV4	"	5k Trim		
C7	Capacitor	2-24 pF Philips 2222 808 00006		
C8,14,15	"	10 pF	Ceramic	
C9,13	"	33 pF	Ceramic	
C4	"	120 pF	Ceramic	
C5	"	180 pF	Ceramic	
C12	"	820 pF	Ceramic	
C3	"	1200 pF	polyester	
C6	"	1500 pF	polyester	
C11	"	0.018µF	polyester	
C10	"	0.056µF	polyester	
C2	"	0.1µF	polyester	
C19	"	4.7µF	non polarised electro	
C16,17,18	"	33µF	10V electro	
C1	"	220µF	6V electro	
IC1,2	Integrated Circuit	CA3130		
IC3	"	LM301		
D1-D5	Diode	IN914, BA318 or similar		
LED 1,2	5023 or similar with panel mounting			
SW1,2	Toggle switch 4 pole 2 positions			
SW3	Rotary switch 2 pole 11 positions			
SW4	Toggle switch 2 pole 2 positions			
M1	Meter	100µA FSD * see text		
PC Board ETI 128				
Die cast Box 6357p				
Two knobs				
One RCA socket				
Eight AA size batteries				
Two- 4xAA size battery holders				
Shield to Fig. 7				

AUDIO MILLIVOLTMETER

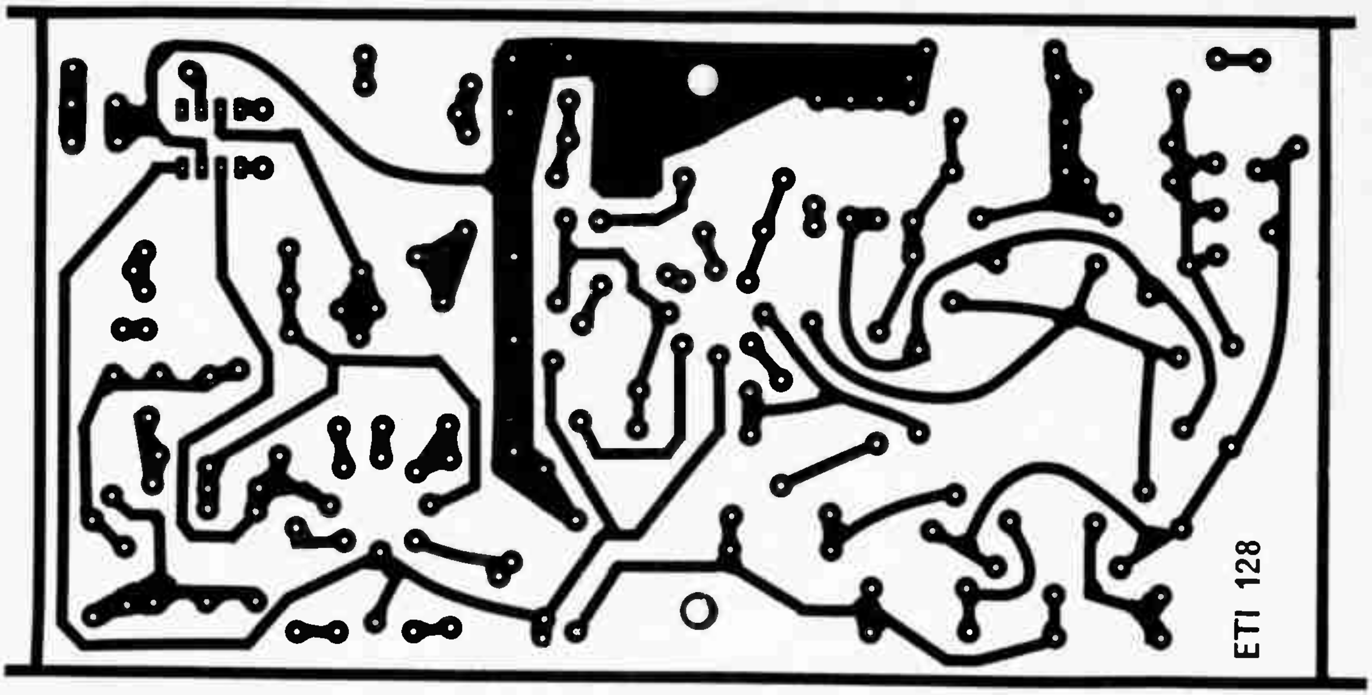


Fig. 5. Printed circuit layout. Full size 170 x 87 mm.

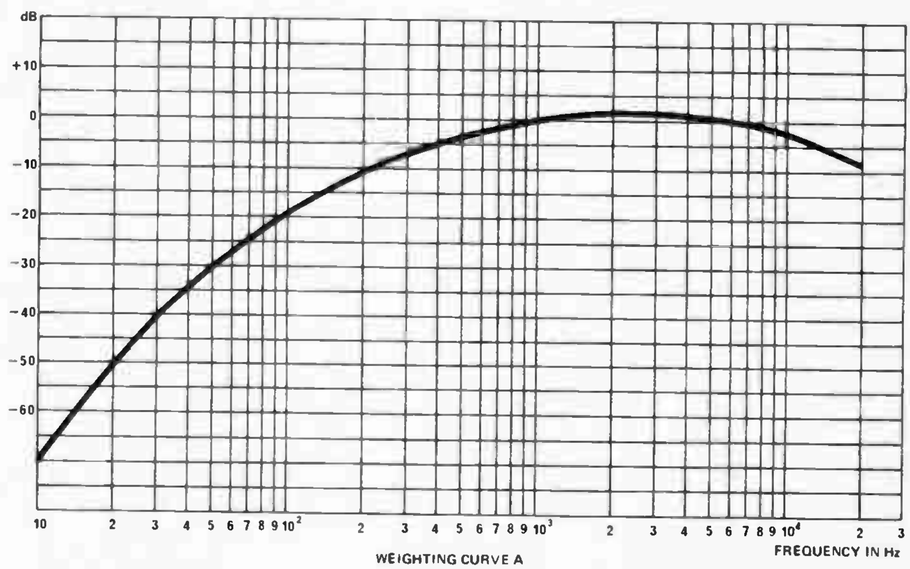


Fig. 4. Curve of 'A' weight response.

control knob would sit down closer to the front panel. Remove any slack in the tinned-copper wires, connecting the switch to the printed-circuit board and ,then solder them to the board. Now remove the printed-circuit board and switch assembly from the front panel. The switch will now be rigidly held onto the board, and the front wafer can now be wired to the board via further tinned-copper links. Make sure that none of these wires is touching.

Add leads to the printed-circuit in the locations shown on the overlay and reassemble the board and switch assembly to the front panel. The components on the front may now be connected to the board by these leads which should be kept as short as possible without placing undue strain on the wires. The only exception to this rule is the wire from SW1a to SW2a which should be kept reasonably well clear of the second pole of SW10. This is best done by running the lead down the front panel along the bottom and then back up to SW2a. Shielded wire should be used where designated on the overlay and wiring diagrams, and this should preferably be of the low capacitance variety.

The LEDs are connected in parallel but in anti-phase, the actual polarities may be determined later if necessary during the calibration procedure.

CALIBRATION

Before commencing calibration, check that the meter performs as it should on all ranges by applying known voltages and checking that a deflection of roughly corresponding magnitude is obtained. Also check that the 'A'-weighted switch appears to work as it should.

1. Short the input, select the 3 mV range and switch on.

2. Allow about 5 minutes for the instrument to stabilize thermally and then adjust RV3 to zero the meter.

3. Select the 10 mV range, dc, and 'flat', and adjust the front panel control RV1 to zero the meter.

4. Remove the short from the input, select the 300 mV range and apply an input having a frequency of less than 500 Hz and a level which gives a convenient indication, eg 0 dB. Change the frequency to somewhere between 10 kHz and 50 kHz making sure that the input level is the same in both cases, and adjust capacitor C7 so that the meter reads the same in both cases.

5. Apply an ac input signal and switch between ac and dc. The reading

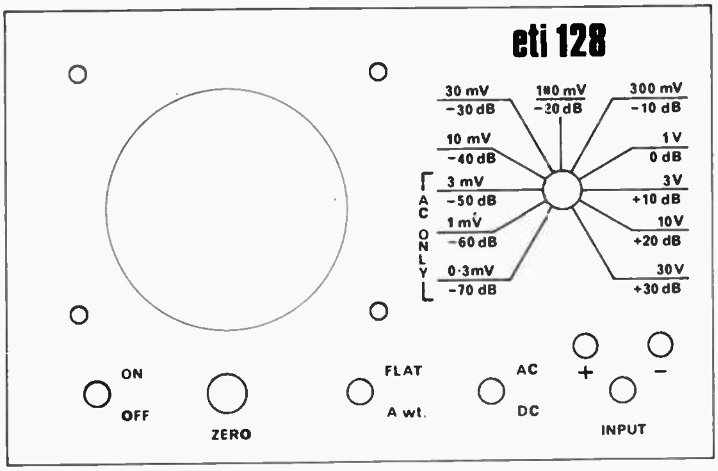


Fig. 6. Front panel artwork. Full size 189 x 121 mm.

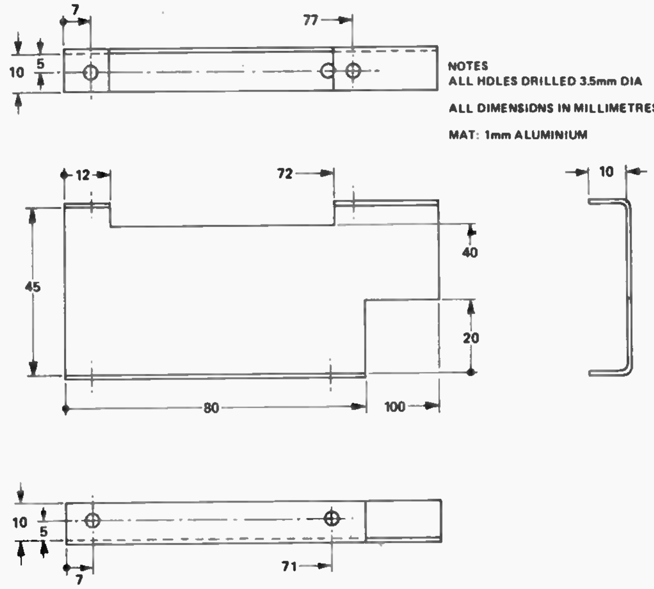


Fig. 7. Details of shield-support bracket.

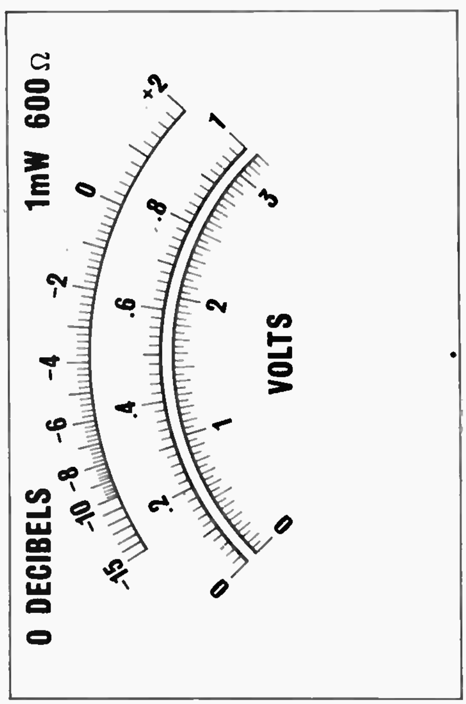
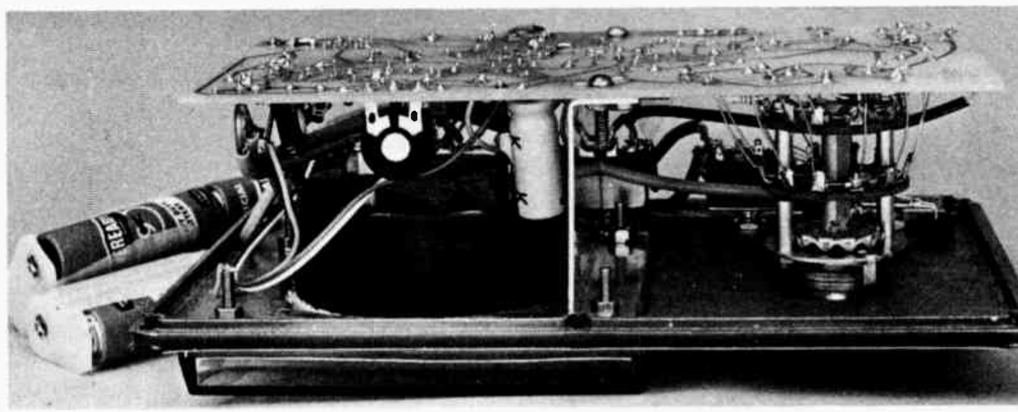


Fig. 8. Artwork for meter (shown full size).



This internal view of the meter shows on the right, how the range switch is wired to the printed-circuit board. Note also the shield.

Note how the shield passes between the earthed, centre contacts of the 'A' weight switch.

on ac should be about 10% higher than on dc. If it is 10% lower the leads to switch SW1b should be reversed.

6. In the ac mode select 'A'-weight and apply a 1 kHz signal of sufficient level to obtain a 0 dB indication on the 1 volt range. Vary the frequency over the whole audio range and check that the response as shown in Fig. 4 is obtained.

7. Go back to 1 kHz and check that zero dB is indicated in the 'A'-weight mode. Now select 'flat' and adjust RV2 to obtain the same reading.

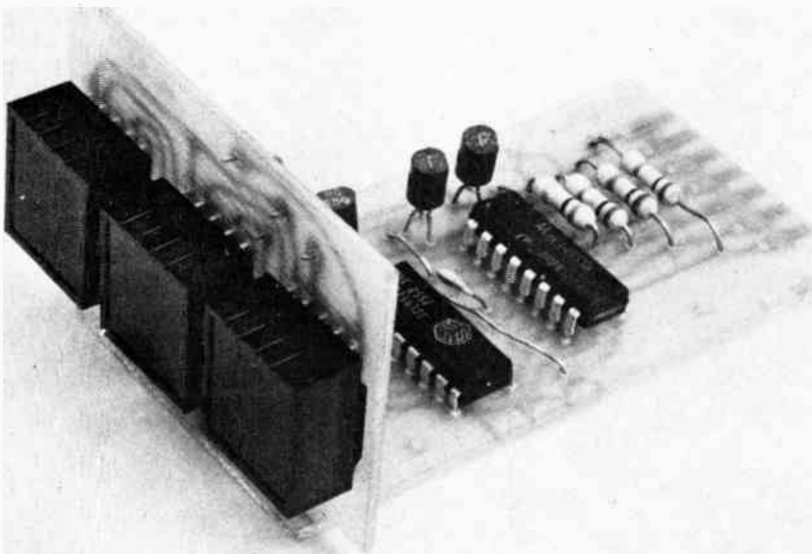
8. Apply an accurately known voltage with the instrument set to the flat and ac modes and adjust RV4 to give the correct reading.

9. Apply a dc input of known polarity and check that the correct LED illuminates. If not, reverse the leads to the LEDs.

This completes the calibration and the instrument should now give accurate readings on all ranges and at all frequencies within the specified range. ●

DIGITAL DISPLAY

This updated version uses bigger and more readily available seven-segment displays.



THE ETI 533 DIGITAL DISPLAY WAS originally published in the July 1975 issue of Electronics Today and has been very popular as a general purpose display module. We are republishing this project for two reasons. Firstly, many people have asked for the details of how to use a larger display. And secondly, the type of display originally specified is now quite difficult to obtain.

We therefore redesigned the display section of the module to accept the Fairchild FND 500 type of display which has 13mm high characters and is quite readily available. Of course the previous design is still quite valid and when the DL 704 and NSN74 type of displays are available they can be used as specified in the original article.

Only the small display board has been changed and it may be fitted to a previously built logic board if necessary. The new display board is, however, wider than the old board and the appropriate room must be available – especially if two modules are to be mounted side by side.

Digital Display

All digital intruments have a common assembly in the display system. Again, almost all instruments require decade counters, stores and decoder-drivers for the display

Normal systems using TTL logic generally have a 7490, a 7475 and a 7447 to drive each 7 segment LED display digit. Hence to build a three-digit display nine ICs are required in addition to three display ICs.

However, complex logic functions are available in CMOS which allow a 3 digit display to be built using only two ICs – and such ICs are available at reasonable cost. One of the devices is a three-digit, decade-counter, store and the second is a three-digit decoder driver. Thus three digit displays can be built which have the following advantages:

1. Small size
2. Low power consumption (120 mA compared to 600 mA in TTL)
3. Wide power supply range (5-15V unregulated).
4. Cost about same as TTL but rapidly decreasing.
5. Immunity to noise is greatly improved.

Disadvantages:

Maximum frequency about 1 MHz compared to 15 MHz for TTL.

Construction

Construction is quite straightforward especially if the printed circuit boards described are used. Since both ICs are CMOS devices, they can be easily damaged by static charges. Hence they should be handled as little as possible, fitted to the board after all other components and soldered using a minimum of heat.

Starting with the display board use a short length of thin insulated wire to form the link which goes between the pins of the FND500 then add the displays.

Next fix lengths of tinned copper wire to each of the four holes on the bottom of the display board. Allow approximately 10 mm of wire to extend from either end of the holes. Bend each wire so that they lie parallel and flush to the surfaces of the display board – do not solder as yet.

On the main printed-circuit board (533A) fit resistors R1,2,3,4, and R9 capacitors C1 and C2 and the three links. now mate the display board to the main board by inserting each of the previously bent wires into its corresponding pair of holes on the main board.

Apply gentle force to the display board until its bottom edge fits snugly against the main board. Solder each of the wires to both the display and main boards to make a sound electrical and mechanical support for the display.

Fit R5,6,7,8, 10, and 11 and, taking care to orientate them correctly, fit Q1, 2 and 3 and IC1 and IC2.

Lastly check that all components have been correctly fitted and all solder joints are good. If possible get someone else to check your final circuit as a final safeguard.

PARTS LIST – ETI 533

R1, 2, 3, 4	Resistor 100 k
R5-11	Resistor see text.
C1	Capacitor 1n0 Polyester
C2	Capacitor 10 nPolyester
IC1	Integrated Circuit MC 14553 (CMOS)
IC2	Integrated Circuit 14511 or 4511 (CMOS)
Q1,2,3	Transistor BC 558 or similar
DISPLAYS	FND500 or similar. Three required.
PC boards	ETI 533A and ETI 533C

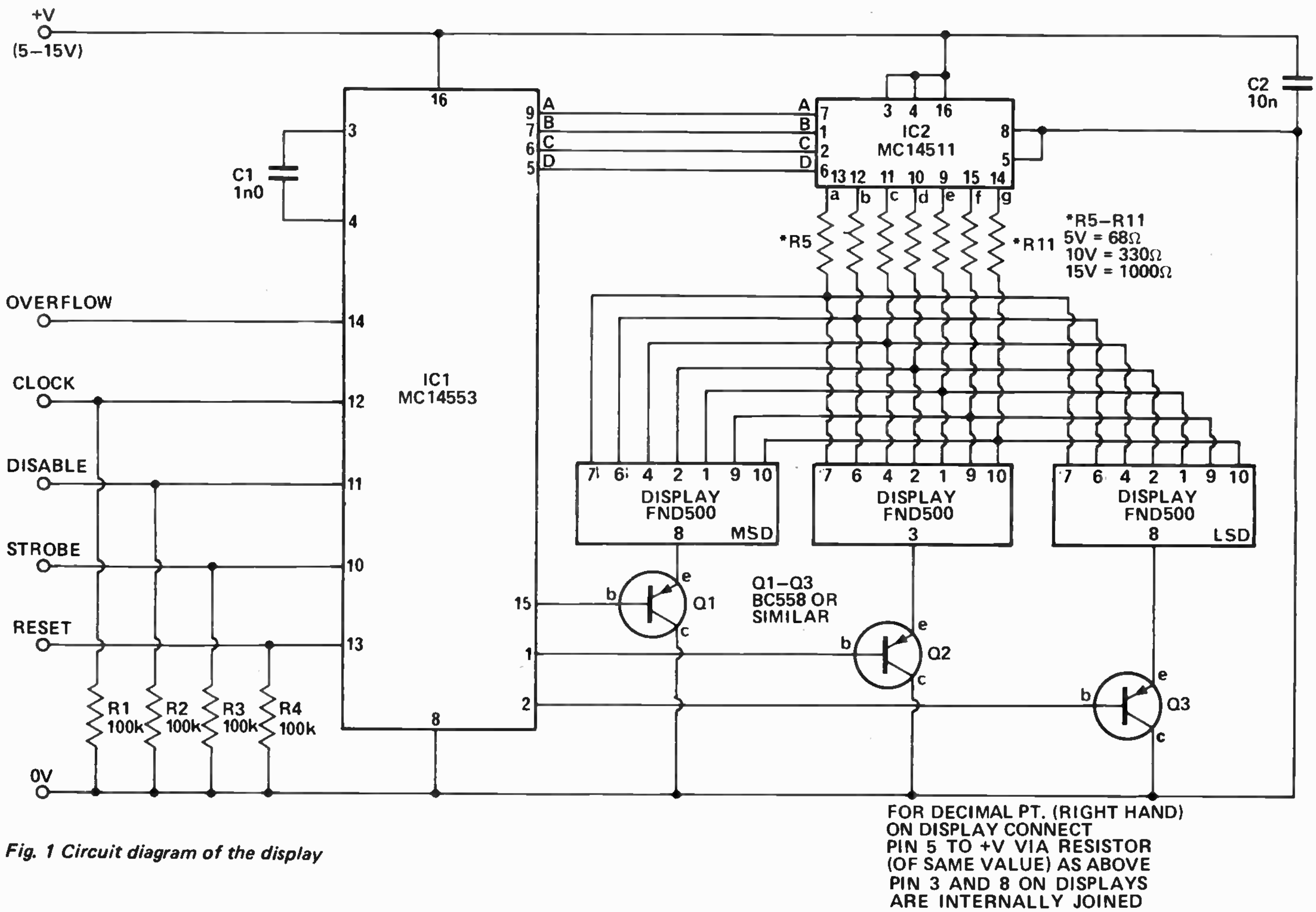


Fig. 1 Circuit diagram of the display

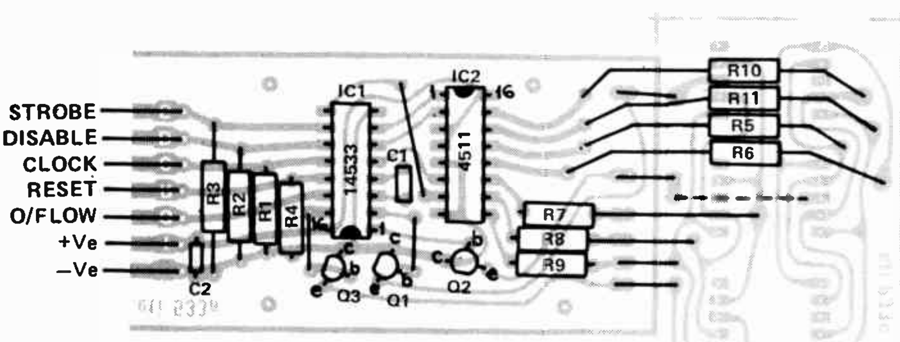


Fig. 2 Component overlay of the logic board.

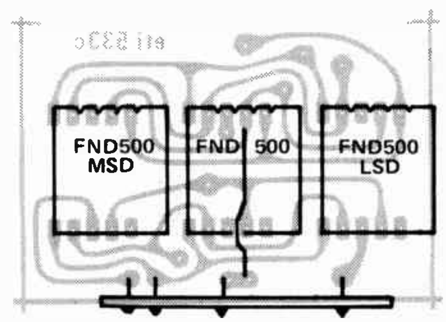


Fig. 3 Component overlay of the display board.

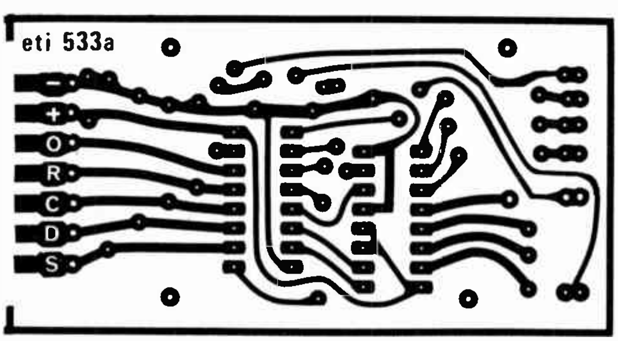


Fig. 4 Printed circuit layout of the logic board. Full size 78mm x 42mm

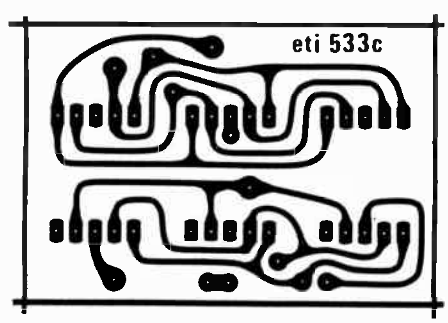


Fig. 5 Printed circuit layout of the display board. Full size 53mm x 36mm.

How it works — ETI 533

The heart of the counter is IC1. This LSI CMOS chip contains a three-digit decade counter, three sets of latches, and a three-digit multiplexer with an internal oscillator. CI is used to set the frequency of this oscillator.

The four input lines to IC1 are used to control the operation of the counter. Since IC1 is a CMOS device R1-4 are used to protect its inputs. Pulses to be counted are fed to the clock input and on a negative transition the value in the counter is increased by one. The schmitt-trigger action of the clock input allows any value of transition time of the input pulse.

The counter operates when there is a low at the disable input (pin 11).

To ensure accurate counting the clock should be low when the disable is brought from a high to a low level. The strobe input controls the loading of the latch. When it is low, data can be accepted for display. However, the strobe input has no effect on the counter, i.e., even with the strobe input high, the counter can still be incrementing.

A high on the reset input clears the counters (to a 000 state) and stops the internal multiplexing oscillation of IC1, and so blanks the display. Returning the reset to a low allows the internal oscillator to start up and all zeros to be displayed.

All inputs are standard CMOS inputs and require a minimum voltage change of from 30% to 70% of supply volts. However it is recommended that a swing from 0V to supply be used to give a satisfactory noise margin. Each input can be considered to be 100k shunted by 8-10 pF. Voltage swing below 0V and above supply are also to be avoided.

The one output available is the overflow (pin 14). This goes positive when the counter is 999 and the clock input is high. When the clock input goes low and advances the counter to all zeros the overflow goes low. This is a CMOS output and will swing between supply rails. It is not recommended that the overflow output be used to drive TTL directly. It can be used to drive a second 533 display to give a 6 digit readout if required.

The internal multiplexer of IC1 allows considerable saving in parts and board space. It allows a three-digit number to be transmitted over a single set of lines and it does this by leaving each digit on the output lines for a short length of time, before replacing it with the next digit. Then after presenting all the digits once, it starts over again and repeats the operation.

IC2 is a CMOS, latch, BCD to seven-segment decoder and driver, however for this application the latch is not used. It converts the 4-bit BCD code into the seven-line code necessary to drive the display segments. It also provides sufficient current to drive the display. If it is required that the display be blanked to save power the track to pin 4 on IC2 should be cut and pin 4 switched to either +V or 0V. If 0 the display will be blanked.

Although IC2 is coupled to all three displays, only one display is lit up at any one time. Thus when it is the turn of the most significant digit to be displayed IC1 presents that number to IC2 which decodes the number and presents it to the three displays, but only Q1 is turned on, so only the left most display lights.

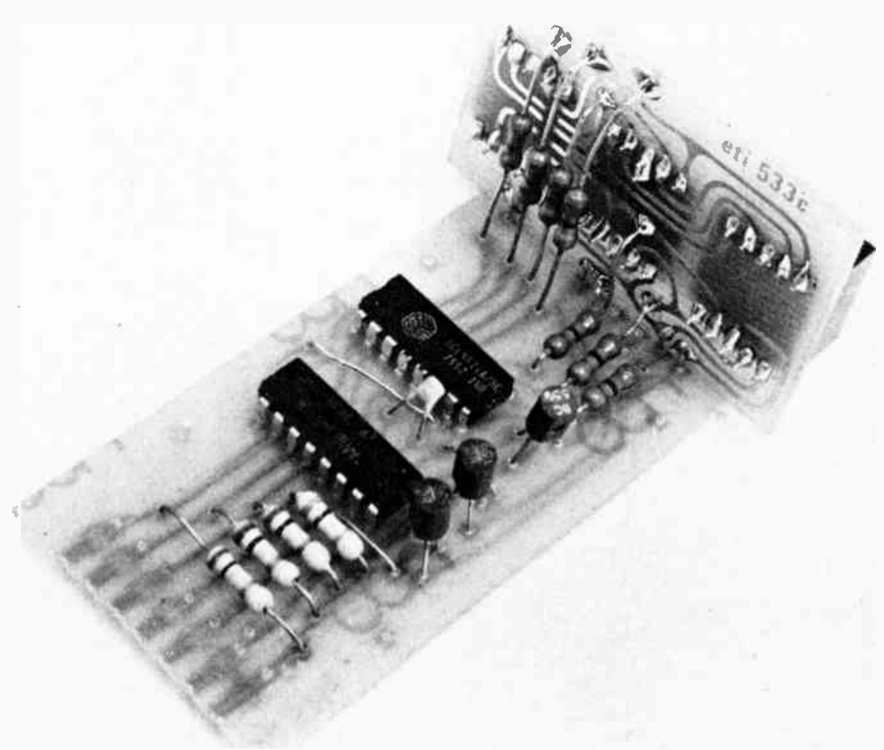
Note that IC1 controls which number is being presented and which transistor is turned on. This is called multiplexing. The switching between displays occurs so quickly that to our eyes the light appears continuous.

Resistors R5 to R11 limit the current to each LED display to a safe level. Three different values have been given for these resistors. Select the value appropriate to the supply voltage that you decide to use, 68 ohms for 5 V, 330 ohms for 10 V and 1k for 15 V. Transistors Q1, Q2 and Q3 also act as current amps since only a limited amount of current can be taken from IC1.

Any voltage from 5V to 15 V can be used to supply the counter, however, a supply voltage of 15 V allows the counter to operate at its highest speed.

USING THE DISPLAY

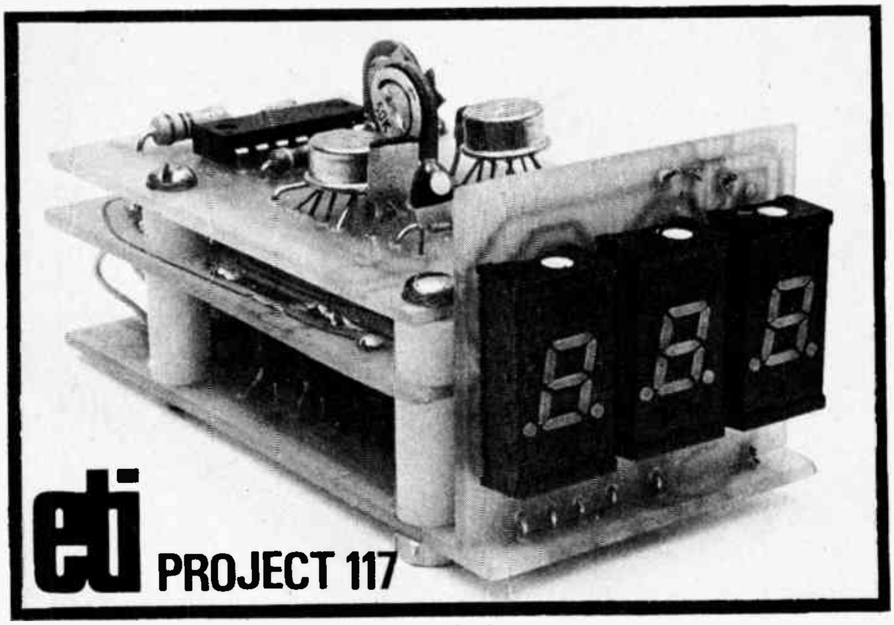
Power Supply	5 to 15 Vdc (do not exceed 16 V)
Clock Input	Counter is advanced on the negative edge of the clock input. Speed of transition is not important.
Disable Input	Must be low to enable clock to advance counter.
Reset Input	Counter is reset to 000 if this input is high.
Strobe Input	If this input is taken high the display will remember the counter state at the time of going high. The counter can still be advanced, or reset without changing the display.
Overflow	This output is used to clock a second module to form a six-digit counter. Or to clock any CMOS circuit as an indication of overflow. The output goes high when the clock input goes high and the counter is 999. It goes low when the counter advances to 000.



Note 1. If long leads are used pickup may occur which causes interference with normal action of the counter. If this is suspected add a 10 nF capacitor across the inputs on the module

Note 2. To use the decimal points provided on the displays connect pin 5 on the displays to the positive supply rail via a resistor which has the same value as those used for R5 to R11. A separate wafer on the range switch should be used to enable the appropriate decimal point. ●

DIGITAL VOLTMETER



Inexpensive unit uses dual-slope technique

THE three-digit display module described on pages 20-22 is readily adaptable to a wide range of applications. The display module has been designed so that an extra circuit board can be mounted above the display board thus enabling a wide range of additional facilities to be 'in-built'.

In this article we describe a simple yet accurate digital dc voltmeter add-on board. It is shown here as a single-range unit which is economical enough to be mounted within other equipment as a panel meter. An input switch and scaling resistors could easily be added to convert the intrument for use on ranges from one volt full scale to 1000 volts full scale.

We have not described the construction of an enclosure as individual requirements will vary widely.

CONSTRUCTION

The display-counter module ETI 533 should be built first using the instructions given on page 20.

Two additional boards are required to complete the voltmeter and the overlays and interconnections are given in Fig. 3. Check that all components, especially the metal case ICs are orientated correctly.

The interconnection wires should be long enough to allow the boards to fold together as shown above. The lower board ETI 533A has the components uppermost, the middle board ETI 117A has the components downwards while the top board ETI 117B again has the components uppermost. It may be necessary to juggle the components slightly on the lower two boards to allow them to fit together closely enough. These two boards are spaced apart with 12mm long spacers while the upper two boards are separated by 6mm insulated spacers. A piece of insulation material should be fitted between the top two boards to prevent the solder joints touching.

Power, 9-15 volts dc, is supplied to the lower board while the input connects to the upper board.

The unit can be either installed in a suitable box or within a piece of equipment. If range switches are required simply change the value of R12 as per Table 1. A suitable box for mounting the unit as a separate instrument is the type PC1 marketed by A & R.

CALIBRATION

Unfortunately to calibrate any voltmeter a known voltage reference or an accurate voltmeter is required for comparison. Two adjustments are provided, one for calibration and the other to compensate for the offset in the integrator IC. For input voltages of 10 V or more the offset potentiometer is not required as the error is within one digit.

This offset potentiometer should be adjusted first by applying a voltage of about one per cent (10 digits) of full scale and adjusting RV2 to give the correct reading. The calibration potentiometer RV1 can now be adjusted by applying an accurately known voltage near full scale.

The meter has a large overrange and voltages up to 250 per cent of full scale can be measured except that the first digit is lost and must be assumed, ie, if you are measuring a car battery on a 10 V range and it reads 3.52 V it is obviously 13.52 V.

MEASURED PERFORMANCE OF PROTOTYPE

Number of digits	3	
Overrange	250% (no indication)	
Dual polarity	No	
Ranges	1, 10, 100 and 1000 V dc	
Accuracy	As adjusted	
Linearity	± 1 digit	
Power supply	9-15 V dc at 120 mA isolatec	
Input impedance	100 k/V	
Overrange Protection		
1 V range	100 V	limited by power
10 V range	500 V	dissipation and
100 V range	500 V	voltage rating of
100 V range	2500 V*	R12
	* input switch permitting	
Reference	5.1 volt zener at constant current.	

DIGITAL VOLTMETER

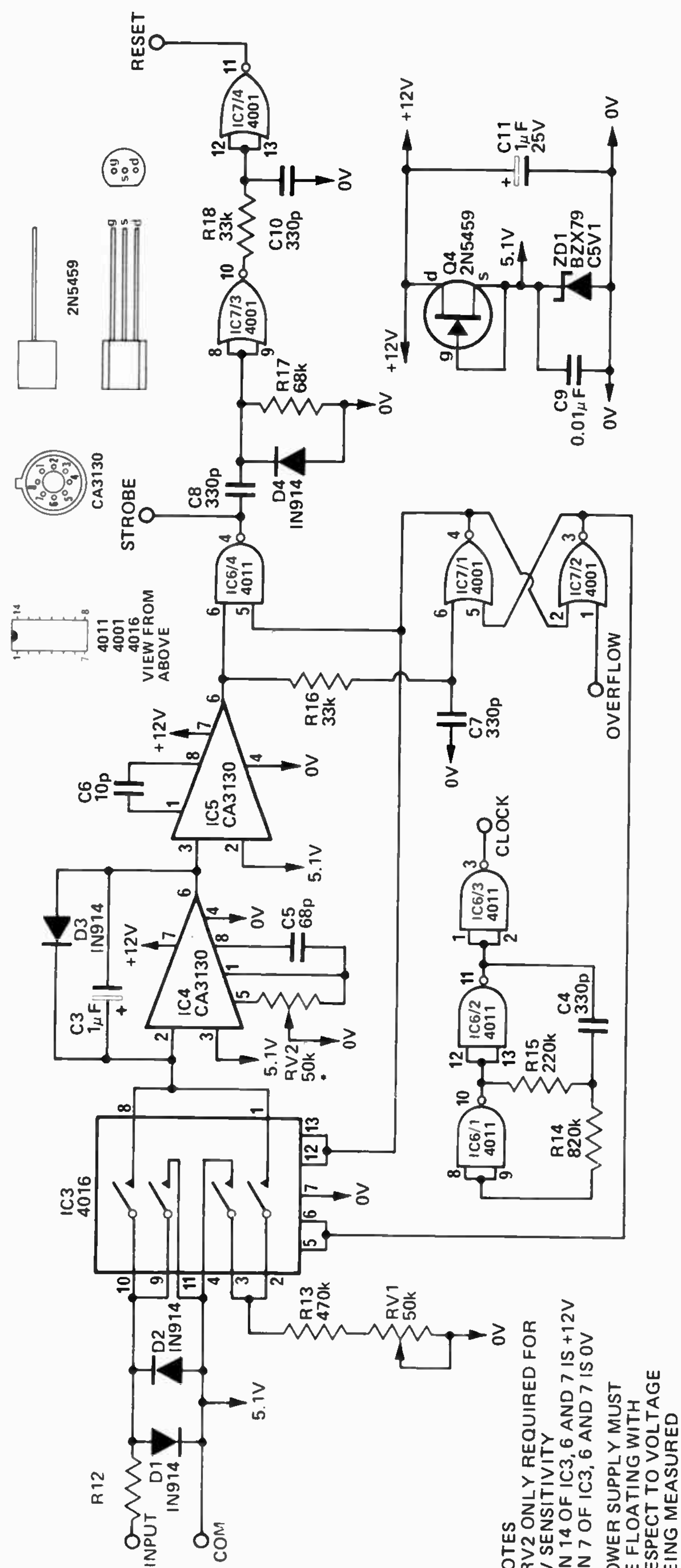


Fig. 1. Circuit diagram of the dual-slope analogue to digital converter. This circuit is used together with the ETI 533 display to make the complete voltmeter.

HOW IT WORKS

The method of analogue-to-digital conversion used is the popular dual-slope integration technique. We chose the dual-slope technique because it is relatively insensitive to component tolerances and gives very linear results with least amount of circuit complexity. The technique was developed by Weston and hence is covered by patents, however, there is nothing to stop individual constructors from using it, nor are there any royalties involved.

The circuit consists of an integrator (IC4 and C3), a comparator (IC5), an input selector (IC3), an oscillator (IC6/1,2,3) an RS flip flop (IC7/1,2), pulse generators for the reset and strobe outputs (IC6/4, IC7/3,4), a voltage reference (ZD1 and constant current source (Q1), and the digital display module on page 74.

The 5 kHz output of the oscillator, which runs continuously, is connected directly to the clock input of the display module and the conversion proceeds as follows. Flip Flop IC7/2, drives IC3 such that it selects either the input voltage via R12 or the reference voltage via R13.

The state of the flip flop is determined by the output state of the comparator IC5 (output high selects input voltage) and the overflow output from the display module (overflow selects reference voltage). If the input voltage is selected the output of the integrator will fall at a rate dependant on the input voltage, and, if the reference voltage is selected the input voltage will rise at a constant rate.

When the integrator output rises above 5.1 volts the comparator output goes high causing the output of IC6/4 to go low (as pin 5 of IC6/4

PARTS LIST

R16,18	Resistor	33k ¼w 5%
R17	"	68k " "
R15	"	220k " "
R13	"	470k " "
R14	"	820k " "
R12	"	See text
RV1,2	Potentiometer	50k Trim type
C6	Capacitor	10pF ceramic
C5	"	68 pF "
C4,7,8,10	"	330pF "
C9	"	0.01µF polyester
C3,11	"	1µF 25V Tantalum
D1,2,3	Diode	IN914 or similar
ZD1	Zener diode	BZ x 79 C5 V1
Q1	Transistor	2N5459 or similar
IC3	Integrated circuit	4016 (CMOS)
IC6	" "	4011 (CMOS)
IC7	" "	4001 (CMOS)
IC4,5	" "	CA3130

PC Boards ETI 117A, ETI 117B

Display Board Complete — Project ETI 533 July 1975 is also high). After about 10 μseconds delay, due to R16 and C7, the flip-flop changes state and the output of IC6/4 goes high again Thus a pulse is generated which is used as the strobe to transfer whatever number is in the decade counters into the store, and hence, to the display. The strobe pulse also triggers a 15 microsecond monostable, IC7/3, the output of which is delayed by 10 microseconds and inverted by IC7/4. This new pulse acts as a reset pulse for the counters setting them to zero.

As the flip flop has now reverted to its original state the input voltage is reselected and the integrator commences to ramp down again repeating the cycle.

Whilst the input voltage is selected clock pulses are gated into the counter and after about 200 milliseconds (1000 clock pulses each 0.2 mS) the counter will be full. The overflow thus generated from the display changes the state of the flip flop and the reference voltage is selected. The voltage across the integrator (referenced to 5.1 volts) at this instant will be proportional to the input voltage. With the reference supply connected the output of the integrator will rise at a predetermined rate and on crossing the 5.1 volt reference level the strobe and reset pulses are generated, the flip flop toggled and the process started again.

The time taken to bring the integrator back to the reference level is proportional to the input voltage and hence the number in the decade counter at that instant is the required reading of input voltage.

The only components which are required to have good stability, if accuracy is to be maintained, are R12, R13 and ZD1. All other components, provided their short-term stability is good, can be almost any tolerance. The integrator capacitor, for example, can have any value between 0.5 microfarad and 2.0 microfarads without affecting accuracy. However variations in the value of this capacitor will affect the over-range capability. The clock frequency may likewise be altered without affecting accuracy however, if the time of 1000 clock pulses is a multiple of 20 milliseconds the voltmeter will automatically reject 50 Hz ripple on the voltage being measured. This however was not considered of great enough importance to warrant special adjustment of the clock frequency which is preset by R15 and C4.

The reference supply is a 5.1 volt zener diode and a FET connected as a constant current source. The 5.1 volts is used as the common and hence, the 12 volt supply for the voltmeter must be left floating and must not be connected to ground or to any other equipment.

Due to the simplicity of the circuit there are some features of the instrument which are not desirable but do not greatly affect the operation of the instrument. Firstly there is no over-range indication and thus if 15 volts is applied to the 10 volt range the instrument will read 5 volts. The unit remains accurate (except for the first digit which is lost) until the integrator clips on its negative swing (about 250% of full scale). The other point is that if the input voltage is negative the comparator, IC5, will remain high and no further strobe or reset pulses will be generated. The effect of this is to freeze the display at the last number. This is not normally a problem as the display goes to zero if the input is disconnected.

Fig.2. Circuit boards used for the converter.

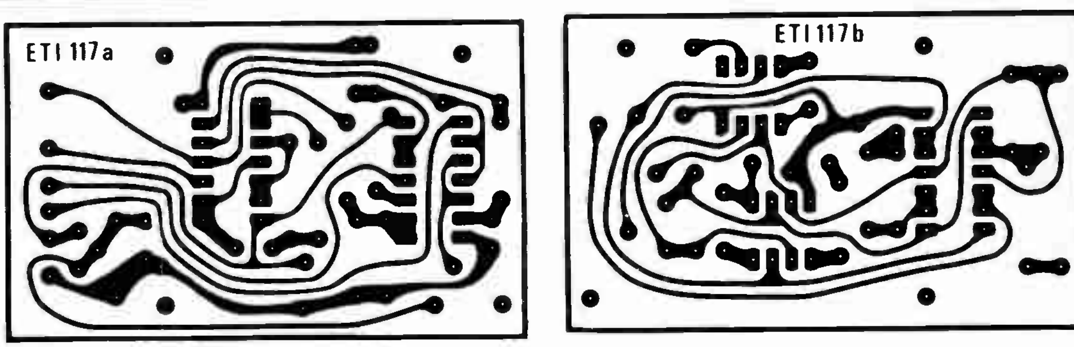


TABLE I

RANGE	VALUE OF R12	
1V	100k	5%*
10V	1M	5%*
100V	10M	5%*
1000V	100M (10 x 10M)	5%*

For multirange meters R12 must be 1% or adjustable.

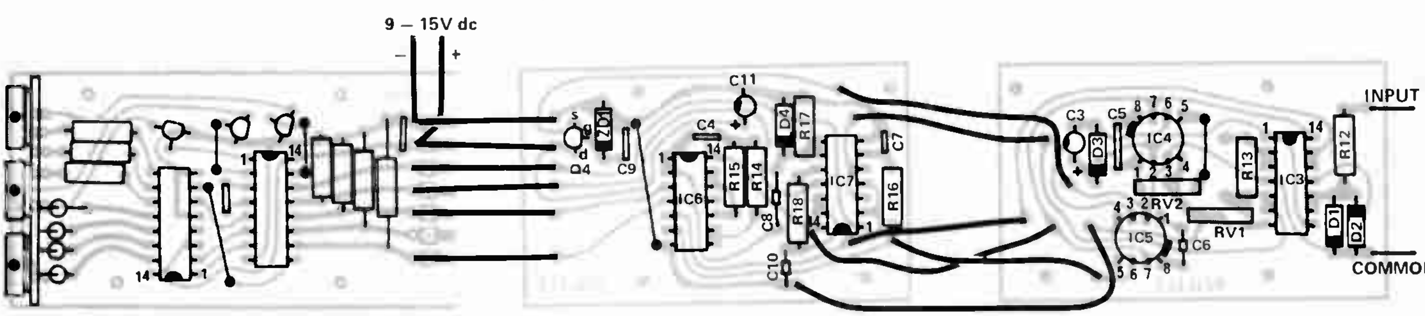
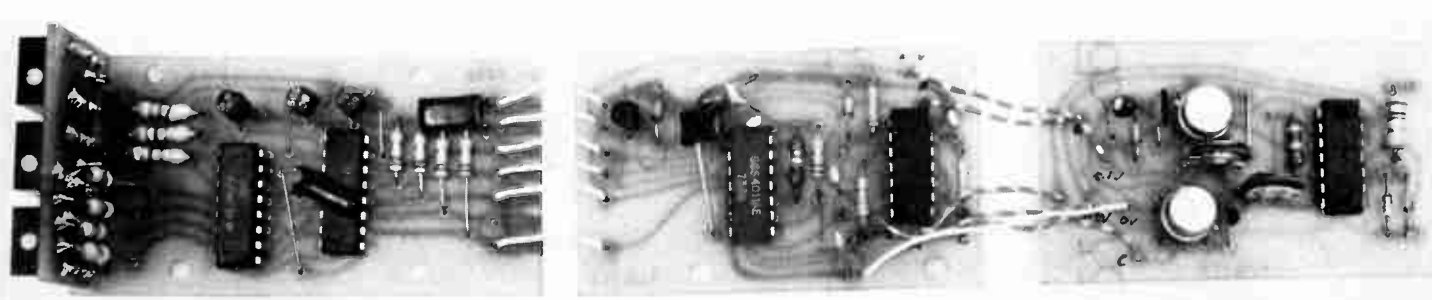


Fig.3. Component overlay of the complete voltmeter.



DIGITAL VOLTMETER

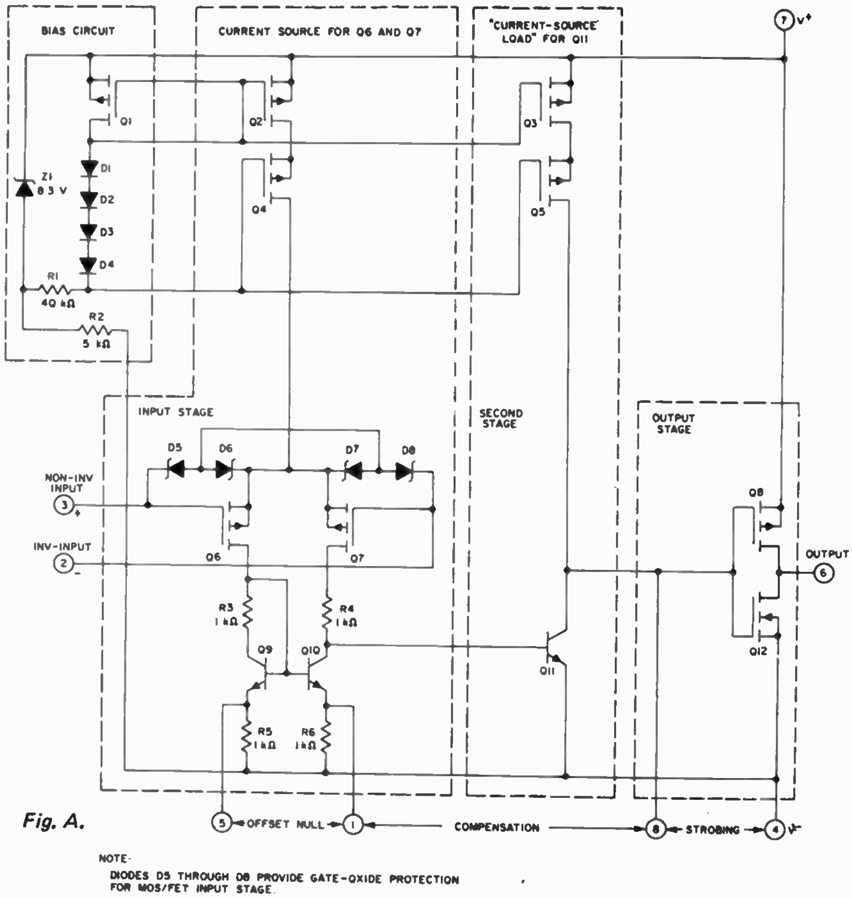


Fig. A.

NOTE

The MC14553 is manufactured by Motorola and distributed by Total Electronics. The MC14511, MC14016, MC14011 and MC14001 are also distributed by Total. However these latter devices are also available from other manufacturers under the numbers 4511, 4016, 4011 and 4001 respectively.

Distributors of these latter devices are Cema, National and AWA (RCA).

THE CA3130 OPERATIONAL AMPLIFIER

For those unfamiliar with this IC we have reproduced the internal circuitry in Fig. A. It is an economical FET input operational amplifier. This IC is unusual as it combines a FET input stage with a bipolar amplifier and a CMOS output stage, all on the one chip! It is a pin for pin equivalent for the 301/741 type of IC and needs a 68 pF capacitor between pins one and eight for compensation. The major differences are a maximum voltage of 16 V between the supply rails (± 8 V) and the extremely high input resistance of 1.5 TΩ (1 500 000 MΩ) and low input current of 5 pA.

eti PROJECT 118

Inexpensive unit works to 1 MHz.

SIMPLE FREQUENCY COUNTER

WE CONTINUE our series of simple units, based on the ETI 533 Display Module, by describing a simple frequency meter. The unit is easy to construct and quite inexpensive.

DESIGN FEATURES

We originally considered that this project would only take a short time to develop — but were we ever wrong! Just about everything that could have gone wrong did so, and we became convinced that Murphy was not only alive and well but was living in Rushcutters Bay.

The first problem was to choose suitable timebase circuitry. As the project had to remain fairly economical to build, the use of a crystal timebase was ruled out. We eliminated a mains referenced timebase because it was considered that the possibility of battery powered operation was a definite advantage. Especially as control tones on the mains can cause problems.

An NE555 timer was tried for the 10 second timebase but it soon became apparent that the device just was not stable enough even when the power supply was regulated. The change in frequency due to supply changes was about 1.5%/volt. And even with a regulator the stability was not good enough to allow more than four digit readings. The error occurs because the NE555 output stage does not go exactly to the supply rails but only 0.6 V away from them. The same applies to the discharge transistor which has a fixed saturation voltage.

We considered many oscillator designs in an effort to find one with a stability of better than one part in 10 000 and ultimately chose the one shown in the circuit diagram. This type of oscillator is well known but is not normally considered to have good accuracy and stability. This is because in a conventional op-amp IC there is normally a base-emitter junction at the output, as in the 555. However in the new CA 3130 device this problem has been eliminated as the output stage is CMOS and appears as a resistance (about 500 ohms) and not as a voltage drop. A further advantage of this IC is the extremely high input impedance which eliminates any inaccuracies due to loading effects. On the prototype the frequency change was less than one part in 10 000 with a supply voltage change of from 8 to 16 volts. The main source of error is now due to the temperature coefficient of R10. The expected error, using good quality metal-film resistors would be around 0.01% per degree C.

The CA3130 IC is also ideal for the input stage, because of its high input impedance, and also because it allows a 0 V reference to be used thus eliminating the centre-tap point normally required for conventional operational amplifier circuits.

(Main text continued page 31)

SPECIFICATION

INPUT IMPEDANCE	470 k // 47 pF
INPUT SENSITIVITY 10 Hz to 10 kHz	< 50 mV rising to 1 V at 1 MHz
RANGES	99.9 Hz to 999 kHz
DISPLAY	3 digits (no overload indication)
ACCURACY	as calibrated.
STABILITY	0.01% can be expected but depends on resistor stability.
OVERLOAD PROTECTION	
up to 100 kHz	50 Vac.
100 kHz to 1 MHz	dropping from 50 Vac to 10 Vac 50 Vdc.
POWER	240 Vac or 12 Vdc at 100 mA.

SIMPLE FREQUENCY COUNTER

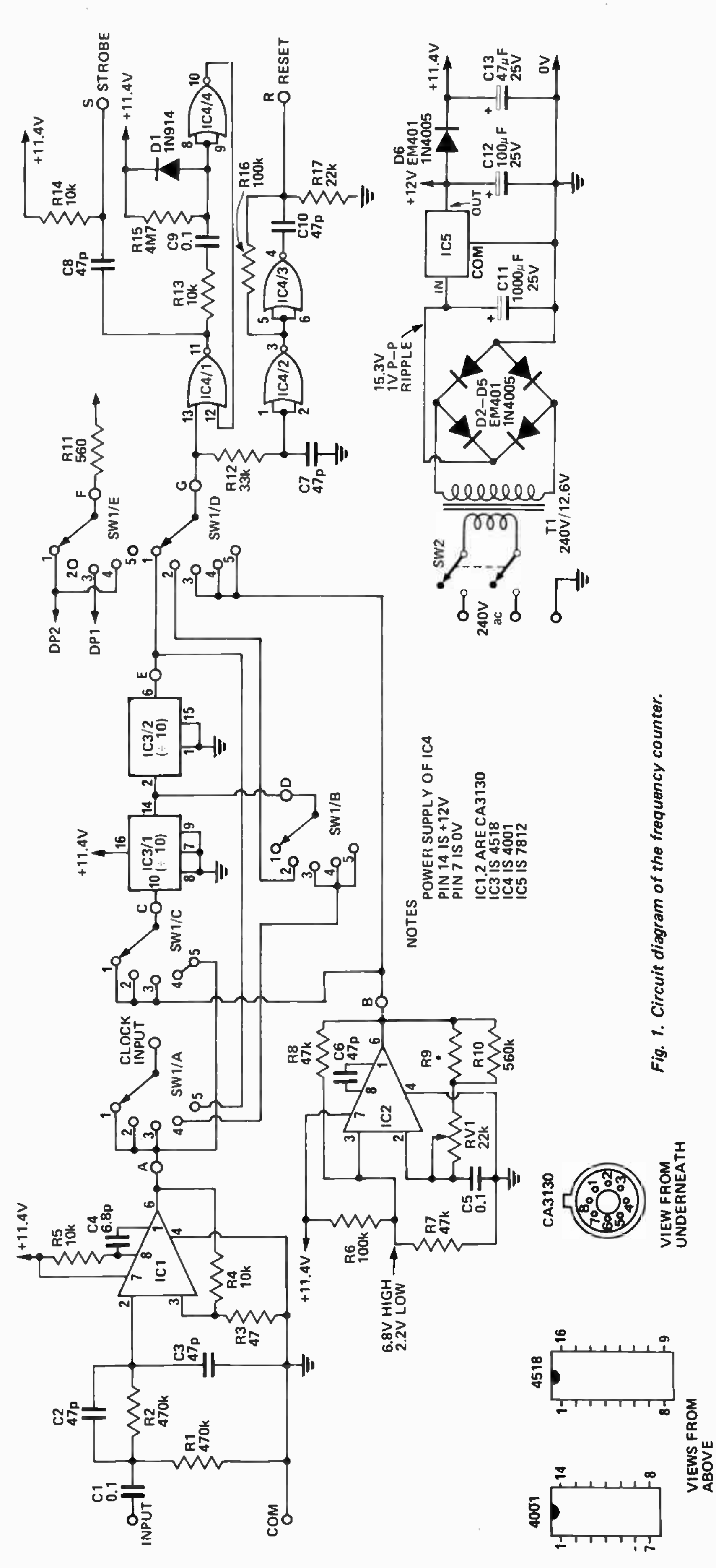


Fig. 1. Circuit diagram of the frequency counter.

PARTS LIST

R3	Resistor	47 ohm	¼W	5%
R11	"	560	¼W	5%
R4,5,13,14	"	10 k	¼W	5%
R17	"	22 k	¼W	5%
R12	"	33 k	¼W	5%
R7,8	"	47 k	¼W	5%
R6,16	"	100 k	¼W	5%
R1,2	"	470 k	¼W	5%
R10	"	560 k	¼W	2%
R15	"	4M7	¼W	5%
R9 (See table 1)				

C4	Capacitor	6.8 pF ceramic
C2,3,6	"	47 pF ceramic
C7,8,10	"	47 pf ceramic
C1,5,9	"	0.1 μF 100v polyester
C13	"	47 μF 25 V electro
C12	"	100 μF 25 V electro
C11	"	1000 μF 25 V electro

D1 Diode IN914 or similar
D2-D6 " EM401, IN4005 or similar

IC1,2 Integrated Circuit CA3130 (AWV)
IC3 " " ·4518 (CMOS)
IC4 " " 4001 (CMOS)
IC5 " " 7812 (plastic pack).

SW1 Rotary switch 6 pole 5 position
SW2 toggle switch DPDT (miniature)
T1 transformer 240 V/12.6 V 150 mA

PC Board ETI 118
Case type PC1 (A&R Soanar) or similar
Display Module — ETI 533
Shield as per Fig. 7.
Front panel as per Fig. 5.
3 plain spacers 6.4 mm long insulated
3 plain spacers 19 mm long
3 1/8 whit. spacers 25 mm long
One 8 way tag strip
3 core flex plug, grommet and clamp pc board pins
3 25 mm long 1/8 whit. screws
9 12 mm long 1/8 whit. screws
6 1/8 nuts.
Two input terminals (red-black)

PARTS AVAILABILITY

The CMOS ICs used in this project are distributed by CEMA and TOTAL.

HOW IT WORKS – ETI 118

The frequency counter may be divided into several basic sections.

a. Input amplifier – Schmitt trigger.
b. 10 Hz oscillator.
c. Two divide by 10 networks.
d. Strobe and reset circuitry.
e. Power supply.
f. Display module (ETI 533).

The input amplifier is a CA3130 connected as a Schmitt trigger. Resistors R3 and R4 provide positive feedback whilst resistor R2 provides protection for the input of the IC. The resistor R5 is used to increase the negative slew rate of the amplifier thus increasing the range of operation to one megahertz.

The 10 Hz oscillator is another CA3130 where positive feedback is applied by R8 and negative feedback by R10. When the output is high the voltage at pin 6 is about 6.8 volts. The capacitor C5 charges via R10, and when it reaches 6.8 volts the output goes low. The voltage now set at pin three is 2.2 volts and the output remains low until C5 has discharged to this point at which the output goes high again. Preset RV1 varies the oscillator frequency by

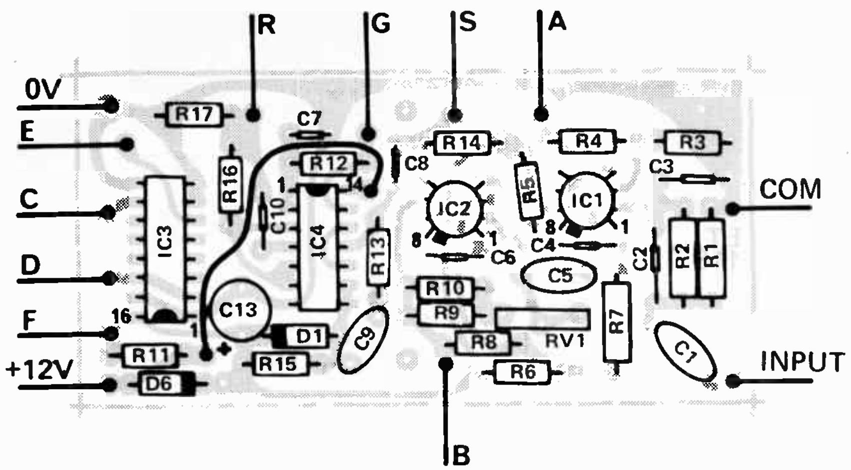
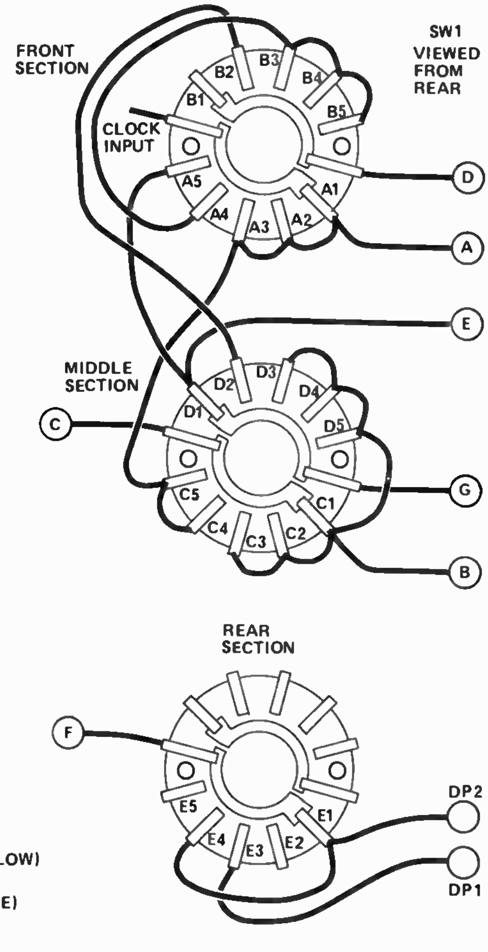


Fig. 2. Component overlay.



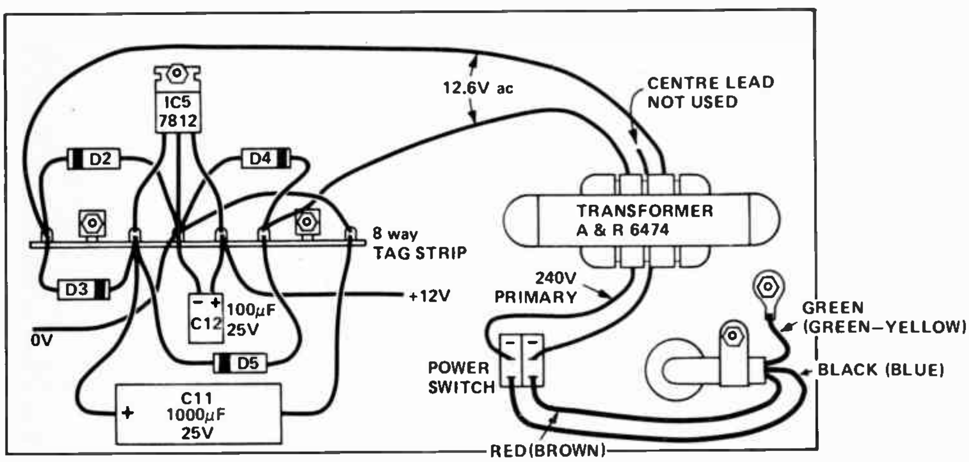


Fig. 3. Wiring of the range switch.

Fig. 4. How the power supply is mounted on the back panel and interconnected.

about 4% and a parallel resistor, R9, is required to set it within the required adjustment range. A higher value preset could be used but it becomes difficult to adjust with accuracy.

The divide by ten circuitry is simply a 14518 IC which contains two decade counters. It can be switched to divide the input frequency (100 k, 1 M ranges) or the timebase (100 Hz, 1 kHz ranges) by means of the range switch SW1.

The timebase, be it ten seconds, one second or 0.1 seconds, is coupled by SW1/d to IC4/1 pin 13. When this voltage goes high the output of IC4/1 goes low and C1 couples a short negative going pulse into the strobe terminal of the display module. After a short time, due to R12 and C7, the output of IC4/3 goes high and C10 couples a short positive pulse into the reset terminal of the display module. When the output of IC4/1 goes low the output of IC4/4 goes high and the output of IC4/1 remains low irrespective of what now occurs at pin 13. After about 350 milliseconds C9 recharges via R15 releasing IC4/1 to the control of the timebase. This procedure removes three out of every four strobe pulses when using the 10 Hz timebase, making the display easier to read.

The resistor R16 is used to raise the steady-state voltage at the reset terminal to about 1.8 volts, thus ensuring that the reset pulse goes high enough to give reliable tirggering. The voltage at the strobe terminal sits at about 10.4 volts due to the 100 k input impedance of the display module.

The power supply is a full-wave rectifier and capacitor filter supply which is regulated down to 12 volts by a 7812 regulator IC. The control circuitry is isolated by a diode D6 and capacitor C13 to prevent any ripple appearing on the 12 volts due to the current drawn by the display module.

The display module contains a three decade counter-store-decoder and display as described on page 22.

To measure frequency all that is needed is to count the number of pulses occurring over a given period of time. If we count the number of input pulses over a one second period we can measure to the nearest one cycle, or one hertz. If a three digit display is used then the maximum reading will be 999 Hz. However if the frequency happens to be, say, 156254 Hz the display will read 254 and ignore the 156. To measure a higher frequency, either a shorter timebase must be used, or, the input frequency must be divided down. For the 10 kHz range we simply use an 0.1 second timebase giving 10 Hz resolution. For the 100 kHz we divide the input by 10 and use an 0.1 second timebase, whilst for the one megahertz range the input is divided by one hundred. For the 100 Hz range a ten second timebase and no division is used.

If we use the one megahertz range to measure our 156254 Hz, we display 156. Switching to 100 kHz we get 563, on 10 kHz we get 625 and finally 254 on the 1000 Hz range, thus the frequency can be read to the nearest hertz but the accuracy depends on the accuracy of the initial setting up and the fact that temperature variations cause an error of one part in 10 000 per degree C.

SIMPLE FREQUENCY COUNTER

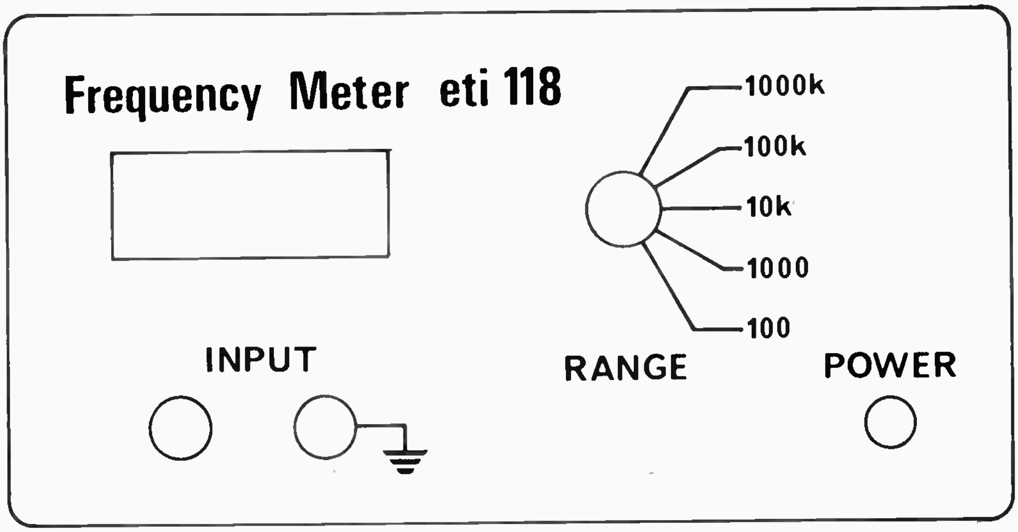


Fig. 5. Front panel of the frequency meter.

Initially the maximum frequency of operation was limited to 200 kHz due to the slow, negative slew-rate of the IC. Looking at the internal circuit of the IC it was decided to increase the bias current in the second stage by adding a resistor between the positive supply and pin 8. This allows the frequency response to be extended to beyond one megahertz. A small compensating capacitor was found to be necessary to eliminate the effects of a small amount of coupling from the 10 Hz oscillator. The resistor to pin 8 also alters the offset voltage but this does not affect the operation of this circuit.

Another problem that occurred was in the strobe and reset pulse network.

Using an 0.1 second timebase the display changes too rapidly for ease of reading. Therefore IC4/4, which is connected as a 350 millisecond monostable, is used to eliminate three out of every four strobe pulses thus making the display more readable. However it was discovered that, when using this delay, the timebase changed frequency by about four parts in 10 000. Since the power supply to the control circuitry was isolated from the display module, the circuitry is mainly CMOS, and the oscillator rejects supply rail change, none of these factors could be suspected as a cause of the trouble. The problem was due to the fact that IC4/4 works in the linear mode and can draw 10 to 20 mA. This modulates the power supply by up to 20 millivolts. The cure is to power IC4 directly from the 12 volts. This explains the use of the link of the board.

Some coupling between the display board oscillator and the input stage occurred and was cured by adding an aluminium shield between the two boards.

To obtain all five ranges with only two divide-by-ten sections necessitates a more complex switch. This was considered to be justifiable as the alternative was to use a switch with one less wafer but add one more 14518 IC.

CONSTRUCTION

The display module should be constructed as described on pages 20 and 21. The value of resistors R5 to R11 should be 560 ohms for operation of the 12 volt supply.

The control board should be assembled with the aid of the component overlay Fig. 2. Use printed circuit board pins for all outputs and for R9 as an aid to later assembly. Make sure that the link between +12 volts and pin 14 of IC4 is installed.

Wire switch SW1, in accordance with Fig. 3, and leave the leads long enough to reach the printed circuit board. Assemble the power supply onto the tag strip and the back panel of the box as detailed in Fig. 4. There is no need to insulate the tag of the regulator from the rear panel as it is the common terminal which should be earthed. The rear panel itself is earthed via the mains cable.

The front panel has to be cut and drilled as shown in Fig. 5. It can be either silk screened with the required

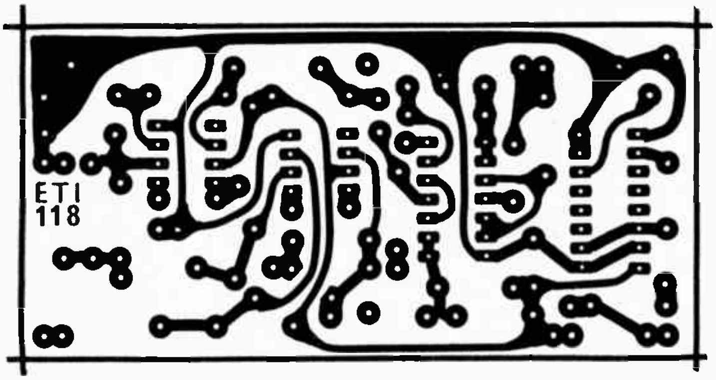


Fig. 6. Printed circuit board for the counter. Full size 90 x 45 mm.

TABLE 1

Frequency with RV1 at minimum 50 Hz input	Value of R9 to allow RV1 to calibrate
48.1 – 50	–
49.8 – 51.8	15 Meg
51.3 – 53.4	8.2 Meg
52.7 – 55.0	5.6 Meg
54.7 – 57.1	3.9 Meg
56.7 – 59.3	3.0 Meg
58.8 – 61.7	2.4 Meg

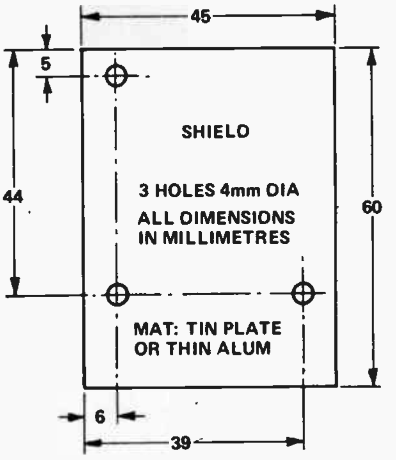


Fig. 7. Drilling details of the shield plate.

range markings etc, or a Scotchcal panel can be used as in our prototype. Scotchcal panels for this project can be obtained for $3.00 (plus SAE) directly from Electronics Today. A piece of polarized or red plastic can be used to protect the displays. If the A&R box, as specified, is used mounting holes will have to be drilled in the base of the box. These can be marked by temporarily installing the front panel, by fitting the three 25 mm spacers to the control module (the left front mounting hole on the module is not used) and sitting the control module in position. It will be found that the right front spacer interferes with a rib in the base of the box. Cut this rib away with a chisel or similar tool such that the spacer can sit flush with the base. Mark the hole positions, remove the module and drill the holes.

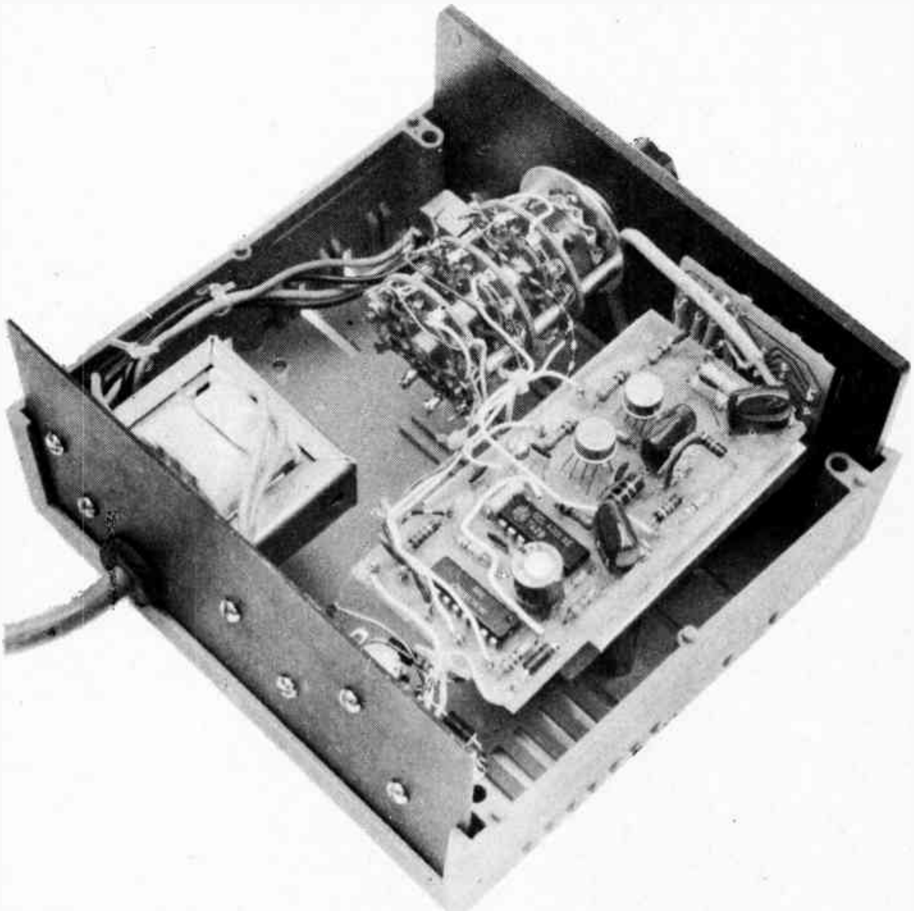
Fit the rear panel in to the box and then mount the input terminals to the front panel (the rear of the screws of the terminals may have to be shortened to clear the display module mounting spacers). The rotary and toggle switches should also now be mounted to the front panel. Connect a short length of coaxial cable to the input terminals (about 150 mm) for later connection to the control board. Connect leads to the positive volt, zero volt, strobe, reset and input clock inputs on the display module and assemble the display board, shield and control board as shown in Fig. 8. Make sure that the spacers do not touch the copper tracks on any of the boards, (except for the front spacer on the control board). If any of the spacers are too close to the tracks add a piece of insulation material under the spacer. The whole assembly can now be mounted in to the box.

On the display module the power rails are taken direct to the power supply whilst the 'reset' and 'strobe' go to the control board and the 'control' to the rotary switch. On the control board the power rails also go direct to the power supply whilst all other leads, with the exception of the coaxial cable, go to the rotary switch.

Finally connect the power switch and insulate it with plastic tape to prevent accidental personal contact.

CALIBRATION

Apply about 6 volts ac at 50 Hz, from the secondary of a power transformer to the input of the counter. Select the 100 Hz range and set the trimpot, RV1, to its minimum resistance position. Wait for the reading to settle (there is about ten seconds between readings) and using this reading look up the corresponding value of R9 from Table 1. Install this resistor and again check the reading, it should now be just under 50 Hz. The trimpot RV1 can now be adjusted to give a reading of exactly 50 Hz. If a more accurate frequency source than the mains is available it can be used instead of the 50 Hz for final calibration. Due to the effect of soldering upon the value of resistors final calibration should be left until several hours after R9 is soldered into position so that the resistor may stabilize. ●



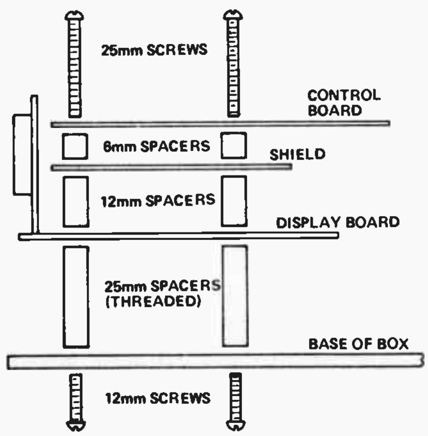


Fig. 8. Assembly of the counter display boards.

eti

HOW TO ORDER

ETI Circuits No 1, the Electronics It's Easy series, and Top Projects Vol 3 are available now from most newsagents or directly from ETI.

Top Projects Volume 4 will be available in mid-June 1977.

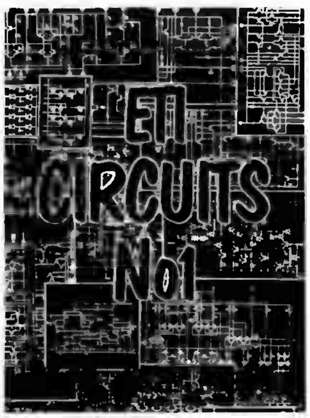
The Transducer and the Synthesizer books are available only from us and a number of kitset suppliers, they are not sold via newsagents.

Send orders to Electronics Today International
15 Boundary St,
Rushcutters Bay
NSW 2011

ELECTRONICS IT'S EASY Vol 1/2/3.

Volumes 1 & 2 now reprinted as a second edition!

Vol 1. The first 12 parts of this very successful series produced as a 100 page book. This volume takes the reader from an introduction to electronics through to operational amplifiers.
Vol 2. The 'middle-third' of the series introduces the reader to more sophisticated techniques, and includes power supplies, waveforms, filters and logic systems.
Vol 3. The final volume covers digital displays and systems, computers, transmission systems, instrumentation and the control of power.
$3.00 per volume plus 40 cents per volume post and packing (60 cents post and packing total if two or three volumes ordered at the same time.)



ETI CIRCUITS NO 1

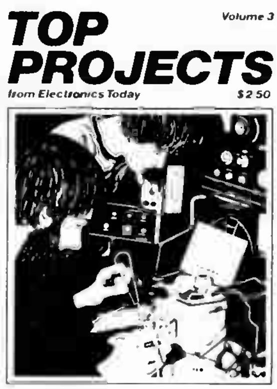
Just published! Available from your newsagent or directly from us. This is a new concept in specials. ETI Circuits No 1 contains over 200 circuits largely based on the Tech Tips and data sheet sections of ETI. Great care has been taken to index each circuit carefully to enable the reader rapidly to locate the item sought. This publication is likely to be regarded as a 'bible' by the experimenter for not only does it contain complete circuits but also ideas and circuit sub-assemblies.
$2.50 plus 40 cents postage and packing.

INTERNATIONAL 3600 AND 4600 SYNTHESIZERS

A totally revised and updated reprint of ETI's phenomenally successful music synthesizer book.
This book has been beautifully printed on heavy art paper and has a sturdy cover varnished for protection.
Available only from ETI and some kit set suppliers $12.50 including postage and packing.

TRANSDUCERS IN MEASUREMENT AND CONTROL

This book is an unusual reprint from the pages of ETI. It is a compilation of a series of articles completed about two years ago — which was so highly thought of by the University of New England that they republished the series splendidly for use as a standard textbook.
Written by Peter Sydenham M.E., Ph.D., F. Inst. M.C., F.I.I.C.A., this publication covers practically every type of transducer and measuring technique and deals with equipment and methods not covered in any other book.
Enquiries from educational authorities, universities and colleges for bulk ordering are welcomed.
$4.50 plus 50 cents postage and packing. Available only from Electronics Today International directly.



TOP PROJECTS Vol 3

Available from newsagents or directly from Electronics Today International.
Published in October 1976.
Contains FM Tuner, 25 Watt Amplifier, Active Crossover, Crossover Amplifier, Booster Amplifier, 50 Watt Power Module, 400 Speaker System, Audio Noise Generator, Crosshatch/Dot Generator, ETI Utiliboard, Linear IC Tester, Dual Beam Adaptor, Impedance Meter, Tone Burst Generator, Digital Display, Digital Voltmeter, Frequency Counter, Logic Probe, Logic Pulser, Switching Regulator Supply, Nickel Cadmium Battery Charger, Radar Intruder Alarm, Intruder Alarm, Colour Organ, Car Alarm, Transistor Connections.
$2.50 plus 40 cents postage and packing.

We regret that due to heavy demand, Top Projects Vols 1 and 2 are no longer available.

Project 133

PHASE METER

By Dr. P.C. Bury, Physics Dept., Victorian College of Pharmacy

This instrument measures phase angles of voltage, current or power from sub-audio frequencies to 100 kHz or beyond. Readout may be either digital or analogue.

THE POWER being dissipated in an ac circuit is one of the more difficult quantities to measure with normal laboratory equipment – unless the circuit is purely resistive. This is because the power dissipated is given by the expression $P = IV \cos\phi$ where I is current, V is voltage and ϕ is the phase angle between them. Theta (ϕ) varies from 90° for an ideal inductance, through 0° for a resistance, to –90° for a capacitance. Since $\cos \pm 90 = 0$, both inductance and capacitance dissipate no power at all. They store it during one half-cycle and release it to the source again during the following half-cycle.

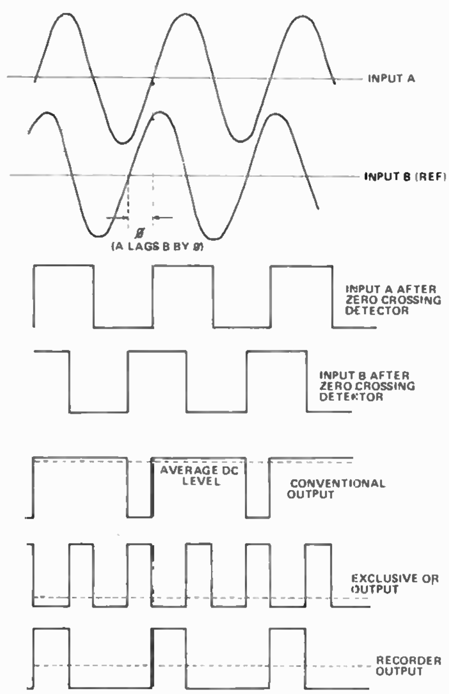


Fig.1. Comparison of the conventional 'flip-flop' method and the exclusive-OR method used in this project.

Therefore, in order to measure power, one either needs a wattmeter – an expensive instrument if any great accuracy is desired – or a knowledge of ϕ, the phase angle. While ϕ can be estimated from a dual trace oscilloscope, this article describes a simple and accurate way of measuring it directly. In addition to power measurements, measurement of the phase difference between two voltages is useful when working on filters, feedback loops and phase-shifting networks: it can be used to measure the Q of an inductor, and hence check for shorted turns, or the loss factor in a capacitor. A further application of growing importance in the audio field is the phase of the sound from individual drive units in a loudspeaker enclosure, or members of an array of loudspeakers.

The phase relationship between two voltages is conventionally measured by detecting when each crosses zero voltage (see Fig. 1) in one direction, and arranging for one voltage to turn a flipflop ON and the other to turn it OFF. The percentage of time that the flipflop is on, and hence the average value of the flipflop output, is proportional to the phase difference between the two voltages. This method has three inherent disadvantages –

(i) Voltages with little or no phase difference can give readings of 0° and 360°, or a reading which varies randomly between these limits.

(ii) Any noise on either signal can cause false triggering and jittery readings.

(iii) Any harmonic distortion can produce a shift in the zero crossing point and hence an error of 0.6° for each 1.0% of distortion.

The method used in the circuit described here is to form the exclusive-OR of the square waves produced by zero crossing detectors from the two voltages.

For those who have not encountered the exclusive-OR (XOR) function before, this is a logic function (in the same way that AND and OR are logic functions) that gives an output (logic 1) if its two inputs are different, but not if they are the same. Thus two square waves which are in phase will produce no output: two which are exactly out of phase will produce a maximum continuous output; and intermediate phases will produce an output proportional to the phase difference (see Fig. 1). This system has the advantage of being almost immune from noise problems since no triggering or latching circuits are involved.

Because the circuit response is symmetrical about 0° and 180°, there are no output discontinuities or ambiguities of reading. However an additional flip-flop is required to sense which voltage is ahead of the other and indicate it. The circuit is implemented with CMOS gates which have the advantage of being able to be used in either linear or digital mode.

CONSTRUCTION

We assume that only the more experienced constructor will build a somewhat specialised instrument of this type, and that they will be capable of assembling, handling the CMOS, with due care, boxing it without step by step instructions. The pc board can be copied from the diagram (Fig. 3), or hopefully will be available through commercial channels. The layout of the components is shown in Fig. 4. Some care is needed to keep the input leads as short as possible as the gain of

Text continues on page 36

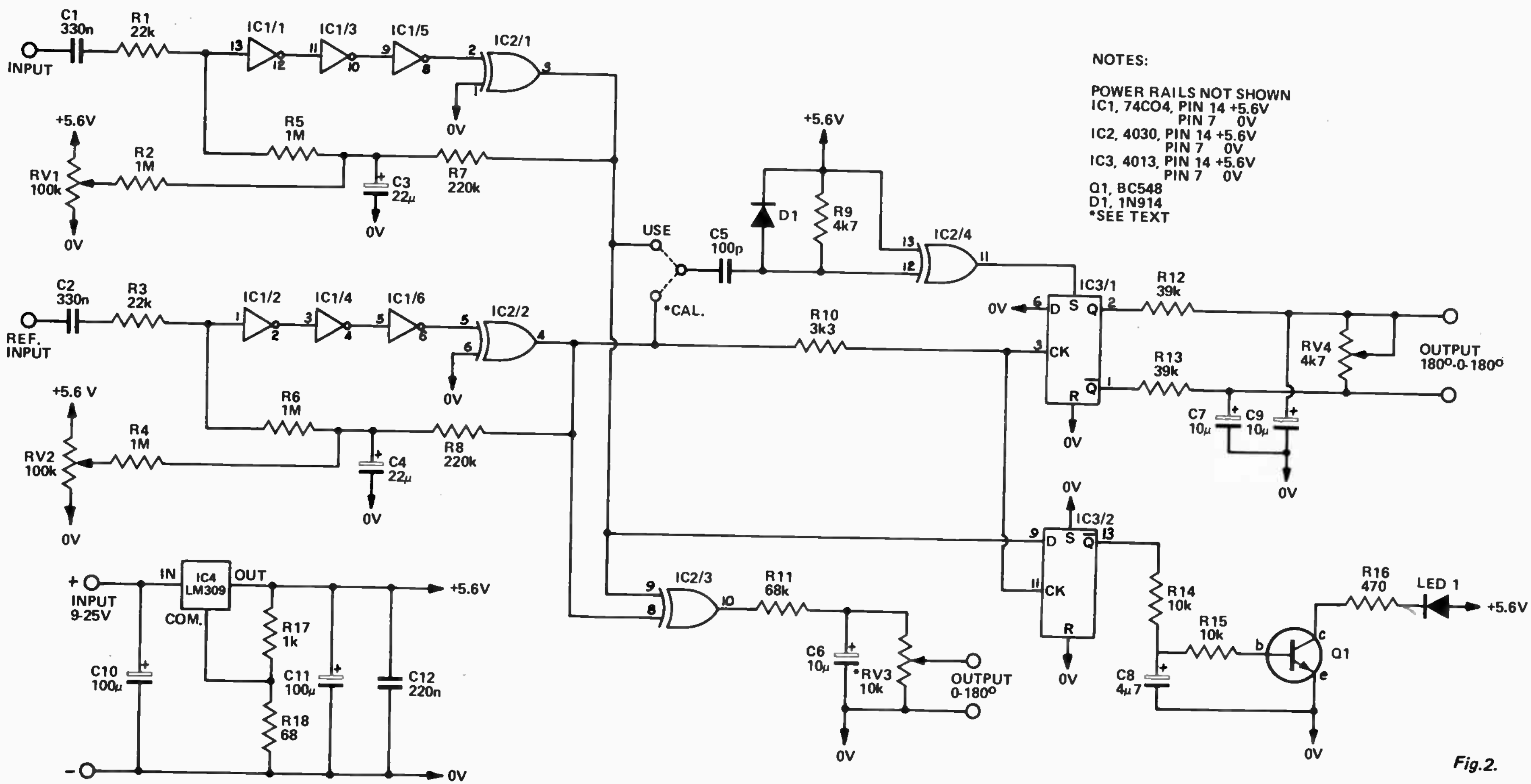


Fig.2.

HOW IT WORKS

The two inputs are first squared. For example the reference input is amplified by gates IC1/2, IC1/4 and IC1/6 (see Fig. 2) and then applied to IV2/2, one of the spare EX.OR gates whose other input is grounded. This conveniently behaves as a Schmitt trigger type of bistable circuit. The average of the output of this gate is formed by R8 and C4, and this is inserted via R6 as the dc level at gate IC1/2.

This produces two important consequences. Firstly it forces the output of IC2/2 to a symmetrical 180°on/180°off condition which is maintained stabely by almost complete dc feedback. And secondstance it may be set to 180 mV for a 180° phase difference and read it on a digital multimeter. Alternatively up to 50 μA can be drawn to give a reading on any suitable meter or multimeter. The use of an external meter is of course a much more economical proposition.

In order to detect which of the inputs is leading the other, the two voltages from the squaring circuits are also fed to the D type flip-flop IC3/2. One voltage is used for the clock input and the other as a data input. This type of flip-flop is really a data latch, and whatever voltage is present at the D input at the moment when the clock voltage changes from low to high is held would leave one flip-flop unused. In fact it turns out that there are two functions that these gates can usefully perform. First, for setting up the input squaring circuits: if the flip-flop is slaved to the squaring circuit, the exact 180° condition can be set when the complementary outputs Q and Q have equal average values. Secondly these gates can be arranged to turn the flip-flop on and off to give a conventional phase meter circuit output. While this does not give as accurate a reading, it does give one which is of opposite polarity for leading and lagging voltages and which can therefore be recorded graphically and unambiguously on an instrument such as a chart recorder. This is therefore desflip-flop is slaved to the reference input as it is set when IC2/2 goes low and reset when IC2/2 goes high, and this enables the 180° duty cycle to be set (see below). When C5 is connected to IC2/1 (the USE position) the flip-flop will have equal outputs at Q and Q if the two inputs are exactly in phase. It produces a positive value of Q relative to Q when the input leads the reference voltage and a negative value when it lags, the average voltage between them being proportional to the phase difference. R12, R13, C7, C9 and RV4 are used to filter this output and set it to some convenient value.

Of the other components, R10 is used to delay the voltage to the clock inputs ly, because we now have a true squaring circuit rather than a zero-crossing detector, all errors due to even-order harmonic distortion are cancelled. R4 and RV2 are used to adjust for input offset and set the exact 180° condition.

IC gates IC1/1, IC1/3, IC1/5 and IC2/1 process the signal from the other channel in an identical manner, and the two squared outputs are fed to gate IC2/3 which is the gate that forms the EX.OR of them. Its output is filtered by R11 and C6 and a voltage proportional to the phase difference of the inputs may be taken from across C6. RV3 is used to set this to a convenient value – for instance until the next clock pulse. Thus if the D input stays low until after the clock input goes high, the output Q will always remain low showing that the D input lags the clock input. The complementary output Q will be high and this is used to turn on the transistor and LED indicating this lag condition. Since any noise arriving at the clock input can cause spurious resetting of the flip-flop, it is preferable to use a clean voltage to drive it. This is why this channel has been designated the reference. Noise on the other channel is almost completely ignored.

These then are the basic EX.OR functional parts of the phasemeter, and this ignated the recorder output.

In operation one input of the EX.OR gate IC2/4 is connected through C5 to the output of one of the squaring circuits. Its other input is tied to the positive rail so that it functions as an inverter with respect to the other input (the EX.OR is a flexible beastie isn't it?). With C5 connected to the output of IC2/2 (the CAL position), when IC3/2 goes negative C5 and R9 differentiate this, and IC2/4 gives a short (½ µsec) positive spike, which is used to set IC3/1. Since the data input of this flip-flop is always low, the clock pulse will always reset it again. Thus the slightly to compensate for the set-up time at the data inputs, and the LM309 regulator and associated components holds the supply voltage constant at just under 6 volts. This is important as the full-scale readings from the outputs is proportional to the supply voltage. The circuit can be run from a 9 V battery and draws about 20 mA with the LED off and 40 mA with it on. Alternatively any power supply that produces between 9 and 30 volts may be used, but it should be a floating supply to simplify the measurement of the phases of currents, and R16 and the voltage rating of C10 should be increased if more than 15 volts are used.

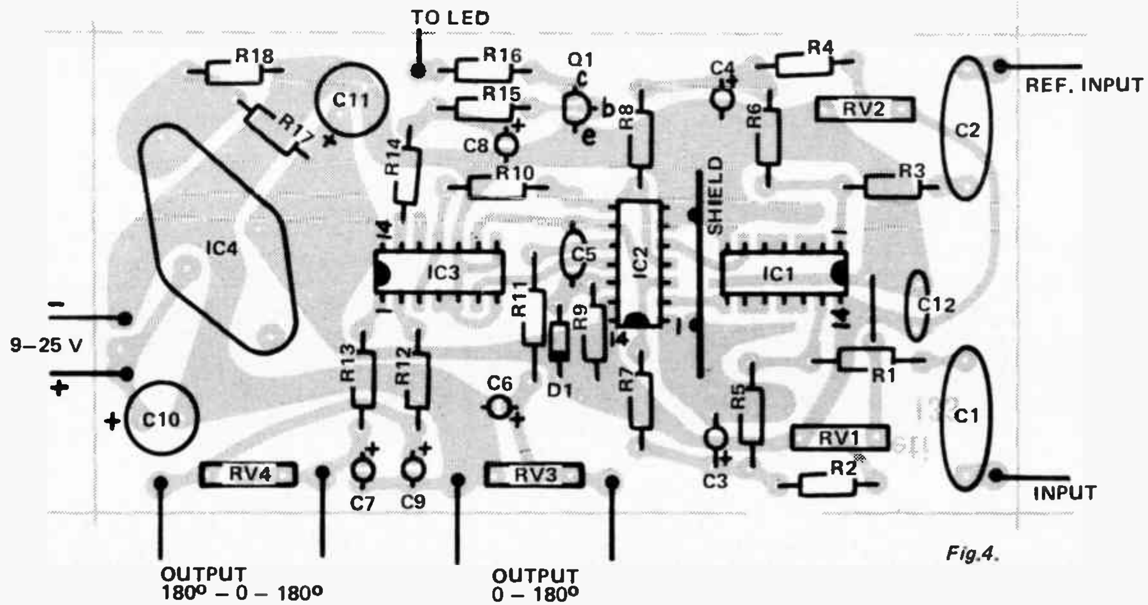


Fig.4.

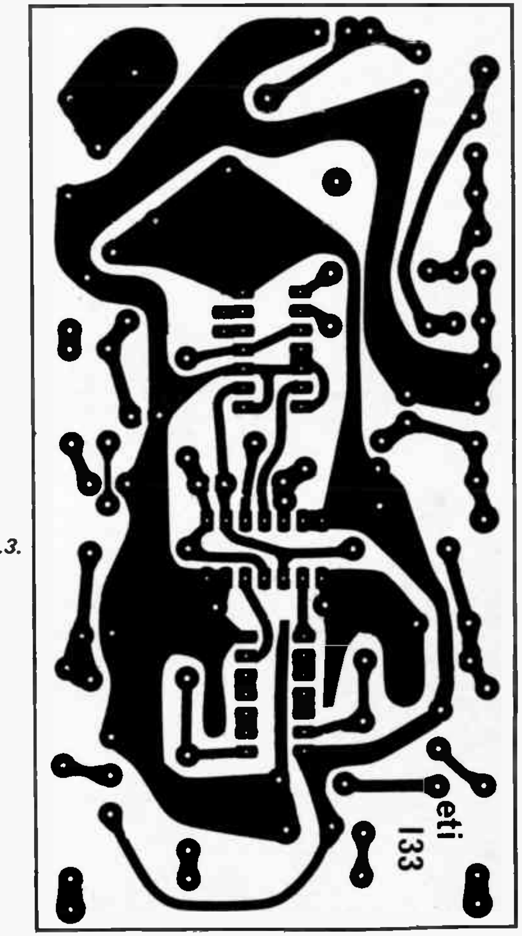


Fig.3.

PARTS LIST ETI 133

Resistors all ½W 5%

R1	22 k
R2	1 M
R3	22 k
R4-R6	1 M
R7,8	220 k
R9	4k7
R10	3k3
R11	68 k
R12,13	39 k
R14,15	10 k
R16	470
R17	1 k
R18	68

Potentiometers

RV1,2	100 k Trim
RV3	10 k "
RV4	4k7 "

Capacitors

C1,2	330 n polyester
C3,4	22 µ 16 V electro
C5	100 µ ceramic
C6,7	10 µ 16 V electro
C8	4µ7 16 V
C9	10 µ 16 V
C10,11	100 µ 25 V
C12	220 n polyester

Semiconductors

IC1	74C04 (CMOS)
IC2	4030 "
IC3	4013 "
IC4	LM309k
Q1	BC548
D1	1N914
LED	

Miscellaneous

PCB ETI 133
Case to suit
Terminals and sockets to suit

PHASE METER

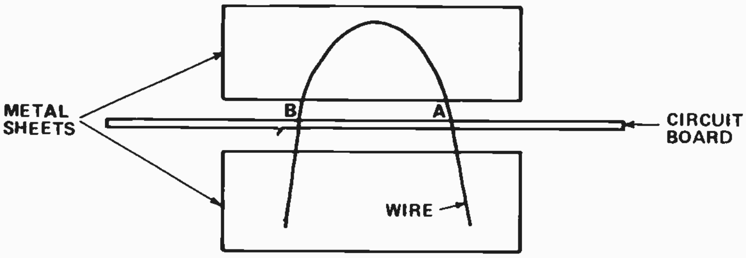


Fig. 5. Details of the shield between IC1 and IC2.

the input stage is extremely high and oscillation can occur if they become coupled to the later stages. To help isolation, small metal sheets, about ¾" x 1½" should be soldered above and below the board between IC1 and IC2. These can conveniently be attached to the link between points A and B in the circuit as shown in Fig. 5. For the same reason, the CAL and USE points should not be taken to a panel-mounted switch but the connection changed on the board itself. We have used two molex pins at these points, marked X in Fig. 4, which work quite satisfactorily.

When the board is assembled, it can be mounted behind the front panel, supported directly by stout wires to the two inputs and the recorder output. Connect a power supply and check that the voltage across C11 is six volts or just under. Calibration and testing are simplified if the leads of C11 and the positive lead of C3 and C4 are left long enough to be able to clip a lead thereon.

CALIBRATION

To calibrate the instrument, first connect capacitor C5 to the CAL position, the meter to be used to the recorder output and a signal of about 100 mV at about 1 kHz to the reference input. Adjust RV1 to give a null reading on the meter. Disconnect C5, leaving the end free, and adjust RV4 to give a convenient reading on the meter to correspond to 180° (eg 180 mV or 45 μA). If this is hard to set exactly, connect a fixed resistance in parallel with RV4 to give better control for any individual meter.

Next connect jumper leads from the positive sides of C3 and C4 to either side of C7 (i.e. one to V+, the other to V−, it doesn't matter which), connect the meter to the main output and adjust RV3 to give a 180° reading (with parallel resistance if needed as in the previous paragraph). Finally, remove the two jumper leads and connect one between the two inputs, connect C5 to the USE position and adjust RV2 slowly and carefully until the LED is just on the point of turning on and off. The meter should now be reading less than half a degree: if not, repeat the calibration procedure.

As a check of proper operation, you should now be able to vary the communal input from millivolts to volts and from sub-audio to over 100 kHz without the phase difference showing more than about one degree. Another excellent test is to connect different signal generators of different frequencies to the two inputs. The output should read exactly 90°, as the signals will be in phase exactly as often as they are out of phase. Our prototype failed this test, reading 92°, and it was only after considerable trouble that we traced this to non-linearity in our trusted (and expensive) multimeter. We guess the moral is to use a digital meter if accuracy is really important. Note that the recorder output is undefined under these conditions.

The high-frequency accuracy is limited by the rise and fall times of the CMOS outputs, by any mismatch in R1 and R3 and their stray capacitances, and by propagation delay differences between the two input and squaring circuits. These, on the two units tested, have been about 50 nsec. This would be equivalent to 1° phase error for every 25 kHz of signal frequency. Thus the meter is usable, but certainly not accurate, up to about one megahertz.

Input protection is provided by resistors R1 and R3 and the internal diodes in the 74C04. We have tested this system to inputs of 80 Vrms before any degradation of the gates occurred, but a value of say 25 Vrms (70 V p/p) should be regarded as a fairly safe working maximum. If IC1 is mounted in a socket, it can be simply changed if accidently overloaded. Under no circumstances can 240 V be applied directly to the inputs!

When using the instrument for measuring phase in 240 V mains circuits, common sense precautions should be observed to ensure no damage occurs to the instrument or the operator! First use a mains testing screw driver to identify the active lead: secondly always switch off the power when connecting or making any alterations to the circuit under test: thirdly make sure that the resistor R makes reliable contact and cannot accidentally become disconnected, otherwise the reference input can get the full mains voltage through the load. Finally use a voltage divider or an oscilloscope X10 probe to reduce the voltage to a safe level.

The circuit shown in Fig. 6 can be set up on an insulating board with a socket for the load to be plugged into. Resistor R is chosen to give a voltage of 1 volt or less when the load current flows through it and must be rated to dissipate a few watts if large currents are to be handled. A value of 0.22 Ω, 5 W is suitable for most situations. And remember that, when set up like this, the instrument reads the phase of the voltage relative to the current.

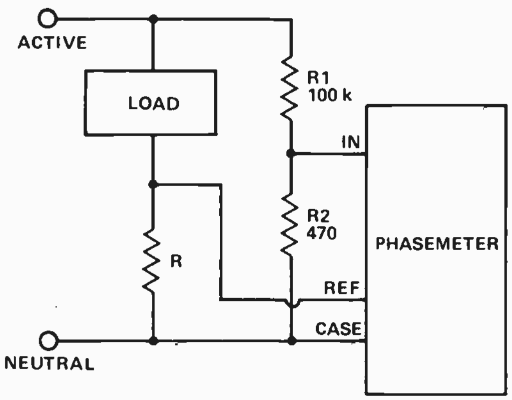
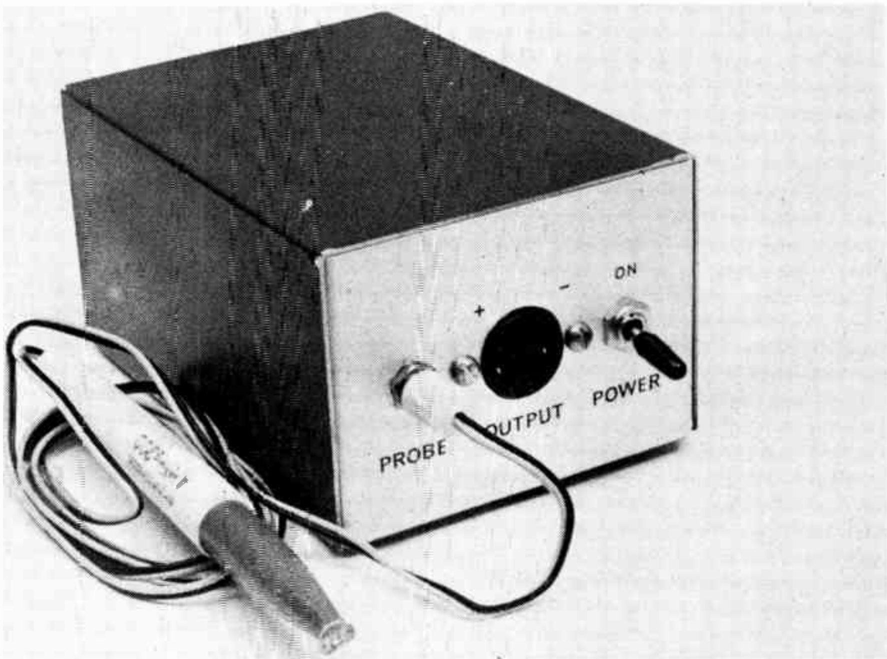


Fig.6. How to use the meter to show the relative phase of mains voltage to mains current.

PROJECT 130 TEMPERATURE METER



Converter connects to any analogue or digital meter.

OUR original design concept for this unit was as a complete instrument based on our ETI 533 digital display using a forward biased diode as the sensor — this generating a temperature-proportional voltage which in turn is supplied to a voltage-to-frequency converter. We planned to use a timebase to generate the necessary strobe and reset pulses. However the cost and complexity of this arrangement was such that we decided against it.

What finally emerged was a simple temperature-to-voltage converter which can be used in front of *any* analogue or digital meter. The converter provides an output of 10 mV/degree which can be either Celsius or Fahrenheit depending on calibration. If a dedicated digital readout is required we suggest that you incorporate the converter with our ETI 118 digital voltmeter.

CONSTRUCTION

Whilst a printed-circuit board is by no means essential, using one certainly makes construction easier and improves the appearance. The potentiometers as shown in our prototype are single turn presets which are quite adequate if an analogue meter is to be used for the readout. However if a digital meter is to be used the extra accuracy of the readout would warrant ten-turn presets being used for RV1 and RV2, as setting accuracy is considerably improved.

The converter quite readily fits into a small aluminium mini-box. Two nine volt batteries are used to power the unit and battery drain is low enough to ensure a life of many months.

A 3.5 mm jack is used to connect the sensor to the unit and the output to the meter is provided via an inexpensive two-pin speaker socket.

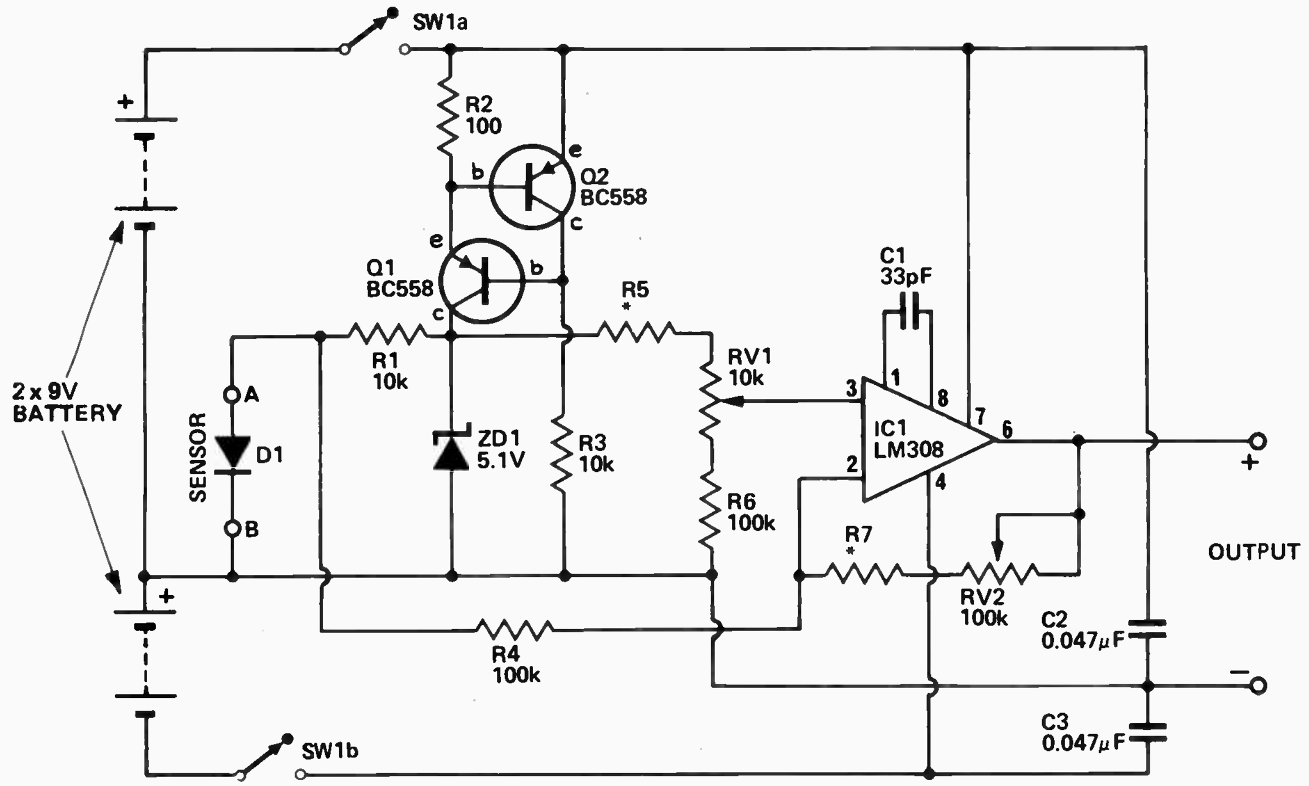
The probe is constructed by mounting the sensor-diode into the tip of a ball-point pen casing, or similar. The method may best be understood by reference to the drawing.

CALIBRATION

To calibrate the instrument, two accurately known temperatures are required. One may be water or oil at room temperature (ice water should not be used as there the temperature may vary several degrees between different points in the solution). The high temperature is best obtained by heating oil or water and allowing it to stabilise at around 80°C. A second smaller heat conductive container filled with water is then immersed in the larger container. This simple procedure prevents errors due to circulating currents in the larger volume of water. An accurate mercury-in-glass thermometer should be used to measure temperatures during the calibration procedure as detailed below.

SPECIFICATION

RANGE	0 to 100°C 32 to 212°F
OUTPUT	10 mV/degree
ACCURACY	± 1°
RESPONSE TIME	3 seconds



NOTE
FOR DEGREES F
R5 = 750k
R7 = 820k
FOR DEGREES C
R5 = 910k
R7 = 470k

Fig. 1. Circuit diagram of the temperature meter.

1. Place the sensor and thermometer into the cool solution, allow a little time for stabilisation, and then measure the voltage from the converter and the temperature. Record these two readings.

2. Place the sensor and thermometer into the hot solution and measure the voltage and temperature as before. The voltage change between the first and second readings should be equal to the temperature change times 10 millivolts.

3. If the voltage versus temperature is not as specified in step 2 adjust RV2 and repeat steps 1 and 2 until it is. Note that varying RV2 changes the voltage at both the hot and the cold positions. It is the correct slope, or rate of change that we are after at the moment.

4. When the correct rate of change has been set as above place the sensor and thermometer into the cool solution and adjust RV1 to obtain a reading of 10 mV per degree. That is if the solution is at 25° C adjust RV1 to obtain a reading of 0.25 V.

Due to the spread of diode characteristics from one device to another the necessarily small adjustment range of RV1 and RV2 may not allow all diodes to be calibrated with the resistor values specified. If this is found to be the case it may be necessary to change the value of R5, R6 or R7.

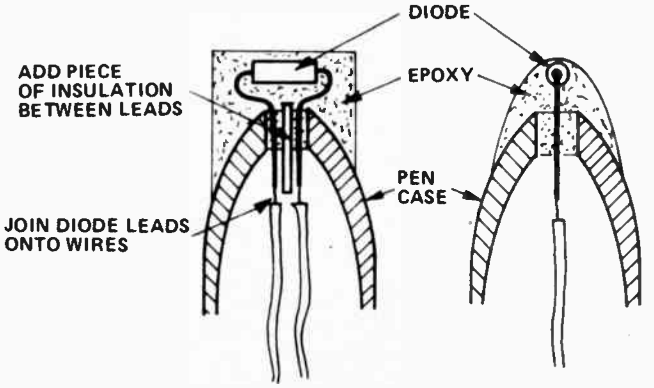


Fig. 2. This diagram shows how the sensor is mounted into a ball-point pen casing or similar.

HOW IT WORKS – ETI 130.

A forward biased diode has a temperature coefficient of about –2 mV/°C. That is the normal voltage across a silicon diode of nominally 0.6 volts will decrease by two millivolts for every degree C increase in temperature. This change with temperature is sufficiently linear over the range of 0 to 100°C to use it as a temperature sensor.

What the ETI 130 circuit does is to amplify this voltage and to provide offset compensation for the normal 0.6 volt drop across the diode.

Transistors Q1 and Q2 provide a constant-current source of about 5 mA into the zener diode ZD1 such that a very stable five volt reference is obtained which is independent of the battery supply voltage. (V supply greater than 6 V.) The forward bias current through the sensor diode is about 0.5 mA as provided by R1. This current is low enough to prevent errors due to self heating of the sensor diode.

The voltage across the sensor diode is amplified by IC1 (a very high input-impedance operational amplifier) whose gain is fixed at the ratio of (R7 + RV2)/R4. The necessary offset is provided by RV1 which is adjusted to cancel the normal 0.6 volt drop across the diode. By selecting the correct values for R5 and R7 as shown on the circuit diagram the indication of temperature in degrees C or F may be obtained.

PARTS LIST

R1,3	Resistor	10k ½W 5%
R2	"	100 ½W 5%
R4,6	"	100k ½W 5%
R5,7	"	See Fig. 1 and test.
RV1	Potentiometer	10k * trim type
RV2	"	100k * " "

*for digital readout a multiturn trim potentiometer is recommended.

C1	Capacitor	33pF ceramic
C2,3	"	0.047µF polyester
D1	Diode	1N914
ZD2	Zener Diode	BZX79C5V1
Q1,2	Transistor	BC558, BC178
IC1	Integrated Circuit	LM308

Metal box
Two 9v batteries
Two pole toggle switch
PC board ETI 130
3.5mm plug and socket
Two pin plug and socket for output

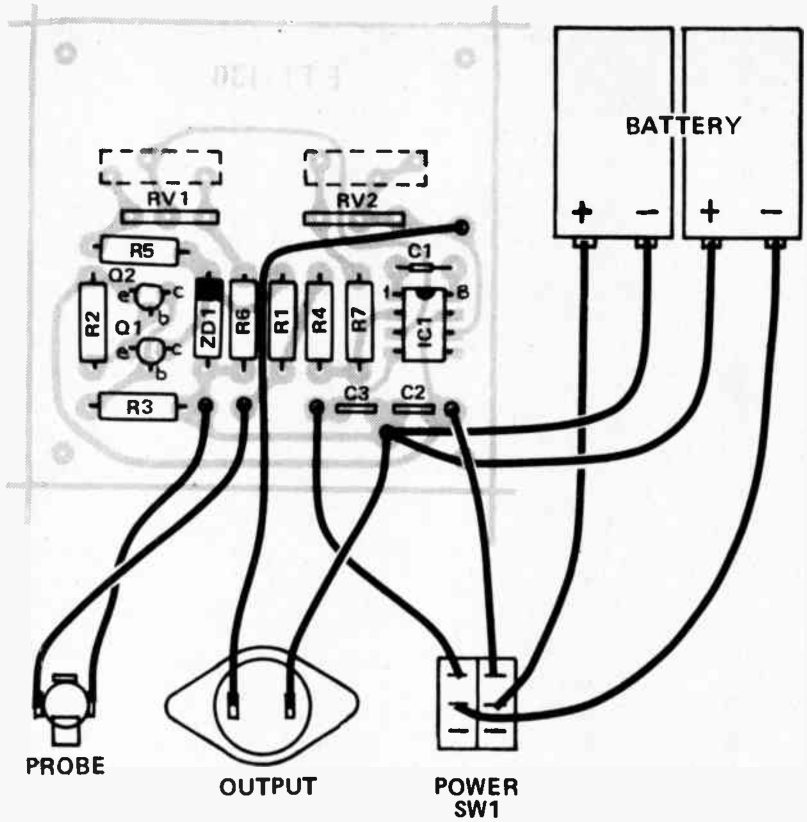


Fig. 3. Component overlay and interconnection diagram.

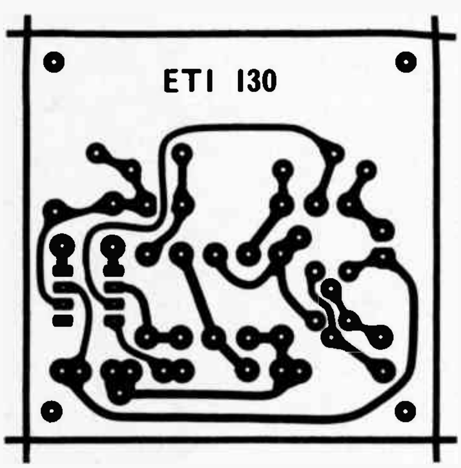
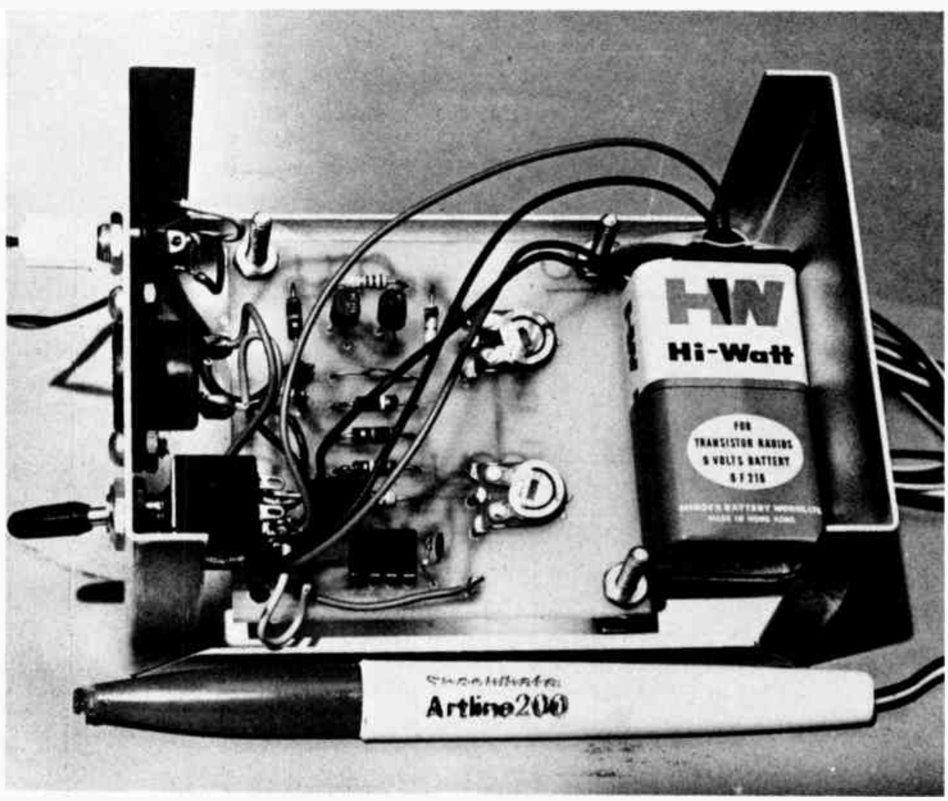


Fig. 4. Printed circuit pattern. Full size 63 x 63 mm.



Internal view of the completed temperature converter. Note also the probe at front.

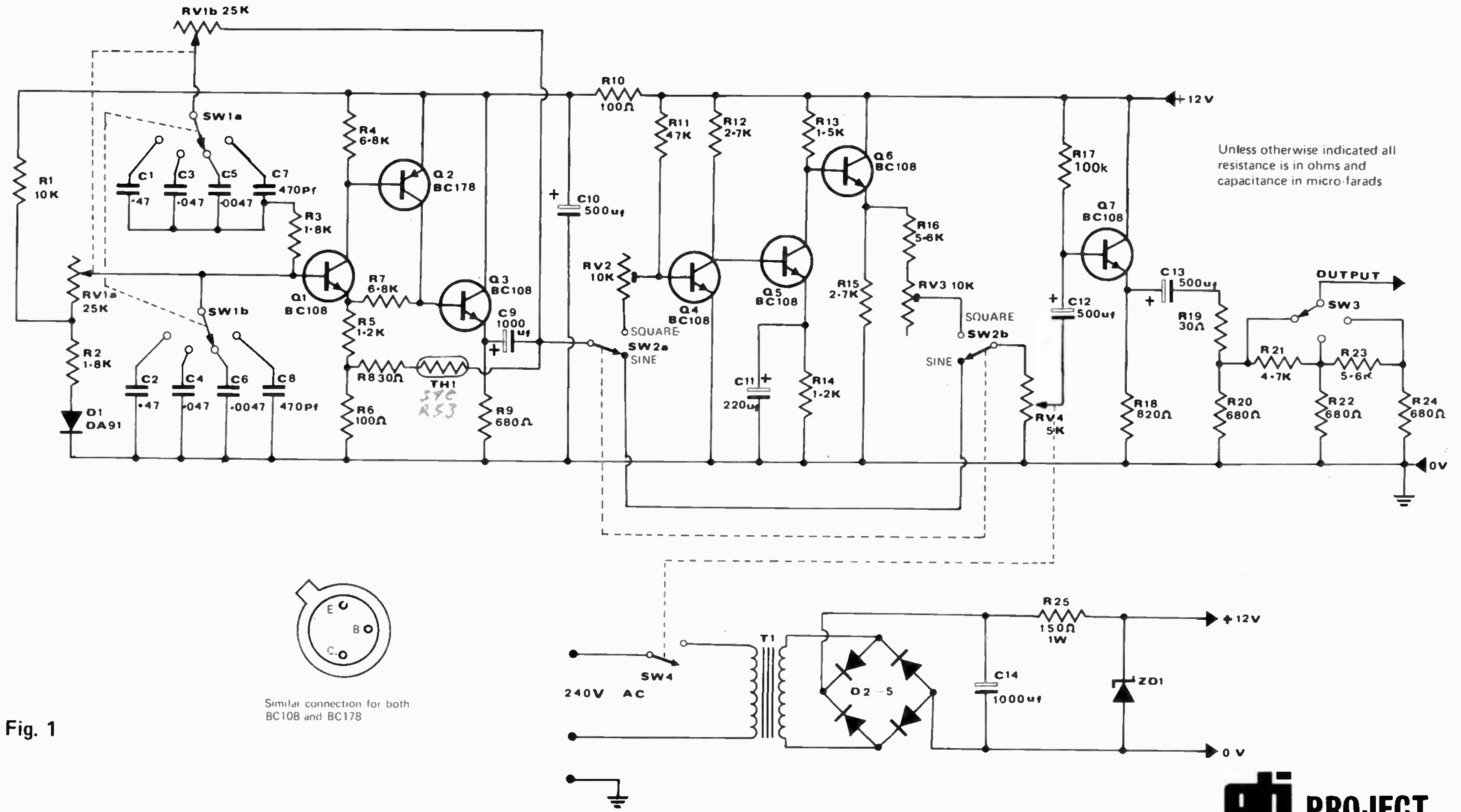


Fig. 1

eti PROJECT 102

Audio Signal Generator

THIS ARTICLE gives all necessary data for constructing a reliable, stable audio signal generator which covers frequencies from 15Hz to 150kHz in four switch-selected ranges.

Both sine and square wave outputs are provided. Sine wave amplitude is one volt rms, square wave amplitude is one volt peak, adjustable by both fine and coarse attenuators in the emitter follower output stage.

The generator uses a total of seven silicon transistors, six of which are npn types and one a pnp type. The Wien bridge oscillator (Q1, Q2, Q3) is a slightly modified version of the well-known Mullard circuit which uses fixed capacitive elements and variable resistance elements in the bridge and includes a thermistor to ensure constant amplitude output. The modifications to the original Mullard Wien bridge circuit have been made to accommodate transistors readily available in Australia.

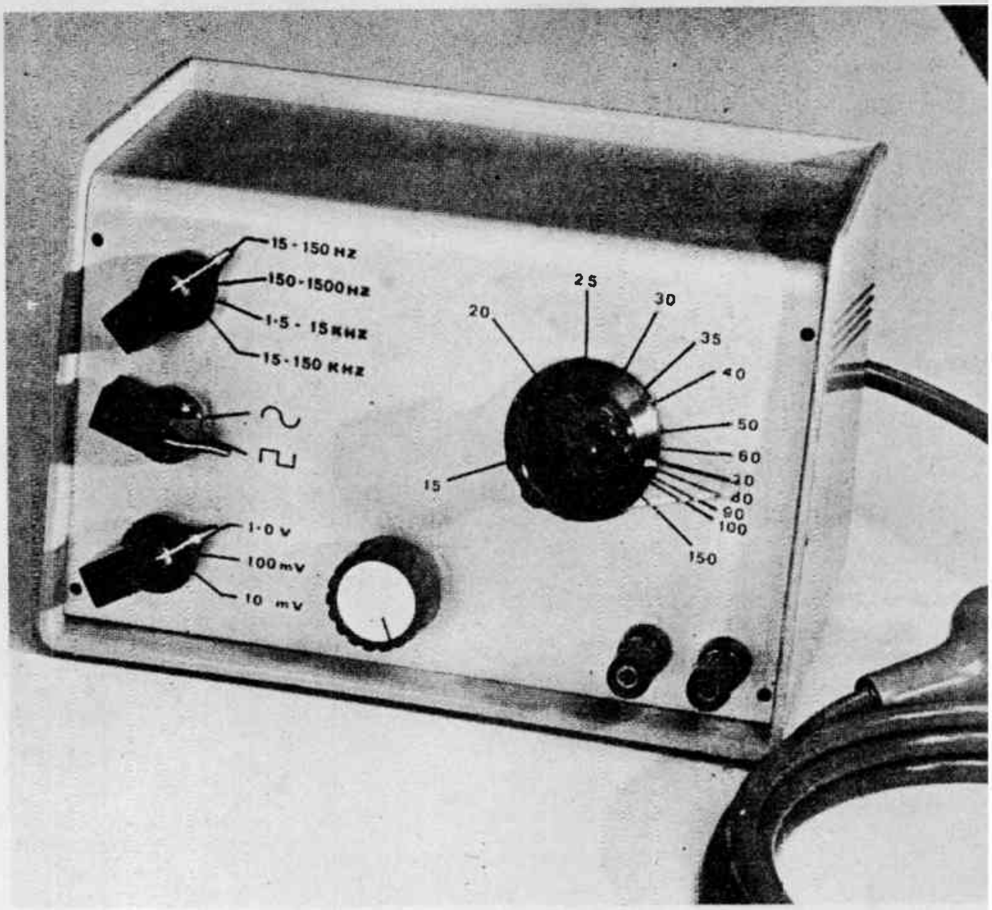
The full circuit is shown in Fig. 1.

Frequency is varied by the ganged potentiometers RV1a and RV1b; these form the resistive elements of the bridge. Constant amplitude output is ensured by the thermistor TH1 in the feedback loop to the emitter circuit of Q1.

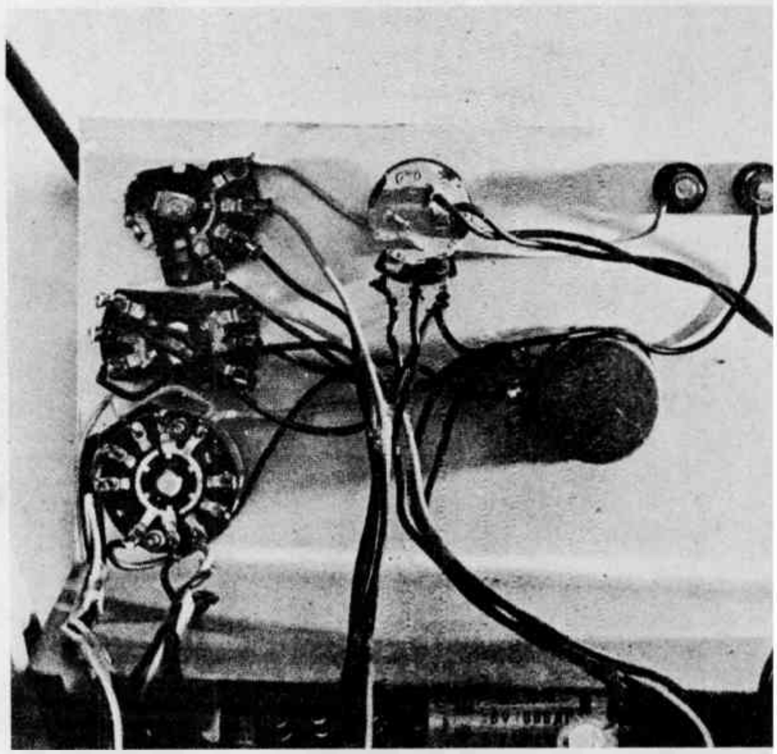
The npn/pnp pair Q1 and Q2 form a high gain amplifier which is coupled to the npn emitter follower Q3.

For sine-wave output the signals are taken via SW2a and SW2b, and the fine attenuator control RV4, to the base of emitter follower Q7. The output of Q7 then goes via the switched attenuation circuit (SW3, R21, R22, R23, R24) which can be used to adjust the sine-wave output to a maximum of 1 volt, 100 mV, or 10 mV rms.

For square-wave signals, the sine-wave output from Q3 is taken to Q4 and Q5, which together form a



Our completed unit


Front panel wiring details.

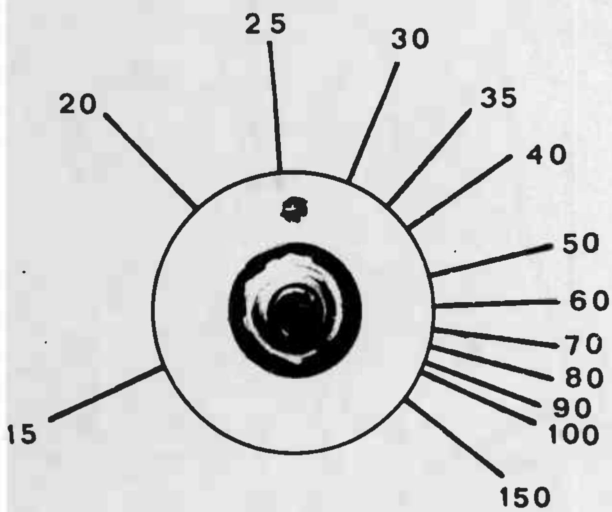


Fig. 5. Calibration scale (full size)

Simply constructed audio signal generator provides adequate performance for home and shop use.

Audio Signal Generator

PARTS LIST

R1	—	Resistor 10k ½ watt 5%
R2	—	" 1.8k " " "
R3	—	" 1.8k " " "
R4	—	" 6.8k " " "
R5	—	" 1.2k " " "
R6	—	" 100Ω " " "
R7	—	" 6.8k " " "
R8	—	" 30Ω " " "
R9	—	" 680Ω " " "
R10	—	" 100Ω " " "
R11	—	" 47k " " "
R12	—	" 2.7k " " "
R13	—	" 1.5k " " "
R14	—	" 1.2k " " "
R15	—	" 2.7k " " "
R16	—	" 5.6k " " "
R17	—	" 100k " " "
R18	—	" 820Ω " " "
R19	—	" 30Ω " " "
R20	—	" 680Ω " " "
R21	—	" 4.7k " " "
R22	—	" 680Ω " " "
R23	—	" 5.6k " " "
R24	—	" 680Ω " " "
R25	—	" 150Ω 1 watt 10%
RV1 (a/b)	—	Dual gang potentiometer 25k linear
RV2	—	Potentiometer preset 10k linear
RV3	—	" " " "
RV4	—	Switch potentiometer 5k linear
C1	—	Capacitor — metallized 0.47 μf, 15 volt
C2	—	" — " 0.47 μf, " "
C3	—	" — " 0.047μf, " "
C4	—	" — " 0.047μf, " "
C5	—	" — " 0.0047μf, " "
C6	—	" — " 0.0047μf, " "
C7	—	Capacitor — metallized 470 pf, 15 volt
C8	—	" — " 470pf, " " (silver mica capacitors may be used for C7 and C8).
C9	—	Capacitor — electrolytic 1000 μf, 16 volt
C10	—	" — " 500 μf, " "
C11	—	" — " 220 μf, " "
C12	—	" — " 500 μf, " "
C13	—	" — " 500 μf, " "
(Note C10—C13 are single ended type capacitors)		
C14	—	" — " 1000 μf, " "
D1	—	diode — OA91 (or equivalent)
D2-5	—	diodes EM401 (or equivalent)
ZD-1	—	zener diode BZY96, 12v., 1 watt (or similar)
Q1	—	Transistor BC108
Q2	—	" BC178
Q3—Q7	—	Transistors BC108
TH1	—	Thermistor type STC type R53
T1	—	Transformer 240 volt to 12 volt 150 ma (Ferguson type PF 2851 or similar)
SW1 (a/b)	—	Switch rotary, two pole, four way
SW2 (a/b)	—	Switch rotary, two pole, two way
SW3	—	Switch rotary, single pole, three way
SW4	—	(rear of switch pot. RV4)
PC ET-006	—	Printed circuit board
Case	—	ATC type plastic case
Various	—	Output terminals, connecting block, mains cable and plug, control knobs.

Errata: Capacitors C1,C2,C3,C4,C5,C6,C7 and C8 should be 100 volts not 15 volts as shown above.

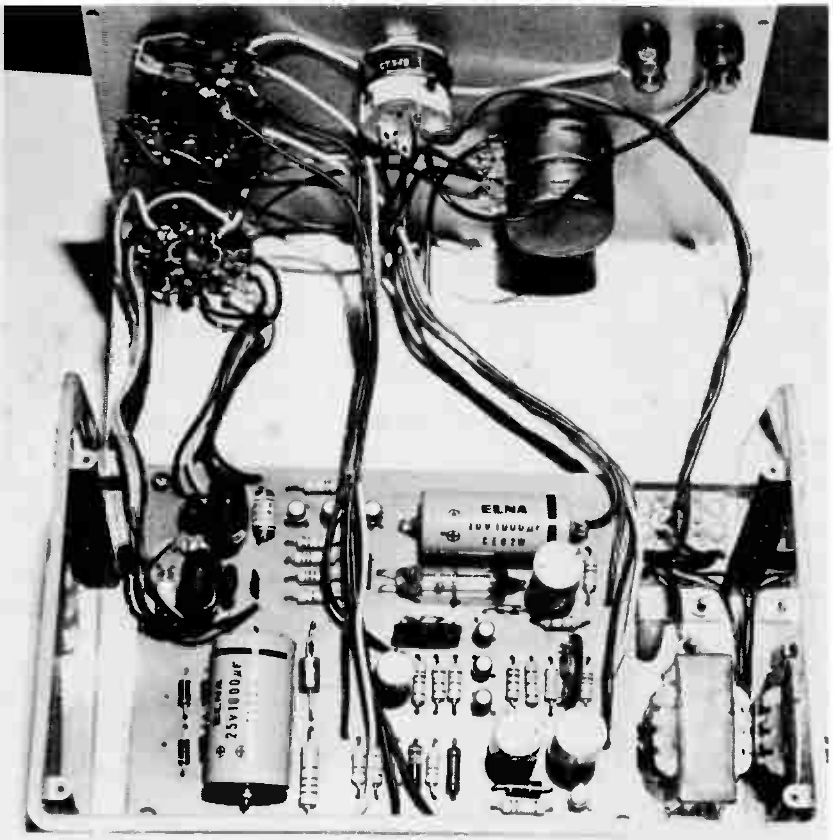
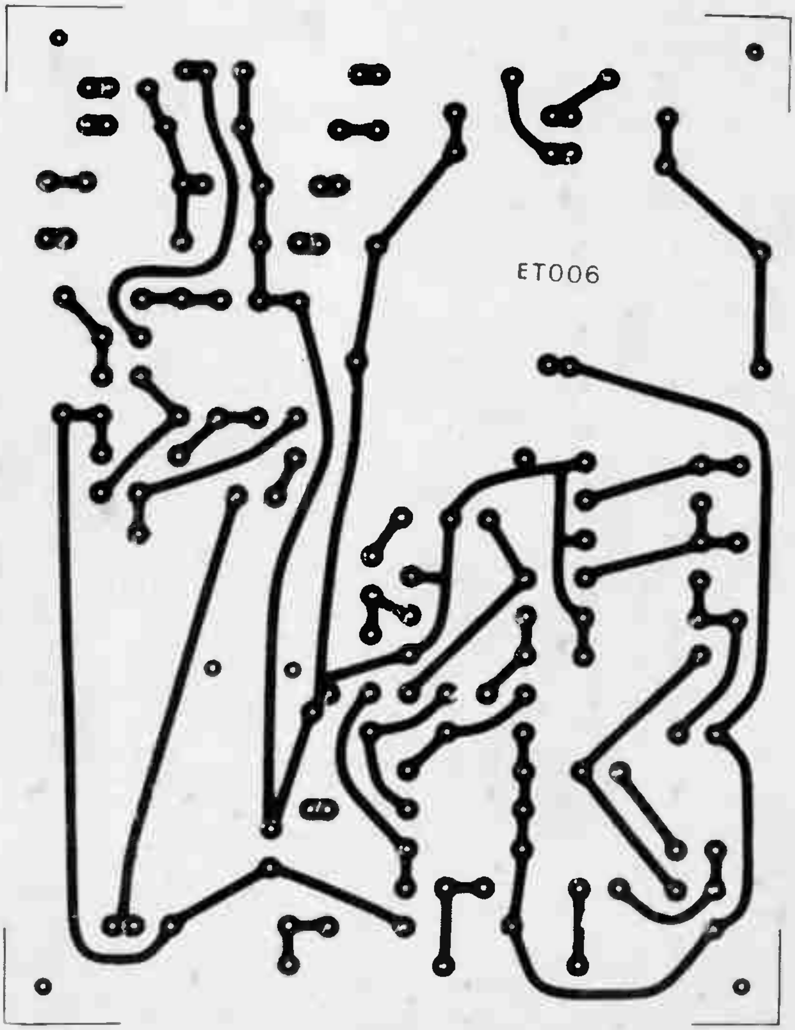


Fig. 4. General layout of components, in our final version we found it was desirable to use screened signal leads between the circuit board and the panel mounted switches.

Fig. 2. Foil pattern — full size



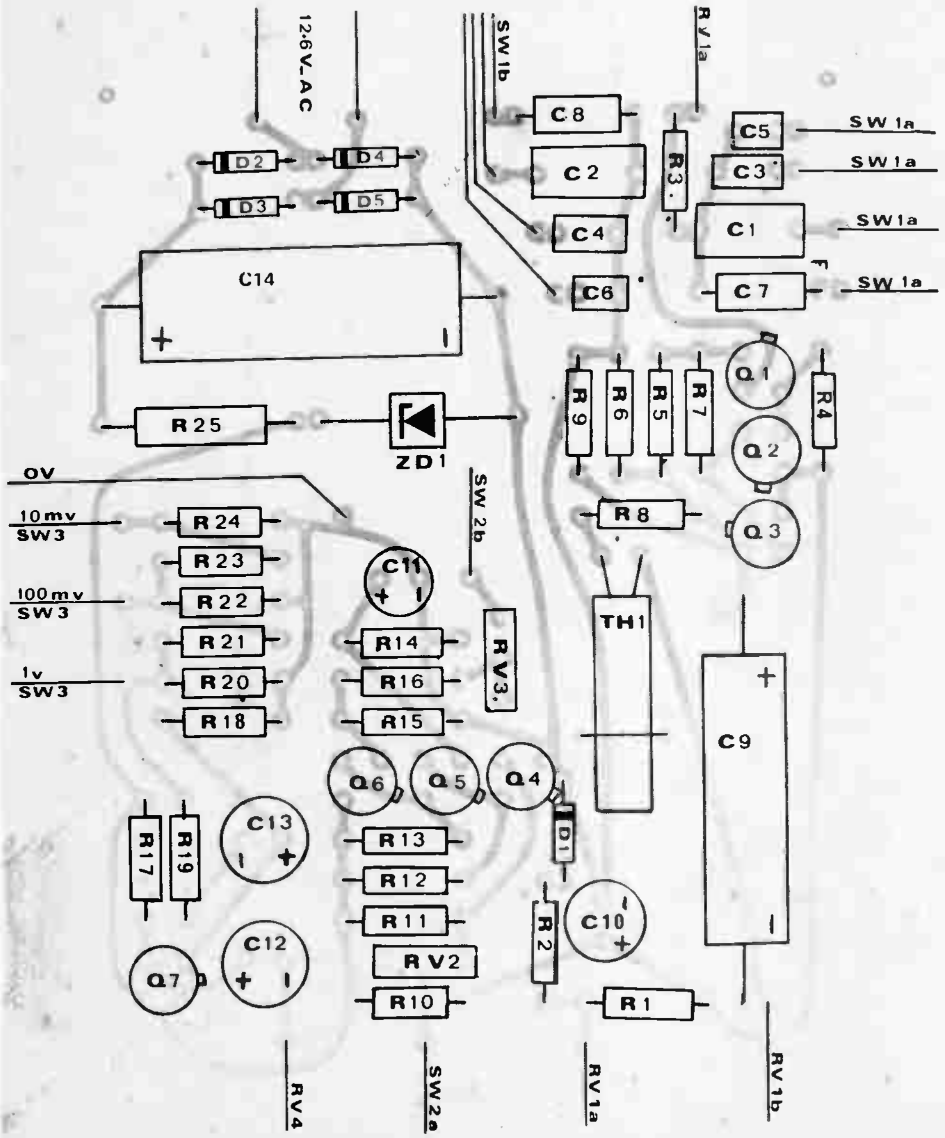
Frequency range 1:	15Hz – 150Hz
Frequency range 2:	150Hz – 1,500Hz
Frequency range 3:	1,500 Hz – 15,000Hz
Frequency range 4:	15,000Hz – 150kHz
(range 4 will in fact extend beyond 150kHz)	
Output variation:	Less than ±1dB from 15Hz – 150kHz.
Distortion:	<1%
Output impedance:	600 ohms.
Sine-wave output:	0-1 volt (rms)
	0-100mV (rms)
	0-10mV (rms)
Square-wave output:	0-2 volts (peak-to-peak)
	0-200mV (peak-to-peak)
	0-20mV (peak-to-peak)
Square-wave mark/space ratio:	nominally 1:1
Square-wave rise time:	less than 1 μsec.

squaring amplifier. The pre-set potentiometer RV2 is used to set the mark-space ratio to 1:1. The squared signal is then taken from the emitter follower Q6, through pre-set attenuator RV3, and then to the output transistor Q7

The preset potentiometer RV3 is adjusted to produce a maximum square-wave output of 2 volts peak-to-peak. The coarse output attenuator will reduce this to either 200mV or 20mV peak-to-peak. Each of three output levels (both sine-wave and square-wave) is then steplessly variable from zero to maximum by means of RV4.

A circuit for a 12 volt power supply has been included in Fig. 1. However, if the generator is used infrequently or for short periods, it can be operated from a 12 volt battery. Current drain is less than 30mA.

Our prototype unit was built on a printed circuit board the foil pattern of which is reproduced in Fig. 2. The layout of components is shown in Fig. 3; compare this with Fig. 4, which is a photograph of the completed unit.



If desired, the unit can be assembled on Veroboard. The layout of components is not too critical, although we have found that it is necessary to use screened leads from the oscillator board to the output level switch, and from the output level switch to the output terminals.

It is essential that, metallized capacitors be used for C1, C2, C3, C4, C5 and C6. Silver mica capacitors may be used for C7 and C8 (470pF). DO NOT use ceramic capacitors.

The frequency range scale illustrated in Fig. 5 will prove accurate enough for most audio work. If greater accuracy is required, the unit must be checked against another oscillator of known accuracy.

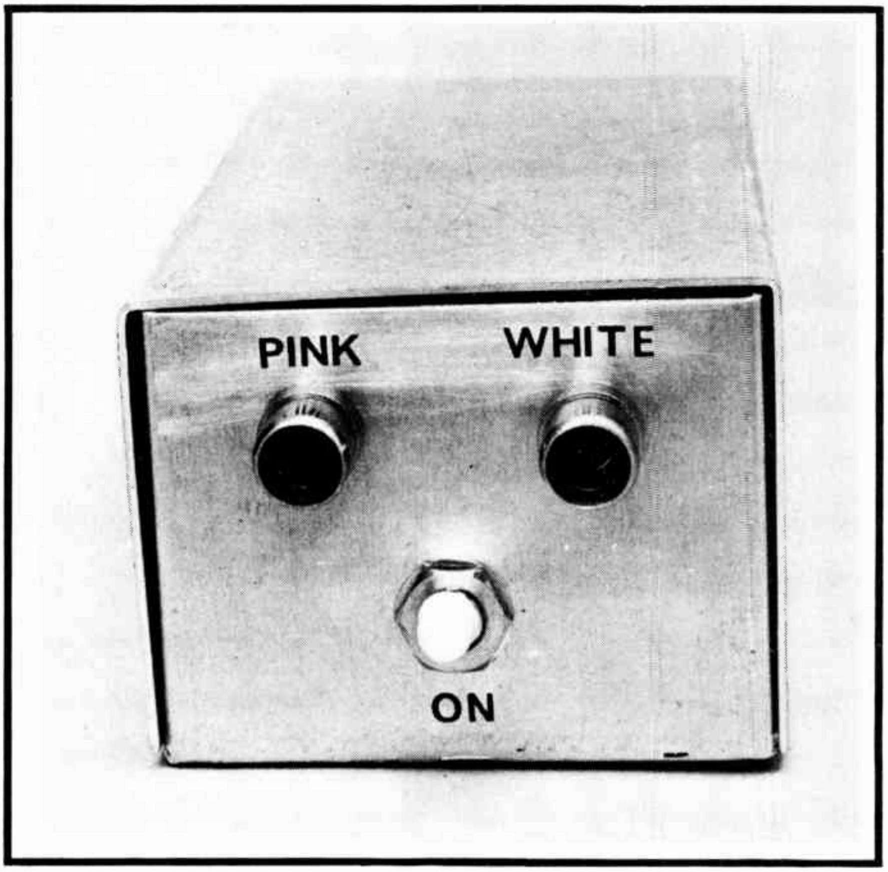
The output waveform should, if possible, be checked with an oscilloscope to verify that a good sine-wave is being obtained and that the square-wave is uniform. The mark-space ratio of the square-wave output should be set to 1:1 by the pre-set potentiometer RV2 and the peak-to-peak output level by RV3. If no oscilloscope is available, RV2 and RV3 should be set to mid-range.

If an audio frequency voltmeter is available, the sine-wave output should be checked from 15Hz to about 100kHz to ensure that reasonably constant output is maintained.

The output signal should remain constant within ±1dB up to 100kHz or so, but there may be changes in level whilst changing frequency. The envelope stability (or the time taken for the amplitude to stabilize after a change in frequency) is mainly a function of the quality of the dual ganged potentiometer.

Performance of the prototype unit is detailed above. ●

AUDIO NOISE GENERATOR



Simple circuit generates both white and pink noise.

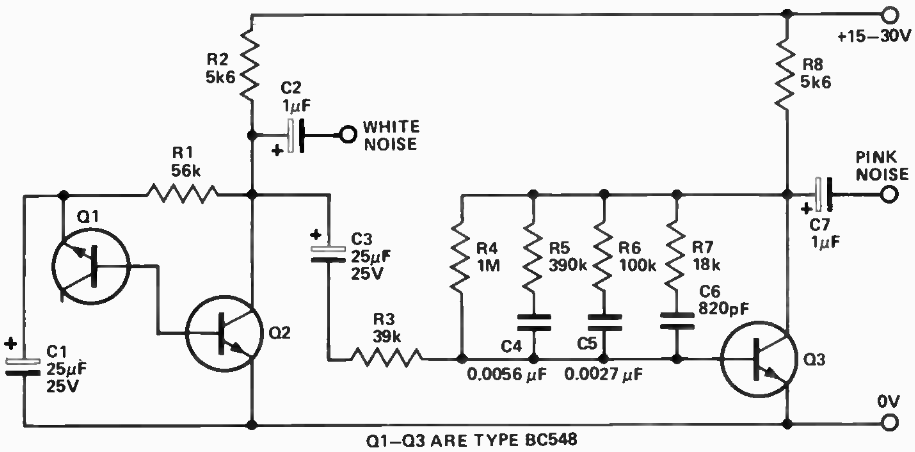


Fig. 1. Circuit diagram of the noise generator.

NOISE is generally an undesirable phenomena that degrades the performance of many measurement and instrumentation systems. It therefore seems strange that anyone should want to generate noise, but this is often the case.

Noise generators are often used to inject noise into radio-frequency amplifiers in order to evaluate their small signal performance. They are also used to test audio systems, and as random signal sources for wind-like effects in electronic music.

There are two commonly used noise source characteristics, 'pink' and 'white'. White noise is so called because it has equal noise energy in equal bandwidths over the total frequency range of interest. Thus, for example, a white noise source would have equal energy in the band 100 to 200 Hz to that in the band 5000 to 5100 Hz.

HOW IT WORKS – ETI 441

In the days when vacuum tubes were in common use the most commonly used form of noise generator was a vacuum-tube diode operated in the current saturation mode. Nowadays noise generators may be very complex indeed. Highly complex digital generators which produce psuedo-random digital noise may cost many thousands of dollars. An example of a simpler type of digital noise source may be found in our synthesizer design (see International Music Synthesizer 4600 ETI December 1973). However for audio work of a general nature the most commonly used, and the simplest, method is to use a zener diode as a noise generator.

Transistor Q1 is in fact used as a zener diode. The normal base-emitter junction is reverse-biased and goes into zener break-down at about 7 to 8 volts. The zener noise current from Q1 flows into the base of Q2 such that an output of about 150 millivolts of white noise is available.

The 'zener', besides being the noise source, also biases Q2 correctly, and the noise output of Q2 is fed directly to the White Noise output.

To convert the white noise to pink a filter is required which provides a 3 dB cut per octave as the frequency increases. A conventional RC network is not suitable as a single RC stage gives a cut of 6 dB per octave. Hence a special network of Rs and Cs is required in order to approximate the 3 dB-per-octave slope required. Since such a filter attenuates the noise considerably an amplifier is used to restore the output level. Transistor Q3 is this amplifier and the pink noise filter is connected as a feedback network between collector and base in order to obtain the required characteristic by controlling the gain-versus-frequency of the transistor. The output of transistor Q3 is thus the pink-noise required and is fed to the relevant output socket.

PARTS LIST — ETI 441

R1	Resistor	56k	½W	5%
R2	"	5k6	½W	5%
R3	"	39k	½W	5%
R4	"	1M	½W	5%
R5	"	390k	½W	5%
R6	"	100k	½W	5%
R7	"	18k	½W	5%
R8	"	5k6	½W	5%
C1	Capacitor	25μF	25V	electro
C2	"	1μF	25V	electro
C3	"	25μF	25V	electro
C4	"	0.0056μF		polyester
C5	"	0.0027μF		polyester
C6	"	820pF		ceramic
C7	"	1μF	25V	electro

Q1-Q3 Transistor BC548, BC108 or similar
PC board ETI 441
CASE
BATTERIES
OUTPUT SOCKETS

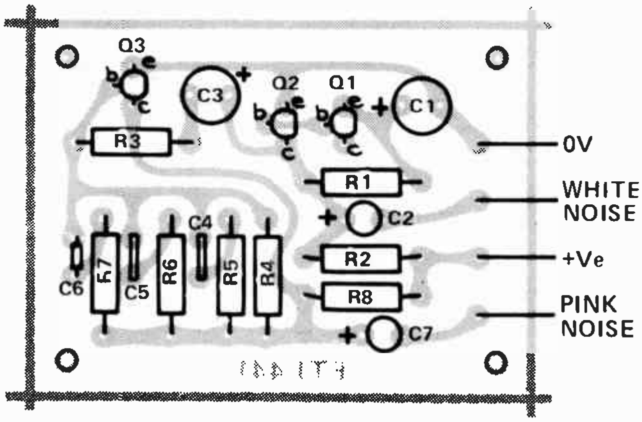
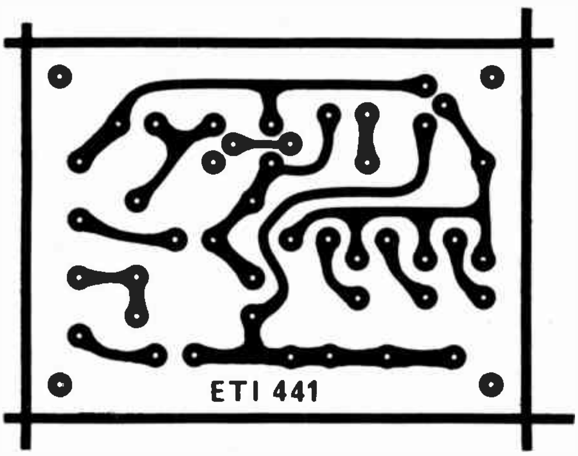


Fig. 2. Component overlay.



Printed circuit layout. Full size 67 x 49 mm.

If white noise is filtered or modified in any way it is referred to as coloured noise or, often more specifically, as 'pink' or 'grey' noise. The term pink noise should be restricted to the noise characteristic that has equal energy per percentage change in bandwidth. For example with true pink noise the energy between 100 Hz and 200 Hz should equal that between 5000 Hz and 10 000 Hz (100% change in both cases).

Pink noise therefore appears to have more bass content than does white noise, and it appears to the ear to have a more uniform output level in audio testing. To change white noise to pink noise a filter is required that reduces the output level by 3 dB per octave (10 dB per decade) as the frequency is increased. The ETI 441 Noise Generator is designed to provide both white and pink noise as required.

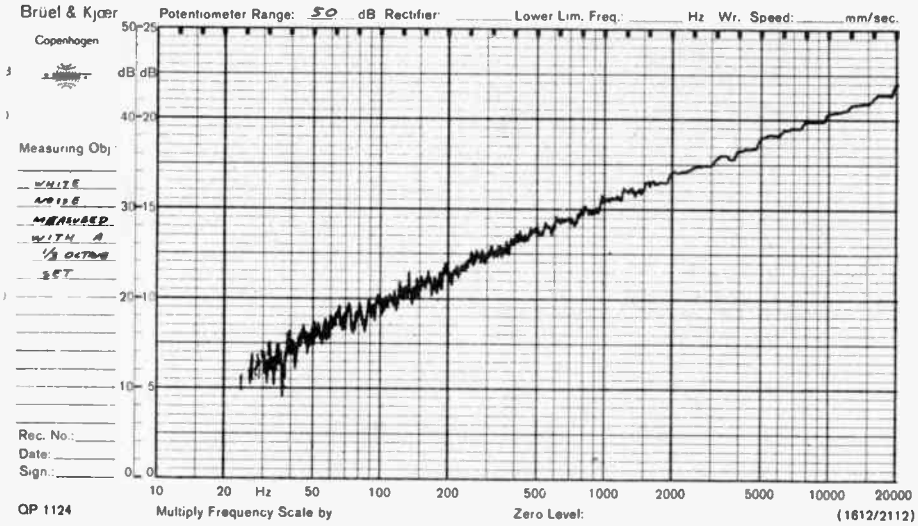
CONSTRUCTION

Construction is relatively simple and almost any of the common methods, such as Veroboard or Matrix board, may be used if desired. For neatness and ease of assembly it is hard to beat a proper printed-circuit board and for this reason we have provided details of a suitable board.

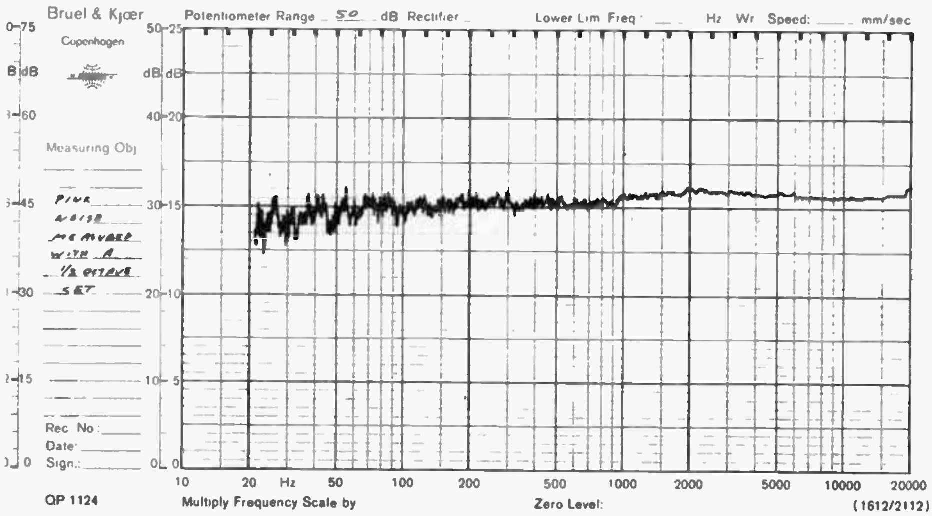
Almost any type of NPN transistor will do for the generator provided that the one used for Q3 has a gain of 100 or more. If BC548 type are used watch out for the two different pin connections used by different manufacturers.

For use as a separate instrument in general experimentation the unit will need to be powered by a pair of nine-volt batteries. However if the unit is to be built into some other piece of equipment, as is often the case, any supply within the equipment which has an output of between 15 and 30 volts dc will be suitable.

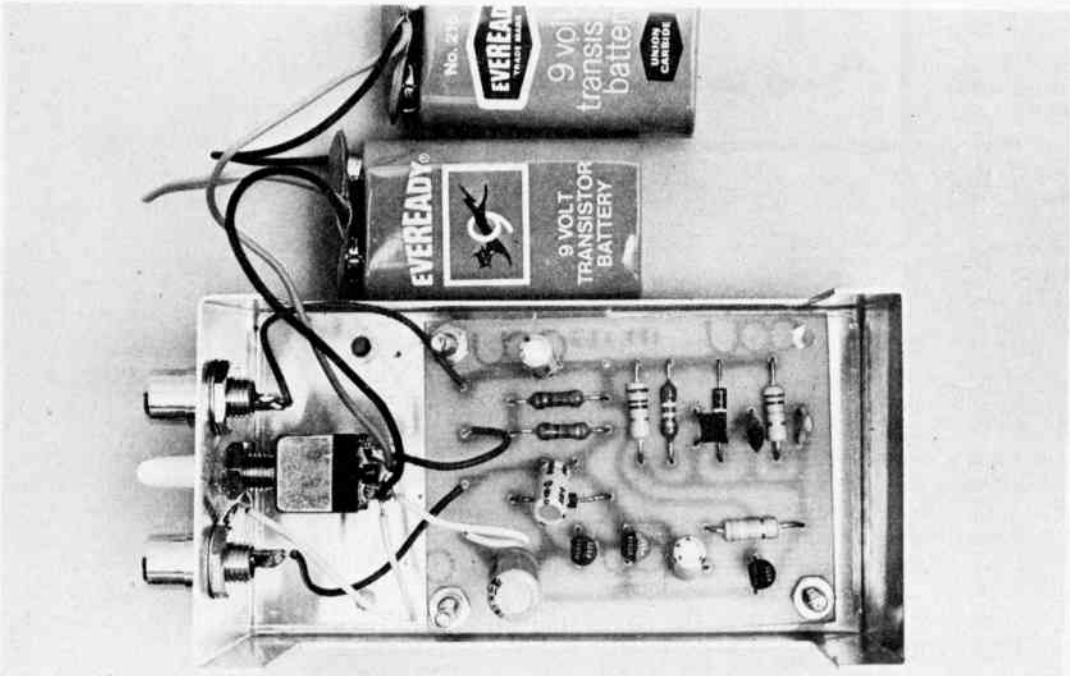
AUDIO NOISE GENERATOR



Amplitude of white noise versus frequency as measured with a one-third octave filter set.



Amplitude of pink noise versus frequency as measured with a one-third octave filter set.



Internal layout of the generator.

PROJECT 124

TONE BURST GENERATOR

A valuable tool for testing loudspeakers.

TONE BURST testing is a technique which is rapidly gaining acceptance in a wide variety of applications. Typical applications are in testing of hydrophones, signal-to-noise in telephone channels, reverberation chamber testing and in the determination of peak distortion in loudspeakers. With loudspeakers, tone burst testing has the further advantage that the speakers may be tested with their maximum peak power level whilst keeping the average sound output level low enough to not annoy the neighbours — a considerable advantage indeed.

Some time ago our audio consultants, Louis Challis and Associates, asked us to build them a tone-burst generator and the resulting instrument has been used by them ever since with much success. The tone-burst test has been mentioned in several speaker reviews and, as a result, many people have asked for constructional details of this instrument.

DESIGN FEATURES

A tone burst must always be an integral number of cycles. If the burst is switched on or off part way through a cycle then undesirable transients will be produced that will mask the test results. Thus the burst must start and end exactly at the zero-crossing point of the sine wave in the burst.

In the original unit, designed for Louis Challis, preset times can be independently selected for the on and off periods of the burst with the exception that the burst time is automatically modified to give an integral number of cycles. The preselected on/off ratio, however, is independant of the burst frequency. To give the required control range, six switched ranges as well as a variable control are provided for both the on and off periods. Other features of the original unit are the ability to start at any point in the cycle as well as the zero crossing point, a phase-inverting switch to select either the positive or the negative half cycle first and an OFF LEVEL control to set a base tone level which is modified when the tone burst occurs. In addition the dc level of the output can be set and a switch is provided to select burst, pure tone or off as required.

When it came to redesigning the unit as a project we decided that many of the features offered by the original design were unnecessary for the user concerned only with testing speakers. Hence the unit has been redesigned in a greatly simplified form.

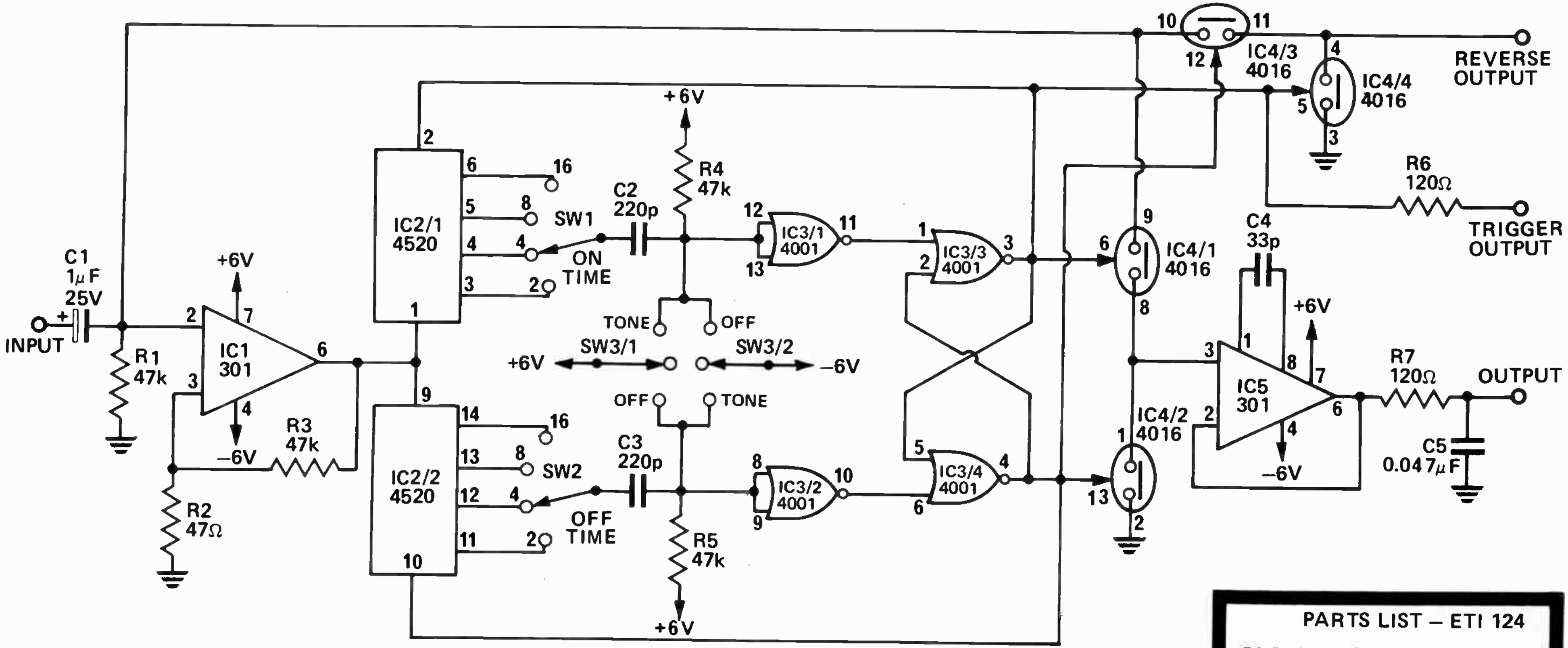
Instead of using monostables to generate variable on/off times we now divide the input with a counter to

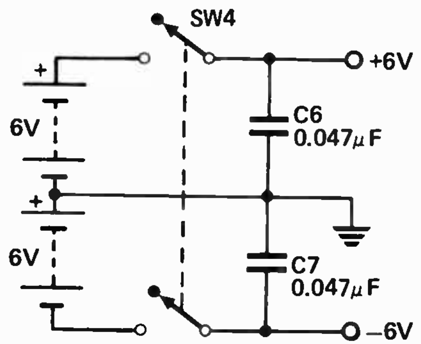
MEASURED PERFORMANCE

TONE BURST GENERATOR.

On Time Cycles.	2,4,8 or 16
Off Time Cycles	2,4,8 or 16
Frequency Response 3 Hz – 300 kHz	+0 –3 dB
Distortion 3 V input at 1 kHz	<0.02%
Input Level	
Maximum	3 V RMS
Nominal range	100 mV to 1 V
Input Impedance	47 k
Output Noise Voltage with no input	<25 µV
Power Supply Current	4 mA

TONE BURST GENERATOR





POWER RAILS OF IC2, IC3, AND IC4 NOT SHOWN
PIN 16 OF IC2 IS +6V
PIN 8 OF IC2 IS −6V
PIN 14 OF IC3 AND 4 IS +6V
PIN 7 OF IC3 AND 4 IS −6V
PIN 7 AND 15 OF IC2 ARE RESET PINS AND −6V

Fig. 1. Circuit diagram.

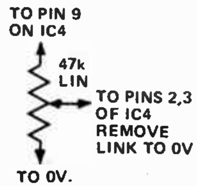


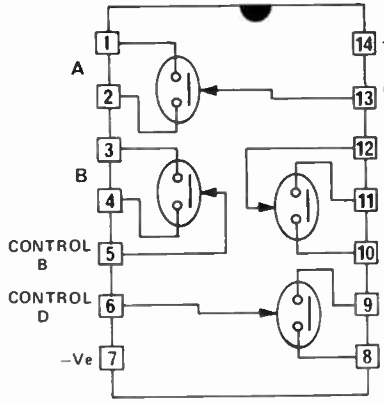
Fig. 4. How to add a potentiometer to the generator for burst-on-tone operation. That is the generator gives a continuous tone level with tone bursts of higher amplitude at intervals.

PARTS LIST – ETI 124

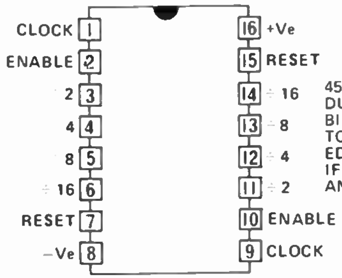
R1	Resistor	47 k	¼W	5%
R2	"	47 Ω	¼W	5%
R3	"	47 k	¼W	5%
R4	"	47 k	¼W	5%
R5	"	47 k	¼W	5%
R6	"	120	¼W	5%
R7	"	120	¼W	5%
C1	Capacitor	1 µF 25V electro		
C2	"	220 pF ceramic		
C3	"	220 pF ceramic		
C4	"	33 pF ceramic		
C5	"	0.047 µF polyester		
C6	"	0.047 µF polyester		
C7	"	0.047 µF polyester		
IC1	Integrated Circuit	LM 301A		
IC2	" "	4520 (CMOS)		
IC3	" "	4001 (CMOS)		
IC4	" "	4016 (CMOS)		
IC5	" "	LM301A		

SW1 Switch 1 pole 4 position rotary
SW2 Switch 1 pole 4 position rotary
SW3 Switch DPDT Toggle with centre off
SW4 Switch DPDT Toggle

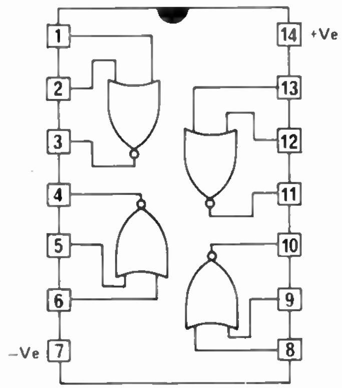
PC Board ETI 124
8 AA size batteries
2 4-way battery holders and clips
Plastic case
Escutcheon
3 single RCA sockets
2 knobs



4016
4066
QUAD ANALOGUE SWITCH. ON IF CONTROL IS HIGH
ON RESISTANCE
4016 TYP 280Ω
4066 TYP 80Ω



4520
DUAL 4 BIT BINARY UP COUNTER
TOGGLE ON POSITIVE EDGE OF CLOCK
IF ENABLE IS HIGH AND RESET IS LOW.



4001
QUAD 2 INPUT NOR GATE

INPUTS		OUTPUT
A	B	
0	0	1
0	1	0
1	0	0
1	1	0

Fig. 3. Pin connections of the ICs used in the generator.

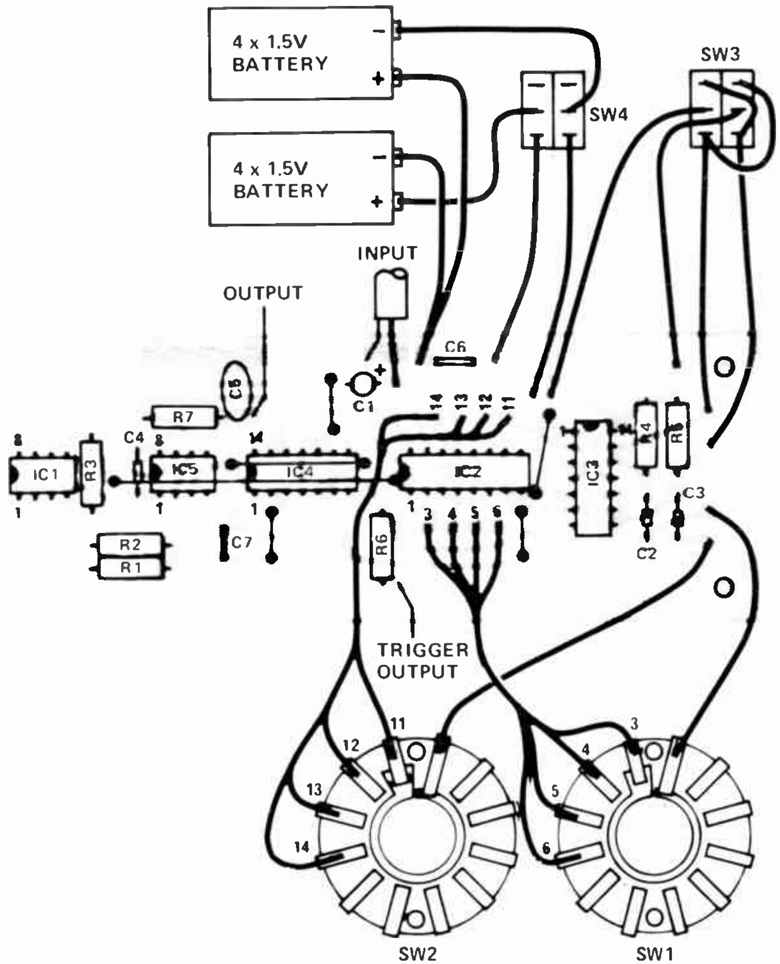


Fig. 2. Component overlay and interconnection diagram. Note that there are six links on the board, including two under IC4, which should be installed first.

HOW IT WORKS – ETI 124

The input signal is squared by comparator IC1 such that the output of the comparator will be high if the input is above +6 mV, and low if the input signal is below −6 mV. Resistors R2 and R3 provide the necessary positive feedback to cause the IC to act as a comparator. The output of the comparator is connected to both clock lines of IC2. If the enable line is high these counters (IC2) will toggle at the input frequency.

IC3/3 and IC3/4 form an RS flip flop where the output must be in either a high or a low state, that is the flip flop has only two stable states. If the output of IC3/3 is high IC2/1 is allowed to clock and, after the number of input pulses selected by SW1 have been counted, the output from SW1 goes low. This low is coupled to the flip flop by C2 toggling the flip flop, disabling IC2/1 and enabling IC3/2. After the number of cycles, as selected by SW2, have been counted the flip flop is again toggled. IC3/1 and IC3/2 are used to square up the pulses generated by C2 and C3 respectively.

The input signal is also coupled to the output buffer, IC5, by the analogue switch IC4/1. When this switch is closed (control signal high) the output of the buffer will be the same as the input. When switch IC4/1 is open IC4/2 will be closed and the output will be held at zero. Since these switches are controlled by the flip flop the output will be the required tone burst.

A trigger output is taken from the flip flop to synchronize an oscilloscope if required. A second output is also available from pins 4/11 of IC4 which is the reverse of the main output.

Switch SW3 forces the flip flop into either of its two possible states thus allowing continuous tone or no output to be selected as required. In the centre position the normal tone burst is obtained.

TONE BURST GENERATOR

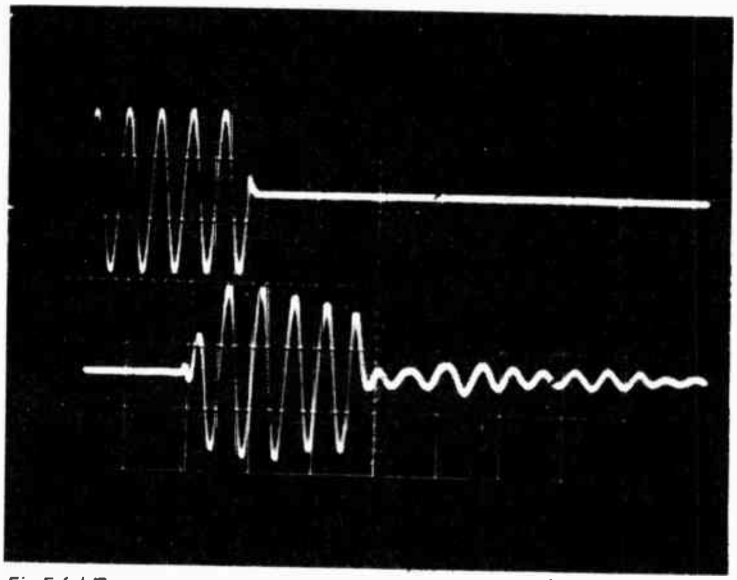


Fig.5 (a) Top trace – the input tone burst of five cycles. (original design).
(b) Bottom trace – the response of a low-cost speaker at 1 kHz. Note the reduced amplitude of the first half cycle and that ringing has added another cycle at the end of the burst. The room reflection can be seen on the trace after the burst.

obtain times that remain in the same ratio regardless of input frequency. We settled for the ability to select 2, 4, 8 and 16 cycles for the duration of either period, as this compromise greatly simplifies the circuitry. We still have the switch to select tone, tone burst or off, but the OFF LEVEL control has been deleted. The latter control may quite easily be added, however, as shown in Fig. 4. . The output dc level control and the starting-point phase change have also been deleted.

Since we only need half of a CMOS 4016 IC, to give the required output, the other half may be used to give an inverse output if required, that is, the reverse output is on when the other is off and vice versa. This output is not buffered or brought out to the front panel. If it is intended to load this output with less than 47 k it is recommended that a 4066 IC be used instead which will handle loads down to 10 k. For loads of lower impedance than this, a buffer such as is on the normal output should be used.

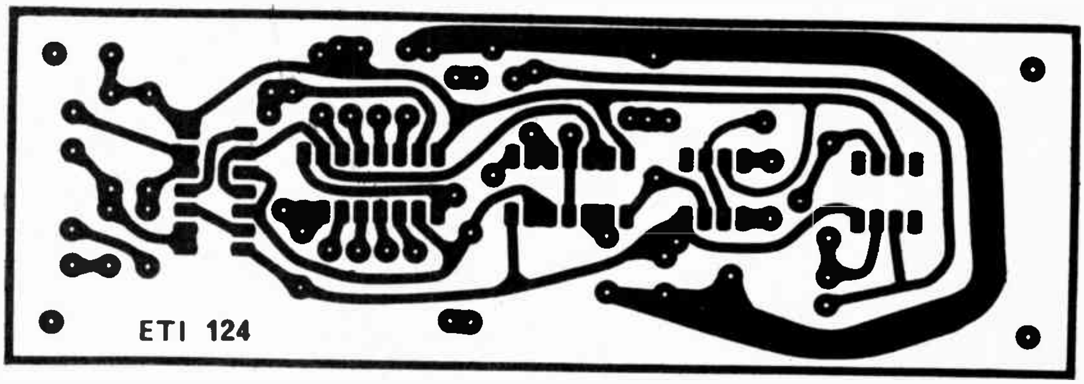


Fig.6. Printed circuit board for the Tone Burst Generator Full size. 142 x 47mm.

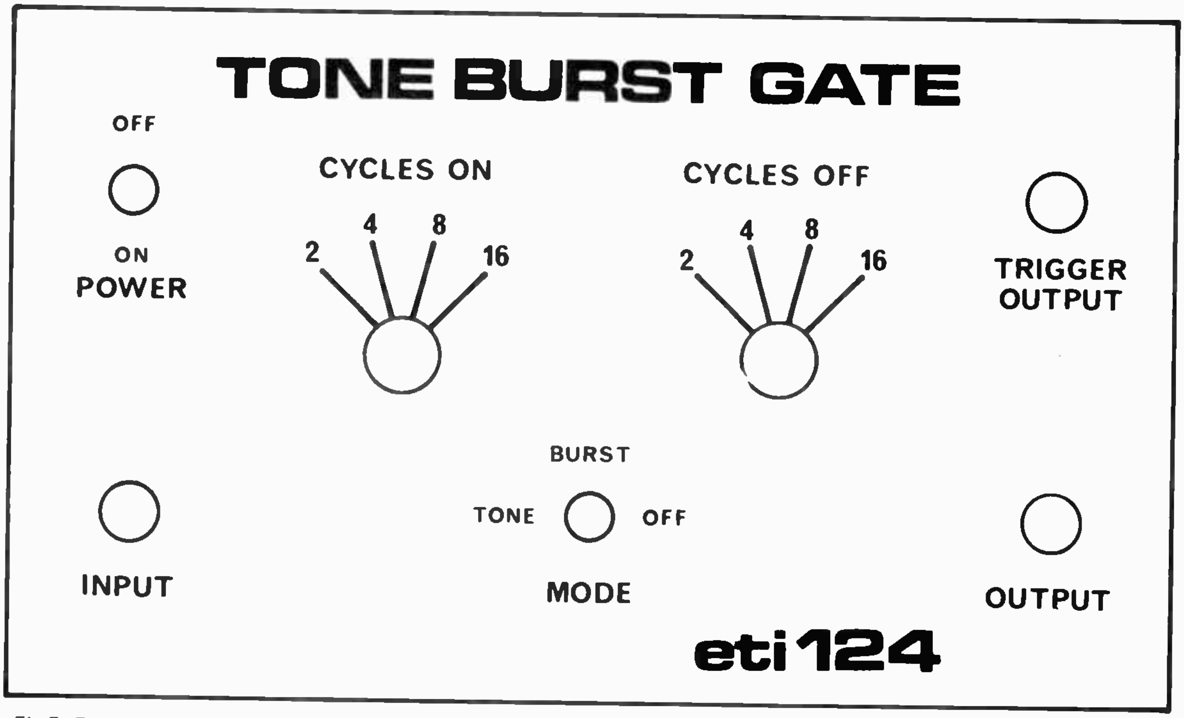


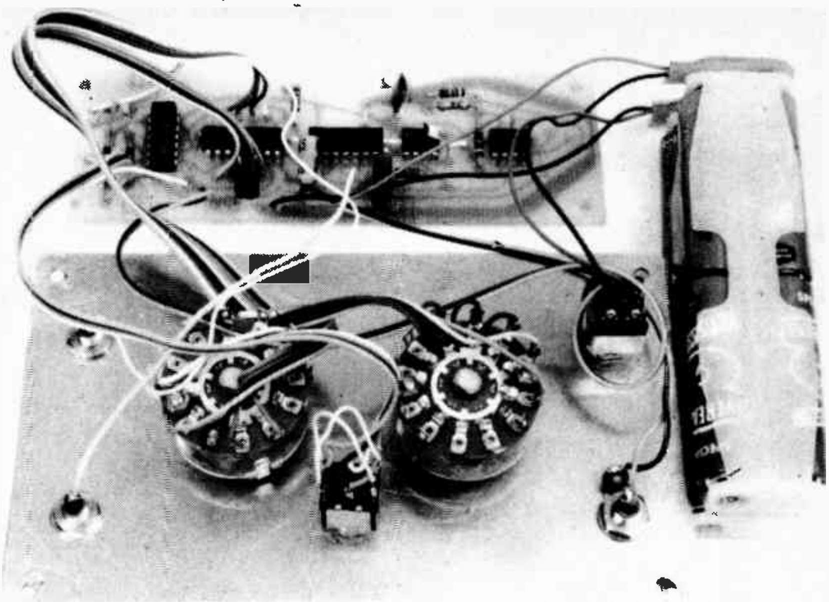
Fig.7. Front panel artwork

CONSTRUCTION

As with any project construction is greatly simplified if a printed circuit board is used. However the layout of the unit is not critical and any other suitable method, such as Veroboard or Matrix board may be used if desired. We strongly recommend that sockets be used for the CMOS ICs, especially if a printed circuit board is not used, as these devices are quite easily damaged when soldering. The use of IC sockets also facilitates later servicing. Also remember that, unlike TTL, all unused inputs of CMOS must be connected to either the positive or negative supply rail.

The plastic box that we used measured 160 x 95 x 50 mm and is very convenient in that the printed circuit may be held in position by sliding it down behind two of the pillars to which the front panel is screwed. The front-panel overlay on the prototype was made from Scotchcal but, as the amount of lettering required is quite small, this may readily be done directly on the panel by hand or with Letraset.

Shielding of the internal wiring is not required providing that the unit is kept away from strong 50 Hz fields. If operation in the vicinity of strong fields cannot be avoided then the unit should be mounted in a diecast box.



How the unit is assembled.

USING THE UNIT

The testing of loudspeakers is very difficult indeed and much effort is still being spent to find test methods which will not only give an accurate understanding of the relative effectiveness of the design, but which will be easy to reproduce.

One of the main problems with speaker testing is that the speaker cannot easily be isolated from its environment. For example, reflections from the walls of a room modify the response, seen by a microphone, no matter where the microphone is placed in the room. If one could eliminate reflections then the situation would be improved considerably, and hence the use of anechoic (echo free) chambers for testing speakers. But such chambers are very expensive to build and consequently not readily accessible to the amateur.

A further problem is in assessing the transient power handling capability of the speaker. Speakers will handle far greater peak transient power than is indicated by their RMS power rating. This is a very important attribute of loudspeakers in handling musical transients. Any attempt to assess this with a sinewave signal may result in the destruction of the speaker due to thermal failure — apart from also being extremely noisy.

The use of a tone-burst generator minimizes both these problems. How this is achieved is better understood by examination of Fig.5. This shows on the upper trace a five cycle 1000 Hz burst that is fed to a loudspeaker. The second trace shows the same burst as picked up by a microphone in front of the speaker. We notice that the burst has been changed by the speaker and an examination of these changes can tell us a lot about the speaker. For example we notice that the first half cycle has not reached full amplitude and this indicates that the speaker would have some difficulty in reproducing high frequency transients. Next we notice that instead of five cycles there are now at least five and a half. This could mean one of two things. Either there is a speaker/room resonance or, the speaker itself is continuing to vibrate after the original excitation has ceased. Which is it? We can determine this by changing the position of the speaker to see if any change occurs in the shape of the burst, if not it is caused by the speaker itself, and if it does then it is a speaker/room resonance. A speaker that lengthens the burst unduly will sound muddy in that region. Of course the speaker must be examined over its whole range to gain a thorough assessment of performance.

It is of course possible to eliminate room reflections simply by performing the tests outside. However unless one lives in a very quiet area, background noise will introduce problems — and your neighbours are unlikely to appreciate the noise that you will generate.

By varying the off period we can also select a ratio where the room reflection, the oscillation seen after the cessation of the burst, does not interfere with the first few cycles of the burst and the response versus frequency of the speaker may then be assessed from the amplitude of the first half cycles that are stable in amplitude. Thus it is possible to gain an appreciation of the frequency response, transient performance and quality in terms of ringing of a speaker by careful use of the tone-burst technique.

The transient power handling capability of a speaker may be assessed by selecting a fairly long off to on ratio for the burst and by feeding the burst to the speaker via a high-power amplifier. If for example an off to on ratio of 8:1 is used then the peak power will be eight times the average power. Thus the speaker may safely be driven to a peak level where a predetermined amount of distortion occurs. Take care that the amplifier is capable of providing the peak power required.

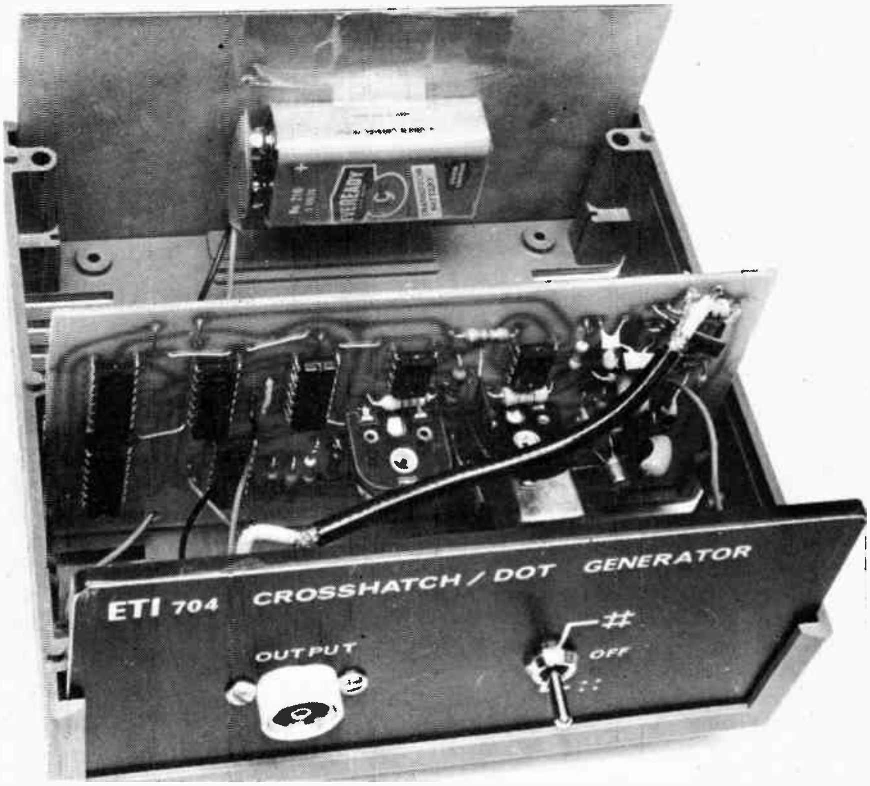
Of course a tone-burst generator may be used for a wide range of testing. We have mainly concentrated in this article on its application to the testing of loudspeakers.

The circuitry of the tone-burst generator may easily be modified for use as a 'silent switch' for A/B speaker testing. The method of doing this is shown on pages 106 and 107.

eti PROJECT 704

CROSSHATCH/DOT GENERATOR

Inexpensive unit for converging colour TV set.



THE COLOUR television picture is created in the receiver picture tube by three separate electron guns – one each for red, green and blue. As these guns cannot be in the same physical position they need to be converged into one spot on the screen.

The process of converging at the centre of the screen is called static convergence and is performed by magnets on the yoke assembly.

However, the screen of the picture tube is not everywhere coincident with the deflection plane and this causes errors when the beam is deflected away from centre. These deflection errors are corrected electronically by 12 or more controls and the process is known as dynamic convergence.

An important part of the process is the use of a crosshatch generator to provide horizontal and vertical lines on the screen. Using the generator, the convergence errors are immediately apparent and the controls on the set are usually labelled with the effect each has on a crosshatch pattern.

In addition to setting up convergence the generator pattern may also be used to set up horizontal and vertical linearity and to orientate the deflection yoke coils on both black and white and colour sets.

Most of the inexpensive pattern generators, which are currently available, produce a video waveform, which must be injected into the correct place in the TV, and require a synchronizing signal from the TV set. Such generators are thus fiddly things to use.

The ETI 704 generator produces a combined horizontal and vertical-sync waveform and this, together with the crosshatch video, is modulated onto a carrier frequency operating in VHF channel 6 (175.25 to 180.75 MHz). Thus to use the generator one simply attaches it to the antenna terminals and selects channel 6.

CONSTRUCTION

Coil L1 should be constructed from 24 gauge B&S enamelled copper wire by winding 6 turns, close spaced, around a former, such as a knitting needle, so that the finished outside diameter of the coil is about 5 mm.

Coil L2 is constructed by winding 4 turns of 24 gauge B&S enamelled copper onto a miniature Neosid former which is fitted with a VHF slug and an aluminium can. Fit links to the board in accordance with Fig. 2 and then the above coils and other components can be assembled to the printed circuit board with the aid of the component overlay. Take particular care with the orientation of ICs and other polarized components. Assemble the CMOS devices to the board last of all and handle them as little as possible. Avoid touching the pins.

Assemble the output socket and switch to the front panel and connect the output of the module to the socket by means of a short length of 75 ohm coaxial cable. The connections to the switch and battery may then be made with ordinary hookup wire.

SETTING UP

Connect the unit to the antenna terminals of a television receiver and select Channel 6. Adjust the coil L2 to obtain the strongest signal on the screen. (This may be totally out of sync at this stage.)

Now adjust RV2 as you would a normal horizontal sync control to obtain vertical lines and then adjust RV1 for vertical sync. Then readjust L2 for clearest picture and make small adjustments to RV2 and RV1 to obtain the most stable crosshatch.

Finally adjust brightness and contrast of the set to obtain white lines on a black background. These adjustments need only be made on initial set up and henceforth the generator is simply attached to the antenna terminals and switched on.

CROSS HATCH-DOT GENERATOR

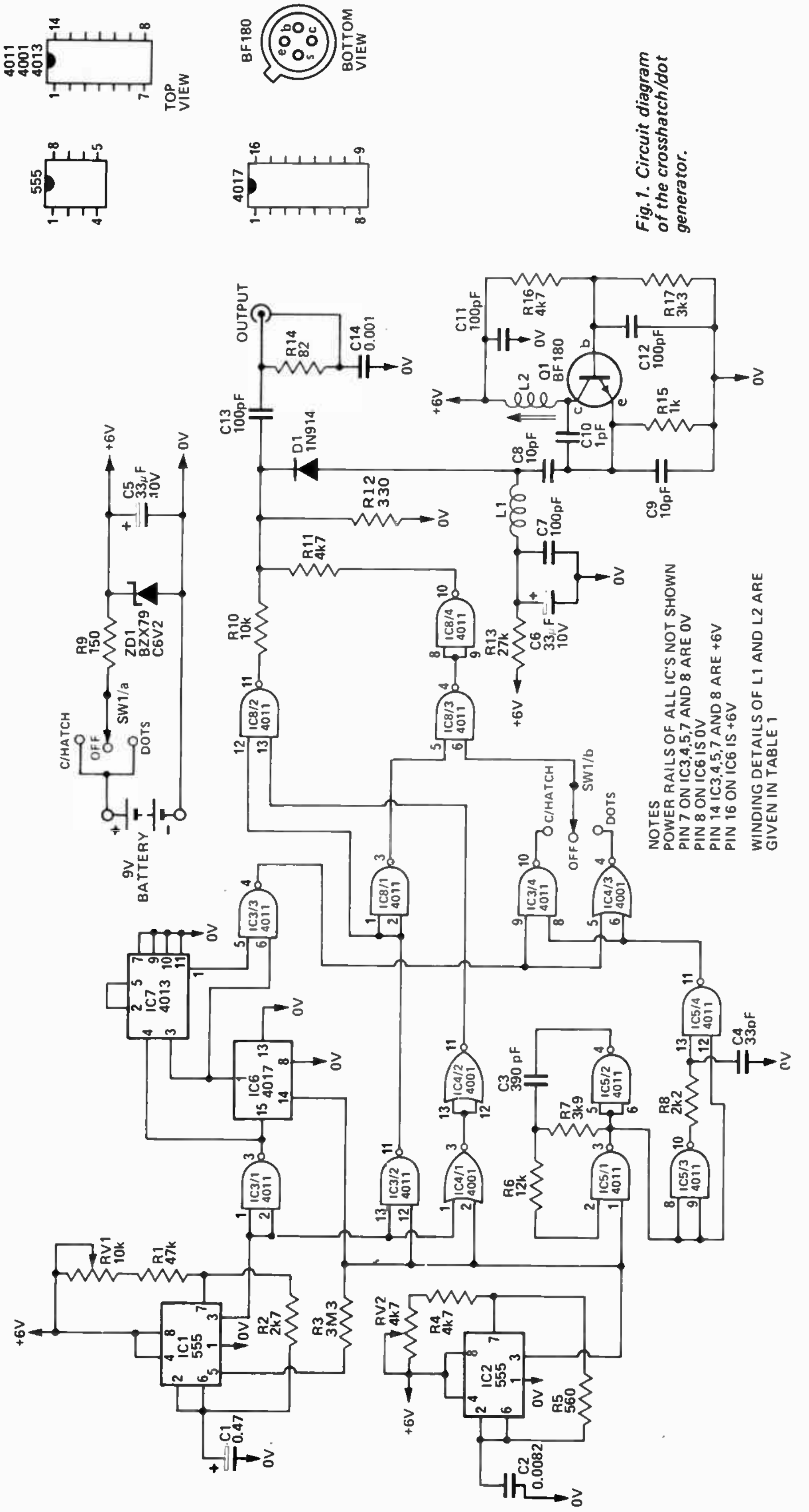


Fig.1. Circuit diagram of the crosshatch/dot generator.

HOW IT WORKS – ETI 704

Two 555 timers, IC1 and IC2, are used as the frame and line sync pulse generators respectively. Integrated circuit IC1 generates one millisecond wide pulses at 20 millisecond intervals (50 Hz) and IC2 generates five microsecond wide pulses at 64 microsecond intervals (15625 Hz). Light synchronization of IC1 to IC2 is achieved by means of R3. Thus both oscillators have to be close to the correct frequency before locking will occur.

Gates IC3/2, IC4/1, IC4/2 and IC8/2 form an exclusive-OR function on these two sync-pulse trains to produce a combined sync-pulse train at the output of IC8/2.

At the end of each line-sync pulse an oscillator, formed by IC5/1 and 2 is gated on, and produces a train of pulses at approximately 240 kHz. The leading edge of each of these pulses triggers monostable IC5/3 and IC5/4 such that a 40 nanosecond wide pulse is generated. Thus approximately 14 40 nanosecond wide pulses are generated between successive line sync pulses. These pulses produce the vertical lines of the crosshatch.

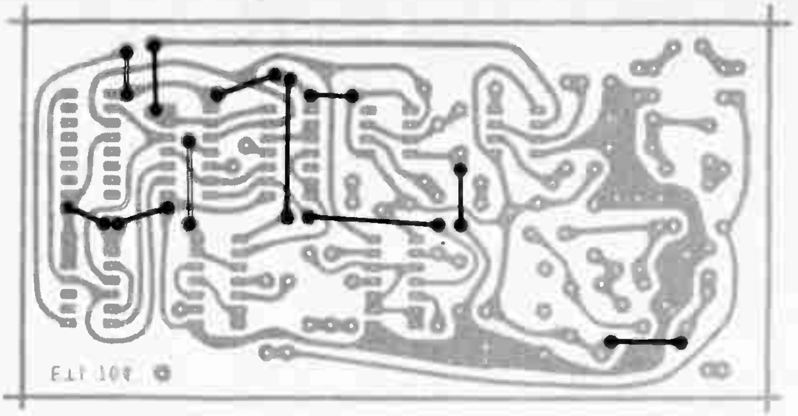
At the end of each frame sync pulse decade counter IC6 is enabled. This is a CMOS Johnson decade counter which provides an output at pin 1 every tenth line sync pulse commencing from the 5th pulse after the counter is enabled. This output is divided by two by IC7 and the output of IC3/3 is therefore low for the duration of every twentieth line period. This output is the horizontal lines of the crosshatch.

The vertical and horizontal crosshatch signals are NANDed and NORed by IC3/4 and IC4/3 respectively to provide either crosshatch or dots as selected by SW1/b. These outputs are inhibited during the line and frame sync periods via IC3/2, IC8/1 and IC8/3. The output from IC8/4 is thus the composite crosshatch video signal.

The composite sync from IC8/2 and the composite video from IC8/4 are summed into R12 by R10 and R11 and form a combined sync and video waveform which modulates the RF from oscillator Q1 via diode D1.

Transistor Q1 and its associated components form an oscillator which runs at around 180 MHz. The output from the generator is therefore a modulated RF signal at channel 6 frequency which is adjustable by tuning coil L2.

The unit is powered from a 9 volt battery which is only on when dots or crosshatch are selected. The 9 volts is regulated down to 6 volts by means of R9 and zener diode ZD1.



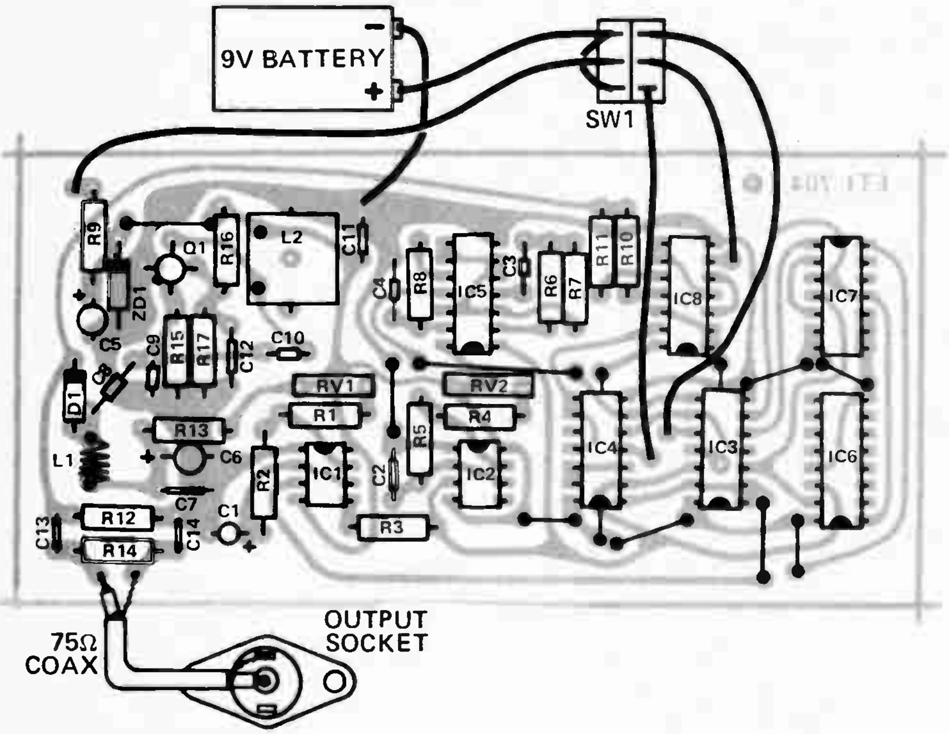


Fig.3. Component overlay.

PARTS LIST – ETI704

R1	Resistor	47k	½W	5%
R2	"	2k7	"	"
R3	"	3M3	"	"
R4	"	4k7	"	"
R5	"	560	"	"
R6	"	12k	"	"
R7	"	3k9	"	"
R8	"	2k2	"	"
R9	"	150	"	"
R10	"	10k	"	"
R11	"	4k7	"	"
R12	"	330	"	"
R13	"	27k	"	"
R14	"	82	"	"
R15	"	1k	"	"
R16	"	4k7	"	"
R17	"	3k3	"	"

RV1	Potentiometer	10k Trim type
RV2	"	4k7 " "

C1	Capacitor	0.47μF TAG Tantalum
C2	"	0.0082μF Styroseal
C3	"	390pF ceramic
C4	"	33pF "
C5	"	33μF 10V electro
C6	"	33μF 10V
C7	"	100pF ceramic
C8	"	10pF ceramic
C9	"	10pf ceramic
C10	"	1pF ceramic
C11	"	100pF ceramic
C12	"	100pF ceramic
C13	"	100 pF
C14	"	0.001μF

ZD1 Zener Diode BZX79C6V2
D1 Diode IN914

Q1 Transistor BF180

IC1,2	Integrated Circuit	NE555
IC3,5,8	" "	4011 (CMOS)
IC4	" "	4001 (CMOS)
IC6	" "	4017 (CMOS)
IC7	" "	4013 (CMOS)

L1 Inductor see text
L2 Inductor see text

PC Board ETI 704

DPDT with centre off toggle switch
75ohm socket
9V battery and connector
Box PC1 or similar (A&R sonar)

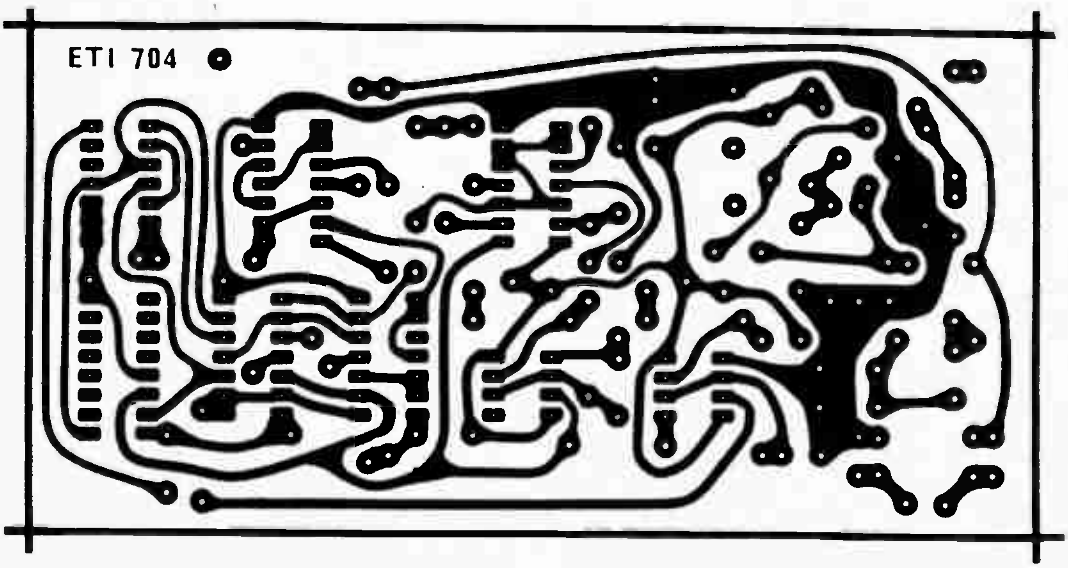


Fig.4. Printed circuit board layout.

RF SIGNAL GENERATOR

simple unit for servicing AM receivers

eti PROJECT 129

AN RF SIGNAL generator is an invaluable instrument for AM radio servicing and alignment — it greatly simplifies alignment and allows each stage to be checked for gain and frequency response.

Three types of signal are required for these purposes. Firstly, we require an audio signal to check that part of the receiver from the volume control (after the detector) to the speaker. Secondly, we need a modulated RF signal at 455 kHz (430 to 480 kHz available for non-standard receivers) for checking and aligning IF stages, and lastly, we need a modulated RF signal in the range 500 to 1600 kHz to check out the RF amplifier and converter.

In addition the level of the generator output should be adjustable so that AGC action may be checked out, and so that optimum levels may be chosen for servicing and gain checks. All the above requirements are met by the ETI 129 generator and, since only one of the available signals is used at any one time, a common level control is used for all these outputs.

In our generator the provision of IF frequencies from 430 to 480 kHz, as well as catering for non-standard receivers, allows receiver IF selectivity to be checked.

CONSTRUCTION

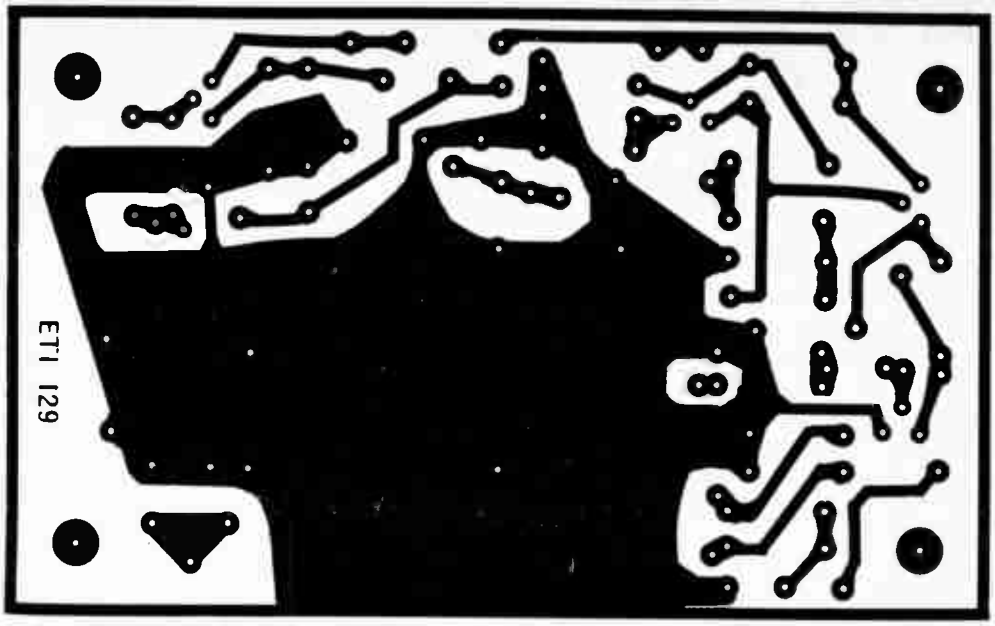
The prototype instrument was mounted in an aluminium box having external dimensions of 145 x 115 x 90 mm. Layout of the circuitry is important and for this reason the printed-circuit board layout provided should definitely be used. Take care when assembling components to the printed-circuit board to correctly orientate capacitors C9, C11 and C15, transistors Q1 to Q4 and diode D1.

The variable capacitor is mounted onto the component side of the printed-circuit board but spaced from it by about 2 mm (an oversized nut may be used). The mounting of the board and variable capacitor assembly to the front panel and reduction-drive assembly may best be understood by referring to Fig. 3. Note that the board is mounted by four standoffs and that rubber grommets are used to allow the board to move slightly — for this reason the screws should not be tightened too much. This method is used to avoid the expense of using a flexible drive to the variable capacitor.

The six-to-one slow-motion drive is mounted to the front panel by two 15 mm long bolts. The drive is spaced back from the front panel by 4 mm long spacers.

The remaining controls are mounted straight onto the front panel as shown in the photograph.

RF SIGNAL GENERATOR



Scotchcal front panels, ready to stick on are available from Electronics Today at $4.00 each. Send stamped addressed envelope – size at least 150 x 120 mm. Address to Scotchcal Offer, Electronics Today, 15 Boundary Street, Rushcutters Bay, NSW 2011.

Printed-circuit board layout. Full size 129 x 80 mm.

CALIBRATION

High Range. Using a conventional AM receiver tune to a station at the top end of the frequency band. Set the pointer of the RF generator to indicate the frequency of the station being received and couple the generator to the receiver.

Adjust capacitor C3 until the signal from the generator can be heard interfering with the station. This will take the form of a whistle which, as C3 is tuned, will go from a high frequency to a low frequency and then back to a high frequency again. The correct tuning point is where the frequency is at its lowest, i.e. in the middle. The level of the generator signal may have to be increased to obtain the correct point with accuracy. This procedure is called

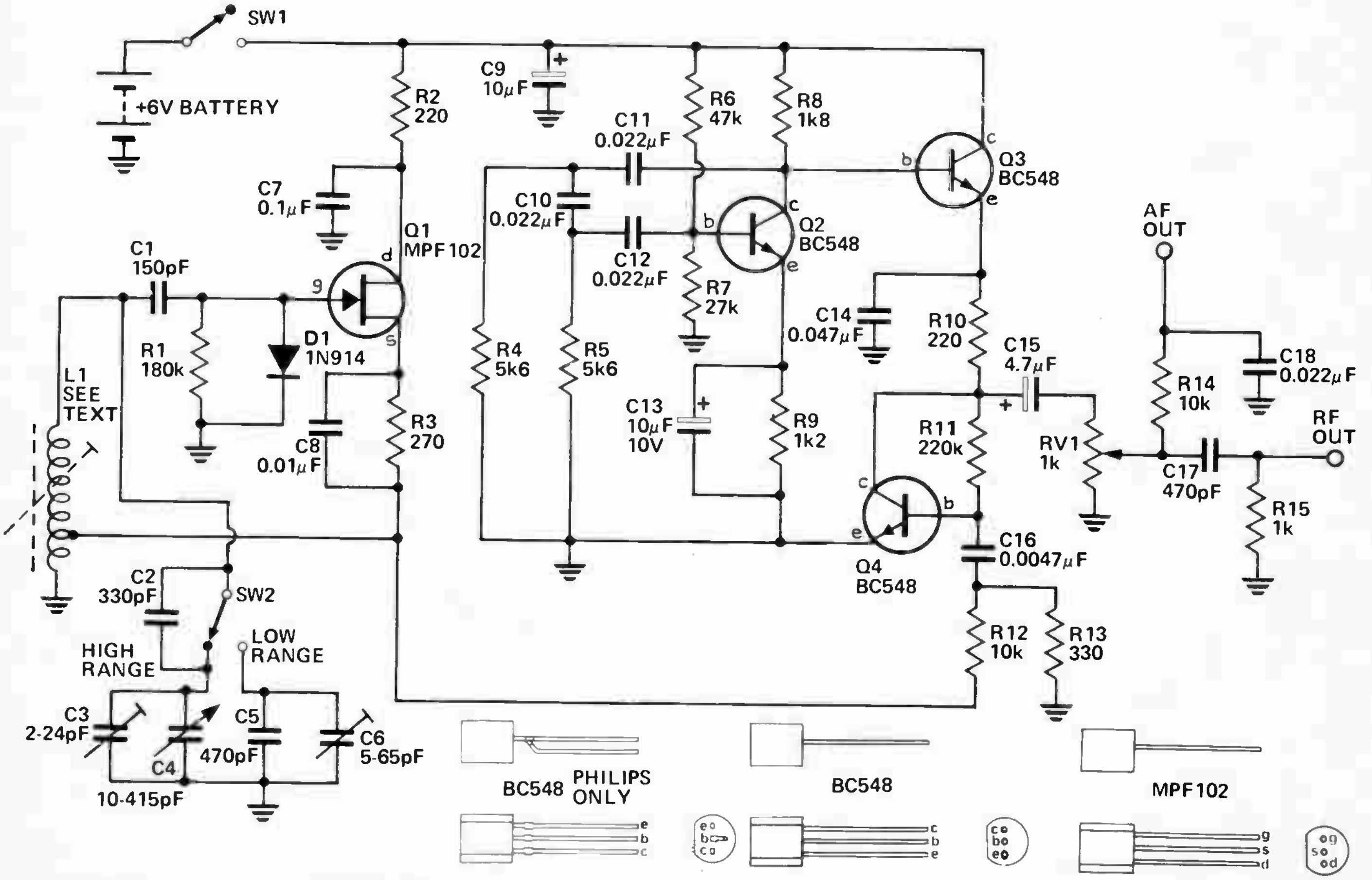


Fig. 1. Circuit diagram of the modulated RF generator.

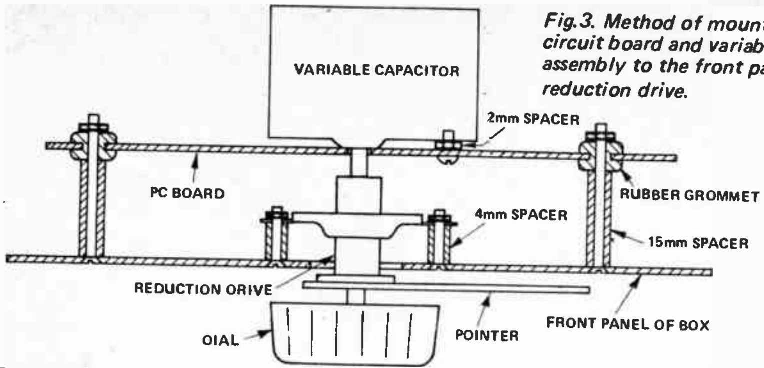


Fig.3. Method of mounting printed-circuit board and variable capacitor assembly to the front panel and reduction drive.

PARTS LIST — ETI 129

R2,10	Resistor	220	¼W 5%
R3	"	270	" "
R13	"	330	" "
R15	"	1k	" "
R9	"	1k2	" "
R8	"	1k8	" "
R4,5	"	5k6	" "
R12,14	"	10k	" "
R7	"	27k	" "
R6	"	47k	" "
R1	"	180k	" "
R11	"	220k	" "
RV1	Potentiometer	1k Lin rotary.	
C3	Capacitor	2-24pF Philips 2222-808-00006	
C6	"	5-65pF Philips 2222-808-01001	
C1	"	150pF ceramic	
C2	"	330pF ceramic	
C4	"	10-415pF Roblan variable	
C5,17	"	470pF Ceramic	
C16	"	0.0047µF polyester	
C8	"	0.01µF polyester	

C10,11	Capacitor	0.022µF	polyester
C12,18	"	0.022µF	polyester
C14	"	0.047µF	polyester
C7	"	0.1µF	polyester
C15	"	4.7µF	10V electrolytic
C9,13	"	10µF	10V electrolytic
Q1	Transistor	MPF102	or similar
Q2,3,4	"	BC548	or similar
D1	Diode	IN914	or similar
SW1	Switch	SPST	toggle
SW2	"	SPDT	toggle
L1	Inductor	See table 1.	

PC board ETI 129
Metal box 145 x 115 x 90 mm
Front panel
6 to 1 Reduction drive.
Four rubber grommets
Four 15mm long spacers*
Two 4mm long spacers*
Two 2mm long spacers*
Three terminals (red, black, green)
Nuts & bolts etc.
*Spacers may be cut from longer sections.

'zero beating'.

Now tune the receiver to a station at the low end of the band. Again set the pointer of the RF generator to the frequency at which the station operates and adjust the slug of L1 for zero beat in the same manner as for the high end.

Repeat the procedure for both the high and low ends of the band until there is no change at either end.

Low Range. The low range should be calibrated after the high range calibration has been completed.

First set C6 to mid position. Then tune in a station on a broadcast receiver that lies somewhere between either 860 and 960 kHz or 1290 and 1440 kHz. These two bands are twice and three times the generator IF band respectively. That is, we are working on the second or third harmonic of the generator respectively. Divide the actual frequency of the tuned-in station by two (for stations between 860 kHz and 960 kHz) or by three for stations between 1290 and 1440 kHz. Now set the pointer on the RF generator to this frequency. Adjust the capacitor C6 for zero beat as detailed for the high range.

Refer to any standard textbook for alignment procedure for AM receivers.

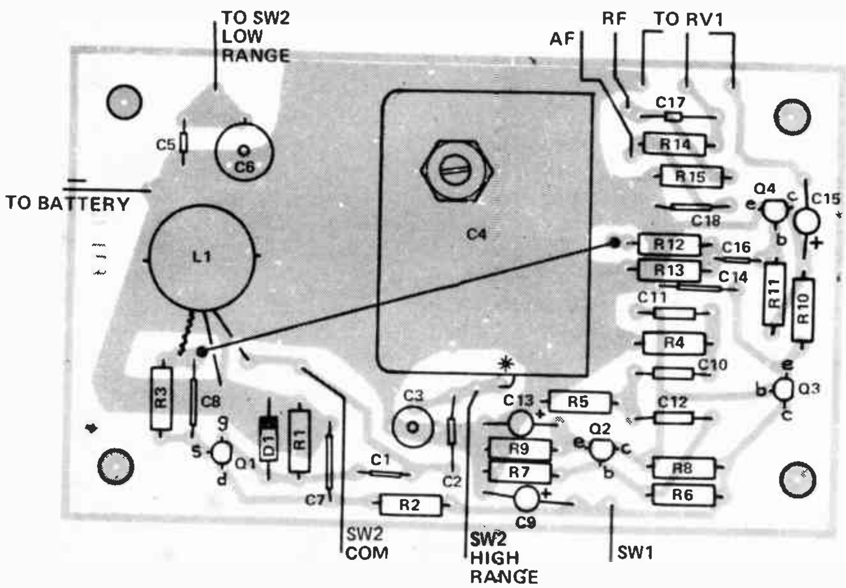


Fig. 2, Component overlay for the RF generator. Note the wire link between R12 and C8 which should be installed before fitting C4. Also note the connection from C4 to the board where shown by the asterisk.

HOW IT WORKS — ETI 129

The circuit may be sub-divided into three basic sections. These are:–

a) RF or carrier oscillator.

b) AF or modulation oscillator.

c) Modulator and buffer amplifier.

a) Transistor Q1 is connected as an Hartley oscillator. The positive feedback necessary for oscillation is provided from the source terminal of Q1 back to the gate via coil L1. The frequency of oscillation is determined by coil L1 in conjunction with capacitors C2 through C6.

Two ranges are provided, 500 kHz to 1600 kHz with L1, C3 and C4; and 430 to 480 kHz with L1, C2, C5 and C6. Diode D1 is used to develop a negative bias across R1 which thus limits the level of oscillation, and hence prevents damage to the gate of Q1.

b) Transistor Q3 is connected as a phase-shift oscillator. The network C10,C11, C12, and R3, R4 provides about 180° phase shift of the signal at the collector of Q3, and the feedback to the base of Q3 is therefore positive – causing oscillation. The frequency of oscillation is about 600 Hz.

c) Transistor Q4 is a class 'A' amplifier which is biased by the half-supply method (R10 and R11) the operating supply being obtained from emitter-follower ,Q3. The output of Q3 is the sine-wave from the phase-shift oscillator. Because of R11 any change at the junction of Q3 and R10 causes a corresponding change in the emitter current of Q4. The emitter current of Q4 therefore varies at the rate of 600 Hz. The emitter resistance of a transistor depends on the emitter current of that transistor and the gain depends on the ratio of the collector load, R10 to the emitter resistance. Since the emitter resistance is varying at 600 Hz, the gain of the transistor will also be varying at 600 Hz and so the RF signal fed to the base of Q4 from Q1 by C16 is modulated by the audio signal.

The signal across R10 is fed to RV1 by C15 and this signal consists of two components – the modulated RF and the audio tone.

After attenuation by RV1 the signals are separated by high and low pass filters to the AF and RF outlets.

RF SIGNAL GENERATOR

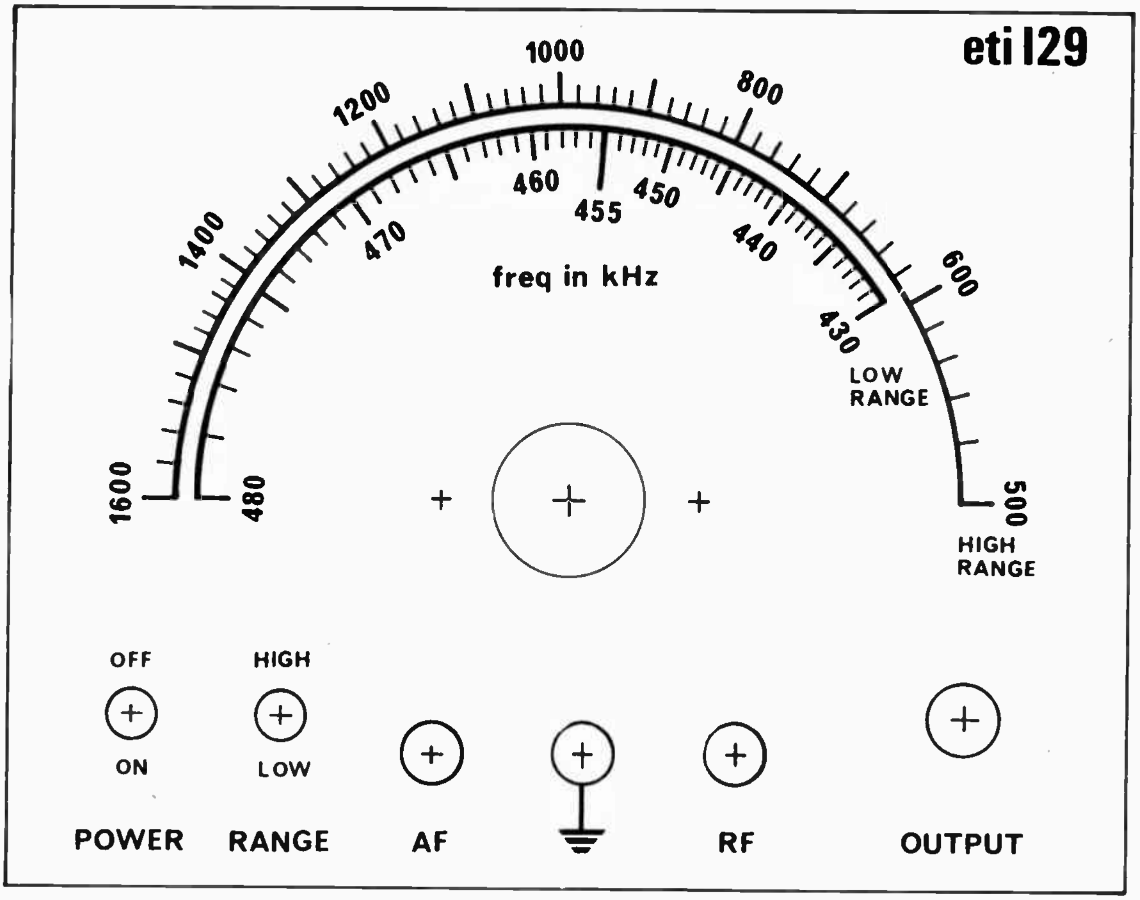


Fig. 4. Front panel artwork. Full size 148 x 116 mm.

COUPLING TO RECEIVER

Method 1. To ferrite rod coil. Connect one end of a length of ordinary hook-up wire to the RF OUT jack and then wrap about two turns of the other end of the wire around and over the aerial coil on the ferrite rod.

Method 2. To an IF amplifier. Connect a wire to the RF OUT jack and to its other end connect a 0.001 capacitor and a 1 k resistor in series. To inject the signal into the IF stage just connect the free end of the resistor to the base of the IF stage transistor.

In both the above methods if insufficient signal level is available an earth connection may also have to be made between the generator and the receiver.

Method 3. Audio testing. Use a length of wire as before but with a series capacitor of about 0.47μF. Note that an earth connection will definitely be required in this case. Once again the best place to inject a signal is straight into the base of the transistor. ●

Internal view of the generator.

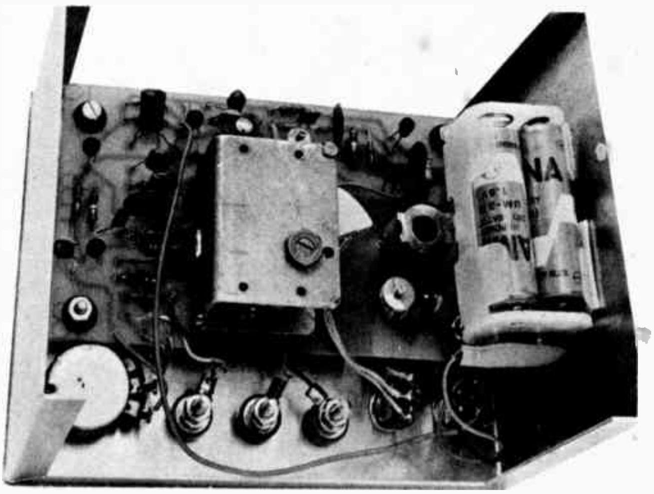


TABLE 1

L1	20 turns 0.5 mm enamelled copper wire tapped at four turns from grounded end.
CORE	Philip potcore P18 series, material 3B7 or 3H1, μe = 220. Part No 4322 – 022 – 24280 or 4322 – 022 – 24080
FORMER	4322 – 021 – 30270
ADJUSTOR	4322 – 021 – 31080
CLIP	4307 – 021 – 20000

One each of core, former, adjustor and clip required to assemble one complete coil.

eti PROJECT 706

MARKER GENERATOR

Accurate crystal-controlled markers for tuning and aligning communications receivers.

A LIMITATION of most low priced communications receivers and conventional radios is that tuning accuracy cannot be guaranteed. This means that when waiting for a short-wave station to come on the air we may well miss the beginning of the transmission because we have been tuned to the wrong frequency. The traditional method of overcoming this problem has been to use a marker generator or crystal calibrator. Such instruments generate a series of accurately known and harmonically related signals which are tuned by the receiver in order to determine the accuracy of the dial. The marker generator may also be used to perform the periodic calibration and alignment required by most sensitive receivers.

Although it is possible to build a generator which could be set to any desired precise frequency this approach is uneconomical. The more practical method is to have an oscillator running at an accurately known frequency and to generate harmonics of this frequency. For example a basic frequency of 1 MHz would have harmonics at 2 MHz, 3 MHz, 4 MHz . . . . and so on.

The marker generator must supply a stable and accurate signal without the necessity of elaborate initial setting up. This requirement leads to the obvious need for a quartz crystal as the basic frequency-determining element.

A slice of quartz crystal has the property that when a voltage is applied to either side of the crystal, the crystal will be mechanically strained and conversely, when mechanical strain is applied a voltage appears across the crystal. The crystal has a natural frequency of resonance and it is thus equivalent to a tuned circuit with a very high 'Q'.

The cheapest available crystals operate at 4 MHz and to obtain the frequency intervals that we require we used CMOS ICs to divide down from the higher frequency. To ensure maximum operating speed the CMOS ICs need to be operated from a 10 volt supply. Some exceptional devices will work on six volts but the level of higher harmonics is then reduced somewhat.

To cover as much of the dial as we can in an effective manner the harmonics should be spaced reasonably close together and should extend to 30 MHz. Ideally a harmonic should fall within the pass-band of a receiver no matter what frequency is tuned. We therefore selected a minimum spacing of 10 kHz as being the most practical. Unfortunately the inaccuracies of many receivers can, in the high bands, exceed 200 kHz thus several harmonics may be within the pass-band at any one time making it impossible for the operator to determine to which harmonic he is tuning. To overcome this problem the marker generator is switchable to provide harmonics spaced at intervals of 4 MHz, 2 MHz, 1 MHz, 100 kHz and 10 kHz.

To produce the series of harmonics required it is necessary to generate a series of very narrow pulses at a repetition rate equal to the spacing required. That is, 10,000 narrow pulses per second will produce the harmonic series 10 kHz, 20 kHz, 30 kHz up through 29 990 kHz and 30 000 kHz.

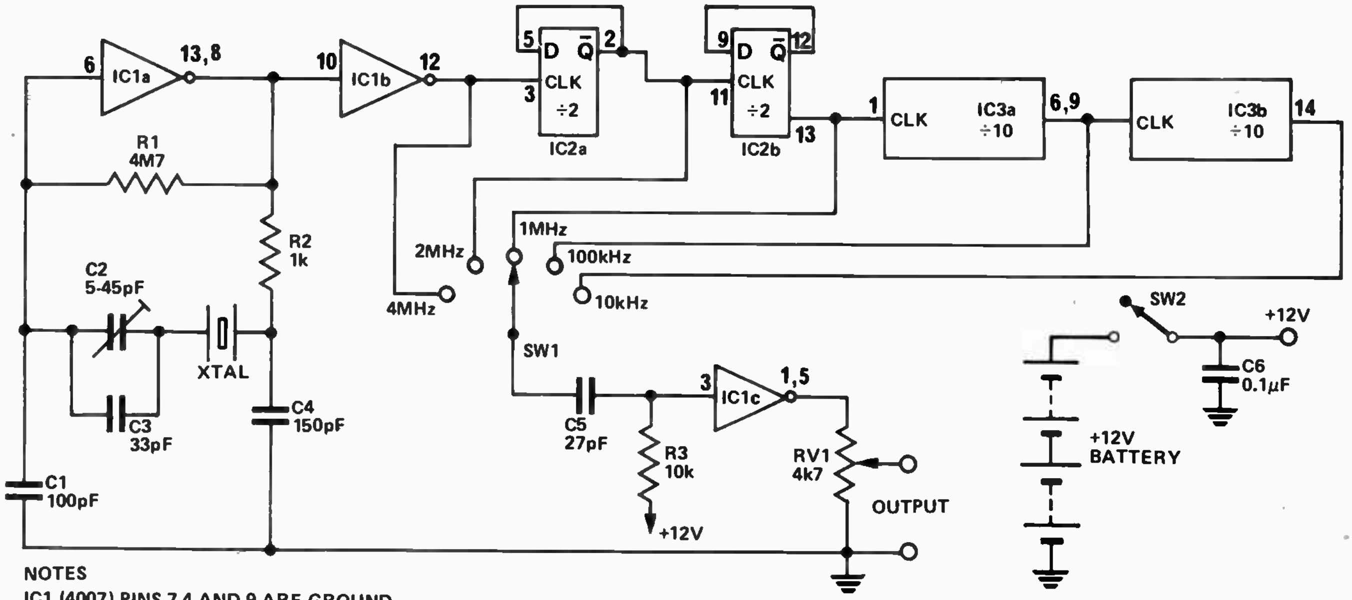
CONSTRUCTION

We mounted our unit in a commercially available aluminium box having dimensions of 150 mm wide by 75 mm high and 100 mm deep. The printed-circuit board is mounted on the rear panel of the box but spaced

SPECIFICATION

Harmonic Spacing	Five switchable outputs 4 MHz, 2 MHz, 1 MHz, 100 kHz and 10 kHz.
Harmonic Range	useable to 30 MHz
Accuracy	dependent on calibration.

MARKER GENERATOR



NOTES
IC1 (4007) PINS 7,4 AND 9 ARE GROUND
IC2 (4013) PINS 6,4,8,10 AND 7 ARE GROUND
IC3 (4518) PINS 8,7 AND 15 ARE GROUND

IC1 PINS 14, 2 AND 11 ARE +12V
IC2 PIN 14, IS +12V
IC3 PINS 2,10 AND 16 ARE +12V

Circuit diagram of marker generator.

HOW IT WORKS – ETI 706.

The marker generator is a constant-frequency oscillator driving into a CMOS divider chain. Switchable outputs from the divider chain are selected to drive a pulse generator.

The oscillator is IC1a in which R1 biases the IC into linear operation. The crystal determines the basic frequency of operation at 4 MHz in conjunction with C1, 2, 3 and 4 which appear to the crystal as one parallel capacitor. The capacitor C2 is used to tune the oscillator exactly to frequency as explained in the text. The resistor R2 adds extra phase shift but also reduces the gain. Thus if the oscillator is slow in starting reducing R2 may help. The output of the oscillator is buffered from the rest of the circuit by IC1/b.

IC2 is a CMOS dual type D flip flop that divides the 4 MHz by four to provide an output of 1 MHz, the 2 MHz also being brought out.

A further dual division by 10 is provided by IC3 which therefore provides outputs of 100 kHz and 10 kHz.

The required output is selected by SW1 and applied to C5 and R3 which differentiate the squarewave output of the divider. The waveform is then amplified and squared by IC1/c to provide an output train of narrow pulses, the amplitude of which may be varied by means of RV1.

from it by four 19 mm long machine screws. Also mounted on the rear panel are the two output terminals. The two switches and the potentiometer are mounted on the front panel whilst the battery holders are clamped to the bottom of the box by means of a clamp made from a scrap piece of aluminium.

With the exception of the ICs, mount all components and fit all links to the printed-circuit board. After checking that all are correct, mount the ICs, double checking their orientation before soldering. Fit all flying leads to the board allowing about 150 mm of free length.

Drill the box with all the required holes and fit all the components such as the switches, the potentiometer and the output sockets. Fit the printed-circuit board to the rear panel with C2 to the top of the box and route the leads to their respective points as detailed in the component overlay. Note that one of the screws through the wafer of SW1 has an earth lug underneath it which is used as the common earth point.

Fit the batteries and connect them up but do not switch on until a final wiring check has been carried out. Ten minutes of your time at this stage could save you the cost of a new set of ICs.

USING THE GENERATOR

Say for example, that we wish to tune a signal that we know to be on 13 250 kHz. First select 4 MHz on the marker generator and connect its output to the aerial socket of the receiver. Tune the receiver to the marker which will be found at 12 MHz (third harmonic of 4 MHz). Once located confirm that it is indeed coming from the marker generator by switching it on and off. Now switch to the 1 MHz markers and tune the receiver upwards to locate the 13th harmonic at 13 MHz. Now select 100 kHz markers and tune upwards through two markers to locate 13.2 MHz. Finally select the 10 kHz

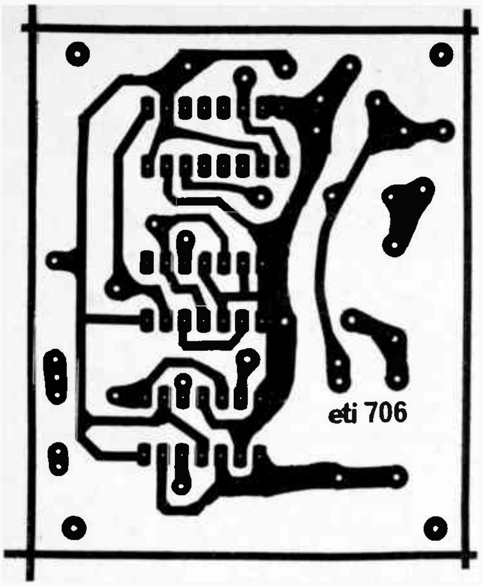
PARTS LIST – ETI 706

R1	Resistor	4M7	¼W	5%
R2	"	1k	"	"
R3	"	10k	"	"
RV1	Potentiometer	4k7	lin	
C1	Capacitor	100pF	Ceramic	
C2	"	5-45pF	Philips 2222 808 91503	
C3	"	33pF	Ceramic	
C4	"	150pF	Ceramic	
C5	"	27pF	Ceramic	
C6	"	0.1µF	Ceramic	
IC1	Integrated Circuit	4007 (CMOS)		
IC2	" "	4013 (CMOS)		
IC3	" "	4518 (CMOS)		

XTAL one 4.0000 MHz quartz crystal 30pF load

SW1 Rotary Switch 1 pole 5 position
SW2 Toggle Switch SPST
One pair of Terminals
PC Board ETI 706
Aluminium Box 150mm, 75mm, 100mm.

Two knobs
Eight AA size batteries
Two 4xAA size Battery Holders
Nuts and Bolts.



Printed-circuit layout. Full size 70 x 58mm.

markers and tune up through a further five markers to locate 13 250 kHz. Note that if this tuning procedure is carefully carried out it is quite simple to locate any position on the dial with great accuracy.

THE CRYSTAL.

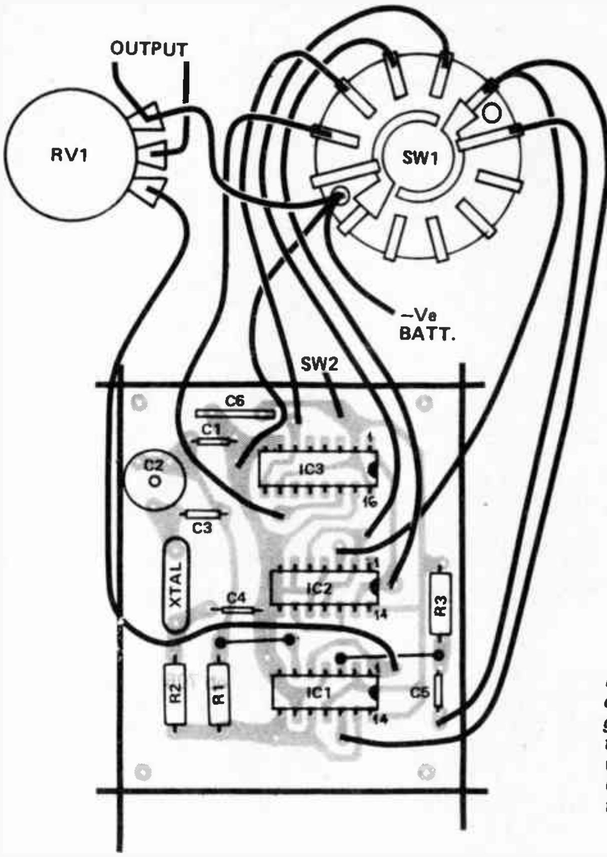
Crystals are supplied to work within specified tolerances. The tighter the tolerances the more expensive the crystal. However the crystal oscillator may be placed exactly on frequency (within small limits) by varying the amount of capacitance in parallel with it.

When purchasing a crystal you must tell the manufacturer what capacitance it will be working with and he will grind your crystal to be within the specified limits when it is used with that particular capacitance. This marker generator has been designed to work with crystals that are ground for 30 pF capacitance.

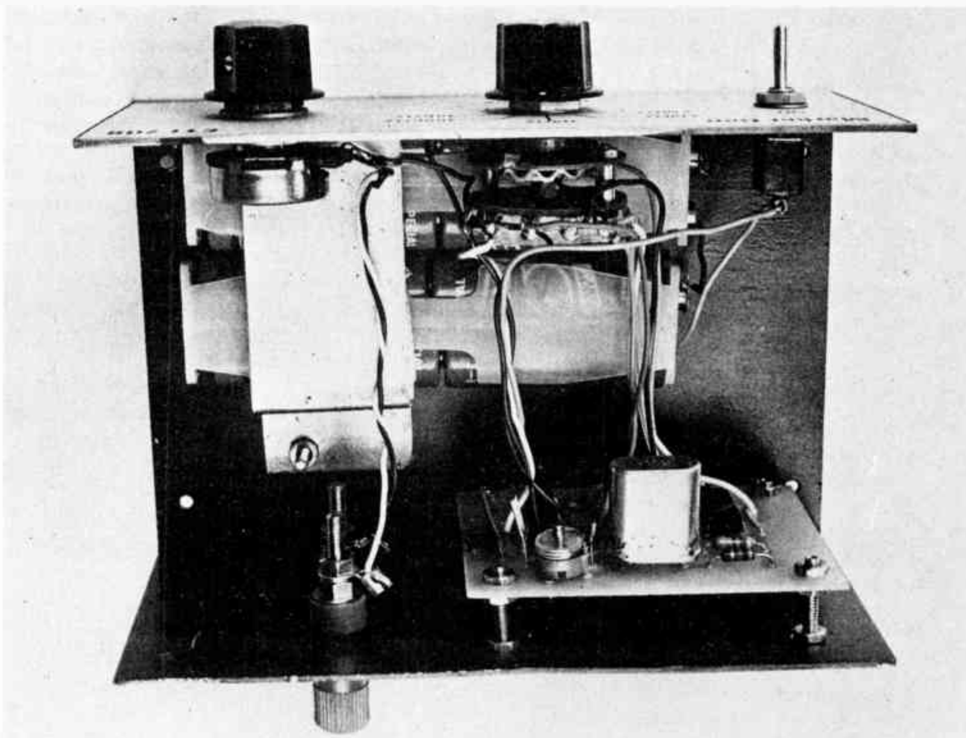
CALIBRATION

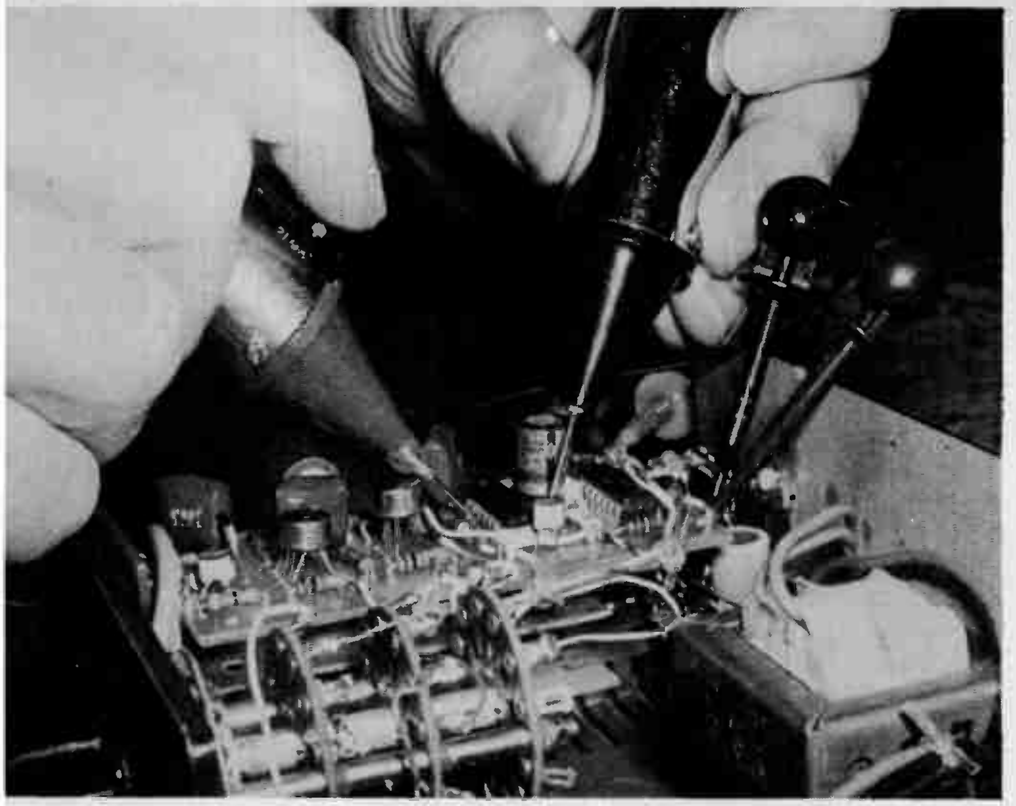
The marker will be sufficiently accurate for most people with C2 set to half value. For those who want greater accuracy the generator must be calibrated against a signal of known accuracy. The PMG transmit a time signal precisely for this purpose and it can be found on 4.5 MHz, 7.5 MHz and on 12 MHz.

The generator may be aligned against one of these frequencies by the zero-beating technique. First tune in the PMG signal and then connect the generator. A whistle will now be heard and C2 should be tuned to the point where the beat frequency has dropped so low that it cannot be heard. The generator is now spot on frequency and it should be noted that this calibration is independent of the receiver accuracy. ●



Internal view of the completed marker generator. Note that the board is mounted with the crystal and C2 towards the top of the box.





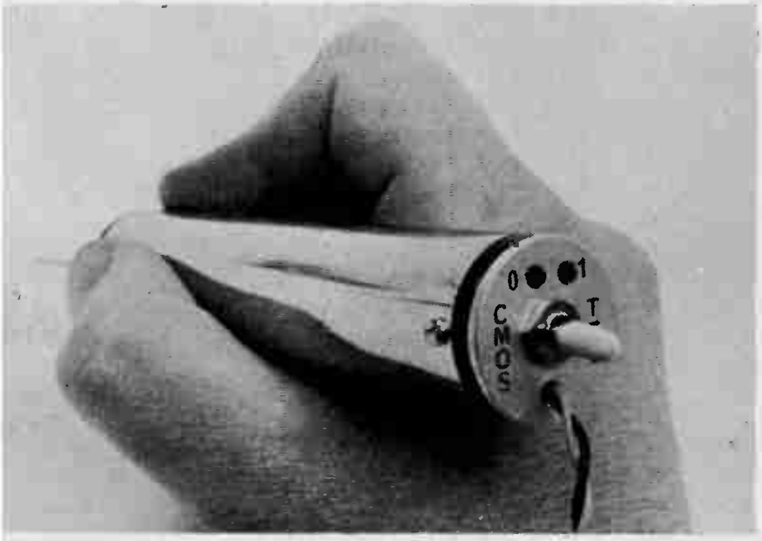
LOGIC PROBE

A basic tool for digital servicing.

THE SERVICING of digital equipment is greatly simplified by the use of a logic pulser and logic probe, for these two instruments enable one to follow circuit operation stage by stage.

THE PROBE

The probe must be capable of detecting pulses as short as 50 nanoseconds (for TTL operation) and make them visible. It was found that readily available linear ICs were not suitable as they are too slow and required dual supply voltages. Neither could CMOS be used as it also is too slow, for testing TTL gates, and its threshold voltages are not consistent. Further, TTL could not be used as it cannot withstand the voltages used with CMOS logic. This virtually means that the only devices that are suitable are discrete transistors.



The logic probe seen from the rear.

HOW IT WORKS

The probe consists of two independent voltage level detectors which, via pulse stretching monostables, drive light-emitting diodes to give a visual indication of the logic state being monitored. Transistors Q1 and Q4 form the low level or '0' detector, transistors Q5 and Q6 the high level or '1' detector whilst the remaining components form the pulse stretching monostables and visual indicators.

The high level detector works as follows. If the input level is below about 2.5 volts (1.3 volts above the level set on R17 by transistor Q5) transistor Q6 will be cut-off. When the input level rises above 2.5 volts, transistor Q6 will turn on, as will Q7, causing LED 2 to light – indicating a '1'. The transition at the collector of Q7 will, at the same time, be passed to Q8 turning it off. The current which was flowing through Q8 will now flow via R22 in to the base of Q7 holding it on even though Q6 may by now have stopped conducting. After fifty milliseconds the charge on C2 will leak away via R19, 20 allowing Q8 to conduct. When Q8 conducts it robs the current from the base of Q7 turning it and the LED off. However should the voltage at the tip of the probe still be present Q6 will still be turned on holding on in turn Q7 and the LED.

Resistors R11, 12, 13 and 14 set the operating conditions of Q5 such that the threshold voltage is optimized for either TTL or CMOS. As CMOS logic works on supply voltages ranging from five to fifteen volts, transistor Q5 has been arranged to track the supply so that the correct threshold is maintained at all times.

The low level detector works in exactly the same fashion except that it is inverted in order to detect pulses which approach within 0.45 volts of the negative line (TTL only). Each PNP transistor and each NPN transistor have been replaced with their complements. In this case Q4 sets the thresholds and the circuit operates exactly as stated for the high detector. Note that the diodes have also been reversed.

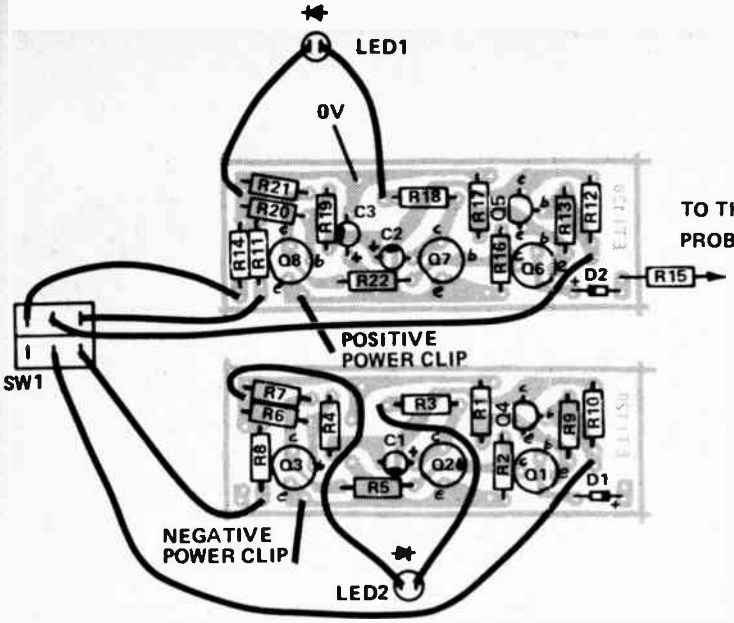


Fig. 3. Component overlays for the two comparators showing interconnection wiring.

PARTS LIST – ETI 120

R3,18	Resistor	680	¼ W	5%
R4,15,19	"	1 k	"	"
R10,13	"	1 k8	"	"
R1,9,12,17	"	2 k7	"	"
R5,14,22	"	3 k3	"	"
R2,16	"	8 k2	"	"
R7,21	"	10 k	"	"
R8,11	"	27 k	"	"
R6,20	"	100 k	"	"
C1,2	Capacitor	0.47 µF	25 V tantalum	
C3	"	10.0 µF	25 V "	
D1,2	Diode	IN914 or similar		
Q1,7,8	Transistor	2N3638, 2N3638A		
Q2,3,6	"	2N3643		
Q4	"	BC179, BC559		
Q5	"	BC109, BC549		
SW1	Switch	Two pole, two position miniature toggle		

PC boards 2 off ETI 120
Probe case (see text)
LED 1, 2 Light emitting diodes 5082 — 4484 or similar
2 Alligator clips or Ezy-hooks

CHARACTERISTICS

PULSER – ETI 121 (See page 65)

- Will source, or sink, up to 500 mA.
- Operates on supply voltages from 5 to 15.
- Suitable for both TTL and CMOS.
- Power supply drain less than 15 mA under worst case conditions.
- Press for '1' release for '0'. High impedance at other times (> 1 M).
- Will drive capacitive loads up to 1000 pF.
- Protected against accidental reversal of supply leads.
- Duration of pulse 500 nanoseconds.

PROBE – ETI 120

- Pulses as narrow as 50 nanoseconds will be detected.
- Stretches narrow pulses to 50 milliseconds for ease of detection.
- Operates on supply of 5 to 15 volts.
- Suitable for TTL or CMOS.
- True '1' and '0' level detectors. Neither LED is alight if the circuit is faulty or the probe is not making contact.
- Current drawn from the circuit is less than 20 microamps.
- Current drawn from power supply (one LED alight) 12 mA on 5 volts, 35 mA on 15 volts.

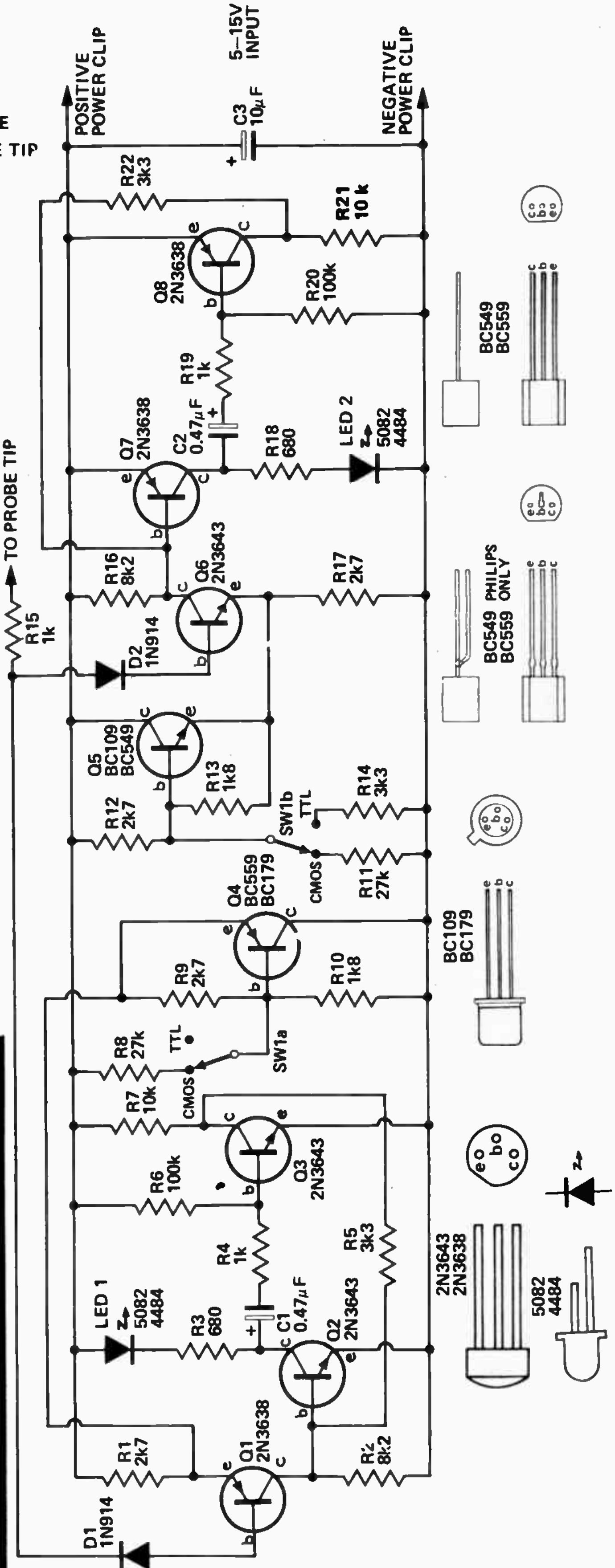


Fig. 1. Circuit diagram of the logic probe

LOGIC PROBE

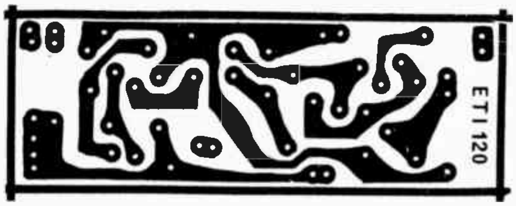


Fig. 2. Printed circuit board for the logic probe (2 required). Full size 23 x 66 mm.

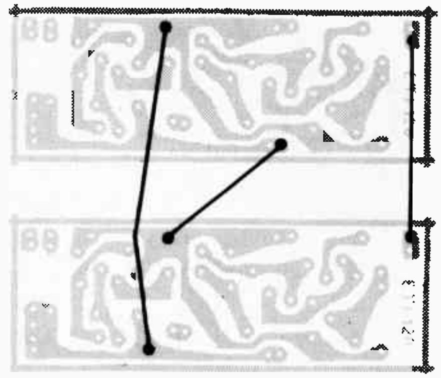


Fig. 4. Linking required between the two boards.

As both high and low logic states must be detected, a discrete transistor voltage-comparator circuit was designed to detect each state separately. These comparators must not load the circuit under test as CMOS is sensitive to current and capacitive loading. In our prototype the current drawn was a maximum of 19.7 microamps for a high, and 10 microamps for a low.

In both comparators the transistors associated with the pulse detector are turned on by an input level that exceeds the comparator threshold.

As transistor turn-on time is much faster than turn-off time, using the transistors in this way ensures the highest possible speed of operation for the particular types of transistors used. Additionally, the delay in turning off assists by lengthening the pulse, thus ensuring more reliable triggering of the monostable on very short pulses.

The input transistors Q1 and Q6 are protected against breakdown, due to excessive base-emitter voltage, by diodes D1 and D2. The diodes are also required to ensure that Q1 and Q6 remain conducting even when the probe tip is taken to the supply voltage.

Transistors Q3 and Q8 are also protected against reverse base-emitter voltages by R4 and R19 respectively.

In operation the probe will light LED 1 if a low level is detected, LED 2 for a high, neither LED if the point being monitored is at ground potential or a poor contact is made with the tip, and both LEDs will light if there is a pulse train present.

A single pulse input will be lengthened, by the monostables, to 50 milliseconds with the pulse polarity being indicated by the LED which is illuminated. Thus even single pulses as short as 50 nanoseconds may readily be detected.

CONSTRUCTION

We built our probe onto two small printed circuit boards and assembled it into a small, commercially-available probe case. The two printed-circuit boards are identical and care should be taken to use the correct overlay for each board as different transistors are used and some components are reversed on the two boards. Note particularly diodes D1 and D2 and capacitors C1 and C2. Also note how the two boards are linked together and that the supply rails are reversed. No difficulty should be experienced if the printed-circuit boards and the component overlay as specified are used.

The probe case used in our prototype was one manufactured by Jabel. The case has a length of 102 mm and an internal diameter of 23 mm. The probe tip, as fitted, is rather large and awkward. We therefore replaced the tip, with a darning needle, as shown in Fig. 6. The fine point of this tip is much easier to use on micro circuitry and, as it is very sharp, it will penetrate varnish etc to make reliable contact. A needle is a little brittle and for this reason it is recommended that a maximum unsupported length of 12 mm be left protruding. Resistor R15 is mounted within the tip and soldered directly to the needle. The other end of the probe is fitted with a plastic stopper which is used to support SW1 and both LEDs. SW1 is also used to hold a small name-plate in position as shown in Fig. 6. Two LEDs are mounted into the end plate, together with SW1, and after soldering leads to the LEDs they should be passed through the holes in the plate, and the plastic end-piece, and secured in position with a drop of epoxy cement. Another hole is drilled in the stopper through which is passed the two supply-voltage leads.

Connect the leads from the stopper assembly to the previously assembled boards. Position the boards together, copper side to copper side, with a piece of insulating material between them. Make sure that the board assembly will fit in to the tube without fouling the sides. Cut a piece of cardboard or plastic 75 x 85 mm, roll it into a tube and fit in the probe body. Now fit the board assembly into the tube – it may be necessary to dress the sides of the boards with a file to obtain a neat fit.

The tip may now be connected and both ends screwed into position. Finally, alligator or, better still, Ezy-hooks clips should be fitted to the supply leads.

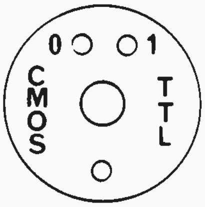


Fig. 5. Artwork for the nameplate on the probe.

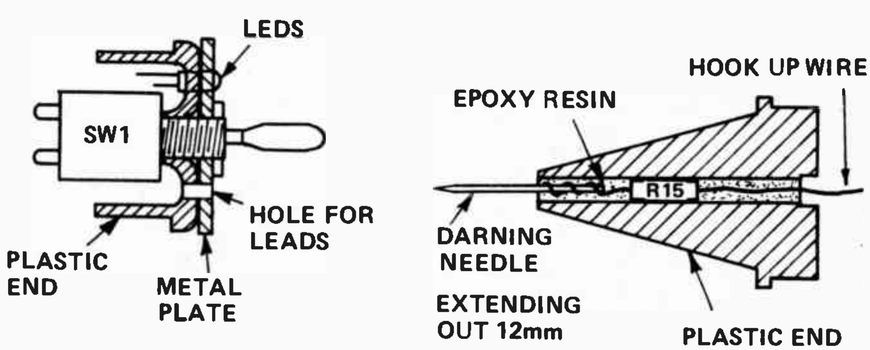


Fig. 6. How the probe ends are constructed.

ETI PROJECT 121

LOGIC PULSER

Companion instrument to the logic probe.

ALTHOUGH the logic probe used alone is a very valuable piece of digital test equipment, it is limited by the fact that it can only observe the logic states that occur naturally within the piece of digital equipment under test.

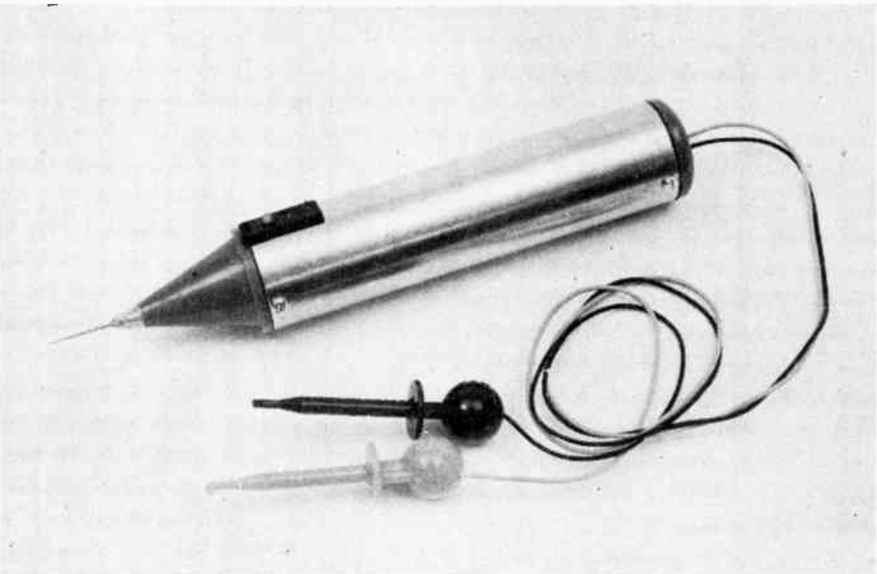
The logic pulser is a further valuable tool that is used in conjunction with the logic probe. It's function is to override the naturally occurring state at the particular circuit node under test. That is, if the circuit node is normally at the '1' state, the pulser will drive that node to a '0' for a very short period when the microswitch is pressed. If the circuit node is normally at a '0', the probe will drive it to a '1' for a very short period when the microswitch is released. Thus it puts a short pulse into the circuit node regardless of it's normal state when SW1 is pressed and released.

A fairly powerful pulse is required to override the normal logic state of a circuit node and care must be taken to ensure that the devices either driving, or being driven from that node are not damaged. This is achieved by making the pulse of very short duration. In our probe the pulse width is 500 nanoseconds. Thus although the pulse is of high current the energy released is insufficient to damage normal logic devices.

The probe must be suitable for driving either TTL or CMOS that is, it must operate from a supply ranging from 5 to 15 volts, it must be capable of operating into loads having a capacitance as high as 1000 picofarads and must supply a current pulse of around half an amp. All these conditions are fulfilled in the ETI 121 Pulser and the prototype has been tested by causing it to generate several hundred thousand half amp pulses without any problems. The probe is quite capable of pulling two (in parallel) high-power TTL 'zeros' to a '1' level and this is the most severe condition it has to meet.

At the same time as providing high level pulses, the pulser should not draw too much supply current as some CMOS supplies may not have much additional capability. Under worst-case conditions the ETI Pulser drew a maximum of 10 mA.

The probe is capable of overriding a normal logic state but is not capable of overriding a point that is connected to ground or to a supply rail. Thus by pulsing a node and at the same time looking at that point with the logic probe it is possible to tell if that point is shorted to either rail.



A basic tool for digital servicing.

The logic pulser combined with the logic probe is thus capable of performing stimulus — and — response testing of both TTL and CMOS logic and of determining the exact nature of a fault at a particular circuit node.

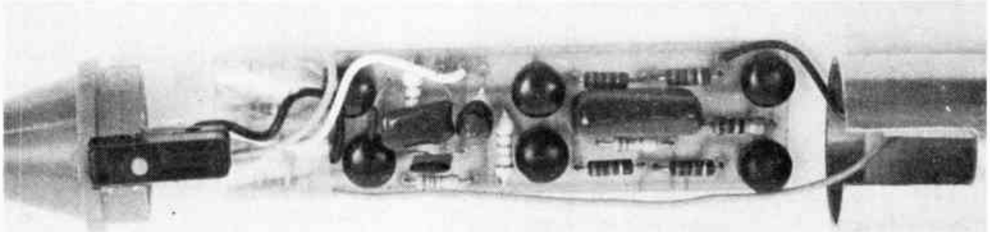
CONSTRUCTION

Construction is greatly simplified if the printed circuit board of Fig. 2, is used. This should have the components assembled to it in accordance with the component overlay. Note particularly the polarity of C1, and the connections of the microswitch such that the normally-closed terminal of the switch is connected to the base of transistor Q1. Also make sure that a red lead is connected to the positive rail of the board, and a black lead to the negative rail, to facilitate later connection.

We used the same probe case for the pulser as for the logic probe. The probe tip is again replaced by a darning needle and the microswitch SW1 is mounted into the plastic-section of the tip as follows. First check the switch to determine what the contact arrangement is. Attach colour coded wires to the switch, to aid later identification, and tape the microswitch into position. Epoxy cement may then be used to fix the switch into place permanently. Now cut a slot into the probe case so that the switch and plastic tip assembly can be inserted into the casing.

Connect the probe tip and microswitch leads to the board and, after insulating the inside of the case with cardboard or plastic as previously described, insert the board into the case. Pass the supply leads through the plastic end piece and then fit both end pieces and secure them in position. Lastly attach Ezy-hooks or alligator clips to the supply leads.

Keep the supply leads as short as is reasonably possible as excessively long leads will degrade the performance of the pulser.



Internal construction of the pulser.

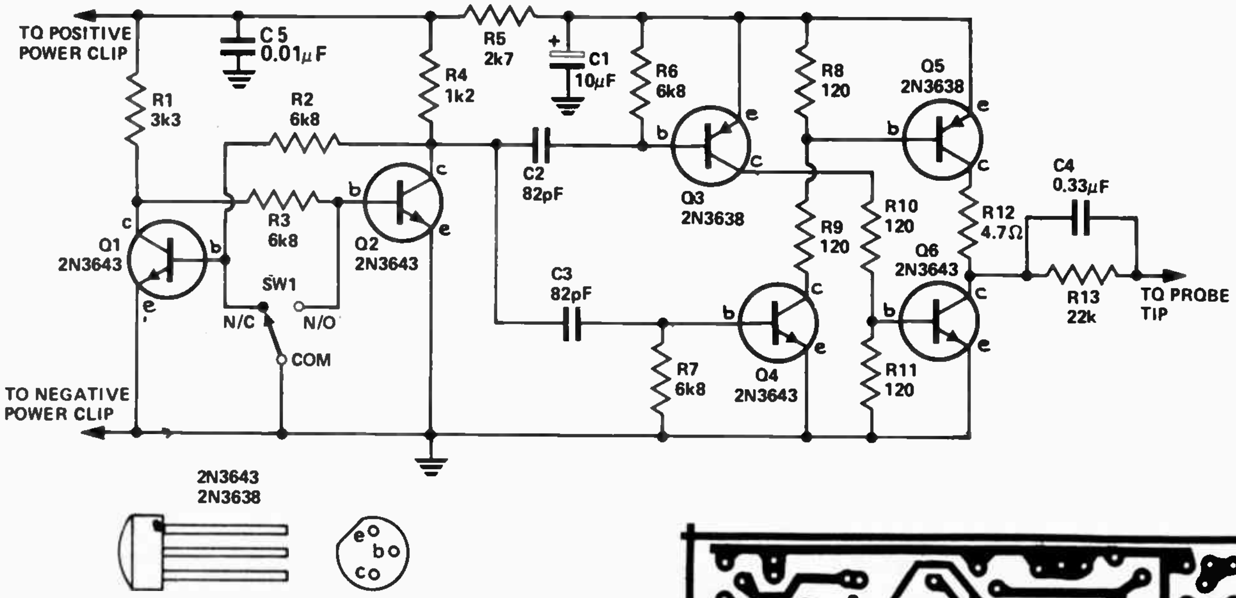


Fig. 1. Circuit diagram of the pulser.

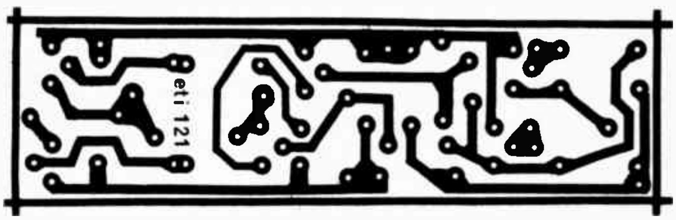


Fig. 2. Printed circuit board for the pulser. Full size 23 x 65 mm.

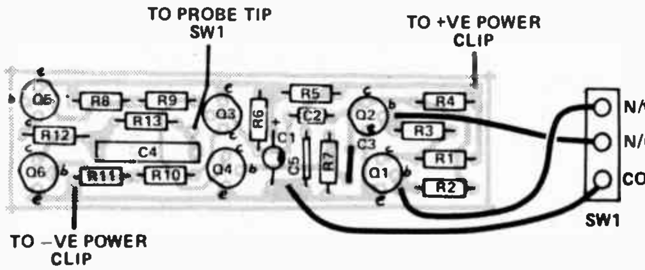


Fig. 3. Component overlay for the pulser.

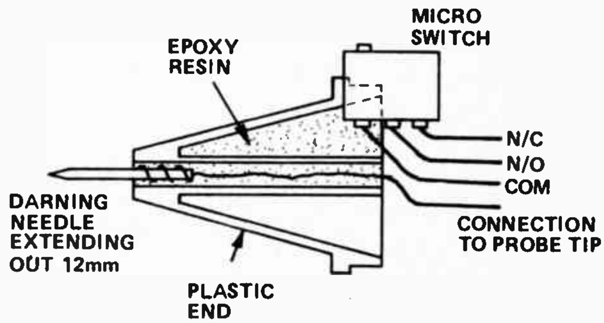


Fig. 4. Construction of the tip for the pulser probe.

HOW IT WORKS

The pulser is activated whenever microswitch SW1 is pressed. This switch controls the state of a flip-flop formed by transistors Q1 and Q2. The flip-flop is necessary to prevent contact bounce of the microswitch from having effect.

The output transistors of the probe, Q5 and Q6, which in turn are controlled by Q3 and Q4 are both normally off. However when the microswitch is pressed Q2 turns off and the rising voltage on its collector is coupled, via C3, to the base of Q4 turning it on. This in turn, turns on Q5 pulling the output to the positive rail. This generates a '1' pulse if the point under test was at a '0' level. Resistor R12 provides a current limit of around 500 milliamps. Due to the small value of C3 the pulse output is only about 500 nanoseconds long, short enough so that there is insufficient energy to damage the device under test.

When the switch is released Q2 turns on and the negative-going edge is coupled to Q3 by C2 turning it on. This turns on Q6 causing the output to be pulled to the negative rail. This gives a '0' pulse which, like the '1' pulse, is only 500 nanoseconds long.

The output from the probe is taken via the paralleled combination of R13 and C4 where C4 carries the current and R13 discharges C4 between pulses. This network protects the probe against the condition where the probe is inadvertantly connected to a voltage which is above or below the logic supply rails.

Resistor R5 isolates the high current pulse from the power supply, capacitor C1 providing the actual current needed.

SPECIFICATION

See probe page 63.

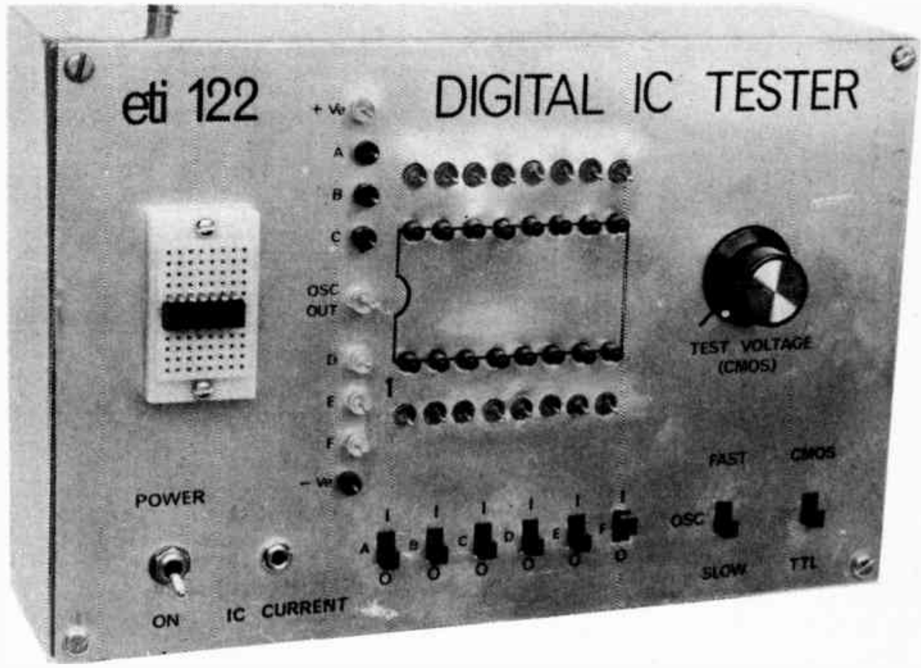
PARTS LIST

R12	Resistor	4.7 ohm	¼W	5%
R,8,9,10,11	"	120 ohm	¼W	5%
R4	"	1k2	¼W	5%
R5	"	2k7	¼W	5%
R1	"	3k3	¼W	5%
R2,3,6,7	"	6k8	¼W	5%
R13	"	22 k	¼W	5%
C2,3	Capacitor	82 pF	ceramic	
C5	"	0.01 μF	polyester	
C4	"	0.33 μF	polyester	
C1	"	10 μF	25 V tantalum	
Q1,2,4,6	Transistor	2N3643 or similar		
Q3,5	"	2N3638, 2N3638A or similar		

1 micro switch miniature McMurdo type 2LM
2 alligator clips or Ezy-hooks
PC board ETI 121
probe case (see text).

eti PROJECT 122 LOGIC TESTER

Test CMOS and TTL with this versatile instrument.



WARNING:

When using the tester, remember that manufacturers recommend that CMOS ICs should not be inserted or removed from a circuit without first switching off the power supply.

EXPERIMENTERS often damage ICs in the process of developing a new circuit and often try a new IC in a circuit that is not working to eliminate that as a possible cause. The result of this is that one usually finishes up with a box full of ICs which are of dubious value. To sort out these ICs one must use a tester that is capable of testing the wide range of differing ICs that are available in the most commonly used families.

Until recently the most commonly used family has been TTL. But CMOS is rapidly gaining widespread usage and any tester, to be of value these days, must be able to test both these families. The ETI Logic Tester is capable of testing both families, and is also capable of being used to breadboard and test simple circuits based on single ICs.

An LED indicator is associated with each pin of the IC under test and these are arranged around the perimeter of a box representing the IC under test. This allows a small card, which has the schematic of the particular IC drawn on it, to be fitted to the front of the tester as an aid to the interpretation of the LED test indications.

CONSTRUCTION

The most expensive single component in the tester, after the transformer, is the case. For this reason we decided to make a wooden case and a plain aluminium front panel. Some people may however wish to mount the unit in a diecast box and for this reason the printed circuit board has been sized to fit in a standard 222 x 146 x 51 mm die-cast box. The following description is for a wooden box specifically, but applies equally well to the metal box.

The printed-circuit board is mounted to the rear of the front panel, copper side to the panel, such that the LEDs and patch pins, mounted on the printed-circuit board, project through the front panel. This greatly simplifies construction as it saves some 48 leads and solder joints. The switches are secured to the front panel by first glueing two pieces of printed-circuit board to the rear of the board and then soldering the switches to the copper side of the board. This procedure avoids the necessity of a multitude of screws passing through the front panel.

The printed-circuit board should be assembled with the aid of the component overlay by fitting all components with the exception of IC1, 5, 6 and 7, and LEDs 1 through 16, and the patch pins. Check that the ICs are orientated correctly as are also C2, 5, 7, 9 and D1, 2 and 3. Now solder these parts into position using the least amount of heat necessary on ICs 2, 3 and 4.

Position the LEDs and patch pins onto the copper side of the board but do not solder them in place as yet. Now fit the board to the front panel so that the pins and LEDs protrude through the panel evenly. Secure the pins and LEDs in position by using a very small drop of five minute epoxy for each, on the component side of the

(Text continued on page 69)

HOW IT WORKS.

The tester consists of four basic sections. The socket for the IC under test, the output level-detect logic, oscillators and switches for the inputs, and the power supply.

The socket for the IC under test has the pins in each row electrically connected to each other. These rows are the groups of five holes which are perpendicular to the central groove on the socket. Each row (ie, each pin on the IC under test) is connected via a 10 megohm resistor to ground to prevent the build up of static charges. The resistors also hold all unconnected inputs at ground potential thus preventing any damage to the IC.

Each row is also connected to a pin on the front panel. Test connections are made to these pins by patchable links from the oscillator and test switches so that the correct test conditions may be set up.

Resistors R19-26 and R43-R50 connect each row (ie pin) to a logic level detector, ICs 5, 6, and 7. These CMOS hex-inverters buffer each pin and drive an LED to indicate the logic state of the pin. When the logic voltage on a pin is high the LED will be alight. Resistors R19 to R26 and R43 to 50 protect the internal diodes of ICs 5, 6 and 7 against the possibility of a pin being taken above the positive supply voltage or below ground potential. Resistors R11 to R18 and R51 to R58 in conjunction with the five volt supply set the operating currents for the LEDs.

A 555, IC4, is used as an astable oscillator which initially charges C8 via R9 and R10 until the 2/3 supply threshold is reached. C8 is then discharged via R9 and pin 7 of the 555 to the lower threshold of 1/3 supply volts. Switch SW6, when operated, puts a larger value of capacitance into the circuit which gives a frequency of about one hertz. This is slow enough so that the eye can follow each logic state transition. The high speed operation is used for checking very long counters and shift registers and can also be used in conjunction with an oscilloscope. The square wave output of the oscillator is made available at a

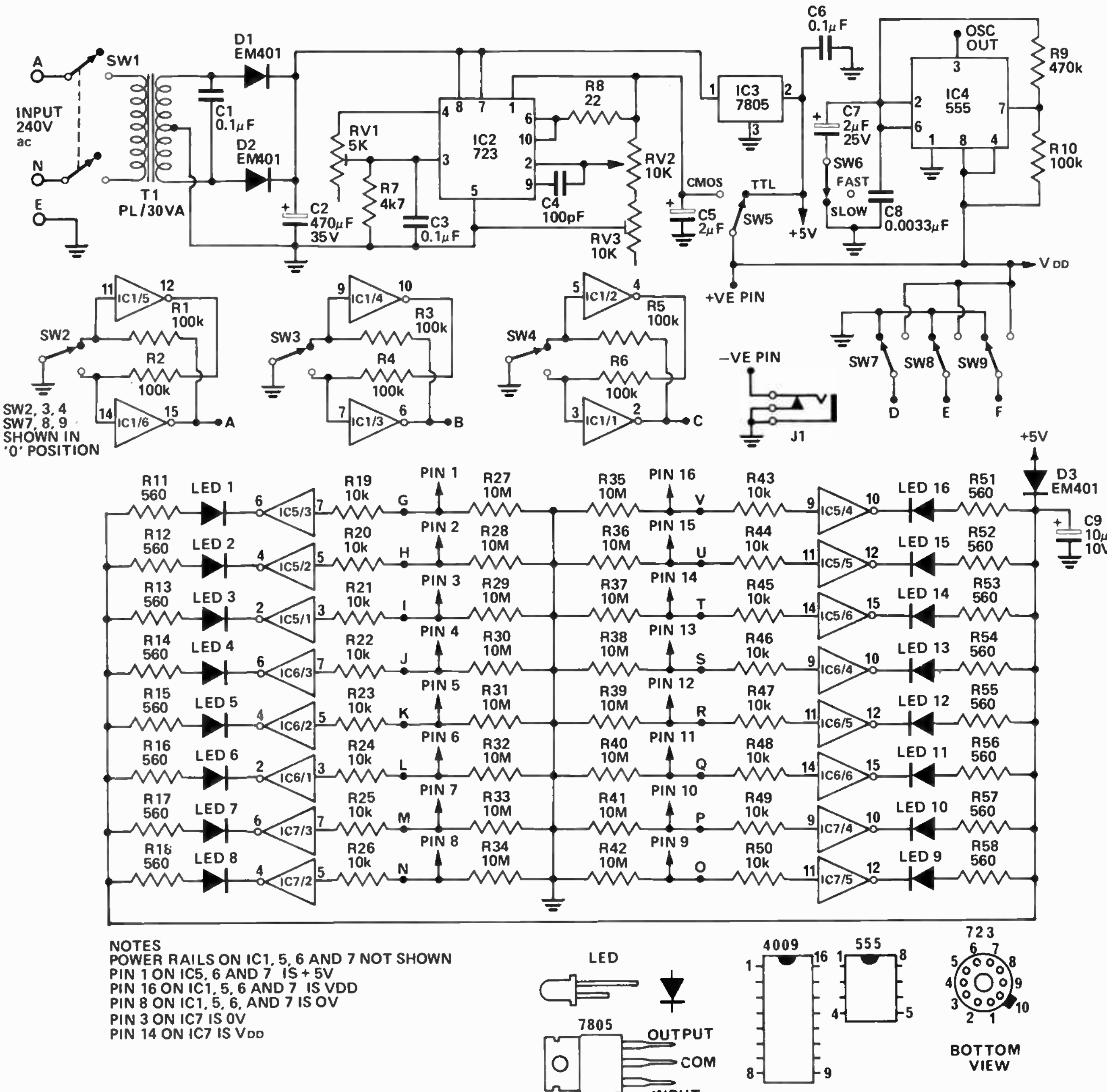


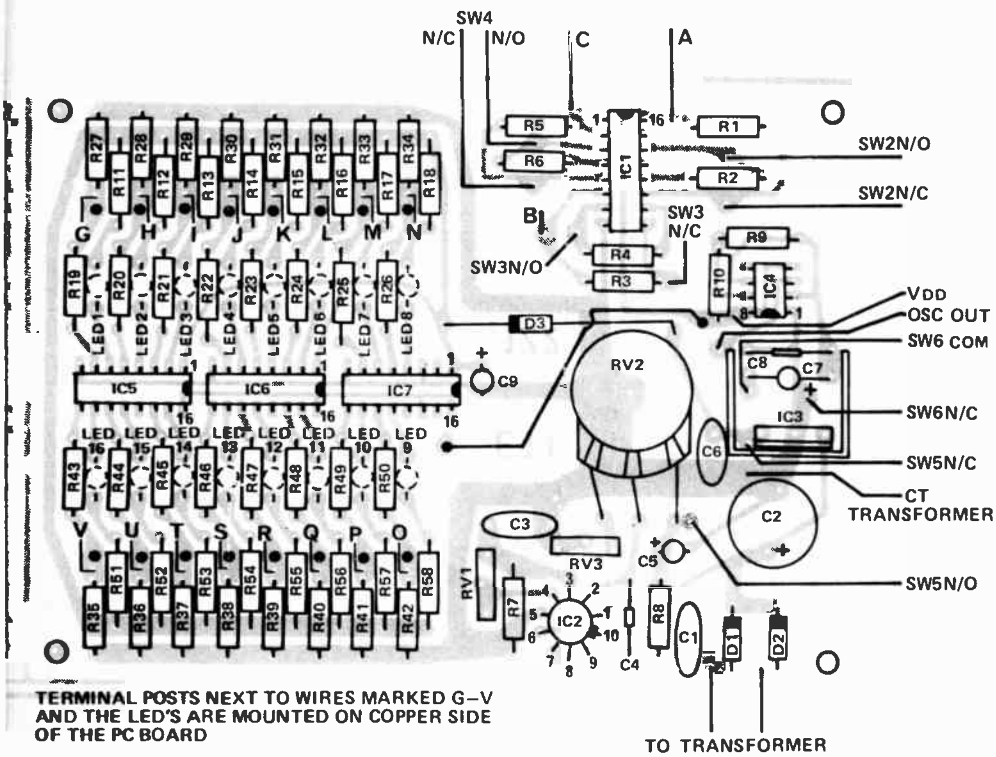
Fig. 1. Circuit diagram of the logic tester.

patch-pin on the front panel.

There are six further output pins on the front panel three of which, D, E and F, are set to negative or positive supply by means of toggle switches. As there is no debounce logic associated with these pins they can only be used to set up static conditions and not for clocking counters and shift registers. The remaining three pins are also programmed by switches but these switches are connected to IC1 which contains three RS flip-flops to effectively remove any contact bounce of the switches. This operates as follows. If initially the input of IC 1/5 is earthed by SW2 its output will be high and hence the output of IC 1/6 will be low. When IC 1/6 SW2 is operated again it earths the input of IC 1/6 sending the output of IC 1/6 and input of IC 1/5 high and the output of IC 1/5 low. Since the input of IC 1/6 is connected to the output of IC 1/5 it is held low even if the contacts of SW2 bounce several times when the switch is operated. Thus the output at A is one single transition from high to low (low to high when next the switch is operated). The output of the three debounced switches are labelled on the front panel as A, B, and C.

In the power supply diodes D1 and D2 full-wave rectify the output from the power transformer. The output from the rectifier is smoothed by C2 and regulated to five volts by IC3. The resulting five volt supply is used to drive the LED indicators and to power the TTL device under test. Integrated circuit IC2, a type 723, is a regulator the minimum output of which is set to five volts by RV1 and the maximum of 15 volts by RV3. Front panel control RV2 allows the output voltage to be adjusted between five and 15 volts. The current limit on the output is set to 30 mA by means of R8. SW5 selects the high current five volt supply for testing TTL or the low current variable supply for CMOS. Terminal J1 in the negative supply lead is provided for checking the current drawn by the IC under test. ●

Fig. 2. How the components are mounted on the pc board.



PARTS LIST — ETI 122

R8	Resistor	22Ω	¼W	5%
R11,18	"	560	"	"
R51,58	"	560	"	"
R7	"	4 k7	"	"
R19,26	"	10 k	"	"
R43,50	"	10 k	"	"
R1,6	"	100 k	"	"
R10	"	100 k	"	"
R9	"	470 k	"	"
R27,42	"	10 M	"	"

RV1	Potentiometer	5 k	Trim type
RV3	"	10 k	" "
RV2	"	10 k	Linear

C4	Capacitor	100 pF	Ceramic
C8	"	0.0033μF	polyester
C1,3,6	"	0.1μF	"
C5,7	"	2μF 25V	electro
C9	"	10μF 10V	"
C2	"	470μF 35V	"

D1,2,3 Diode EM401 or similar
LED 1 — LED 16 Light Emitting Diodes RL4484 or similar

IC1,5,6,7 Integrated Circuit 4009 (CMOS)
IC2 " Circuit 723 (metal can case)
IC3 " Circuit 7805 (TO-220 case)
IC4 " Circuit 555

J1 Jack small earpiece type

SW1 DPST toggle 240V rated
SW2-SW9 miniature slider switch 2 pole 2 position

PC BOARD ETI 122

IC Socket SK20 see text
Wooden case see text
Transformer 240 V primary 30 V CT secondary or 2 x 15 V windings PL30/20VA
25 patching Pin McMurdo type FT-1 feed throughs front panel
3 core flex and plug
heatsink for IC3 (see Fig.6)

(Text continued from page 67)

board. Do not glue the LEDS to the front panel. Once the glue has set, carefully remove the board from the front panel and then solder the LEDs and pins into position. Fit 250 mm long leads to the board for later connection to the switches and power transformer and then, using a minimum amount of heat, solder ICs 1, 5, 6 and 7 into position.

If using the recommended test socket prepare it by removing the paper from the rear of the socket, cut the paper in half and then remove about 12 mm from each side. The paper is then replaced on each side so that leads can now be soldered to the metal forming the pins of each row. The front panel must also be cut out so that these leads may be passed through. Now affix the socket to the front panel and install the printed circuit board. Mount the transformer into the base of the box and interconnect the board and switches etc.

The wooden box was constructed from 12 mm thick pineboard such that the outside dimensions were 225 x 148 x 70 mm. We finished our box with coloured Estapol high-gloss enamel which resulted in a very pleasing final appearance.

DESIGN FEATURES.

There are several design requirements which must be met in a unit which is designed to test both CMOS and TTL devices. These may be summarized as follows.

1) The unit must be capable of correctly testing both types of logic.

2) Simple gate functions should be tested by go no/go checks and complex functions such as counters and shift registers should also be reliably checked.

3) There should be the least possible chance of damaging the device during testing.

4) CMOS ICs must be testable with a variety of supply voltages.

5) A clock oscillator and a means of setting up the input conditions must be provided.

One of the major design difficulties with a unit such as this is coping with the many different pin configurations

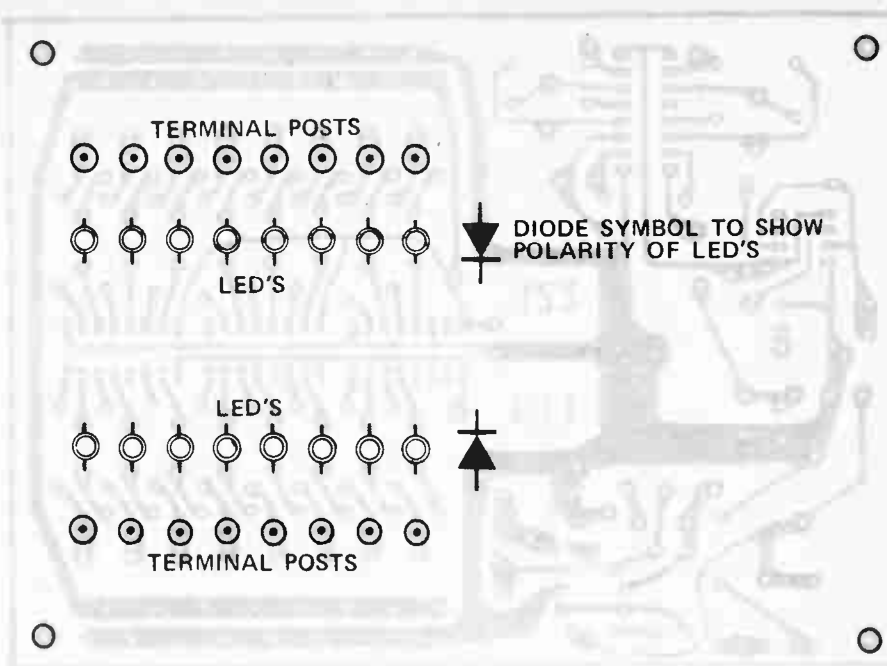


Fig. 4. Positioning of LEDs and terminal posts on the copper side of the printed-circuit board.

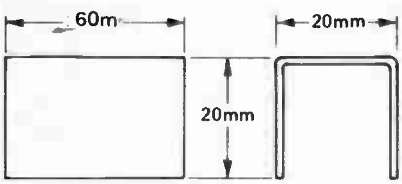


Fig.6 Heatsink for IC3. The IC is mounted (by a screw) through a 3.2 mm hole in the base of the heatsink (see photograph of inside of unit).

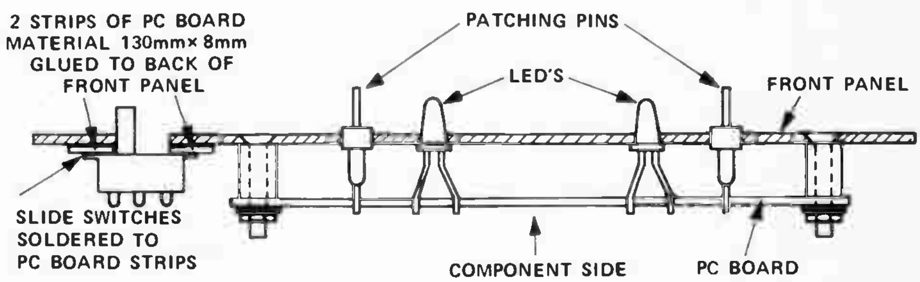


Fig. 5. How the front panel and printed-circuit board are assembled.

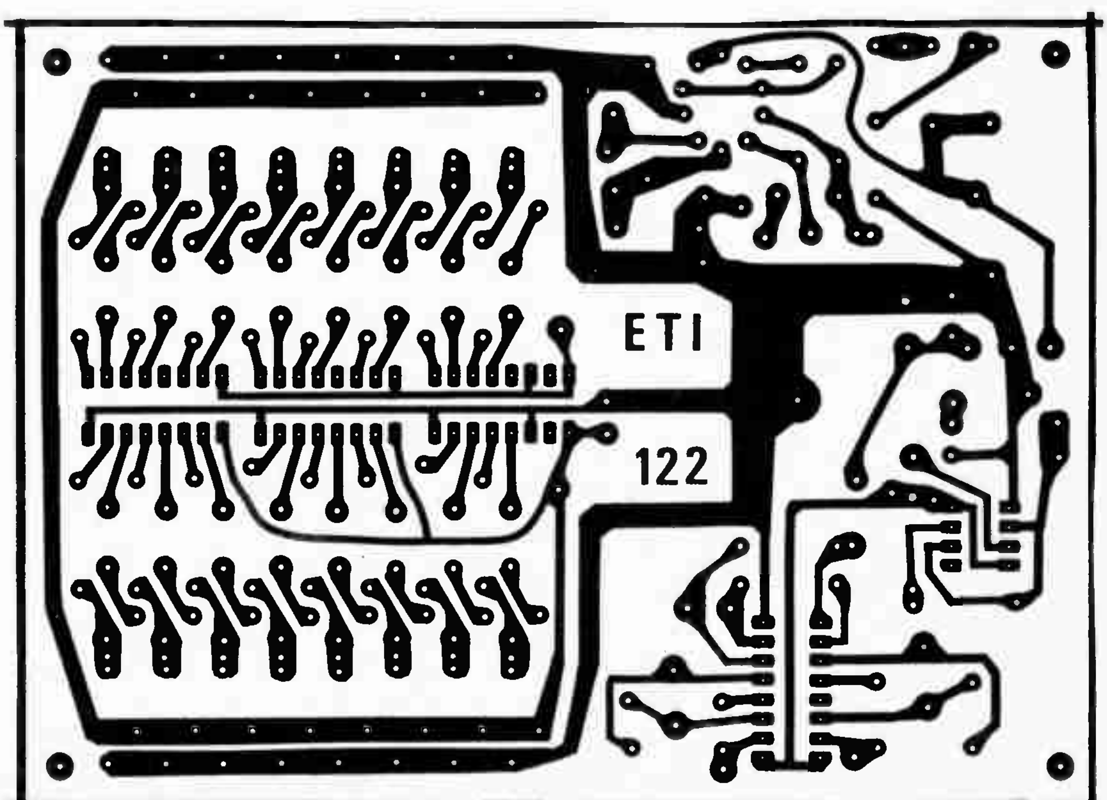


Fig.7. Printed circuit-board artwork. Full size 142 x 104 mm.

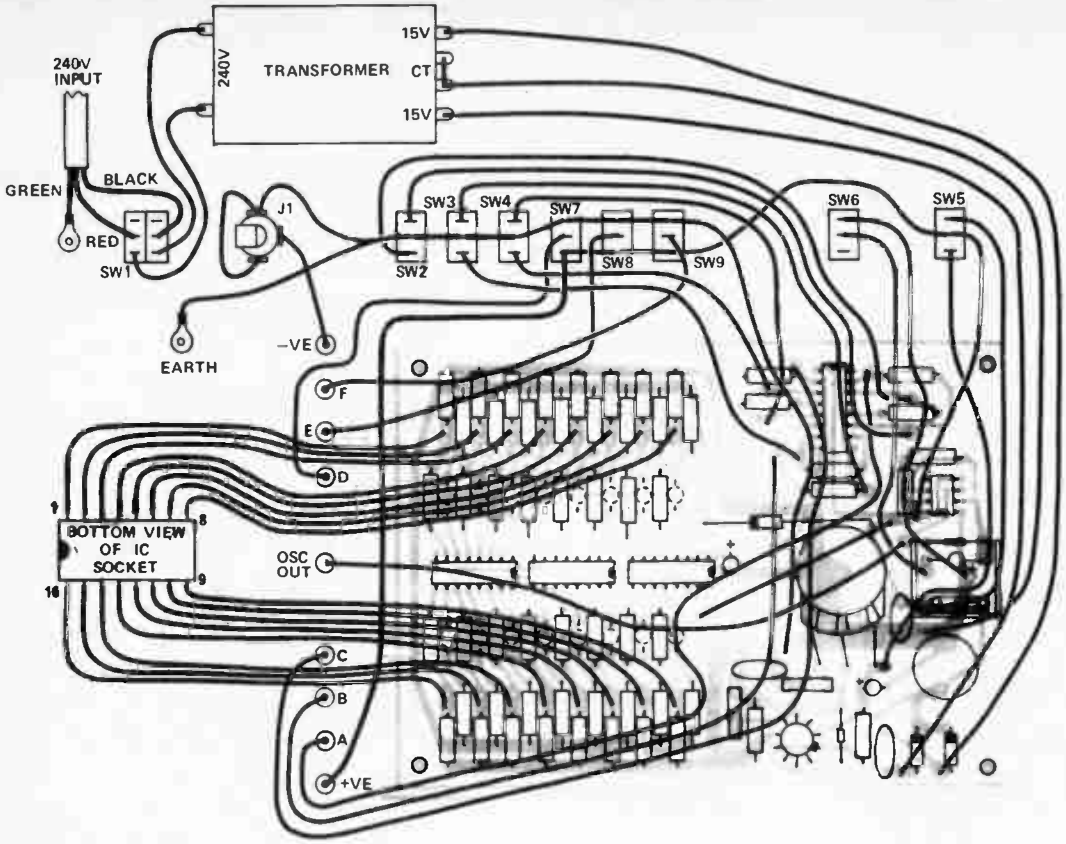
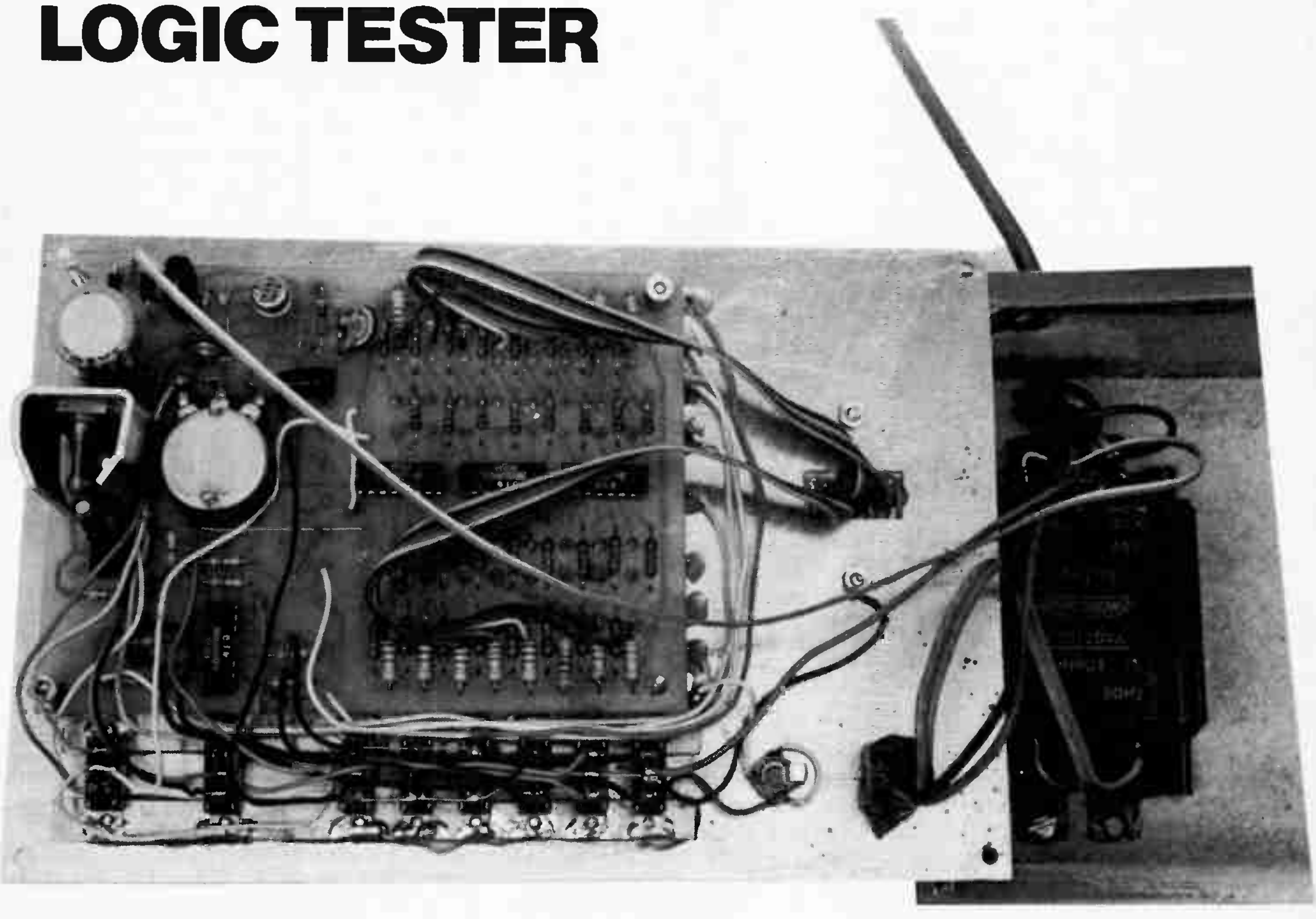


Fig. 3. Wiring diagram of complete unit.

LOGIC TESTER



LOGIC TESTER

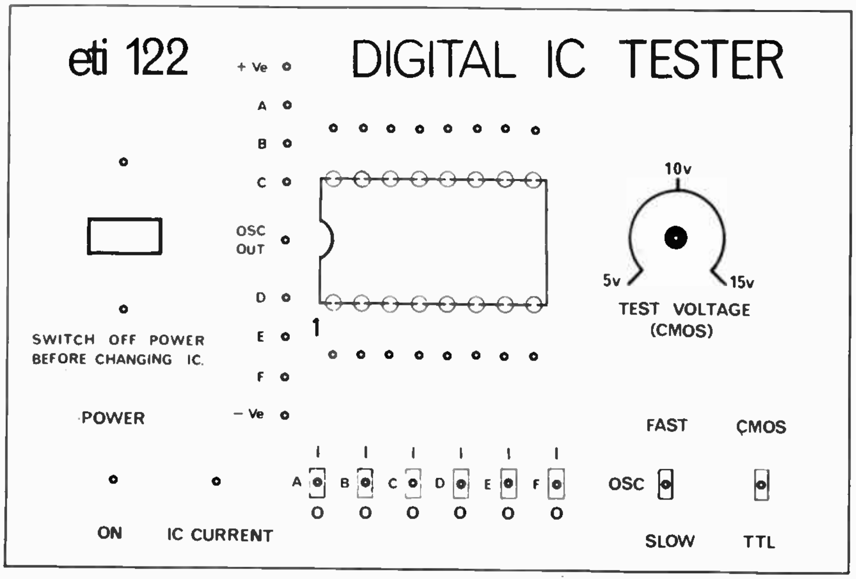


Fig.8. Front panel artwork (shown half-size — full size should be 223 mm x 148 mm).

of the differing functional requirements (eg a shift register versus a two-input NAND gate) of devices within the one family, as well as those between different families. A multi-way switch could be used for each input pin but would greatly increase the expense of the unit. A good alternative is to use patchable links, and this is the approach that we have chosen to use in our unit. In addition we have used a small breadboard socket as the test socket, rather than a standard 16 pin dual-in-line socket, as this allows us to improvise special test circuits for the more complex logic ICs, and the means to breadboard simple circuits.

The need for a variable power supply for CMOS testing presented two additional problems. The first of these was the danger of plugging a TTL IC into the unit when it is set up for CMOS and for some higher supply voltage than the five volts required for TTL. Secondly the LEDs used for monitoring each pin would draw more current as the supply voltage increased. The current ratio could be as high as four to one and a corresponding variation of LED intensity would occur. To overcome this problem it was decided to provide a second supply of five volts to operate the LEDS which will also provide the higher current required by TTL for its operation. The other supply is a variable one for testing CMOS and is not capable of supplying more than 30 mA. Thus a TTL gate inadvertently connected to this supply would not be damaged.

The regulator used for the five-volt supply is a three terminal IC which has built in current limiting and thermal shutdown. It will not therefore be damaged by a short circuit due to testing a faulty IC. It is not possible to construct a discrete design, as cheaply, that has the same performance.

Next we need a device that will detect the state of each pin on the device under test and drive an LED to indicate that state. The device has to be driven by TTL and CMOS outputs, that is, by voltages anywhere between 5 and 15 volts. A suitable IC is the CMOS 4009 IC which has six inverters in one package. Each inverter will monitor a pin without drawing appreciable current. The 4009 is also designed to translate logic levels. Thus we may use it to monitor a 5 to 15 volt input level at its input but provide a five volt signal only at its output.

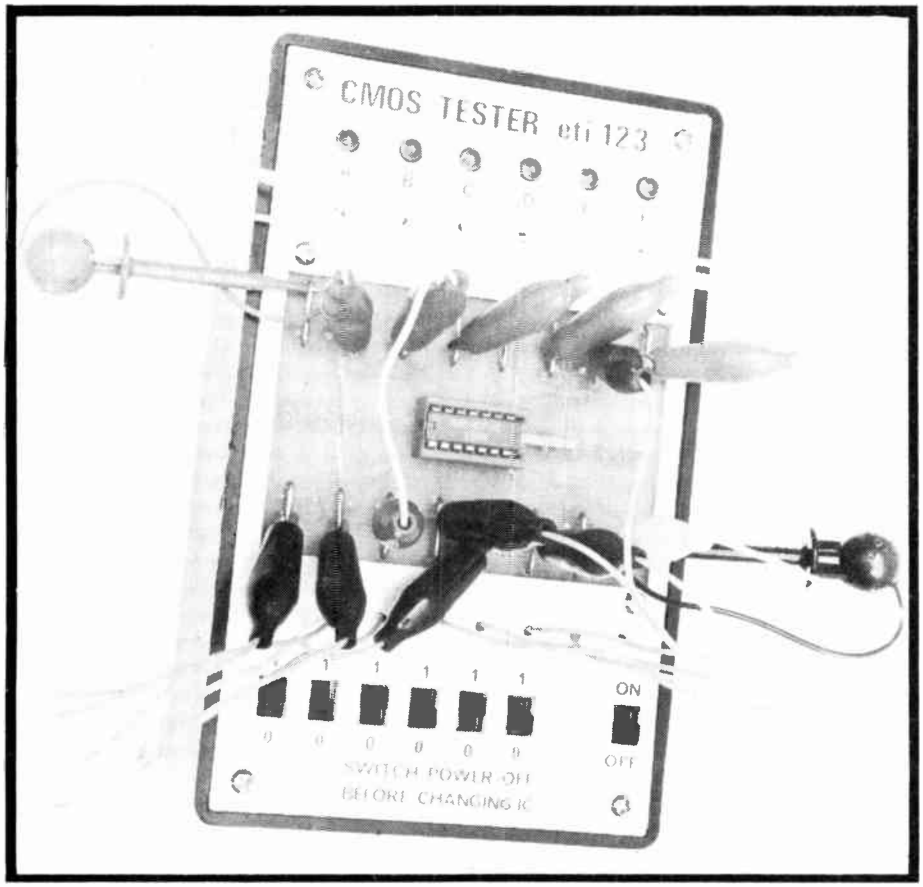
Switches are provided which have debounce logic associated with them. This is necessary so that single bounce free rise and fall transitions can be generated for the testing of more complex logic. The debounce logic must be capable of operating on 5 to 15 volts and of sinking at least two milliamps for TTL tests. The 4009 IC with its high output current capability was again considered to be most suitable for this task.

We would also like to have used the 4009 as the oscillator, but RCA do not recommend using CMOS that has a high output capability in a linear mode as the power dissipation of the device may be exceeded. The oscillator must provide pulses that swing between the positive and negative supply rails (in order to drive CMOS) and must be capable of sinking the two milliamps required by TTL. It must also be capable of operating on supply voltages of 5 to 15 volts. Since the standard CMOS devices cannot provide the current requirement it was decided to use a 555 IC as the oscillator.

CMOS devices should not be operated with inputs left floating as some devices may drift into the linear mode and be destroyed by excessive power dissipation. For this reason a 10 megohm resistor is connected between each pin, on the test socket, and ground. These resistors also conduct away any static charge that may build up.

SIMPLE CMOS TESTER

An inexpensive unit for the hobbyist.



PROJECT 123

NOW THAT the use of CMOS logic is becoming widespread there is an obvious need for a simple CMOS tester suitable for the hobbyist. In last month's issue we described a sophisticated tester for both CMOS and TTL. That particular instrument is very versatile but may be too expensive for many budding experimenters and we have therefore designed this simpler instrument to cater for their needs.

A simple CMOS tester, although being inexpensive, must be capable of performing the majority of tests required for CMOS logic without causing any damage to the ICs under test or being damaged itself. It must also use only those components which are readily available to the average home constructor. The ETI 123 Tester fulfills all these requirements.

The tester circuitry draws very little current except for that drawn by the LEDs. Even the LEDs only draw current whilst a device is actually under test. For this reason we thought that the expense of a mains power supply was unwarranted and chose to use batteries instead. For those who would rather operate the unit from a mains derived supply, one capable of supplying anywhere between 5 and 12 volts at up to 40 milliamps will be suitable. Another major expense, that of providing a large number of programming switches to set up the test conditions, has been alleviated by using flying leads fitted with alligator clips to connect to the IC under test.

Several steps have been taken to prevent damage to the IC by the tester and conversely, damage to the tester by the IC. Firstly each pin of the test socket is fitted with a static discharge resistor to earth. A current limiting resistor, R 37, is in series with the supply so that the tester is protected against damage due to possible excessive current into an internal short in the test IC. This limiting resistor also ensures that current through the input-protection diodes on the IC does not exceed the specified limit of 10 mA.

Only readily available components are used in the tester and, in fact the ICs used are available from at least four different manufacturers.

To test simple gate functions, eg NAND gates, NOR gates, we need at least four switches and a logic level detector but for the more complex functions, eg multipliers, we need at least six switches and six level detectors. A clock — pulse generator is required for the testing of flip flop and other clocked devices. This pulse generator must be free of the contact bounce that is typically encountered with mechanical switches. For this reason we used a pair of CMOS NAND gates wired as an astable multivibrator to generate a continuous train of pulses. This may be used to increment counters and to shift data in shift registers. As it is a CMOS circuit it is perfectly suited to driving other CMOS devices.

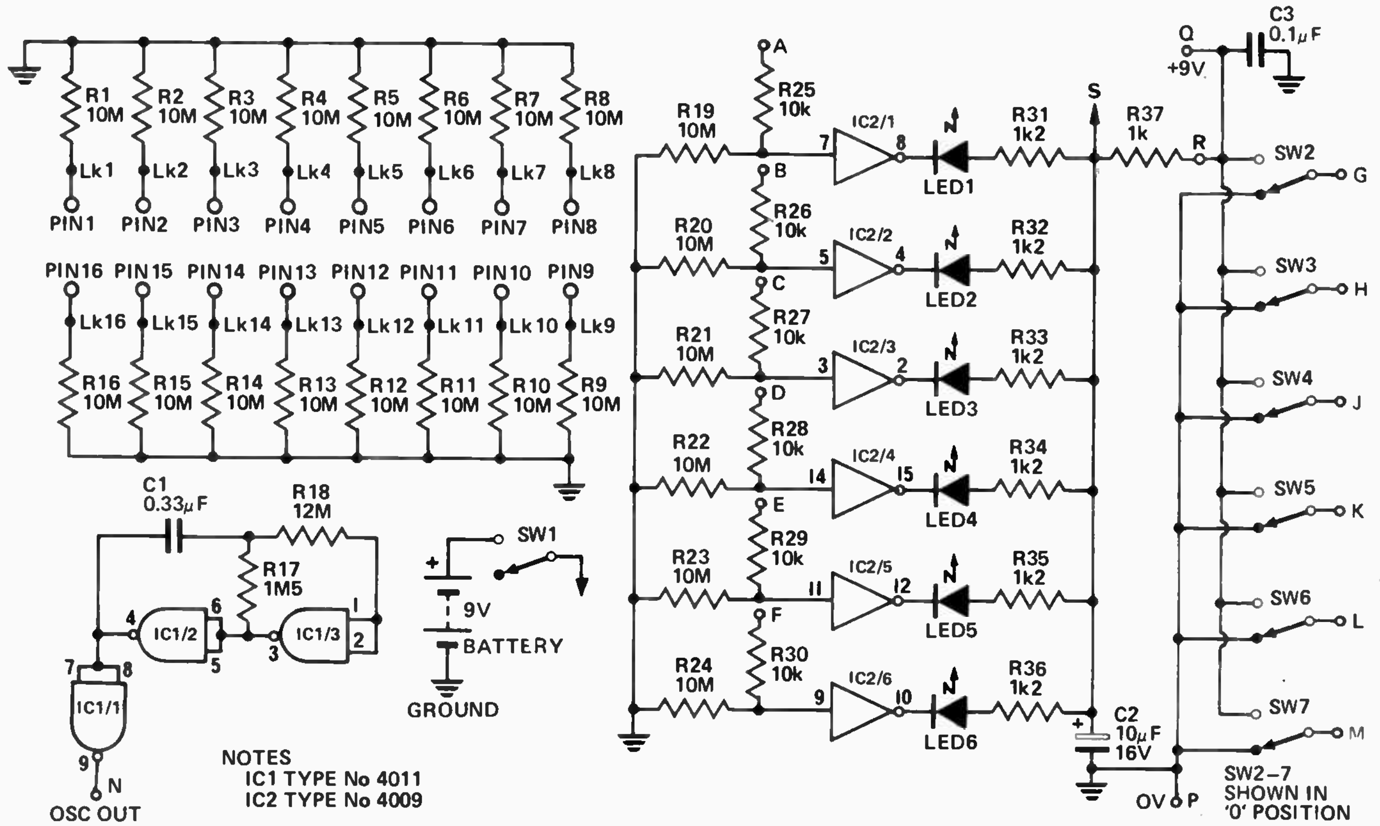
CONSTRUCTION

We recommend that the printed-circuit boards as specified be used as construction is thereby greatly simplified. The printed-circuit boards should be assembled as detailed in the component overlay diagrams. Switches SW1 to SW7 should be mounted by first glueing two strips of printed-circuit board to the front panel (copper side out). The switches may then be soldered to the copper side of the board. This procedure avoids the necessity of having 14 screw heads visible on the front panel.

The test socket is mounted on the non-copper side of board 123b. This board also carries links Lk1 to Lk16 which connect directly to the pins of the test socket. These links are also mounted on the non-copper side of the board and should be of reasonably heavy gauge tinned-copper wire, and should be installed such that sufficient room is under the link to enable test leads to be attached to them by means of alligator clips or Eazy hooks. Resistors R1 to R16 are mounted on the *copper* side of this board so that they are not visible when the board is bolted to the front panel. The top two screws, nearest to the LEDs, should be 18 to 25 mm long so that board 123a may also be mounted on them later.

On board 123a, mount and solder in position on the component side of the

SIMPLE CMOS TESTER



NOTES
IC1 TYPE No 4011
IC2 TYPE No 4009

POWER RAILS ON IC1, 2 NOT SHOWN
PIN 7 on IC1 IS 0V
PIN 12, 13, 14 ON IC1 IS +9V
PIN 8 on IC2 IS 0V
PINS 16 AND 1 ON IC2 IS +9V

POINTS A-Q GO TO ALIGATOR CLIPS

POINT Q IS +9V FOR IC UNDERTEST
POINT P IS 0V FOR IC UNDERTEST

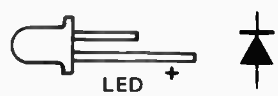


Fig. 1. Circuit diagram of the CMOS tester.

HOW IT WORKS – ETI 123

The ETI 123 CMOS tester can be described in three separate sections. Firstly there is the test socket for the device under test. The test socket is mounted on a printed circuit board which also holds a 10 megohm static-discharge resistor to protect each pin of the IC. Each IC pin is also connected to a surface mounted link by which connections can be made to the IC.

The next major section of the tester contains detectors which monitor the voltage at each pin of the IC. Each detector consists of a CMOS inverter which derives an LED indicator. When the voltage at the input of the inverter is greater than half the supply voltage the LED will be alight. Conversely the LED will be off when the voltage at the input to the inverter is below half supply voltage. Resistors R19 to R30 protect IC2 against static charges and from the condition where a detector has no input. Resistors R31 to R36 set the operating currents for the LEDS.

The final section contains switches SW2 to SW7 and a clock oscillator. The output of the switches can be either 0 volts or +9 volts that is, a logic '0' or a logic '1'. These outputs are made available at test leads which may be connected to the IC under test as required. To protect the tester against internal shorts on the IC under test, and incorrect connections, R37 has been inserted in series with the supply rail to limit the current that may be drawn to a level which cannot cause any damage.

IC 1/2 and IC 2/3 are wired as an astable multivibrator where the frequency of oscillation is determined by the time constant of C1 and R17, whilst R18 is used to protect the input of IC 1/3 from any voltage excursions past the supply rails. IC 1/1 is used as an inverting buffer and the output of the circuit is made available at the front panel by means of a lead and alligator clip.

PARTS LIST — ETI 123

R37	Resistor	1k	¼ Watt 5%
R31-36	"	1.2k	" "
R25-30	"	10k	" "
R17	"	1.5M	" "
R1-16	"	10M	" "
R19-24	"	10M	" "
R18	"	12M	" "

C3 Capacitor 0.1µF polyester
C1 " 0.33µF "
C2 " 10µF 16 electrolytic
IC1 Integrated Circuit 4011 (CMOS)
IC2 " " 4009 (CMOS)
LED 1-6 Light Emitting Diode RL 4484 or similar
SW1-7 Miniature slider switch 2 pole 2 position.
IC Socket 16 pin DIL (preferably with IC removing slide)
Case 160 x 90 x 50 mm plastic box with aluminium front panel UBI
Alligator clips (15)
Battery 9 volts (6 size AA cells)

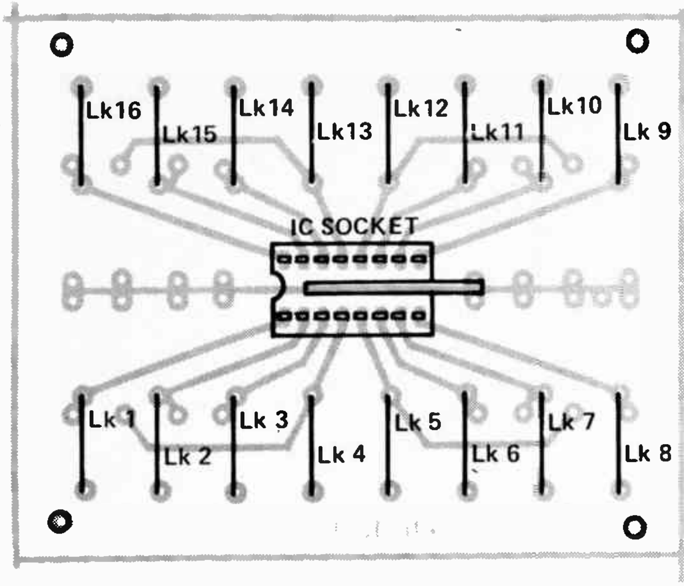


Fig. 2. Component overlay for the test-socket board ETI-123b, non-copper side.

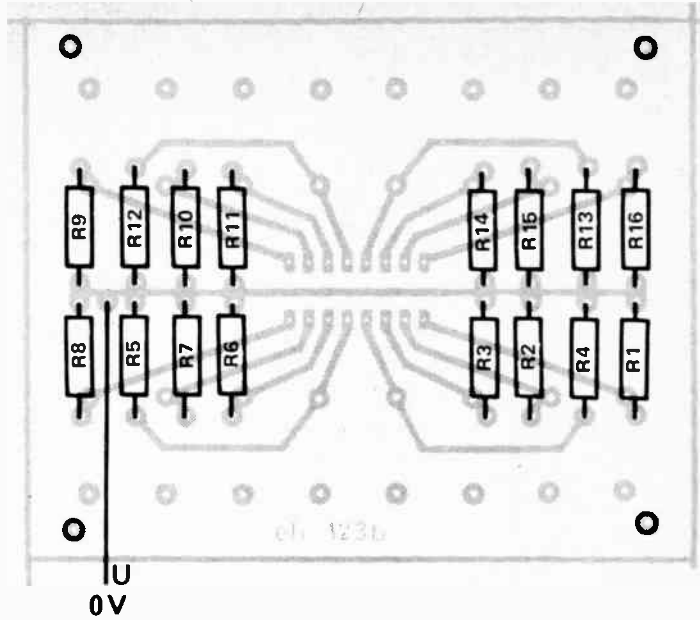


Fig. 3. Component overlay for the copper side of board ETI-123b.

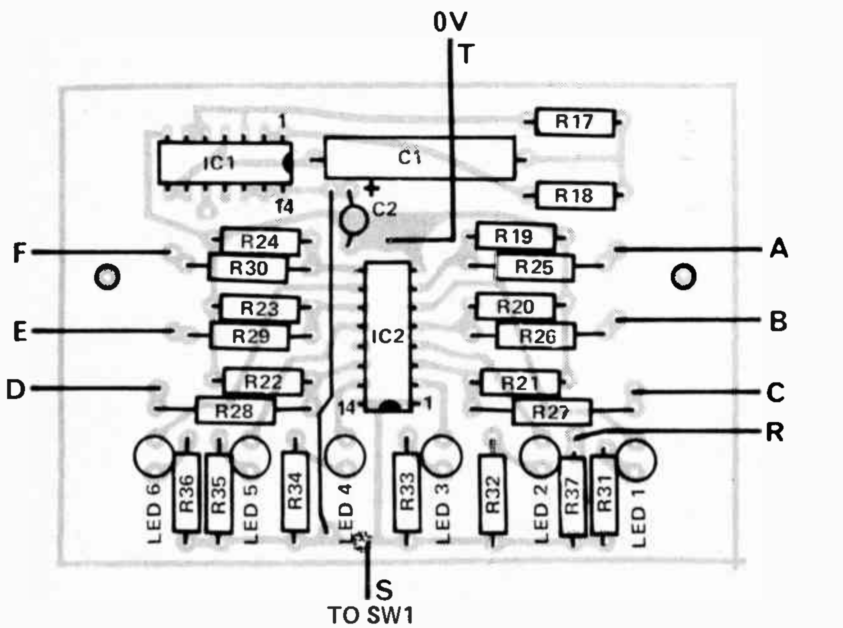


Fig. 4. Component overlay for board ETI-123a. Note that C1 may need to be mounted on reverse side, and that the LEDs should be mounted as detailed in the text.

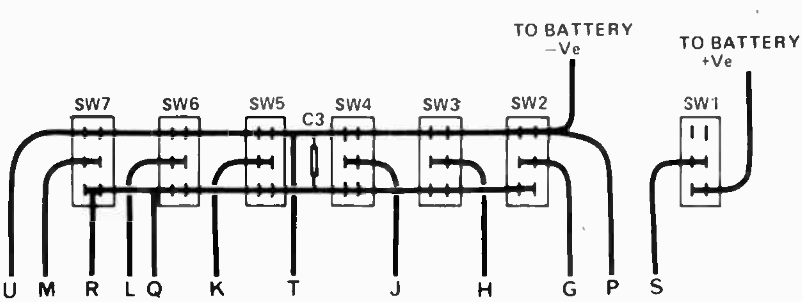


Fig. 5. Switch interconnection diagram. Note that C3 is mounted across one of the switches.

board, all components with the exception of the LEDs and capacitor C1. As C1 needs to be a polyester type it may be physically too big to be mounted on the component side without fouling the front panel and should therefore be mounted on the copper side. The LEDs should be inserted in their positions but not yet soldered. Temporarily mount the board in position such that the LEDs protrude through their correct holes in the front panel. Keeping the front panel face down, solder the LEDs into the board. Remove the board and solder 150 mm lengths of hookup wire to the points marked A to F on the overlay and pass these leads through the corresponding holes in the front panel. Do the same for the leads G, H. J. K L M P and Q from switches SW2 to SW7 using a different coloured wire to that used previously. These wires should also be passed through the appropriate holes in the front panel.

Finally solder alligator clips or Eazy hooks to the ends of all these leads and connect supply and earth leads to the 123b board. Check both boards for wiring errors or errors in component insertion before bolting board 123a in position. The battery may then be connected and the unit is ready for use.

Note that if the type UB1 box is used as in our protype the top corners of the 123a board may have to have the corners trimmed off at 45 degrees so that the board will fit in the box

SIMPLE CMOS TESTER

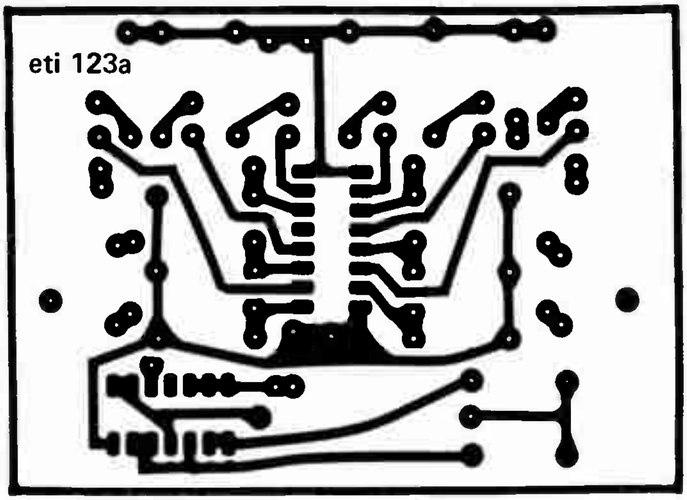


Fig. 6. Printed-circuit board layout – ETI 123a. Full size 88 x 63 mm.

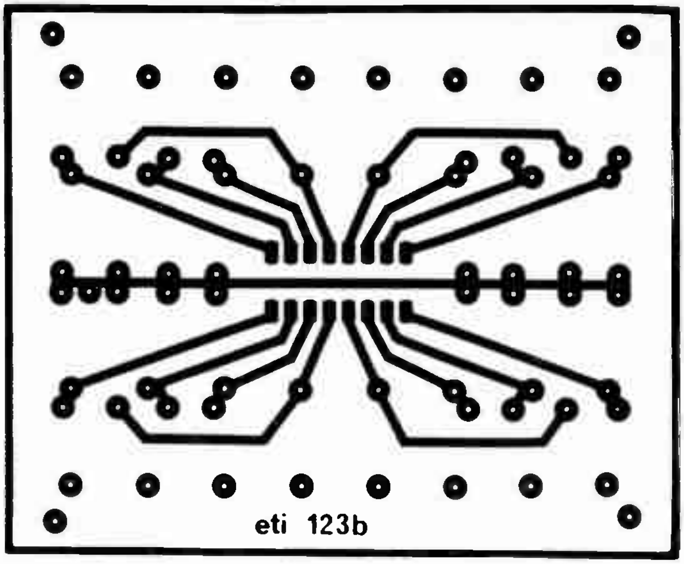


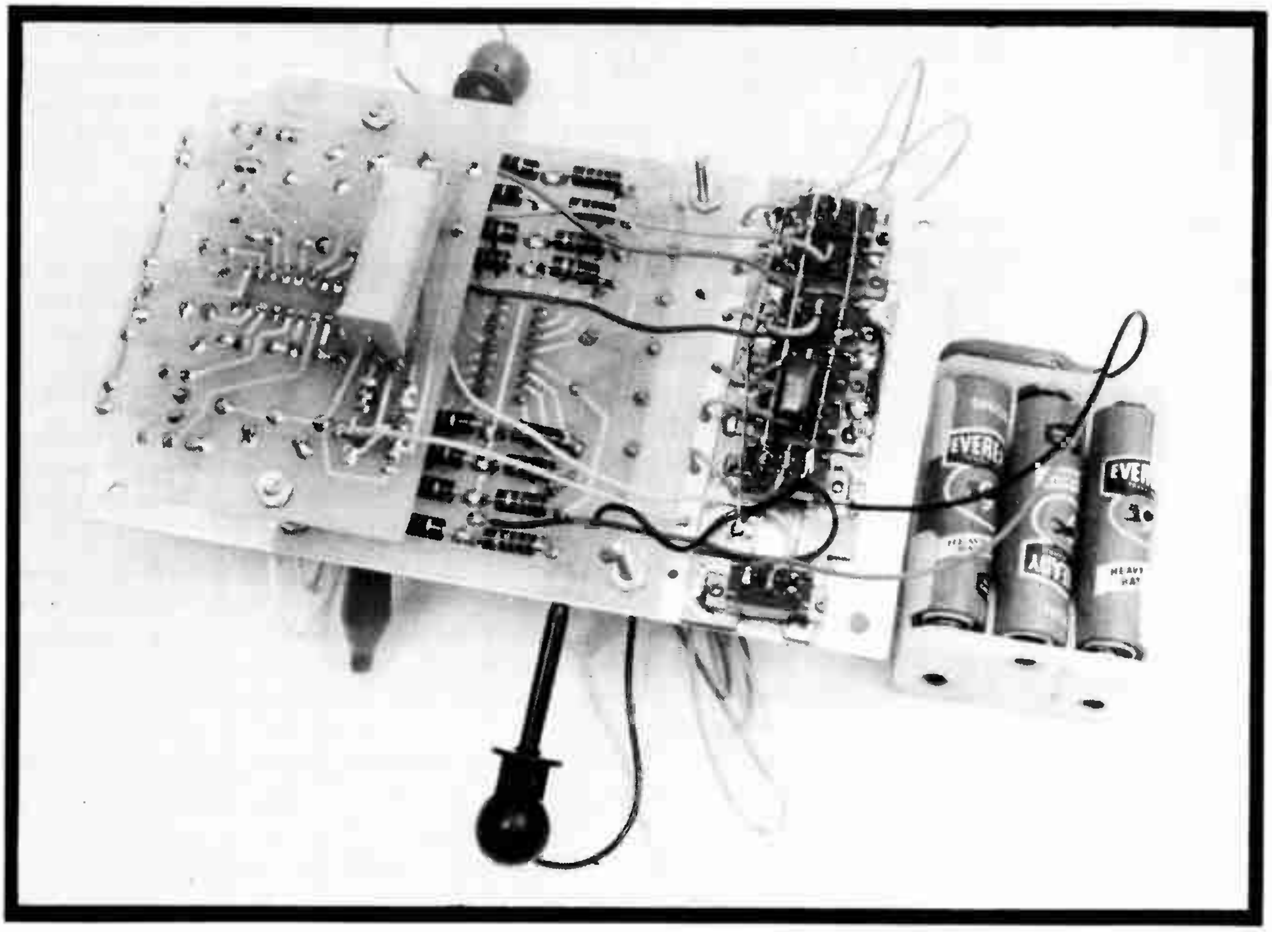
Fig. 7. Printed-circuit board layout – ETI 123b. Full size 88 x 71 mm.

without fouling the mounting pillars for the front panel.

OPERATION

Before testing or inserting any IC make sure that the power is switched off. Set up the operating conditions for the IC to be tested either by consulting the manufacturers data or by duplicating the conditions under which the IC will be used in the circuit.

Next insert the IC to be tested into the test socket and connect the power supply leads to the links for appropriate pins of the IC. Double check these connections to make absolutely sure that these connections are correct. Reversed power connections will destroy the IC. Switch on the tester and use the input switches to systematically apply all the possible input conditions to the IC whilst noting that the output conditions of the IC are as they are supposed to be. ●

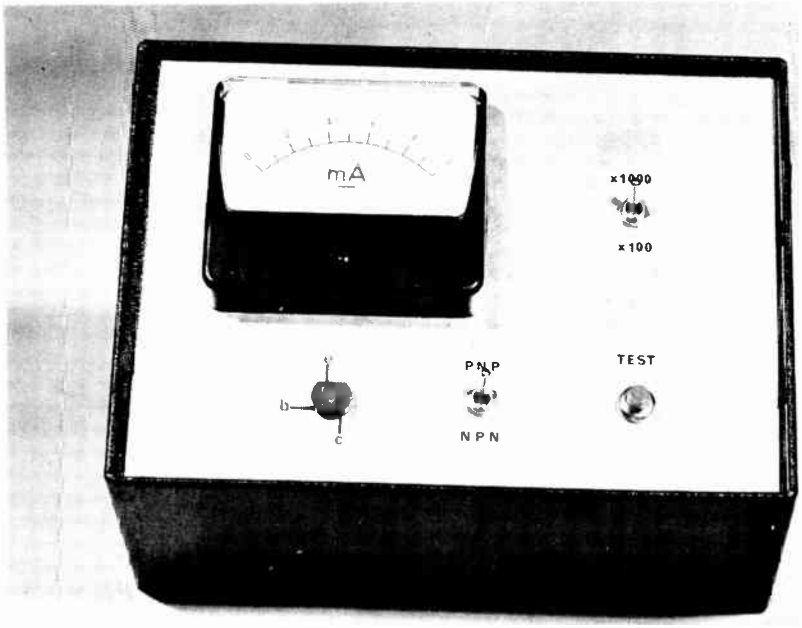


Internal view of the tester. Note how the top board is mounted (see text).

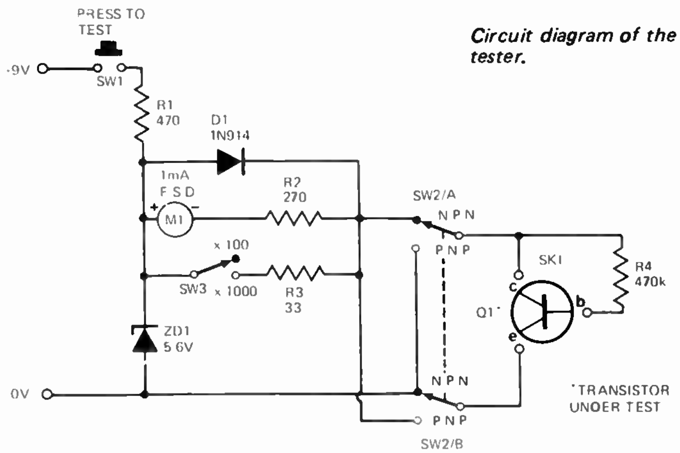
TRANSISTOR TESTER

PROJECT 222

Measure and test your transistors with this easily built device.



The transistor tester mounted in a metal case.



Circuit diagram of the ETI transistor tester.

EXPERIMENTERS will frequently use the same transistors in a whole sequence of experimental circuits, for recovering and re-using such components saves considerable outlay.

But semiconductors are easily damaged — by incorrect operating conditions — or by excessive application of heat when soldering.

Only too often a malfunctioning experimental circuit will be checked and rechecked before one realises that a transistor is dead.

A transistor tester will save hours of such frustrating and unproductive effort.

Transistors can often be bought cheaply in bulk — usually in unmarked and untested lots — or recovered from old computer boards. Here again a transistor tester will prove invaluable in eliminating the faulty bits.

The simple transistor tester described in this project not only sorts out the good from the bad but indicates also the approximate gain (β) of the transistor. This is a most useful feature for those circuits where transistors need to be matched. Two ranges of gain (beta) are provided, 0-100, and 0-1000. The tester may also be used to check transistor polarity. ●

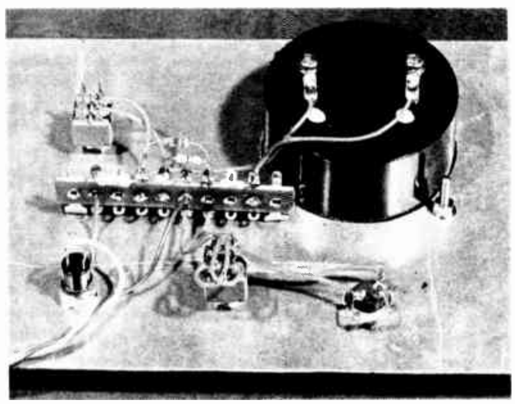
PARTS LIST — Transistor Tester — ETI 222

- R3 Resistor 33Ω ½ watt 5%
- R2 Resistor 270Ω ½ watt 5%
- R1 Resistor 470Ω ½ watt 5%
- R4 Resistor 470k ½ watt 5%
- D1 Diode IN914
- ZD1 Zener diode BZY88C5V6
- SW1 Push button push-to-make
- SW2 Switch toggle DPST
- SW3 Switch toggle SPST
- 9V battery
- M1 Meter 1mA movement
- SK1 Socket T05 transistor type
- Metal case or minibox

HOW IT WORKS

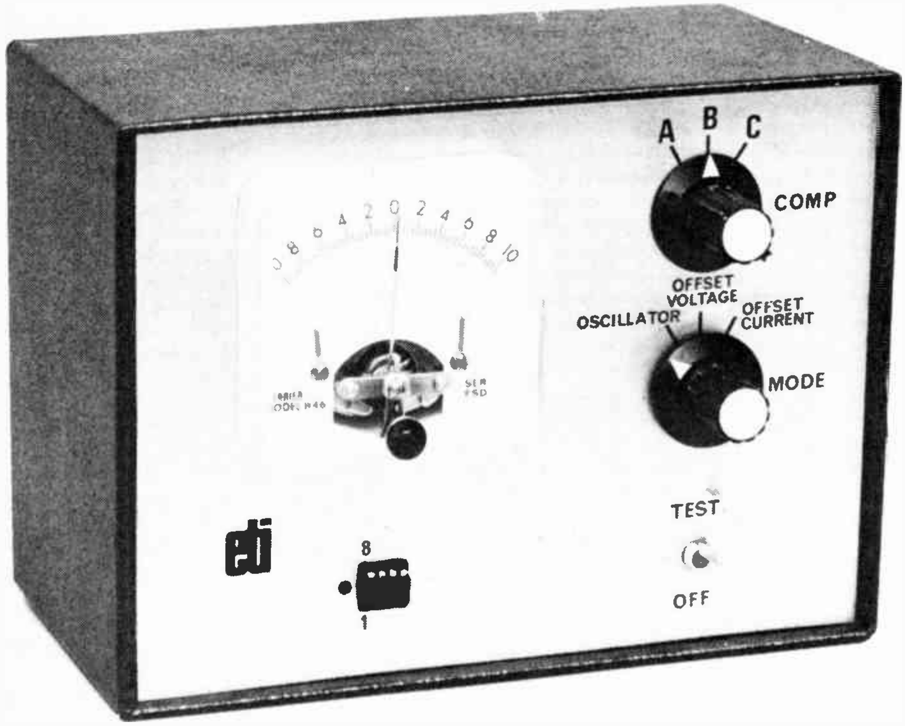
Operation of the tester is very simple. The meter, M1, monitors the collector current of the transistor under test whilst R4 supplies a current of about 10 μA into the base of the test transistor. Thus, on the 100β range, the maximum collector current will be 1 mA and, on the 1000β range, 10 mA. Switch SW3 therefore changes the meter sensitivity according to the beta range selected.

The meter is protected by means of D1 against damage due to test transistors being shorted. The zener diode ZD1 stabilizes the battery voltage to 5.6V.



The construction method may readily be seen from this photograph of the back of the front panel.

LINEAR IC TESTER



Test all commonly available operational amplifiers for three vital parameters.

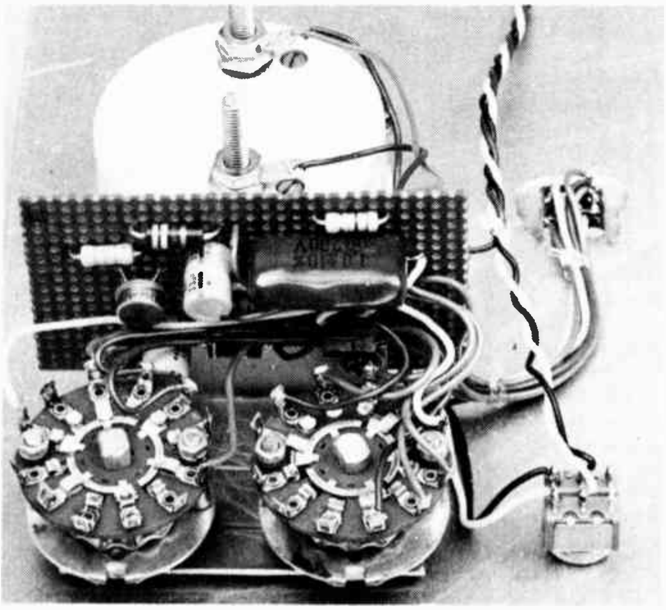
ETI PROJECT 115

LINEAR integrated circuits are available today at prices little higher than those of discrete transistors. As they offer far better performance parameters, and greater versatility than transistors they are being used in new designs in ever increasing numbers.

Most linear ICs are now built into a standard 8-pin, dual-in-line plastic pack, have the same pin connections and very similar characteristics. Hence as the only real difference is in the associated frequency compensation network, a universal, linear — IC tester is quite a feasible proposition.

The tester, described here provides a quick check of vital operating parameters. Checks are provided for offset voltage (max ±10mV), offset current (max ± 1000 nA) and of operation in an actual circuit configuation.

It is a most valuable instrument; saving an experimenter time that would otherwise be spent tracing down faulty ICs.



Showing the internal construction of the tester. Note that matrix board holds the majority of the components.

CONSTRUCTION

We chose to mount our circuitry on a small piece of matrix board, rather than a printed circuit board, as there are relatively few components used.

Make sure that IC1 is orientated correctly (note pins 1, 5 and 8 are not used). The wires from the compensation switch (SW2) should be as short as possible in order to minimise the chance of unstable operation.

The test socket should be glued into place (taking care not to get glue down the pins) and, after the wires to the socket are soldered on, these should also be held to the panel with glue or a metal clamp.

The wires to the socket must be supported in some way, as detailed above, to prevent the rather fragile pins breaking off.

HOW TO USE

The parameters of commonly-available ICs are detailed in Table 1. An IC on test should not exceed these figures. Those that do exceed these values may not operate correctly in some circuits and should be discarded.

To test an IC, plug it into the test socket making sure that it is orientated correctly. Select the appropriate equalization as detailed in column 4 of Table 1 and switch the unit on. Select 'OSCILLATOR' mode and observe that the meter should sweep up and down the scale at about 1 Hz.

Now switch to 'OFFSET VOLTAGE' mode and read the meter which is calibrated at 10 mV full scale deflection.

Next switch to 'OFFSET CURRENT'. In this mode the meter is calibrated at 1000 nA (1 microamp) full scale deflection.

Discard any IC that does not oscillate or has excessive offset current or voltage.

TABLE 1

TYPE	MAX OFFSET CURRENT	MAX OFFSET VOLTAGE	COMPENSATION
301	50 nA	±7.5 mV	C
307	50 nA	±7.5 mV	A
308	1 nA	±7.5 mV	C
709	500 nA	±7.5 mV	B
741	200 nA	±6 mV	A
748	200 nA	±6 mV	C
777	20 nA	±5 mV	C
1456	30 nA	±12 mV	A

PARTS LIST ETI 115

R1,2	Resistor	100	5% ½W
R3,5	"	2.2M	" "
R4	"	1 M	" "
R6	"	22 k	" "
R7	"	1.5 k	" "
R8	"	3.9 k	" "
R9	"	33 k	" "
R10	"	150 k	" "
C1	Capacitor	1μF polyester	
C2	"	0.0047μF polyester	
C3	"	33pF	ceramic
C4	"	220pF	"
C5,6	"	10μF 16V electrolytic	

IC1	integrated circuit	μA741
M1	meter	0.5 mA — 0 — 0.5 mA Ferrier Instruments model B46 or similar
SW1	Switch	2 pole, 3 position rotary
SW2	"	2 pole, 3 position rotary
SW3	"	2 pole on-off toggle.

Metal box approx. 150 x 180 x 90mm. (Aust. Trans 70-50-40) or similar.
2 x 9V battery (type 216 or similar).

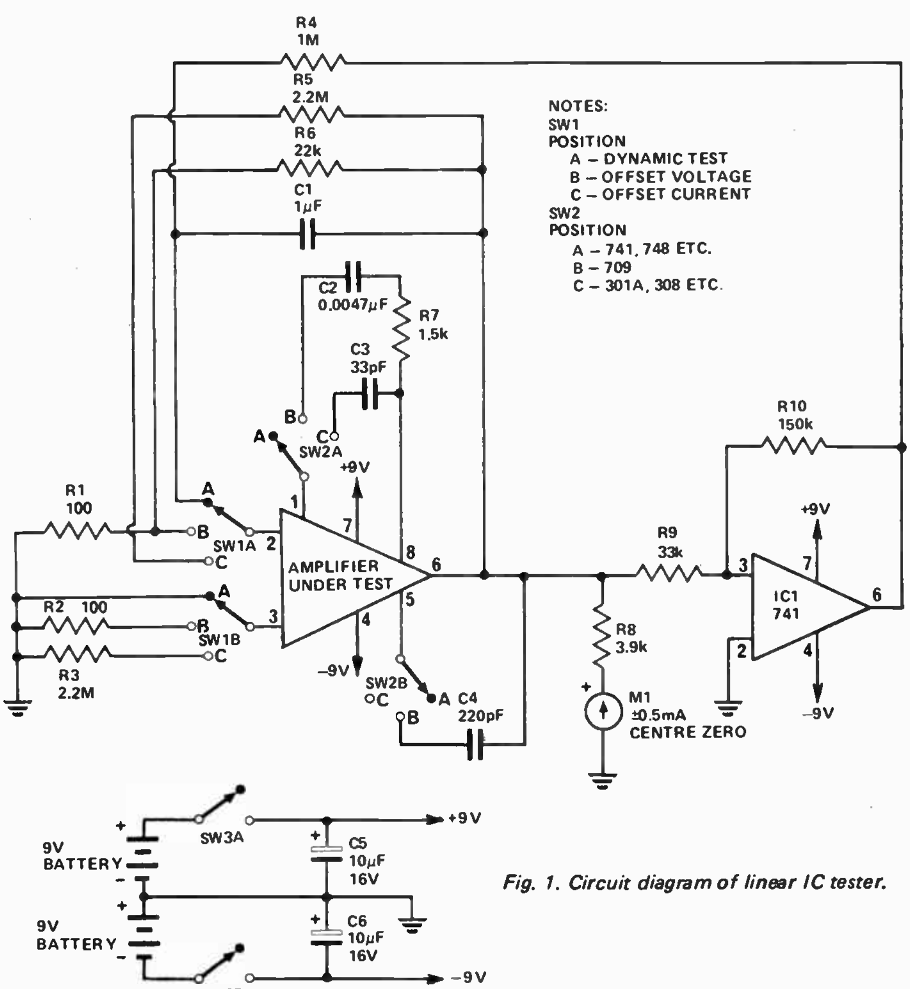


Fig. 1. Circuit diagram of linear IC tester.

HOW IT WORKS – ETI 115

Centre-zero meter M1, via resistor R8, indicates the output voltage from the IC under test. The frequency compensation components for the particular IC under test are selected by SW2, and the test mode is selected by SW1.

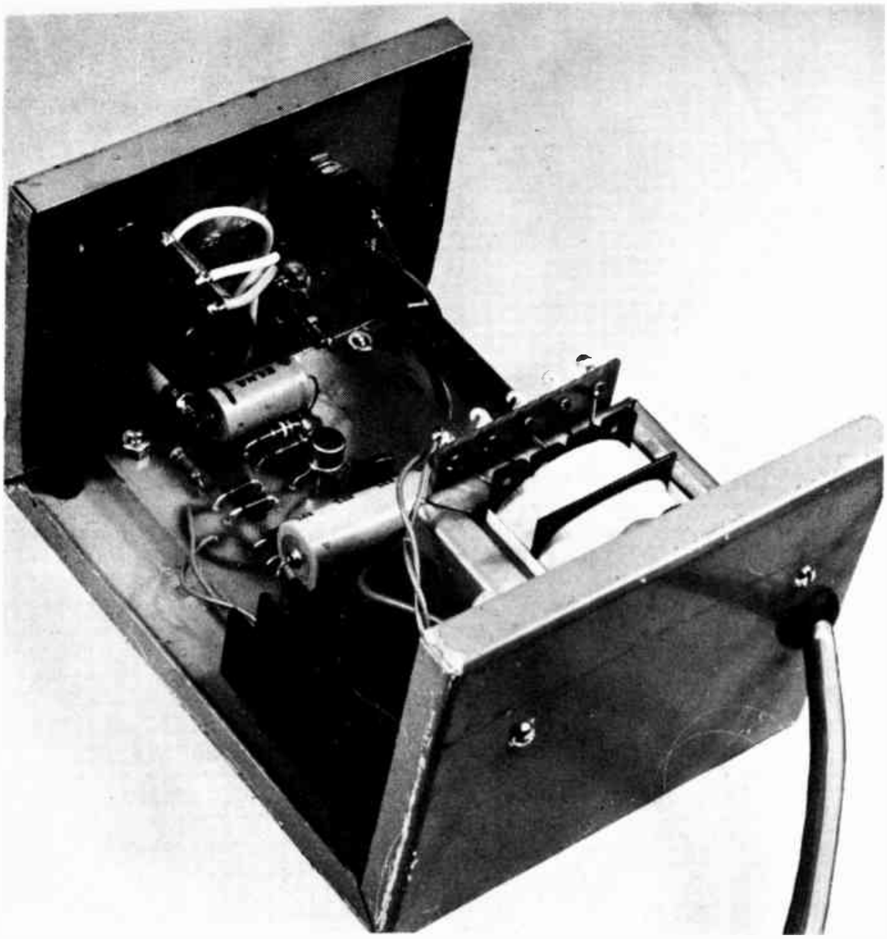
In position "C", of SW1, a 2.2 megohm resistor is connected from the output (pin 6) of the IC under test to the inverting input (pin 2), and a 2.2 megohm resistor from the non-inverting input (pin 3) to ground. Current is drawn by both pin 2 and pin 3 of the IC and, if these currents are equal, the output voltage will be zero. Any difference in input currents will therefore be indicated as an output voltage on meter M1.

In position B the resistor from pin 6 to pin 2 is reduced to 22k and a 100 ohm resistor, R1, is connected from pin 2 to ground. This results in the IC having a voltage gain of 220. Resistor R2 is also made 100 Ω so that offset current does not affect the operation in this mode. Hence the IC will now amplify any offset voltage between pin 2 and pin3 (that is, it is operating in the linear mode) by 220 and the meter deflection will be proportional to the offset voltage.

If either offset voltage or offset current are excessive the meter will read off scale and the IC should be discarded.

In mode A the IC is connected as a triangular wave oscillator having an operating frequency of 1Hz. Integrated circuit IC1 is connected as a Schmitt trigger where the output of the Schmitt goes high if its input drops below -1.5 volts, and will go low if the input exceeds 1.5 volts. The output of IC1 is taken, via a 1 megohm resistor, to the input of the IC under test and the output of the Test IC becomes the input of the Schmitt trigger. An integrating capacitor, C1, is connected across the IC under test. The effect of this is to cause the output of the test IC to rise at 7 volts per second until +1.5 volts is reached. At this point the Schmitt operates and the output of the test IC now commences to fall at the same rate. When -1.5 volts is reached the direction reverses again and the cycle repeats. Thus we have an oscillator with a frequency low enough to be followed by the output meter as an indication of correct operation.

IC POWER



IN June 1971 we published constructional details of a 'Logic IC Power Supply' which was specifically intended to power RTL and TTL circuitry. Since then we have received a surprisingly large number of requests to provide details of a similar unit with an extended voltage range.

Here then are details of a simple yet versatile power supply capable of delivering 1 amp up to 10 volts and ½ amp up to 15 volts.

The unit may readily be adapted to operate over other voltage and current ranges.

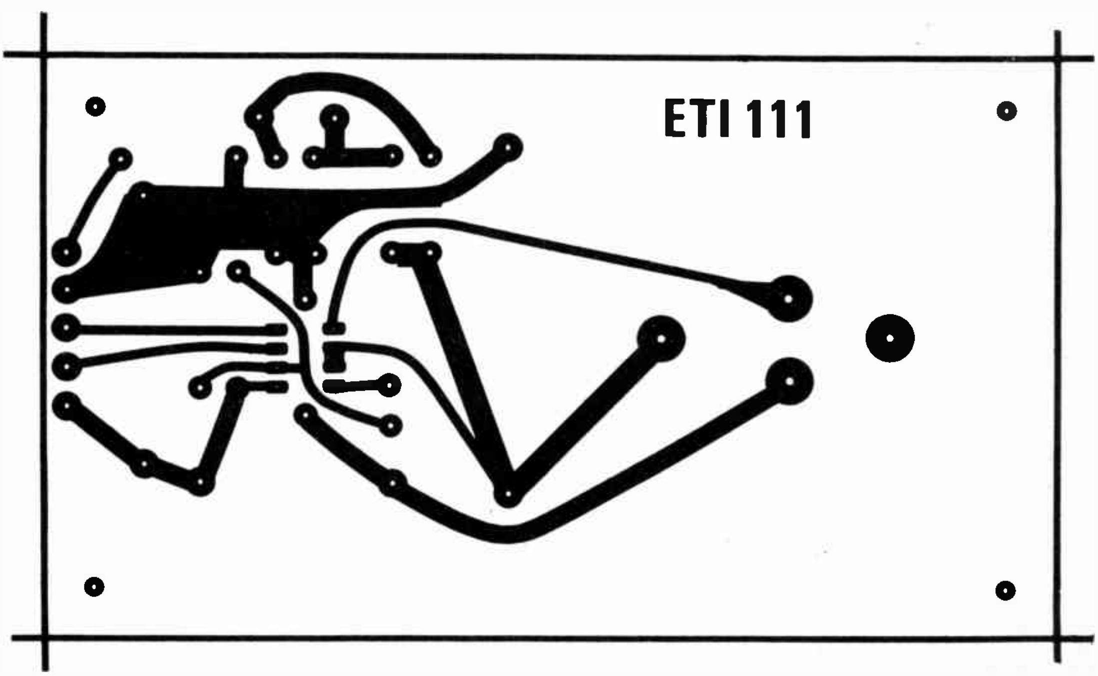
As with the previous unit, refinements such as output voltage and current metering, variable current limiting etc. may be added to the basic circuit.

VOLTAGE REGULATOR IC

The control circuit of this supply is formed by the integrated circuit precision voltage regulator – shown as IC1 in Fig. 1. This IC is now produced by a number of companies including SGS, Fairchild and Motorola (respective type numbers are included in the parts list for this project).

The integrated circuit is a monolithic voltage regulator constructed on a single silicon chip using the planar expitaxial process. The device consists of a temperature compensated

ETI PROJECT 111



Foil pattern for logic power supply (actual size).

SUPPLY

Simple, adjustable power source has innumerable applications.

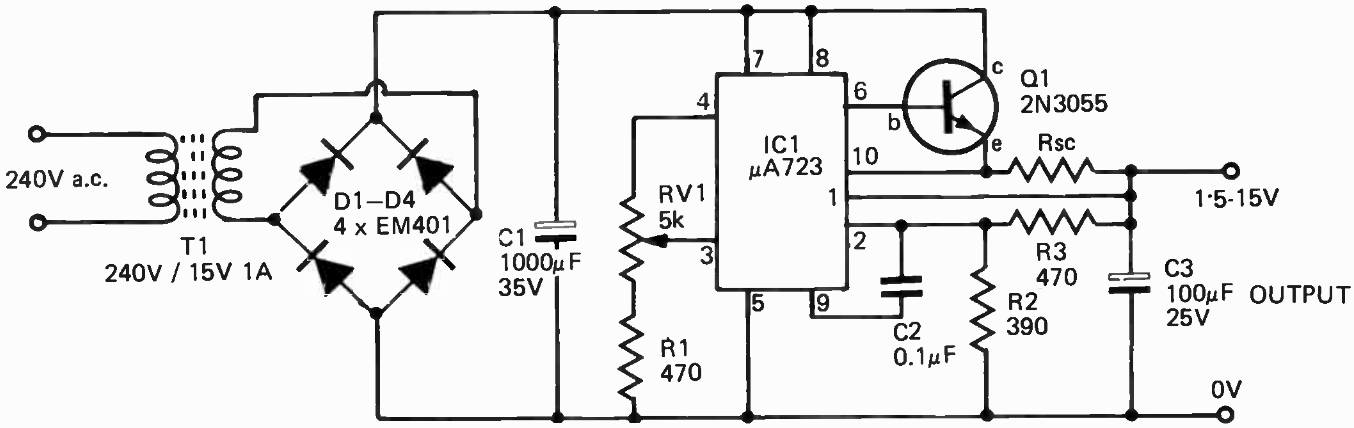


Fig. 1. Circuit diagram of regulated supply.

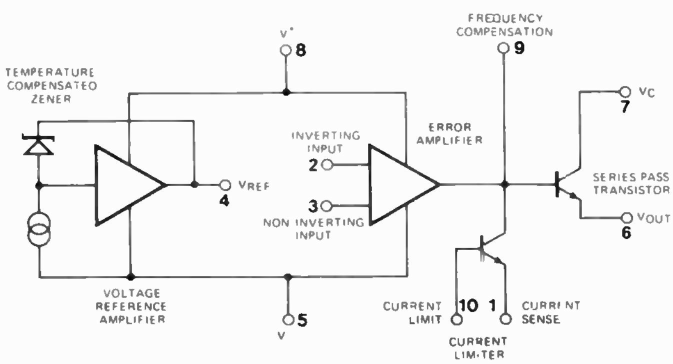
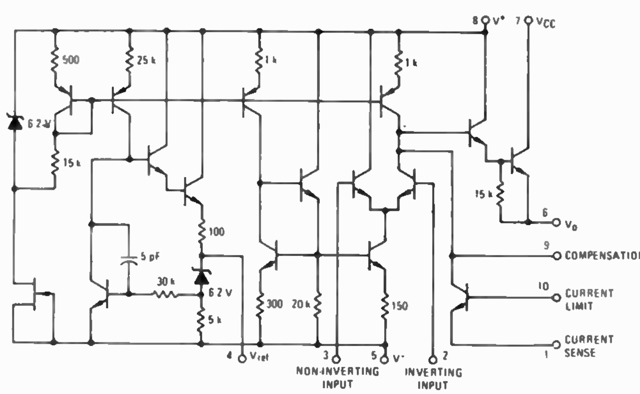


Fig. 2. Simplified schematic of uA723.



Circuit schematic of IC uA723.

reference amplifier, error amplifier, power series-pass transistor and current limiting circuit. Additional external npn and pnp pass elements may be used when output currents exceeding 150mA (from the IC) are required. Provision is made for adjustable current limiting and remote shut-down. In addition to this the IC features low standby current drain, low temperature drift and high ripple rejection.

CONSTRUCTION

Our prototype unit was built on an epoxy glass board, however the constructional method is not critical and the unit may alternatively be built on matrix board, tag strips etc.

The power transistor is mounted on a 2" strip of extruded heatsink which in turn is located on the printed circuit board by the same screws that locate the transistor. One of these screws is

HOW IT WORKS

Figure 2 shows a simplified equivalent circuit of IC1. The voltage reference amplifier produces (typically) 7.15V at pin 4, this voltage has a maximum temperature coefficient of 0.015%/°C.

The V_{ref} voltage is taken to potentiometer RV1 which enables it to be varied between 0.7V and 7.15V. The error amplifier (within the IC) drives a power transistor (also within the chip), and this in turn drives the external series pass transistor Q1.

The output of Q1 is divided by R2 and R3 (÷2.2) and this voltage provides the feedback signal for the error amplifier. Hence the output voltage will be approximately 2.2 times the voltage on RV1.

Current limiting is determined by the voltage drop across RSC. If this exceeds 0.6V, the current limit transistor within the IC becomes forward biased and bypasses any further increase in drive current from the output stage.

The max. output voltage and current of this unit is a function of the transformer, filter capacitor, and the heatsinking of Q1. The prototype unit used a 15V centre tapped 1A transformer (AR 2155) and this provided 1A up to 10V and ½A at 15V. The drop in output current is due to rectified dc voltage decreasing on load. If a higher voltage transformer is used – or one with a higher current rating, thus providing better regulation – then higher output currents may be expected.

The maximum output voltage may be altered by changing the ratio of R2 and R3. Note that the maximum no-load voltage across C1 should not exceed 35V.

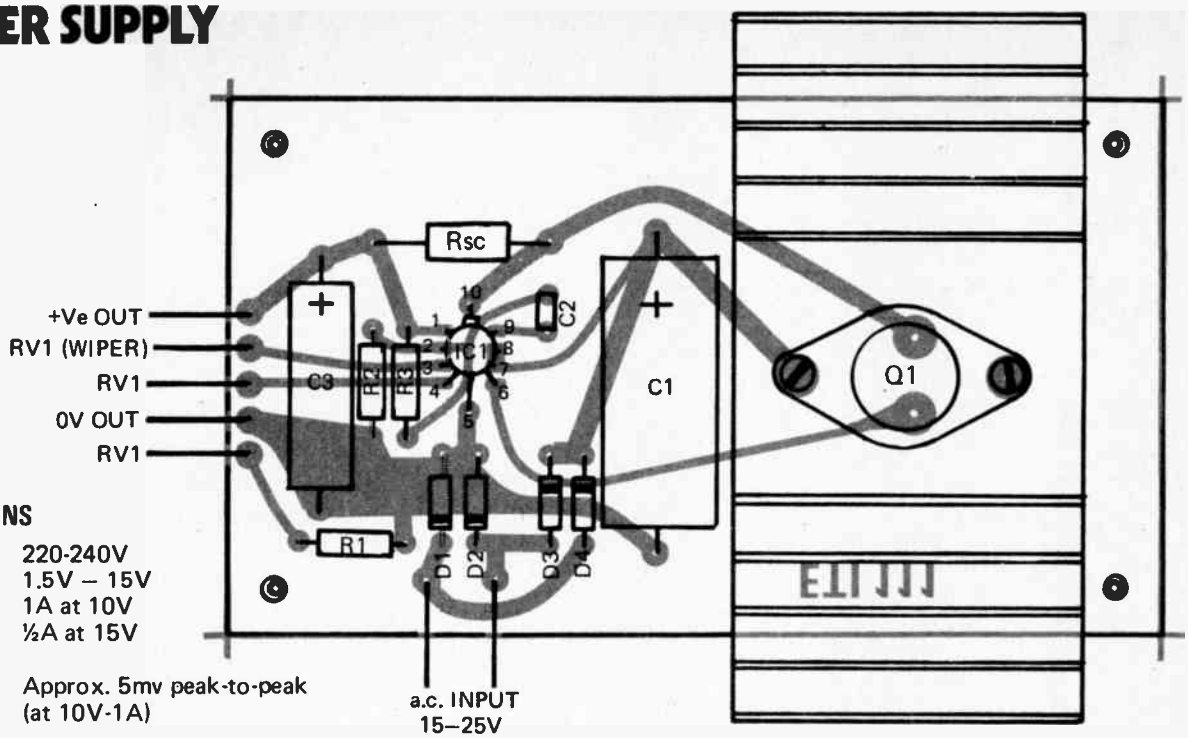


Fig. 3. Layout of components on circuit board.

SPECIFICATIONS

Input Voltage	220-240V
Output Voltage	1.5V – 15V
Output current	1A at 10V ½A at 15V
Ripple	Approx. 5mv peak-to-peak (at 10V-1A)
Regulation	0-100mA = 4mV (10 Volt) 0-1.0A = 20mV (10 Volt)

also used for the electrical connection for the collector of the transistor.

The IC may be soldered directly into the circuit – ensure that the device is correctly orientated – and avoid excess heat. Recommended maximum lead temperature during soldering is 300°C.

A load sensing resistor (RSC) is used to provide overload protection. In our prototype we used a short length of resistance wire cut to length to limit the current to the desired value. An interesting alternative is to substitute a 20 ohms 5 Watt wire-wound potentiometer for RSC. This enables the current limiting facility to be steplessly varied. With this feature the user can start experimenting with a very low current limit and then increase the current when the circuit is operating correctly.

The basic circuit described in this article can be modified to provide other ranges of voltage and current. The main design limitations are that the voltage across the IC must not exceed 40V and that the output current from the IC must not exceed 150mA, or 800mW of power.

Transistor Q1 (2N 3055) is capable of dissipating up to 115 watts but if power levels of this magnitude are envisaged then a second transistor should be added, in a Darlington pair configuration, to transistor Q1. This will reduce the loading on IC1. A arger heat sink will also be required.

INCREASED RIPPLE REJECTION

The integrated circuit chosen for this project has a typical ripple rejection of 74 dB. This is more than adequate for most applications. However by additional filtering at the non-inverting input (pin 3), the ripple can be even further reduced. A typical performance, using a 4.7μF capacitor across the non-inverting input and V_{ref} is approximately 86 dB. ●

RSC – TYPICAL VALUES

Value of RSC	Current Limiting
10 ohms	65mA
1 ohm	650mA
0.5 ohms	1.4A
0.2 ohms	3.2A

PARTS LIST ETI 111

R1	resistor 470 ohm ½W 5%
R2	resistor 390 ohm ½W 5%
R3	resistor 470 ohm ½W 5%
RSC	see text
RV1	potentiometer 5k linear
D1-D4	diodes EM401, 1N4005 or similar
Q1	transistor 2N3055
IC1	integrated circuit uA723 (or SGS L123, or MC1723CG) (metal can types)
C1	capacitor 1000 uF 35V electrolytic
C2	capacitor 0.1 uF 100V
C3	capacitor 100 uF 25V electrolytic
T1	transformer 240V primary 15-20V sec @ 1A

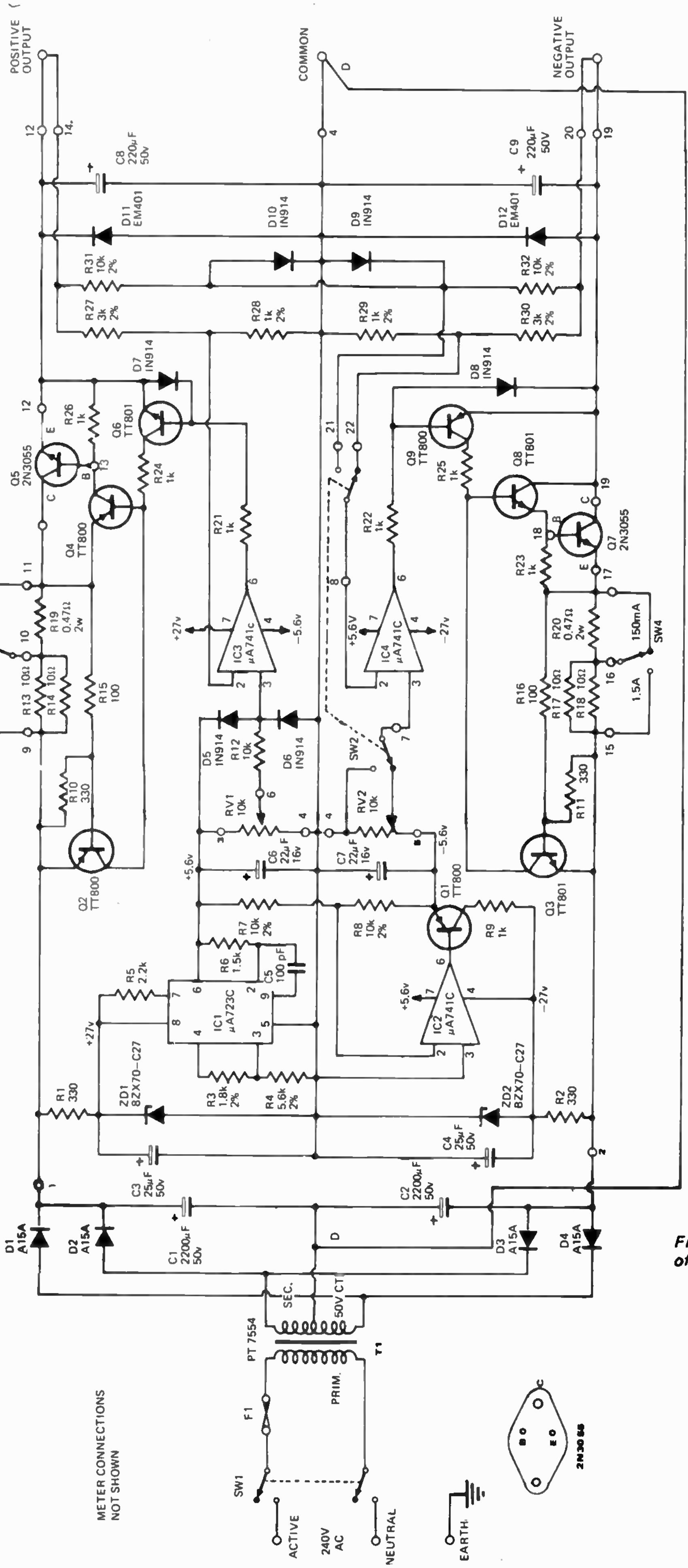
PC board ETI 111
on-off switch, terminals, knob, 3 core flex and plug, metal box approximately 4½ x 3½ x 6 etc.

DUAL

By BARRY WILKINSON

Specifically intended for powering experimental integrated circuit projects, this power unit features independent positive and negative supplies — but with automatic tracking when required.

PROJECT 105



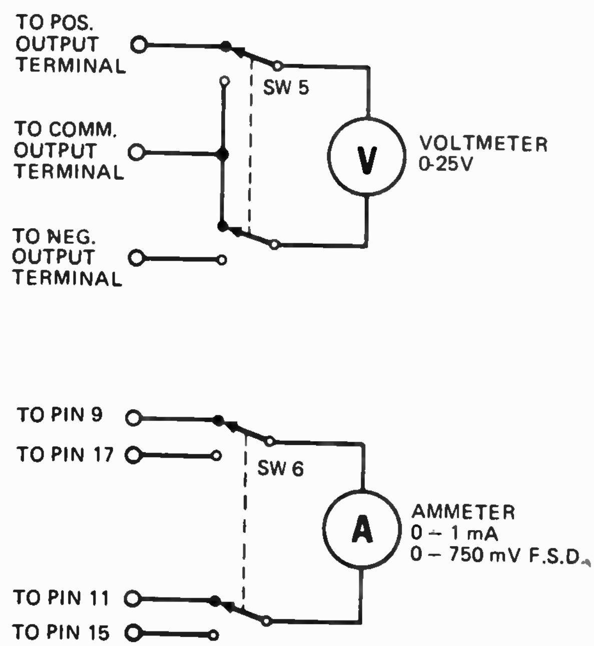


Fig. 1. Circuit diagram of complete unit.

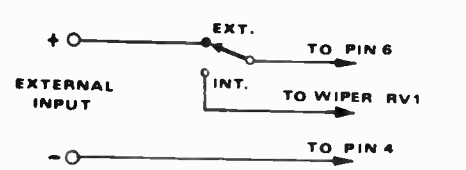
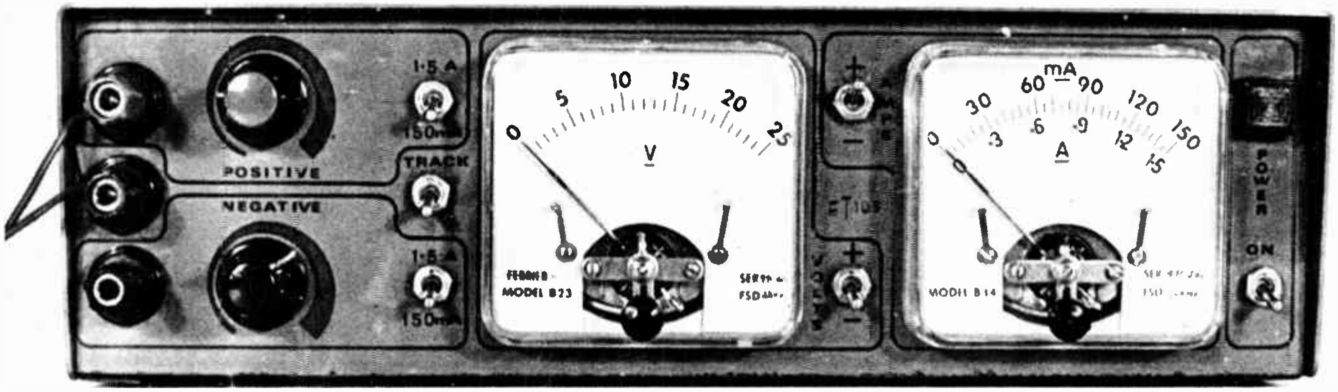


Fig. 2. Circuit modifications for externally programming positive rail (see text).

POWER SUPPLY



UNTIL recently, integrated circuits were priced beyond the reach of the average enthusiast, and very few constructional projects specified their use.

But as with so many electronic components, mass production and wide-spread commercial acceptance has resulted in quite dramatic price reductions, and integrated circuits costing twenty or thirty dollars a couple of years ago, are now readily available for well under two dollars. Many cost less than a dollar.

From the enthusiast's point of view this is a most exciting development for it opens up the possibility of constructing far more ambitious projects than were previously feasible using discrete components. Many such projects are featured in Electronics Today International.

But there is one minor drawback to integrated circuits and this is that many of them require both positive and negative power supplies. These supplies must also have a better level of line and load regulation than was previously necessary.

The power supply described in this project has been designed specifically for this purpose. It is intended for both the serious enthusiast and the professional development engineer.

As may be seen from the specifications, its performance is equivalent to many commercially built units at many times the price.

The unit has two outputs, one positive, and one negative — each separately adjustable from zero to 20 Volts, or settable in such a way that the negative supply automatically tracks the positive supply.

CURRENT LIMITING

Both the unit, and your experimental circuits, are protected against damage by current limiting networks incorporated within the power supply.

A panel mounted switch is used to select the maximum desired current at either 190 mA or 1.80 Amps. If this level is reached, the output voltage will drop and current will be held at the selected limit.

For the professional user of this unit,

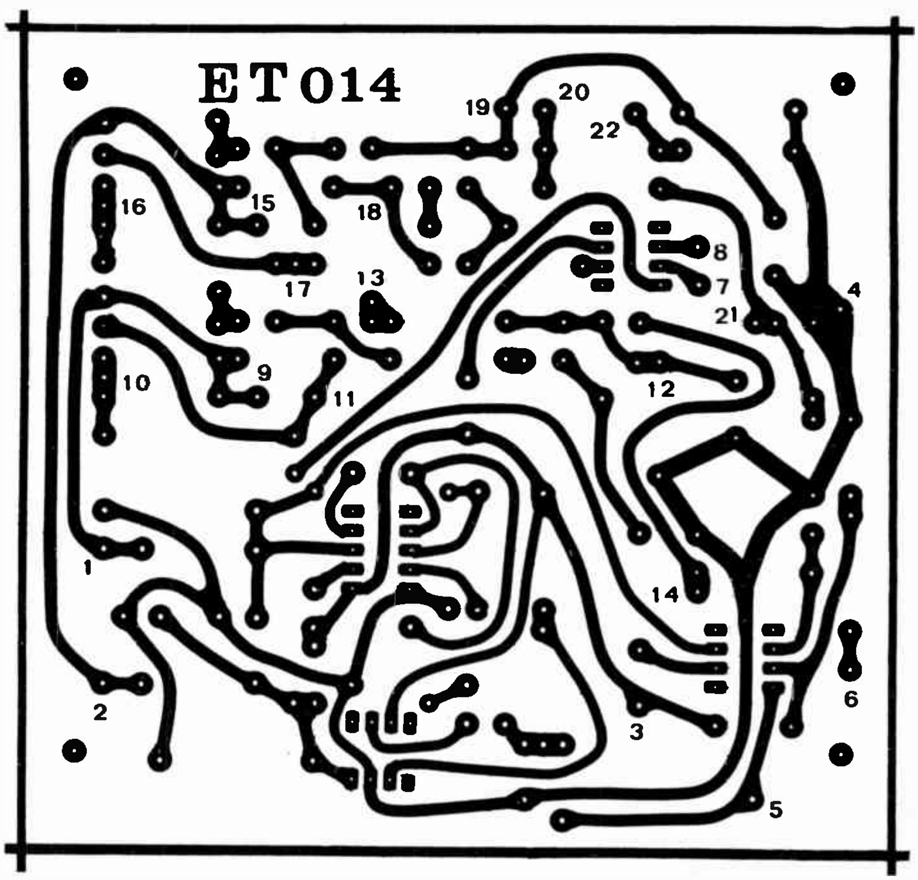


Fig. 3. Foil pattern of printed circuit board (full size).

DUAL POWER SUPPLY

provision has been made for the positive regulator to be externally programmed. The necessary wiring changes are shown in Fig. 2.

Due largely to the use of externally mounted heatsinks, and the use of integrated circuits in the control and voltage reference circuits, the complete power supply unit is quite small and compact. Yet despite this, the internal layout is spacious and all major components are readily accessible.

CONSTRUCTION

Construction is reasonably straightforward if work progresses in the correct manner. The unit may be assembled on matrix board, but we strongly recommend that the correct printed circuit board be used. The foil pattern of the p.c. board is shown in Fig. 3.

Assuming that the printed circuit board is used, commence construction by inserting the pc board pins into the positions numbered on the board. These pins should be inserted with the flange (if flanged) on the component side of the board. All external wiring to and from the printed circuit board will be attached to these pins on the foil pattern side of the board.

When installing the integrated circuits ensure that they are orientated correctly before soldering. (Note that Fig. 4 shows all components, including integrated circuits, as seen from the component side of the board.)

Small heatsinks are fitted over transistors Q4 and Q8. Ensure that these do not contact any other component by mounting them about 1/8" above other nearby components.

When all components have been mounted on the board, recheck for correct orientation and polarity.

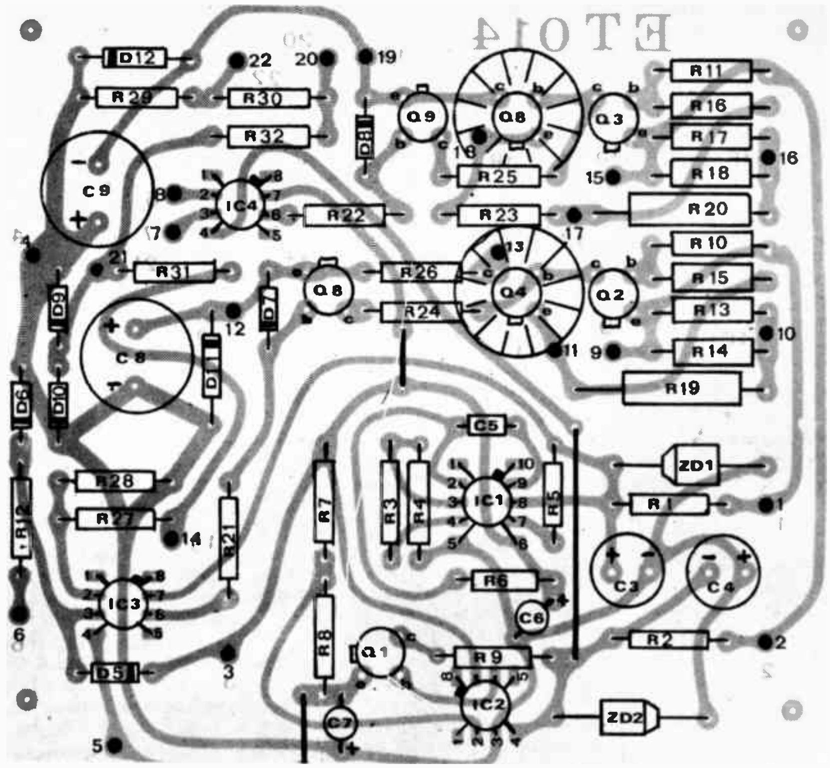


Fig. 4. How the components are mounted on the printed circuit board. Compare this with Fig. 3.

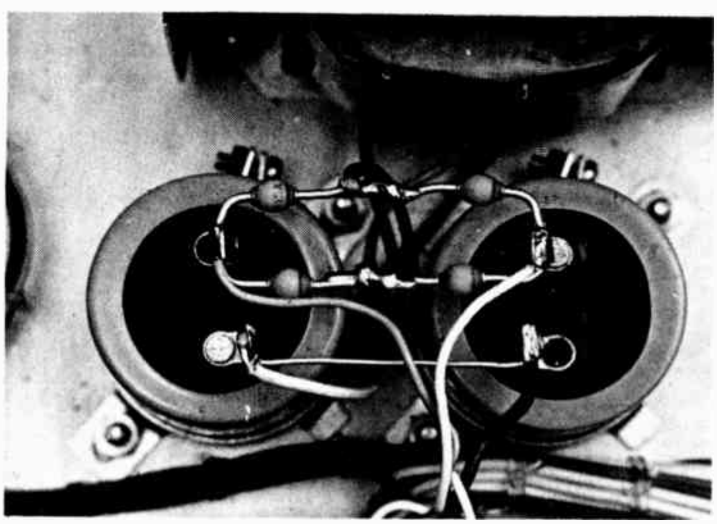
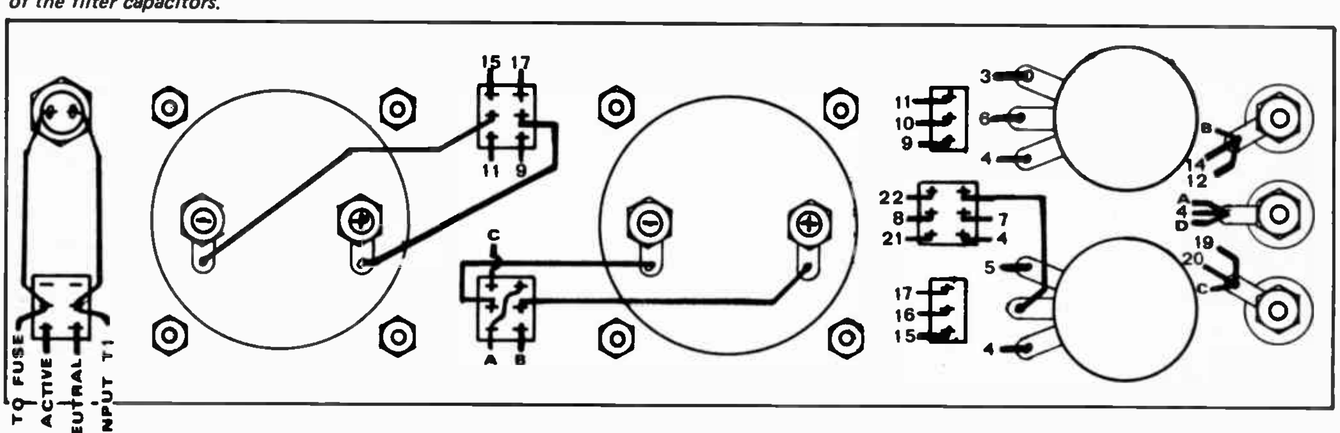


Fig. 5. Diodes D1 – D4 are mounted on top of the filter capacitors.

Fig. 6. This drawing shows front panel wiring details. Wires A, B and C are interconnecting wires on the front panel. Wire D goes to the common of the filter capacitors.



Now mount the transformer and the filter capacitors onto the chassis. Locate diodes D1 – D4 on top of the filter capacitors as shown in Fig. 5.

The heatsinks must now be drilled to take the two 2N3055 output transistors. Carefully remove any burrs from around the holes and then mount the transistors preferably using McMurdo type 2210-01-01 anodised insulating washers. If available, use a smear of silicon grease between transistors and the heatsinks – this will further improve heat transference. Finally, check insulation between the transistor and the heat sink, and then fit the McMurdo type 9151-09-01 transistor covers.

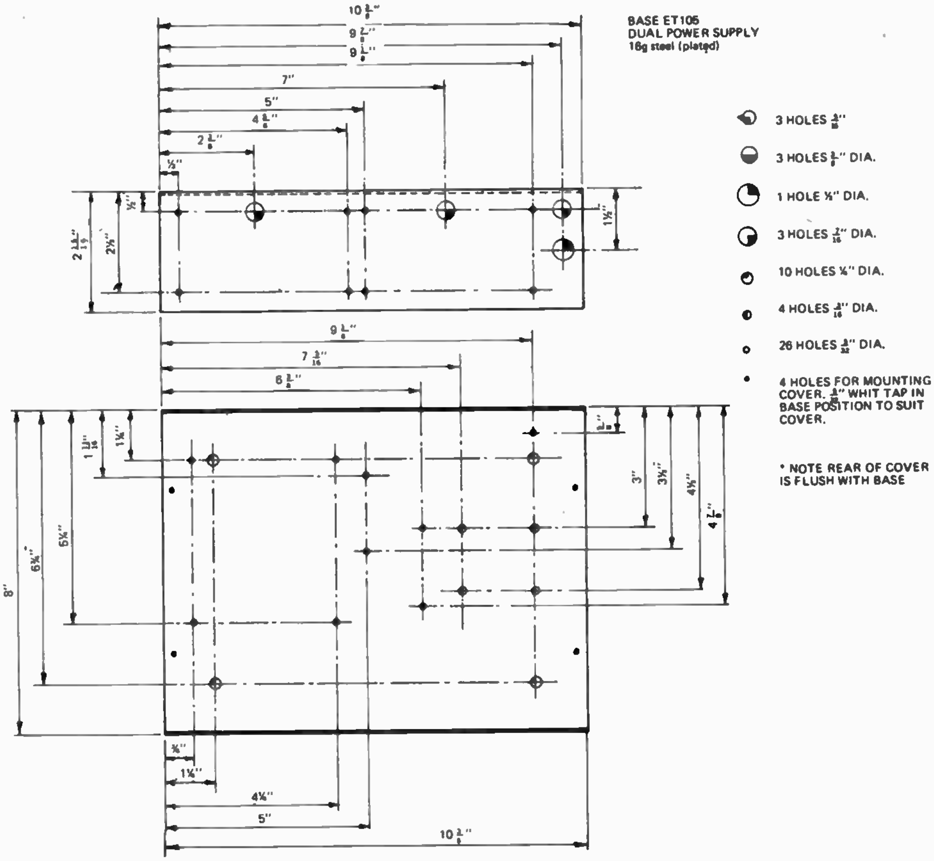
On our prototype unit we constructed our front panel by sandwiching a line drawing between the chassis and a piece of smoked perspex. This provides a very professional looking appearance. An even better finish can be obtained by using an anodised aluminium panel, and these may be available from parts suppliers.
obtained by using an anodised aluminium panel, and these may be available from parts suppliers.

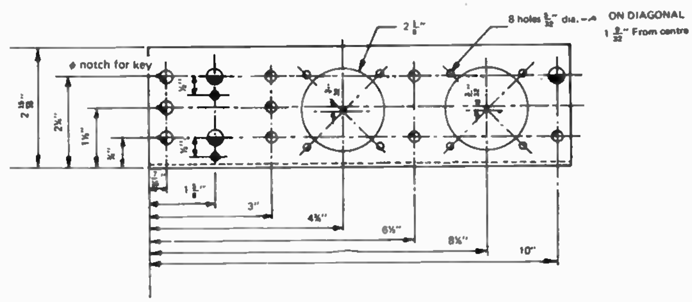
Having determined the method of finishing the front panel, assemble all the relevent components onto the panel.

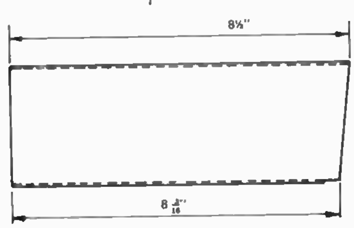
Wires should now be attached to the pins on the underside of the printed circuit board. Insulated 14/0076 wire should be used for this purpose. Two wires should be attached to pins 9, 12, 17 and 19, three wires attached to pins 11 and 17, and four wires attached to pin 4. All wires should be either colour coded or marked so that they may be clearly identified.

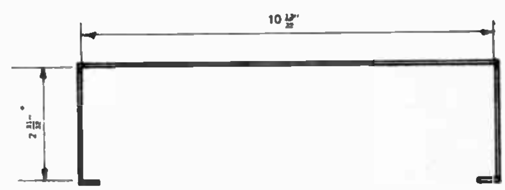
The printed circuit board should now be mounted onto the chassis and the wires loomed to their respective destinations. Note that one each of wires 11, 12, 17 and 19, together with wires 13 and 18 go to the back of the unit and to the heat sinks. Wires 1 and

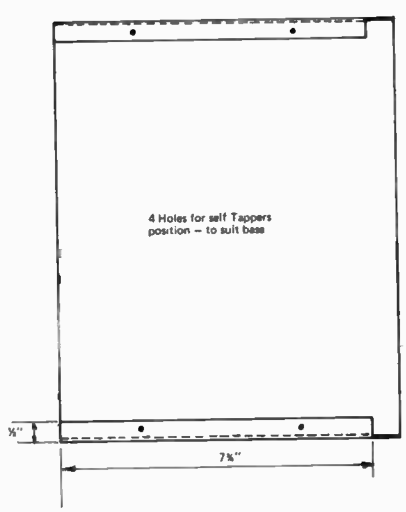
BASE ET105
DUAL POWER SUPPLY
18g steel (plated)











COVER ET105 DUAL POWER SUPPLY
18g steel

SPECIFICATION – POWER SUPPLY – ET 105

Output Voltage	0 – 20 Volts positive 0 – 20 Volts negative
Output Current	0 – 1.5 Amps
Current Limiting	190 mA and 1.80 Amps
Meter Ranges	
(current)	150 mA and 1.5 Amps
(voltage)	25 Volts
Line Regulation	better than 1 mV for 15 Volt input voltage change
Load Regulation	less than 10 mV drop from no-load to full-load
Ripple	less than 2 mV peak to peak
Output Impedance	7 mΩ @ dc – 1.5 kHz 14 mΩ @ – 3 kHz 56 mΩ @ – 15 kHz 200 mΩ @ – 100 kHz

NOTE: The power supply is short circuit proof but shorts in excess of 30 seconds should be avoided due to excessive power dissipation in the transistors.

DUAL POWER SUPPLY

PARTS LIST ET105

R1	— resistor 330 ohm
R2	— " "
R3	— " 1.8k, 2%
R4	— " 5.6k, 2%
R5	— " 2.2k
R6	— " 1.5k
R7	— " 10k, 2%
R8	— " 10k, 2%
R9	— " 1k
R10	— " 330 ohms
R11	— " 330 ohms
R12	— " 10k
R13	— " 10 ohms
R14	— " 10 ohms
R15	— " 100 ohms
R16	— " 100 "
R17	— " 10 "
R18	— " 10 "
R19	— " 0.47 ohms, 2 Watt, ASW2
R20	— " 0.47 " " "
R21	— " 1k
R22	— " 1k
R23	— " 1k
R24	— " 1k
R25	— " 1k
R26	— " 1k
R27	— " 3k, 2%
R28	— " 1k, 2%
R29	— " 1k, 2%
R30	— " 3k, 2%
R31	— " 10k, 2%
R32	— " 10k, 2%
	(all resistors are ½ Watt 5% unless otherwise stated. The 2% resistors are Pye type TR5 or equivalent)
C1	— capacitor, 2200 μF, 50 Volt, Elna type RG
C2	— " " " " " " " "
C3	— " 25 μF, 50 Volt, Elna type RB
C4	— " " " " " " " "
C5	— " 100 pF.
C6	— " 22 μF, 16 Volt, tag tantalum type
C7	— " " " " " " "
C8	— " 220 μF, 50 Volt, Elna type RB
C9	— " " " " " " " "
D1 — D4	— diodes type A15A or equivalent.
D5 — D10	— " " " 1N914 " "
D11 — D12	— " " " EM401 " "
ZD1	— zener diode type BZX70 C27
ZD2	— " " " " "
Q1	— transistor type TT800
Q2	— " " "
Q3	— " " TT801
Q4	— " " TT800
Q5	— " " 2N3055
Q6	— " " TT801
Q7	— " " 2N3055
Q8	— " " TT801
Q9	— " " TT800
IC1	— integrated circuit type μA 723C
IC2	— " " " μA 741C
IC3	— " " " μA 741C
IC4	— " " " μA 741C
	(all the above ICs are metal can type).
SW1	— miniature switch, double-pole changeover, 240 Volt, Plessey C & K type 7201 or similar.
SW2	— " " " " " " " "
SW3	— " " single pole changeover, C & K 7101
SW4	— " " " " " " "
SW5	— " " double pole changeover, C & K 7201
SW6	— " " " " " " "
T1	— transformer, A & R type PT 7554, 50 Volt, centre tapped, 1.5 Amp.
RV1	— potentiometer, linear, 10k, Plessey type E or equivalent.
RV2	— " " " " " " "
Sundries	—

TO5 Heatsinks, 2 off, McMurdo TXBF 032 025 CB Power transistor heatsinks, 2 off, Mullard 35 DB 3C drilled to suit. Two transistor covers, McMurdo 9151 09 01. Two anodised insulating washers, McMurdo type 2210 01 01. One set of metalwork. One front panel. 240 Volt neon panel light. Three terminals. Two potentiometer knobs. One fuse holder for size 00 fuse. One 1 Amp size 00 fuse. One 3 core flex and plug. One cable clamp. One printed circuit board ET 014. Twenty two pc pins McMurdo type 5737 54 08. Three grommets. Four rubber feet. Four ¾" spacers. 14/0076 connecting wire (insulated) various screws, washers, nuts etc.

Voltmeter	— 25 Volts fsd, 2½" square, Ferrier type B 23 or equivalent.
Ammeter	— 750 mV fsd, 1mA, scaled 1.5 Amps and 150 mA. Ferrier type B34 or equivalent. (when ordering, specify that meters should be scaled for steel panels).

2 go to the filter capacitors and a wire D comes from the common of the filter capacitors up to the loom and to the common terminal on the front panel.

The front panel can now be wired as shown in Fig. 6.

The wires to the heat sink mounted transistors are taken through the grommetts provided, and the already assembled heat sinks mounted into position.

Complete all remaining wiring taking care that all leads carrying 240 Volts are adequately insulated. The mains lead must enter the case through an insulating grommett and the lead must be securely anchored to the case. It is

HOW IT WORKS

The mains input voltage is reduced and isolated by transformer T1. The 25-0-25 Volt output from the transformer is then rectified and filtered by diodes D1-D4, and capacitors C1 and C2 to provide an unregulated ± 40 Volt dc supply.

Series regulators are used in the main control system. The two regulators – one for each supply – are almost identical in operation, therefore only the positive regulator will be described in detail.

The series pass transistor Q5, is mounted on an external heat sink. Transistors Q4 and Q6 provide current amplification for Q5 giving the combination a total current gain exceeding 50,000. The voltage gain is approximately unity.

The main reference supply is generated by IC1 which is a precision voltage regulator. The reference level required is obtained by potentiometer RV1 which is connected across the 5.6 Volt regulated output from IC1.

Power for the IC voltage reference is supplied by R1, ZD1 and C3. This maintains a constant voltage across the IC, eliminating variations due to changes in mains voltage. The 27 Volt supply from this circuit is also used to supply power to IC3.

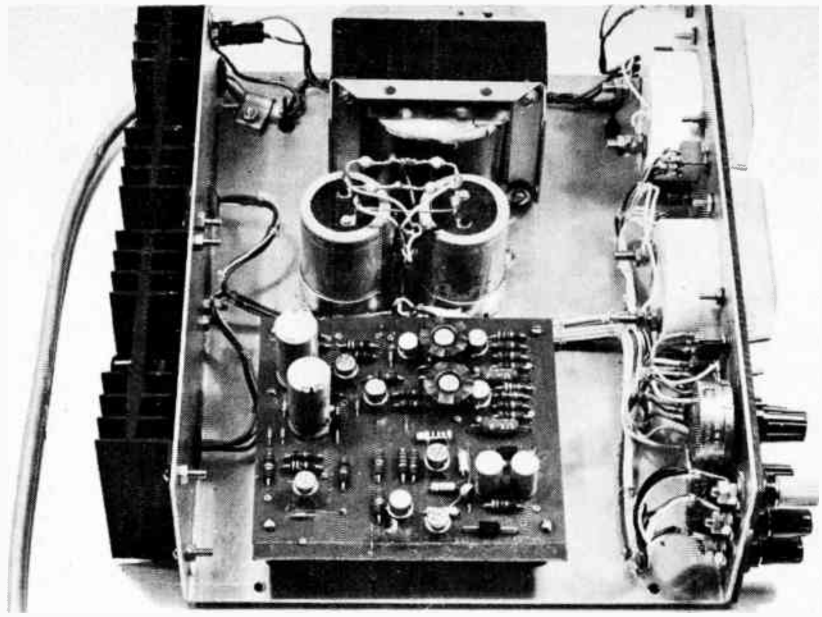
The reference for the negative supply is obtained from operational amplifier IC2 which is connected so as to track the positive reference supply. The 5.6 volt output from this circuit is just as accurate as the output from the main regulator. Power for this operational amplifier is supplied from a 27 volt zener which is also used to supply IC4.

The power supplies for IC1 are + 27 Volt and 0 Volts; for IC2 and IC4, the supplies are + 5.6 Volts and - 27 Volts; for IC3, +27 Volts and - 5.6 Volts.

Resistors R27 and R28 divide the output voltage by four. This voltage not sufficient merely to tie a knot in the mains cord — this is a dangerous practice.

The supply should now be ready for use, but before connecting to the mains, recheck all point-to-point wiring and all soldered connections.

One point that may not be commonly realised is that meters are calibrated specifically for one panel material. A meter calibrated for mounting on a steel panel may be as much as 30% out if it is mounted on an aluminium panel — and vice-versa. We recommend that a steel chassis is used for this project — but if you decide to use aluminium notify the meter supplier accordingly. ●



is compared against the voltage set by RV1 by operational amplifier IC3. The output of IC3 controls the series regulator configuration, and hence the output voltage. The action of IC3 is to keep the two voltages at its input at the same level. Thus, the output voltage will be four times the input voltage, and virtually independent of load current.

When load current approaches the level set by the limit switch, transistor Q2 becomes forward biased sufficiently to cause it to conduct. This bypasses current from the base of Q4 and causes IC3 to lose control of the output. If the load continues to increase, the output voltage will fall and the current will remain effectively constant.

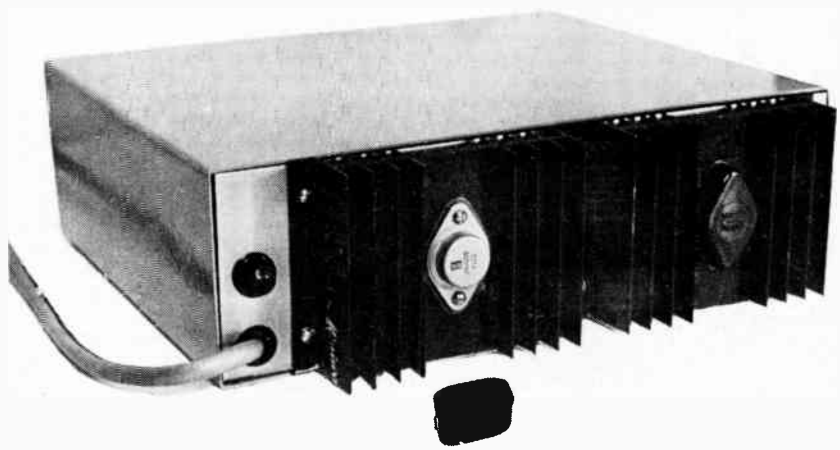
The negative regulated supply works in the same manner when the power supply is used in the normal mode.

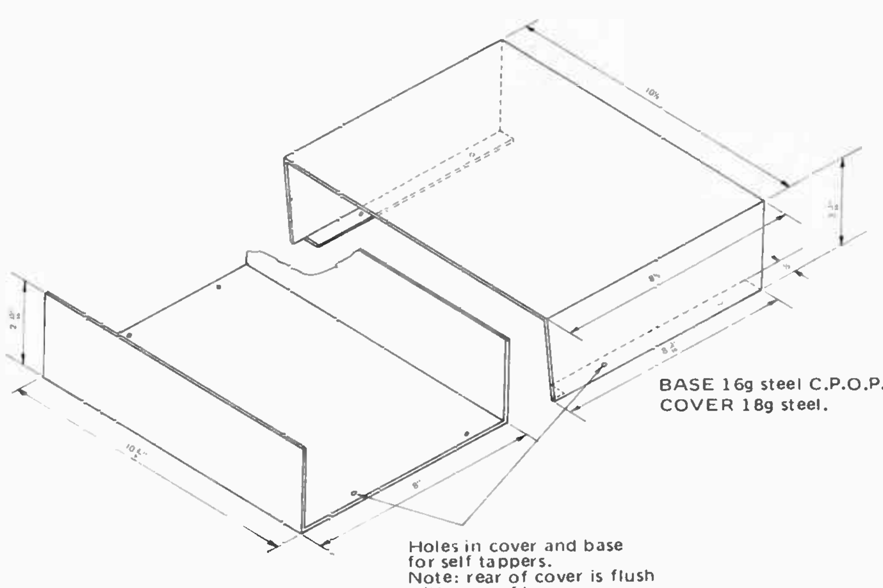
A tracking mode of operation is also included, and in this mode the negative supply tracks the positive supply. If, for example, the positive regulator is set at + 14.5 Volts, then the negative supply will automatically be set at - 14.5 Volts.

In the tracking mode of operation, the reference voltage is zero volts and the voltage used as the output voltage reference is the centre voltage of the two supplies. And since the IC tries to maintain both inputs at the same level (in this case zero volts) then the two output voltages must be of the same value.

Diodes D7 through D12 are used to protect the integrated circuits and output transistors against various forms of misuse, including shorting the positive and negative outputs together.

Provision has been made in the design for externally programming the positive regulator. If this facility is required, alter the wiring as shown in Fig. (Resistor R12, and diodes D5 and D6 protect the IC when this mode of operation is employed).

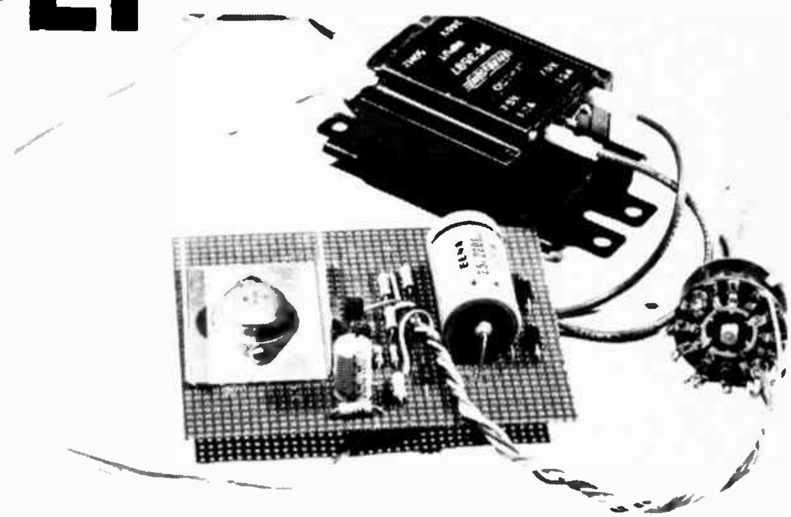




PROJECT 221

BASIC POWER SUPPLY

Simple regulated supply provides 4.5-12 volts at 400 mA maximum.



THIS LITTLE power supply provides a range of switch selectable output regulated voltages from 4.5 to 12 volts. The supply will provide up to 400 mA and the output can withstand a short circuit without damage. It is therefore ideal for the experimenter or for use with high drain appliances.

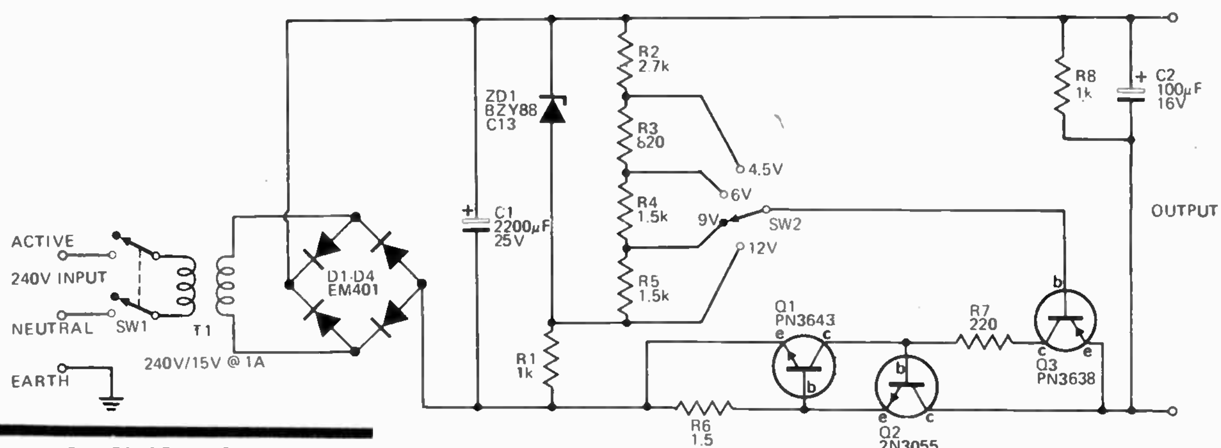


Fig. 1. Circuit diagram of the regulated power supply.

SPECIFICATION

Nominal output voltage 12 V, 9 V, 6 V, and 4.5 V
Output current 0 – 300 mA
Current limit approx. 500 mA
* except when modified for use with organ.

PARTS LIST
POWER SUPPLY ETI 221

R6	Resistor	1.5 ohms ½ W 5% (2 x 1.5 ohms in parallel for organ)
R7	"	220 ohms ½ W 5%
R3	"	820 " " "
R1, 8	"	1k " "
R4, 5	"	1.5k
R2	"	2.7k
Q1	Transistor	PN3643 or similar
Q2	"	2N3055 "
Q3	"	PN3638 "
D1-D4	Diode	EM401 or similar
ZD1	Zenerdiode	BZY88C13 "
T1	Transformer	240V/15V @ 1A
SW1	DPST	240V switch
SW2	4 position single pole switch	
	heatsink for Q2	
C1	Capacitor	2200μF 25V electrolytic
C2	"	100μF 16V "

Piece of matrix board.

HOW IT WORKS

The 240 V mains voltage is reduced to 15 volts by transformer T1, and this secondary voltage is then full-wave rectified by rectifier bridge D1-D4.

The output of the bridge rectifier is filtered by C1 to provide approximately 20 volts dc.

The series combination, of Zener diode ZD1 fed by resistor R1, provides a stabilized voltage of around 13 volts which is applied across the voltage divider R2, R3, R4 and R5. Thus a series of reference voltages are generated for the regulator, where the positive rail is fixed and the negative rail is the one that is varied.

Transistor Q3 is an emitter follower where the output (emitter) is about 0.6 V higher (more positive) than the base. The base voltage is selected by SW2 from one of the tappings on the reference-voltage divider. Since Q3 cannot handle the required output current, it drives Q2, a power transistor, which *can* handle the required load.

When the load exceeds 400 mA (approximately), the voltage drop across R6 forward biases Q1 which turns on and shunts current away from the base of Q2. Thus the regulator loses control and the output voltage falls, limiting the current to 400 mA. As the power dissipated in Q2 under short-circuit conditions is around 10 watts, Q2 must be fitted to a heatsink. Additionally, resistor R7 limits the current supplied by Q3 to a safe value (for Q3) under short circuit conditions.

If a fully variable supply is required, a 10 k potentiometer should be used in place of the voltage divider. The wiper of the potentiometer is then fed directly to the base of Q3.

EXPERIMENTER'S POWER SUPPLY

This power supply is suitable for the experimenter. It has fully adjustable output voltage and current limiting. A single meter can be switched either to voltage or current while an LED will indicate an overload.

THIS ECONOMICAL POWER SUPPLY replaces the ETI 111 supply published some years ago. The 111 gave an output voltage variable between 1.5 and 15 V, this project gives the full range 0 to 15 V. In addition this supply features metering (or you can use the calibrated scale on the second version if you don't have a spare meter) to enable accurate setting of voltage or current.

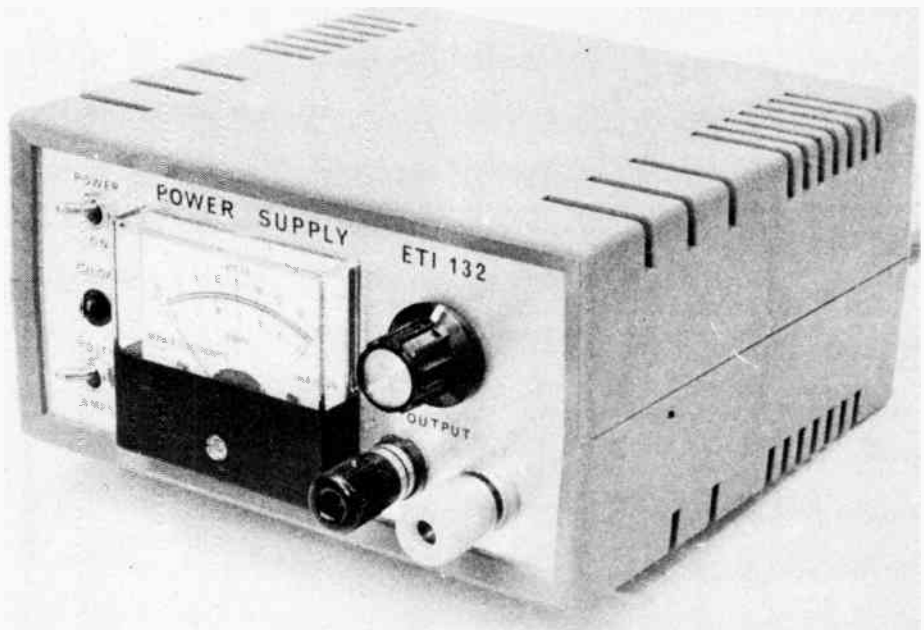
The 132 is attractively housed in a plastic case and Scotchcal front panels are available to give the unit a professional look.

Construction

Commence by assembling the pc board with the aid of the component overlay diagram. The main filter capacitor C1 is normally a chassis-mounting type, but we mounted this satisfactorily by passing the lugs through the large holes in the pc board, bending them flush with the copper and soldering. Check the polarity of the capacitor before fitting, as it cannot be seen later. The transistor Q3 is fitted, along with its heatsink, with the two mounting screws. No insulation is used between the transistor and the heatsink but pass a small piece of tubing over the base and emitter leads where they go through the heatsink, to prevent shorting. If the meter is not required RV3, RV4 and R10 are not used.

The front and rear panels can now be drilled. Note that the mounting bracket of the transformer has to be cut back about 12 mm on one end to allow it to fit easily. If a scotchcal panel is used it

(Continued on page 93)



SPECIFICATION ETI 132

Output Voltage	0-15 V variable
Output Current	0-1 A
Current Limit	approx 1.2 A
Load regulation	35 mV 0 to 1 A load
Line regulation	20 mV 220 to 260 V input
LED indication of current overload	

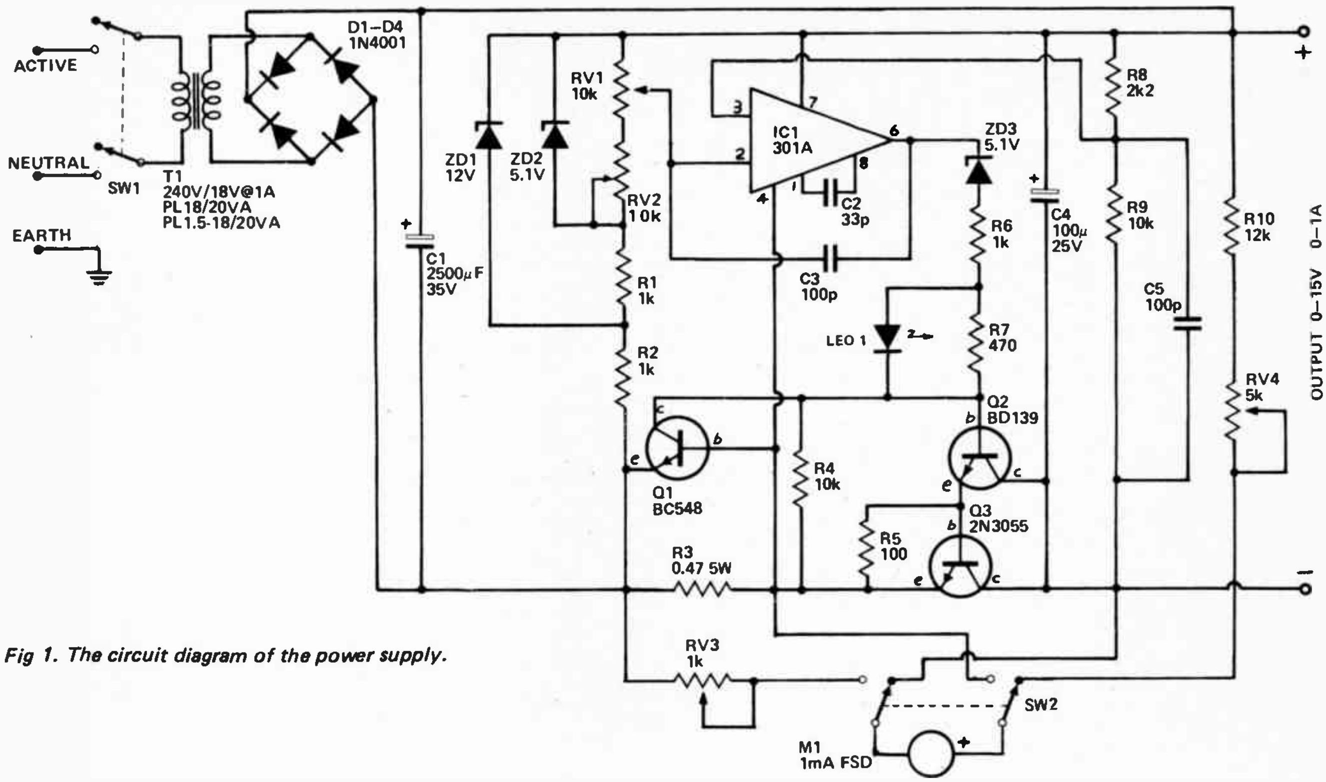


Fig 1. The circuit diagram of the power supply.

How It Works

The 240 V mains is reduced to 18 V in T1. This 18 V ac is then rectified by D1-D4 and filtered by C1 to give about 25 volts dc (on no load). The voltage reference for the supply is ZD2, which gives about 5 V dc. However, due to the large variation in voltage across C1 (caused by load changes) additional regulation is used, incorporating ZD1, and the two circuits give the stability required.

The regulator is a 'series-pass' type with the positive rail common and the negative rail variable. We have done it this way to achieve outputs down to 0 V. The comparator IC (LM301) cannot work with its input less than about 2 volts above the negative rail, but it can work with the inputs at the positive supply rail. However this will not work with all types of op amp – so do not substitute the 301 with a 741 or similar.

The output of ICl controls the output transistors, Q2 and Q3. A level-shifting zener ZD3 is used in the output of IC1 as its output cannot swing low enough. The output voltage is divided by R8 and R9 and is taken to IC1 which compares it to that set on RV1. IC1 then adjusts the drive to the output stage until the two voltages are the same. RV2 is used to compensate for variations in the voltage of ZD2.

In the event of an overload the voltage drop across R3 will forward-bias Q1, which will bypass current away from the output transistors. This causes the output voltage to fall, the comparator sees this error, and the output of IC1 goes to the positive supply rail (trying to compensate). Q1 however will continue to bypass any extra current, holding the output current constant at about 1.2 A. However, the additional current out of IC1 will forward bias LED 1 and it will indicate the overload.

With such high gain in the circuit additional frequency stability is needed and C3 and C5 provide this. For metering, we simply use a 1 mA movement meter and measure the voltage across the output (via R10 and RV4) and across R3 (current).

Setting Up

1. Without Meter – With this version we rely on the potentiometer to be linear. In practice it is not linear at the two ends of its travel. Calibration is done by adjusting the knob position and RV2.

Set the output to one volt and position the knob to read one volt. Now turn the knob to 15 V and adjust RV2 to give 15 V output. Recheck the 1 V setting and repeat the procedure, if necessary.

2. With Meter – Connect the output to an accurate voltmeter and turn the pot to maximum. Adjust RV2 to give 16 V. Adjust RV4 until the meter reads 16 V (with RV2 switched to volts). Now connect a load and an ammeter. Set 1A on the ammeter and then adjust RV3 until the power supply meter reads 1 A.

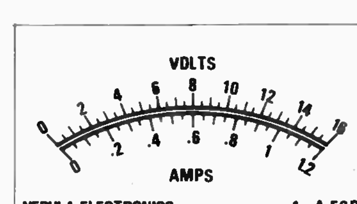


Fig 2. The meter scale used.

PARTS LIST ETI 132

Resistors			
R1,2		1 k	½ W 5%
R3		0.47 Ω	5 W 5%
R4		10 k	½ W 5%
R5		100	" "
R6		1 k	" "
R7		470	" "
R8		2k2	" "
R9		10 k	" "
R10		12 k	" "
RV1	Potentiometer	10 k lin rotary	
RV2	"	10k Trim	
RV3	"	1 k "	
RV4	"	5k "	
Capacitors			
C1		2500 μ electro type RG	
C2		33 p ceramic	
C3		100 p "	
C4		100 μ 25 V electro	
C5		100 p ceramic	
Semiconductors			
D1-D4	Diodes 1N4001		
ZD1	Zener 5.1 V 400 mW		
ZD2	Zener 12 V 400 mW		
ZD3	Zener 5.1 V 400 mW		
LED 1	LED with mounting clip		
Q1	Transistor	BC548	
Q2	"	BD139	
Q3	"	2N3055	
IC1	Integrated circuit LM301		

Miscellaneous
PCB ETI 132
Transformer 240 V – 18 V 2A
Pl 18/20 VA or PL 1.5-18/20 VA
Case PC1
Power cord and clamp
Heat sink DSE H-3400
Two 2 pole 2 position 240 V Toggle switches
Two terminals
Meter 1 mA FSD scaled 0-16 V, 0-1.2 A
Knob

*If meter is not required delete RV3, RV4, R10, the meter and one switch

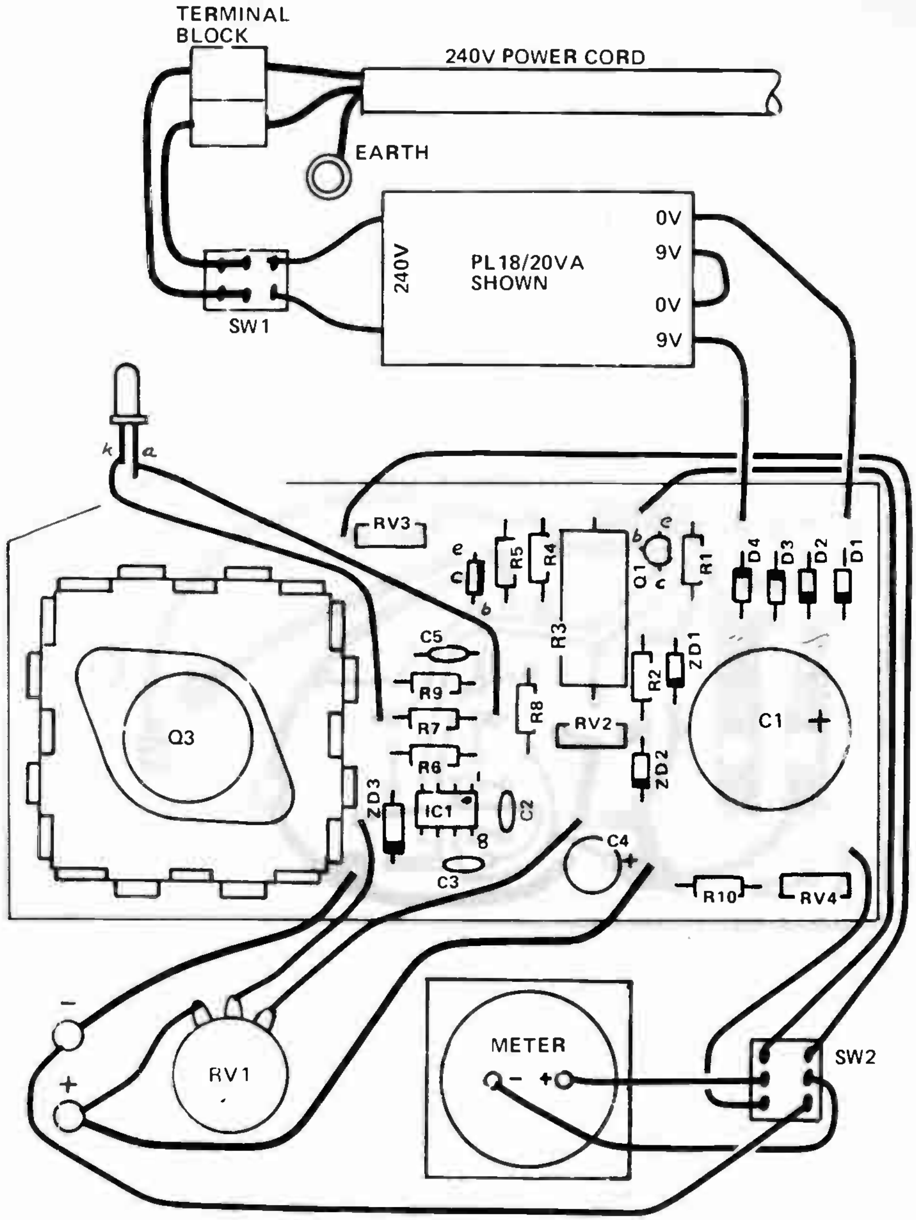


Fig 3. The component overlay and interconection diagram.

The photo on the right shows the second version of the power supply — where the voltage is set using a calibrated pot rather than a meter.

can be fitted before drilling and used as a template. Take care, however, not to scratch the panel.

Assemble the front and rear panels and wire the unit accordingly to Fig 3.

The wires to and from the power switch can pass the pc board via the chamfer on the lower left hand side. Other wires from the pc board to the front panel can be connected onto the copper side of the board.

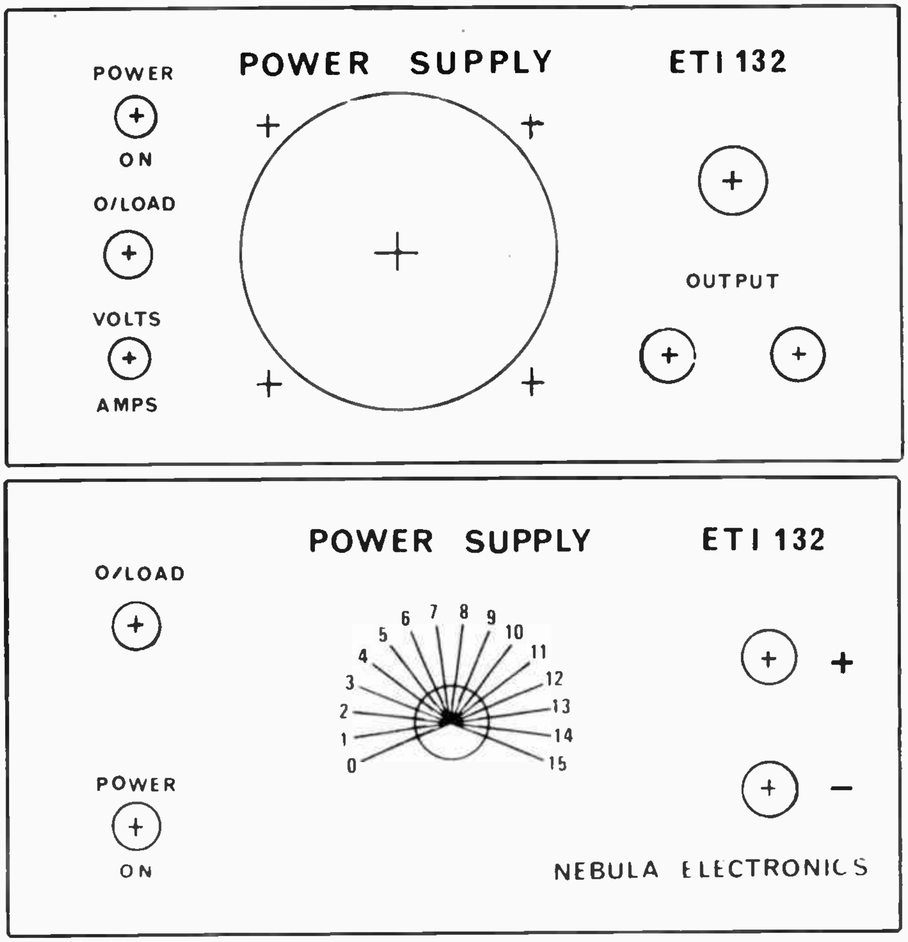


*Fig 4. Front panel layouts
Full size 131 x 66 mm.*

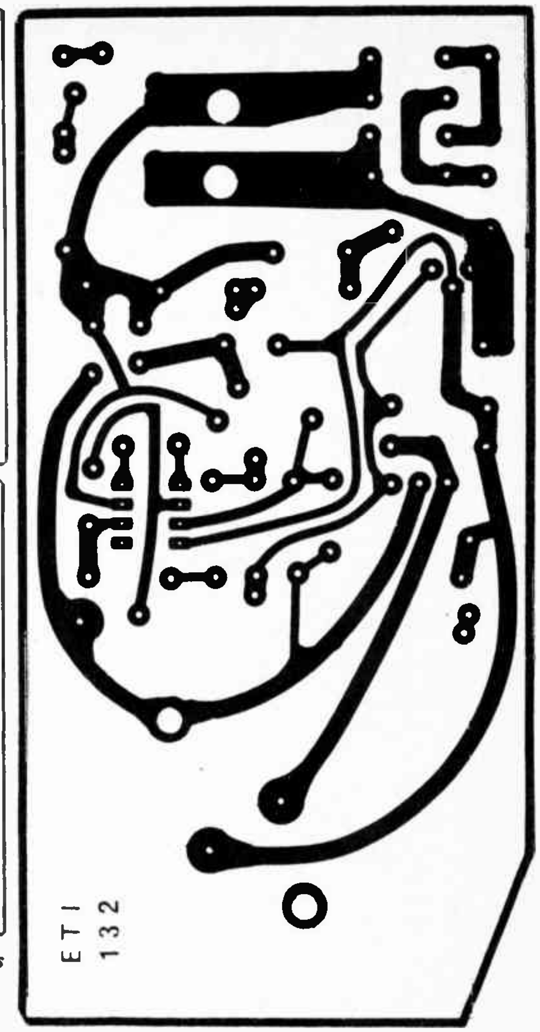
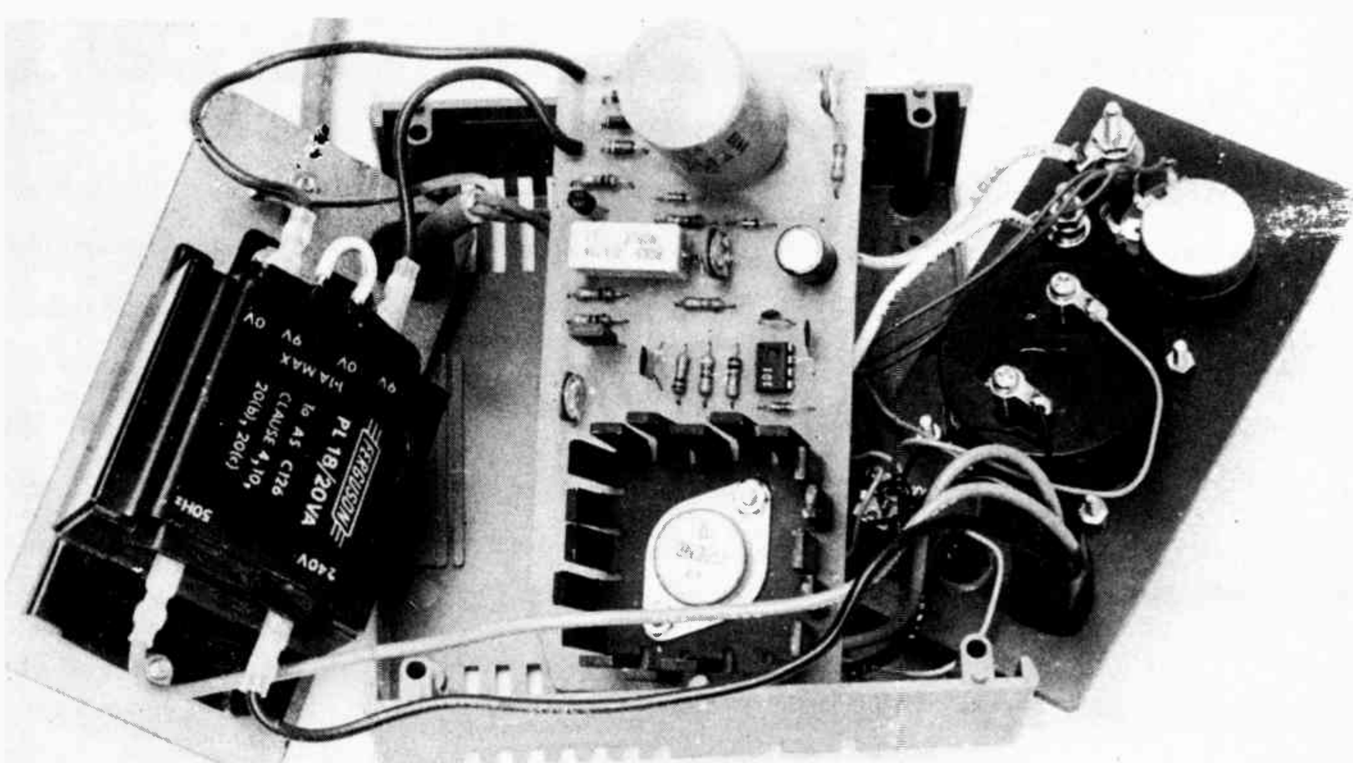


*Fig 5. Printed circuit layout.
Full size 132 x 66 mm.*

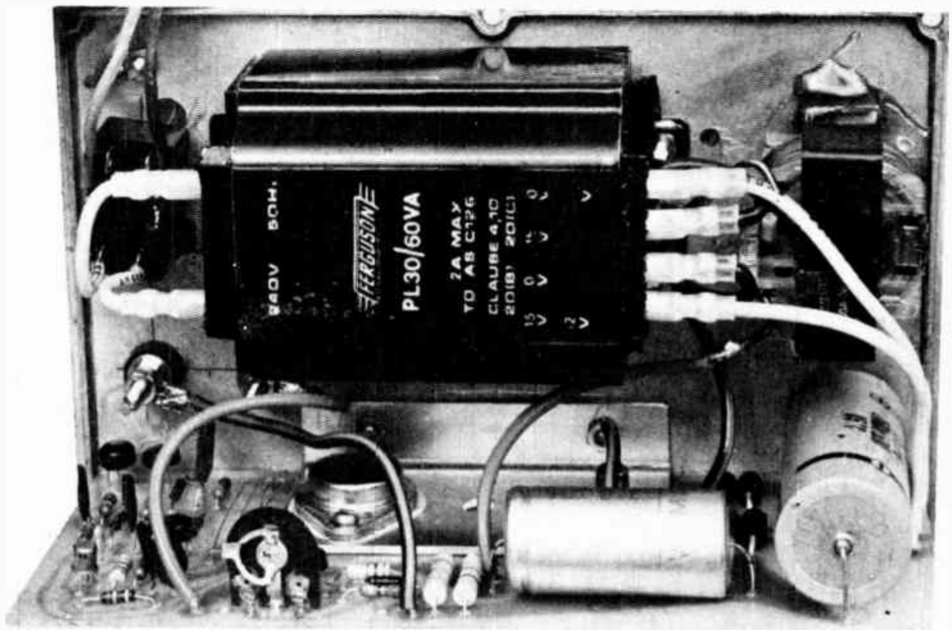
SCOTCHCAL OFFER

**Scotchcal panels ready to stick on are available from Electronics Today at $3.00 each. Send order together with a stamped addressed envelope – size at least 150 x 120 mm.
Address to Scotchcal Offer 132, Electronics Today, 15 Boundary Street, Rushcutters Bay, NSW, 2011.**



SWITCHING REGULATOR SUPPLY

ETI PROJECT 119



Drive those TTL circuits with this 5 volt 10 amp (max) supply.

WHILST the introduction of CMOS has lowered the power requirements of digital equipment using it, many large scale systems, because of cost and availability, are still designed around TTL logic. For such systems a five-volt supply having a capability of up to 10 amps is often required.

The choice of power supply for a system depends very much on the output requirements. In very low power applications a shunt regulator consisting of a series resistor and a zener may be entirely adequate. For medium power systems however a series-pass transistor regulator is normally used.

Whilst the series pass regulator is very good with regards to ripple and regulation the specification of the transformer is critical if the supply efficiency is to be above 50%. In a larger system this can be a very important factor.

With a switching regulator the requirements on the transformer are greatly relaxed and an efficiency of 70% or more can readily be obtained with mains-input variations of from 160 to 260 volts.

A fourth type is the switch-mode supply where the mains voltage is first rectified and filtered. The rectified mains then drives a high-frequency inverter which employs a ferrite transformer. Regulation is obtained by controlling the inverter and by this means very high efficiencies may be obtained. Nearly all the components in such a system work at mains voltage and hence for safety reasons this approach was not used in our project.

CONSTRUCTION

All components, with the exception of the transformer and the choke are best mounted on a printed-circuit board such as the one specified. The choke should be wound as detailed in Table 2 with four layers close wound of 14 gauge B&S wire. Due to the dc current in the choke an air gap is necessary to avoid saturation. The easiest method of adjusting this gap for best performance is to run the supply at the maximum current required and adjust the gap by inserting that thickness of insulation between the cores which gives minimum ripple voltage. We found that a 3 mm gap was required at 10 amps for a ripple of 50 mV peak-to-peak.

The prototype was mounted in a ▶

TABLE 1
Comparison of typical series and switching regulators

	SERIES	SWITCHING
Output Voltage	5 V	5 V
Output Current	10 A	10 A
Efficiency		
240 V in	50%	70%
260 V in	40%	70%
Ripple Voltage	< 5 mV p-p	50 mV p-p
Regulation 0-10 A	< 0.05 V	0.3 V
Input Voltage	240 ± 10%	160 to 260 V
Transformer Secondary	8.5 V @ 12 A	20 to 30 V @ 80 VA
Diodes Required	10 A	3 A
Filter Capacitor	33 000 μF	2 200μF
Short Circuit Current	15 A	15 A

SWITCHING REGULATOR SUPPLY

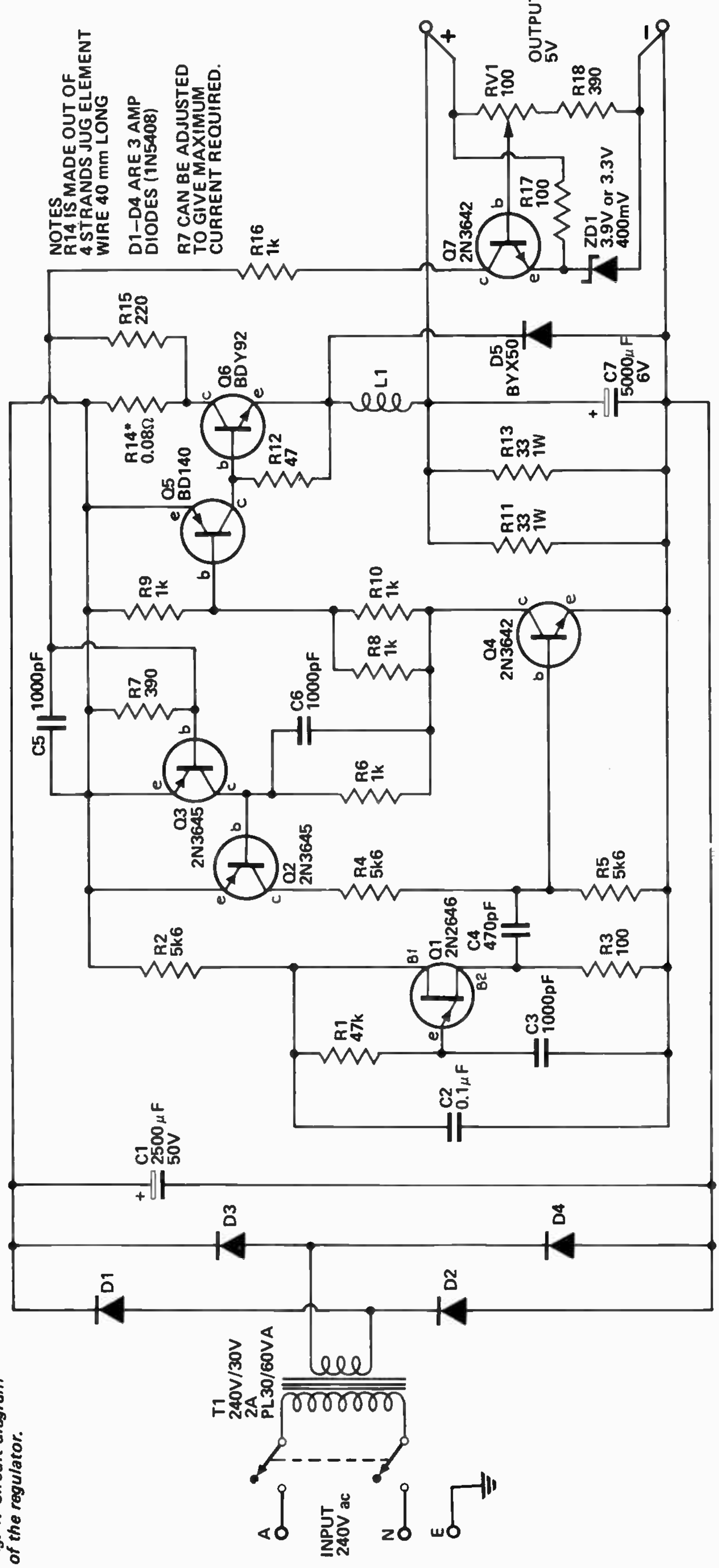


Fig. 1. Circuit diagram of the regulator.

die-cast box which acted as the heatsink as well as a shield to prevent the radiation of RFI generated by the switching action of the supply. If another form of box is used a heatsink must be added to the transistor-diode bracket for cooling.

An external LC filter will reduce the ripple even further if required. For example a series choke of 20 turns of 1.6 mm wire on a 10 mm ferrite rod and a parallel combination of 1000 µF electrolytic and 0.47 polyester capacitors external to the box will provide considerable extra ripple attenuation.

PARTS LIST – ETI 119

R14	Resistor	0.08 Ω *		
R11,13	"	33 Ω	1W	5%
R12	"	47 Ω	½W	5%
R3,17	"	100 Ω	½W	5%
R15	"	220 Ω	½W	5%
R7,18	"	390 Ω	½W	5%
R6,8,9	"	1k	½W	5%
R10,16	"	1 k	½W	5%
R2,4,5	"	5k6	½W	5%
R1	"	47 k	½W	5%

RV1 Trim Potentiometer 100 Ω

C4 Capacitor 470 pF ceramic
C3,5,6 " 0.001 µF polyester
C2 " 0.1 µF polyester
C1 " 2500 µF 50 V electro
C7 " 5000 µF 6V electro

D1-D4 Diode 3 Amp 100V IN5408 or similar
D5 " BYX50-200

ZD1 Zener Diode 3.3 V or 3.9 V 400 mV

Q1 Transistor 2N2646 or similar
Q2,3 " 2N3645 or similar
Q4,7 " 2N3642 or similar
Q5 " BD140 or similar
Q6 " BDY92 or similar

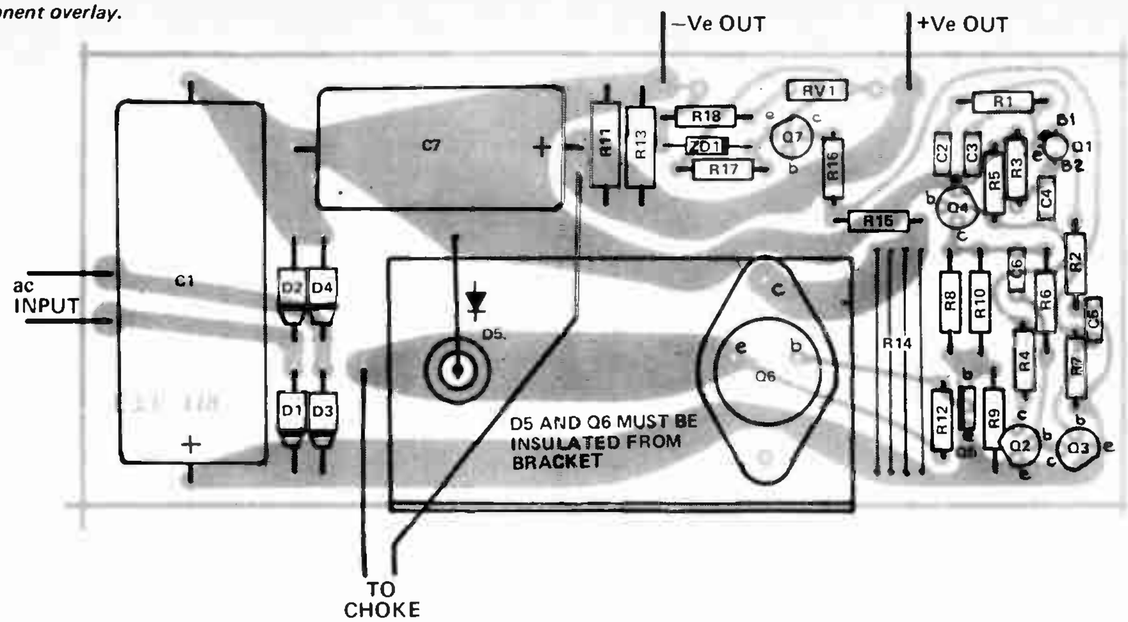
L1 choke see Table 1.

T1 Transformer 20V — 30V @
60VA (7.5 Amp output)
75VA (10 Amp output)

SW1 Toggle switch 2 pole 240 V rated.

Heatsink bracket to Fig.3
Diecast Box 6357p
PC Board ETI 119
Insulation kit for Q6 and D5
* R14 is made out of 4 strands of jug element each 40 mm long.

Fig. 2 Component overlay.



HOW IT WORKS – ETI 119

IN a conventional series regulator power supply the resistance of a series transistor is controlled in order to maintain the correct output voltage. The series transistor dissipates considerable power and therefore at very high load currents series regulators are quite inefficient. In the switching regulator a series transistor is still used but does not operate in its linear range. Instead it switches ON and OFF at high speed such that the load is alternately connected and disconnected to a supply voltage that is higher than that required across the load. By controlling the ratio of ON to OFF time we effectively control the average voltage as seen by the load. For example if it is on for 25% of the time the average output voltage will be 25% of the input. Thus by controlling the ON/OFF ratio the output voltage may be stabilized whilst dissipation in the series transistor is very greatly reduced.

However since most loads do not like their supply to be in the form of a square wave an LC filter is used before the load to pass only the dc component.

Referring to the main circuit diagram we see that transistors Q5 and Q6 are used as the series switch. L1 and C7 form the output filter. Due to the inductance of the choke a flywheel diode is required, not only to protect the transistor, but to provide proper operation. When the switch is on, the load current flows through the transistor, the choke, and into the capacitor and the load (Fig. A). When the switch is opened the load current must continue to flow through the choke and this is done via the flywheel diode D5 (see Fig. B). The current through the choke will thus rise during the on period and fall during the off period. The current never falls to zero except at very low load currents and the average is the same as the load current.

The operating frequency is set by the UJT Q1 which runs about 20 kHz; the higher the operating frequency the lower the ripple voltage on the output. However as the operating frequency goes up so also do switching losses in both transistor Q6 and diode D5. The 20 kHz was chosen as a compromise. It is high enough not to be audible but low enough to keep these losses to a minimum. A fast transistor and diode are still required however. For example if an MJ802 transistor is used the power losses increase by 5 to 10 watts at 10 amps output current.

When the UJT fires the pulse generated is coupled into the base of Q4 by C4 turning Q1 on. This, inturn, turns on Q2 and the switch Q5/6. When Q2 turns on Q4 also turns on and both latch on. If the current through Q6 rises above about 12 to 14 amps Q3 will turn on robbing current from the base of Q2 allowing both it and Q4 to turn off. This also turns off the output switch Q5/6. This is the current protection circuitry.

A voltage proportional to the output is provided by RV1 to Q7 for comparison to the voltage of ZD1. If Q7 is turned on sufficiently it will also turn on Q3 thus unlatching Q2/4 and turning off the output switch. Once the supply has stabilised this action will control the on time of the switch in each cycle of the 20 kHz, such that the output voltage is maintained at a voltage as set by RV1 in a smooth and even manner.

We used a 240 V to 30 V 2 A transformer, which is adequate for supply currents of up to 7.5 amps, however any transformer having an output of 20 to 30 volts and a power rating of 60 VA would do. If up to 10 amps output is required then a transformer with a rating of 75 to 80 VA would be required.

It is also possible to supply the regulator from a dc supply of 10 to 40 volts. If the voltage available is less than 20 volts R2 should be replaced by a link to ensure that the UJT operates correctly.

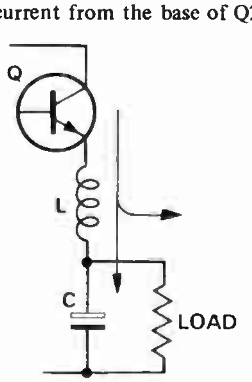


Fig. A. Current paths with switching transistor on.

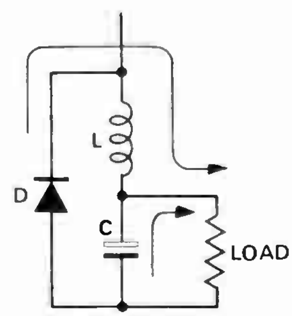


Fig. B. Current paths with switching transistor off.

SWITCHING REGULATOR SUPPLY

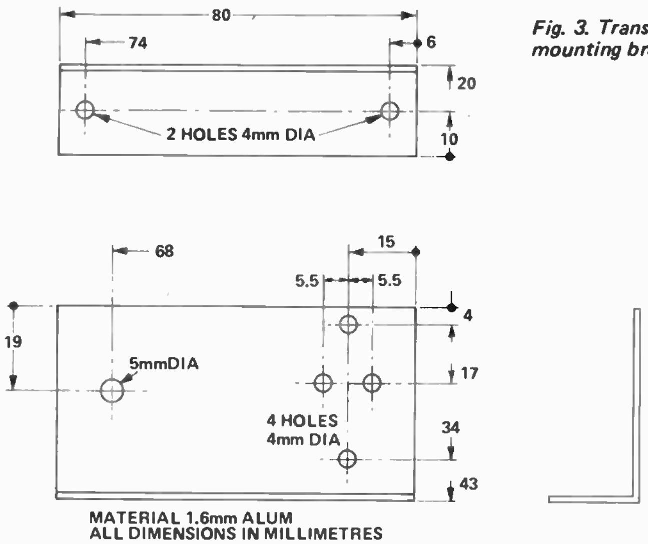


Fig. 3. Transistor/diode mounting bracket.

TABLE 2 Choke winding details.
CORE Philips E core 4322-020-34720 two required
FORMER Philips 4322-021-31830 or 4312-021-23622 one required
Four layers close wound of 1.6 mm wire core gap 3 mm (see text).

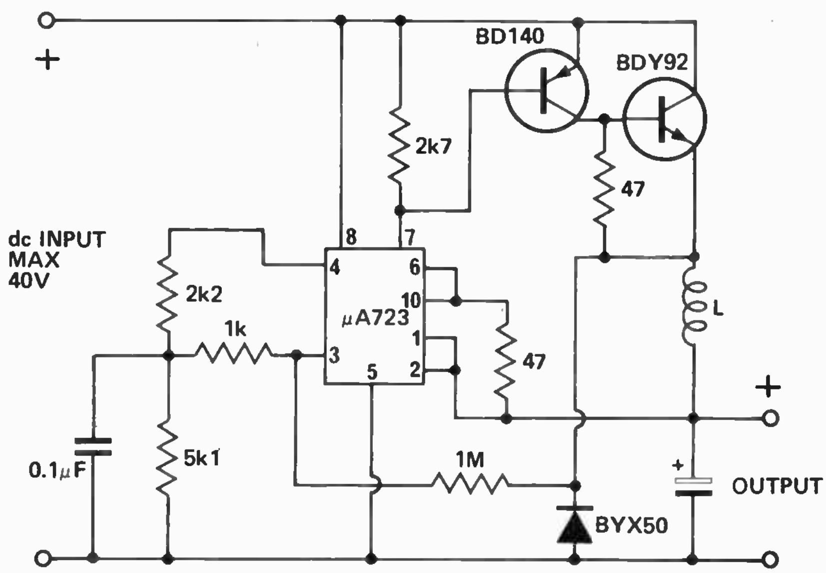
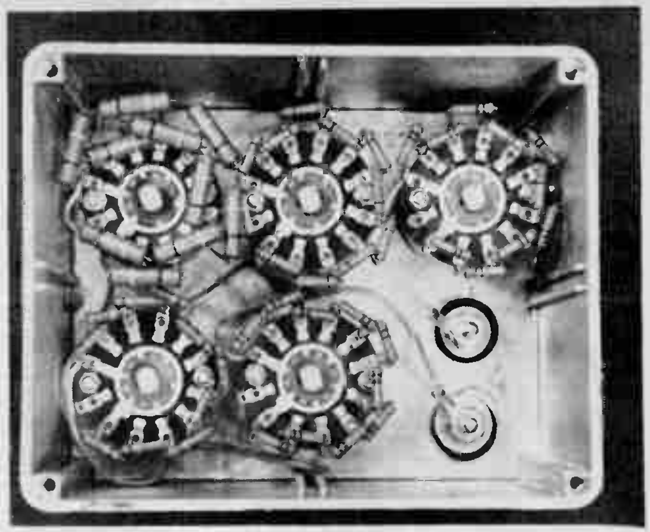


Fig. 4. This circuit recommended by a components supplier is simple but lacks short-circuit protection. Such protection is difficult to add to this circuit.

Fig. 5. Printed circuit-board layout. Full size. 178 x 78 mm.

The decade resistance box.

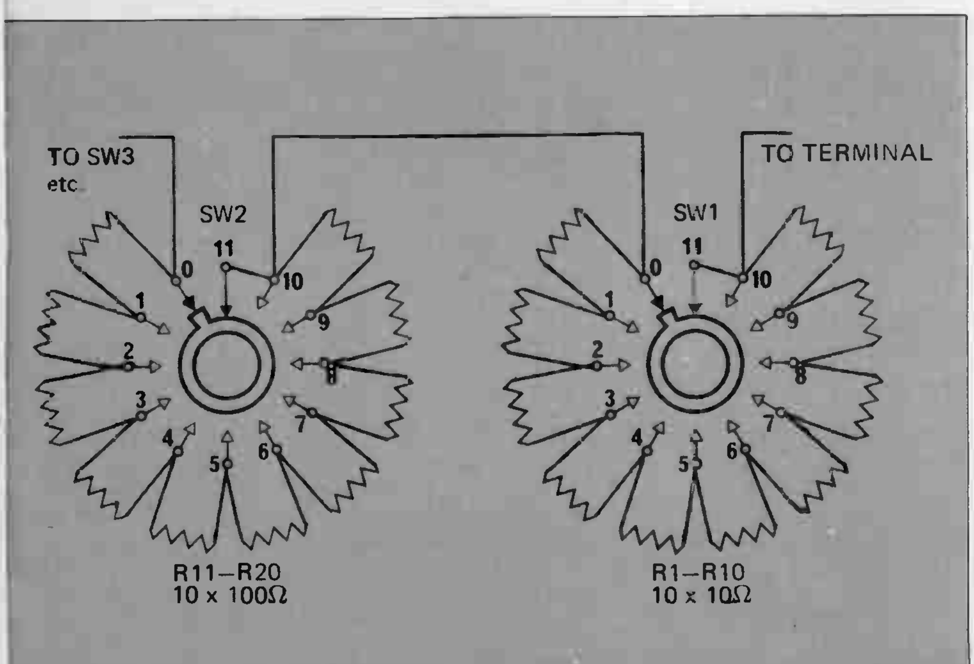


Interior of the completed unit.

DECADE RESISTANCE BOX

PROJECT 108

A versatile and accurate variable resistance unit for experimenters



A DECADE resistance box is an extraordinarily useful device – enabling the user to select precisely the resistor value needed when that breadboard mock-up doesn't quite match the theoretical calculations: there being nothing more time consuming than the 'unsolder and try another' technique.

Another use for a decade box is as a precision variable resistor in experimental measurements – and to meet this requirement we have provided ten-ohm resolution and have specified 2% accuracy resistors. These resistors are available at reasonable cost – you could of course go to 1% resistors but such accuracy is rarely needed for most applications.

Fig. 1. The method of interconnecting the switches.

PARTS LIST

- R1 – R10 Resistor 10Ω 2% ½ watt Philips type MR30 or equivalent.
- R11 – R20 Resistor 100Ω 2% ½ watt Philips type MR30 or equivalent
- R21 – R30 Resistor 1000Ω 2% ½ watt Philips type MR30 or equivalent
- R31 – R40 Resistor 10kohm 2% ½ watt Philips type MR30 or equivalent
- R41 – R50 Resistor 100kohm 2% ½ watt Philips type MR30 or equivalent
- SW1 – SW5 Wafer switch, single pole, 11 position OAK type F or similar.
- diecast box 4¾" x 3¾" x 2", ITT type 043B00 or similar.
- 2 binding post terminals and 5 knobs.

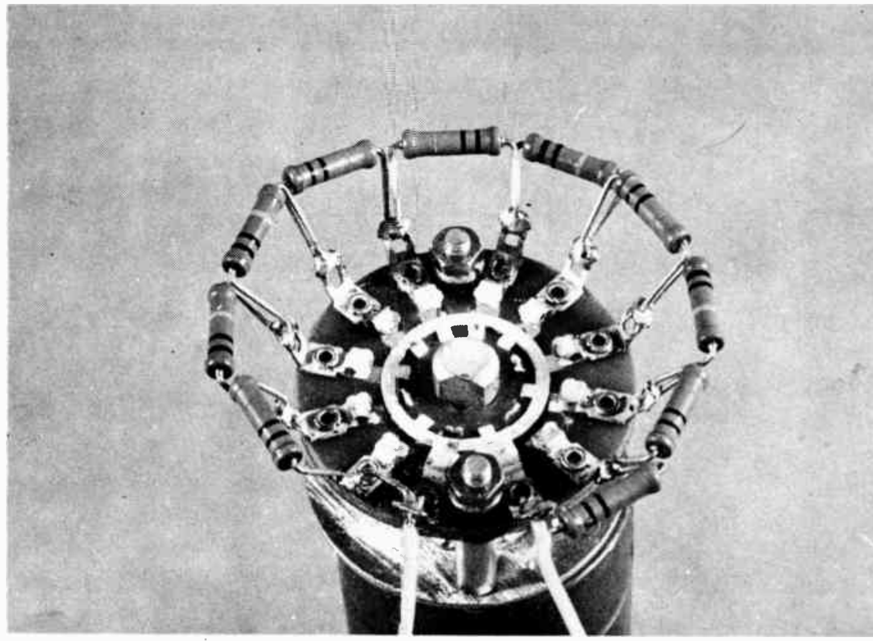


Fig. 2. The resistors are mounted to the switches as shown.

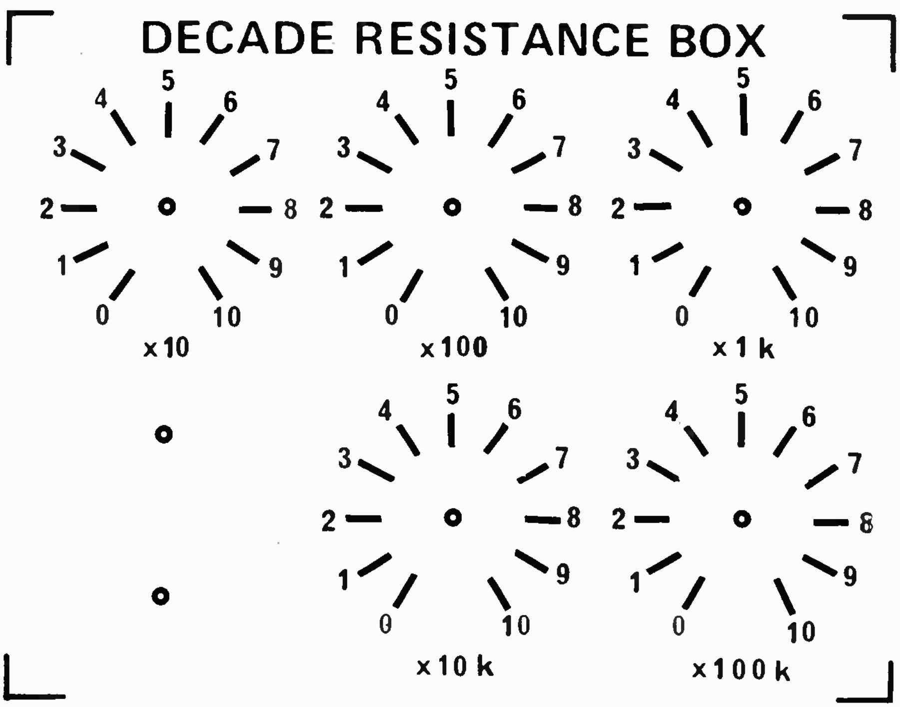


Fig. 3. Front panel overlay (full size).

CONSTRUCTION

Assemble the resistors to the switches as shown in the photograph, R1-R10 to SW1, R11-20 to SW2 and so on to SW5.

Fit the switches to the metal box and ensure that the resistors are clear of the metal box sides. If there is insufficient clearance a piece of manila-folder cardboard will provide the necessary insulation.

Connect all the switches in series, as shown in Fig.1, and then connect the wiper of SW1 to one of the input terminals with one single strand from a piece of flexible hook-up wire. This will act as a fuse and protect the switch contacts from damage due to shorts. Finally connect pin "O" of SW5 to the other terminal.

In our prototype we disassembled the switches and filed away the stops thus allowing continuous rotation in either direction. Be very careful, if you decide to do this, not to damage the switch. Remove the wafer assembly taking care not to apply pressure to the rotating wiper section. Remove the circlip retaining the shaft and withdraw the shaft/clicker plate assembly. The stop may then be removed with a file and the switch reassembled. ●

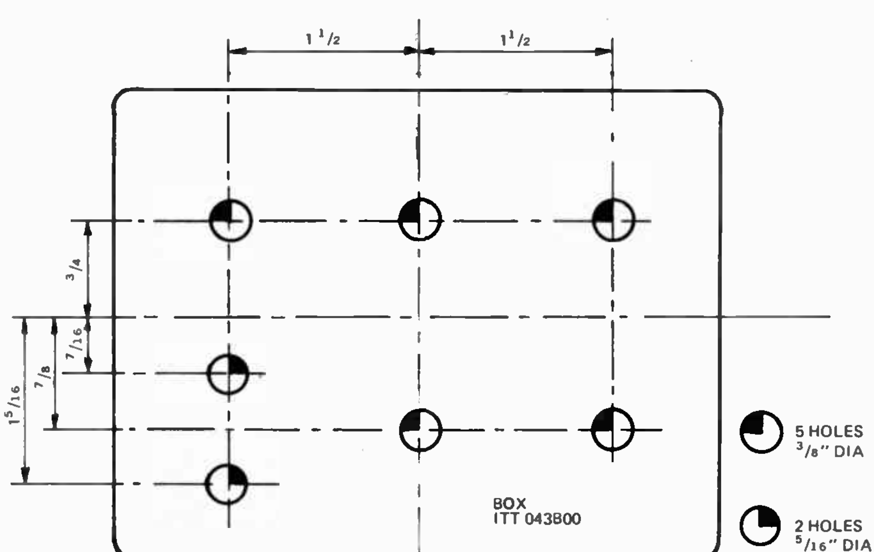
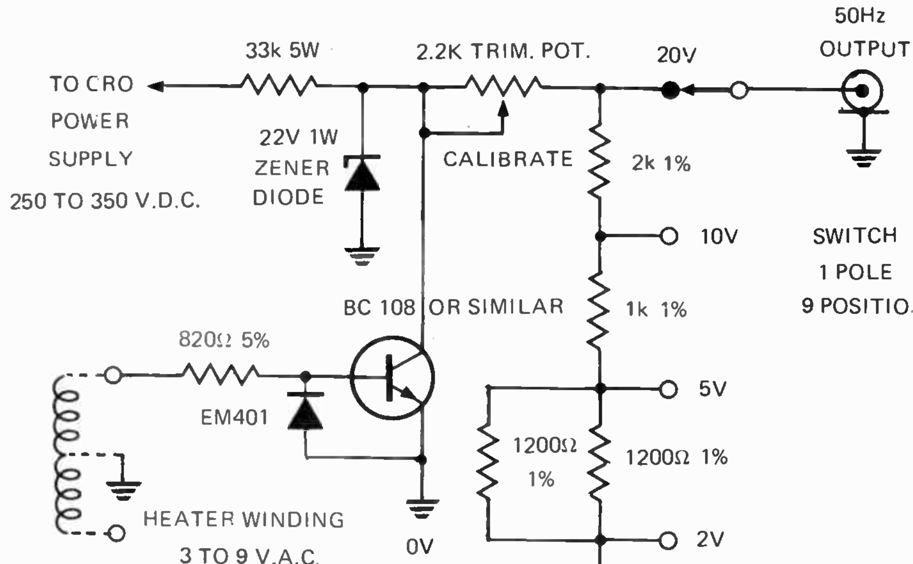


Fig. 4. Drilling details for the diecast box.

OSCILLOSCOPE CALIBRATOR

PROJECT 106

This simply-constructed voltage calibrator can be built into practically any existing oscilloscope.



THIS simple calibrator enables 50 Hz square waves of exact amplitude to be displayed on an oscilloscope.

The calibrator can be added to existing oscilloscopes, or built as an external accessory. It eliminates measuring errors due to gain controls, or probe dividers, as a calibration signal is obtainable by inserting the probe tip directly into the calibration output socket and checking the displayed calibration signal against the calibration control switch setting.

The oscilloscope time-base accuracy can also be checked with this calibrator — the 50 Hz square-wave signal is derived from the mains thus providing a stable 20 millisecond period.

The calibration voltage is derived from a 22 volt zener diode; this voltage is chopped at 50 Hz by the BC 108 transistor, trimmed to exactly 20 volts by the calibration potentiometer and applied across a chain of precision resistors.

The consumption of the unit is negligible and is energized by the power supplies of the oscilloscope to which the unit is fitted.

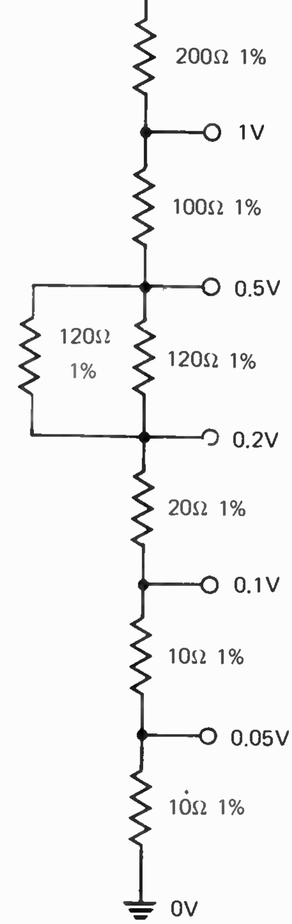
It is obviously impossible to give installation instructions for each individual make and type of oscilloscope — however all that is required is to locate an HT rail carrying between 250 and 350 volts for the main divider supply, and the filament transformer for the chopper supply. The diode connected between the BC 108's base and emitter protects against excess voltage at this point.

The divider earth point and the emitter of the BC 108 chopper transistor should be taken to the same point. All wiring should be shielded or located away from existing oscilloscope wiring to avoid cross talk.

The 50 Hz calibrator output should be taken to a binding post mounted on the front panel of the scope. The rotary calibrator switch should also be mounted on the front panel. If a miniature switch is used for this function no problem should be found finding a suitable location — the existing panel lamp mounting can often be utilized for the purpose.

The calibration potentiometer is best mounted internally. Once set it rarely needs further adjustment.

Initial calibration is performed by disconnecting the collector of the BC108 transistor and then using an accurate dc voltmeter to set the top of the divider chain at exactly 20 volts. ●

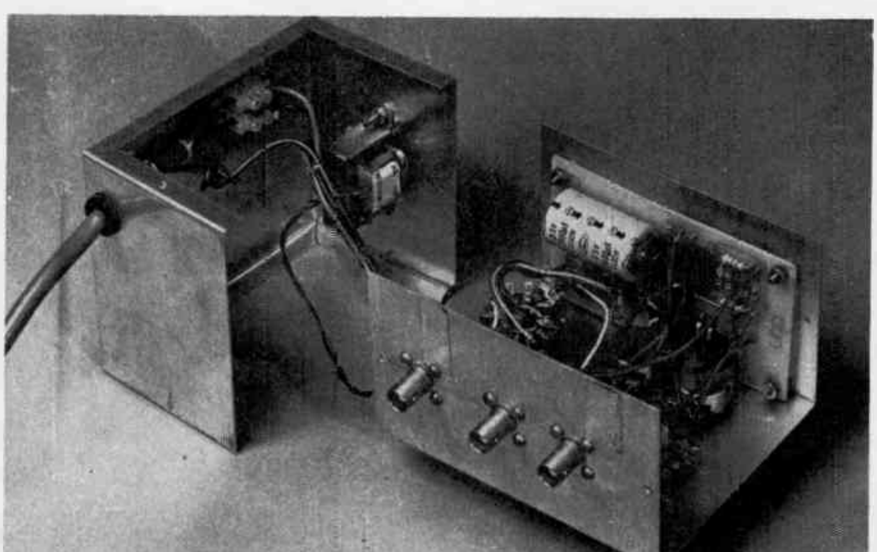


Circuit of complete oscilloscope calibrator — note that some resistor values in the divider chain are obtained by parallel resistors of higher value.

PROJECT 114

DUAL BEAM ADAPTOR

Simple unit converts single beam CRO to dual beam operation.



THE oscilloscope, next to the multimeter, is perhaps the most useful test instrument. Indeed, for any serious experimental work an oscilloscope is indispensable. Unfortunately they are expensive beasts, and whilst an experimenter may well afford a simple, low-frequency single-beam type, a dual-beam version (at $300 or more) is usually beyond his means.

Nevertheless a dual-beam facility is most convenient, for it allows comparison of two different signals, for wave-shape or timing, and makes obvious, differences which otherwise would not be discernable.

The simple dual-beam adaptor described here, whilst not providing *all* the capabilities of an expensive dual-beam CRO, will however, cover most experimenter's requirements.

It is a low cost unit which allows two inputs of similar amplitude to be displayed simultaneously on separate traces. Frequency response of the unit is sufficient to allow observation of signals up to about 1 MHz.

CONSTRUCTION

Most of the components are mounted on a printed circuit board. However, if desired matrix or veroboard may be used.

Be careful to orientate the polarised components correctly, as shown on the component overlay. Wiring to the sockets and switches should be as short as possible. Note that C3 and C4 are mounted on the input switches and C5 is mounted on the output socket.

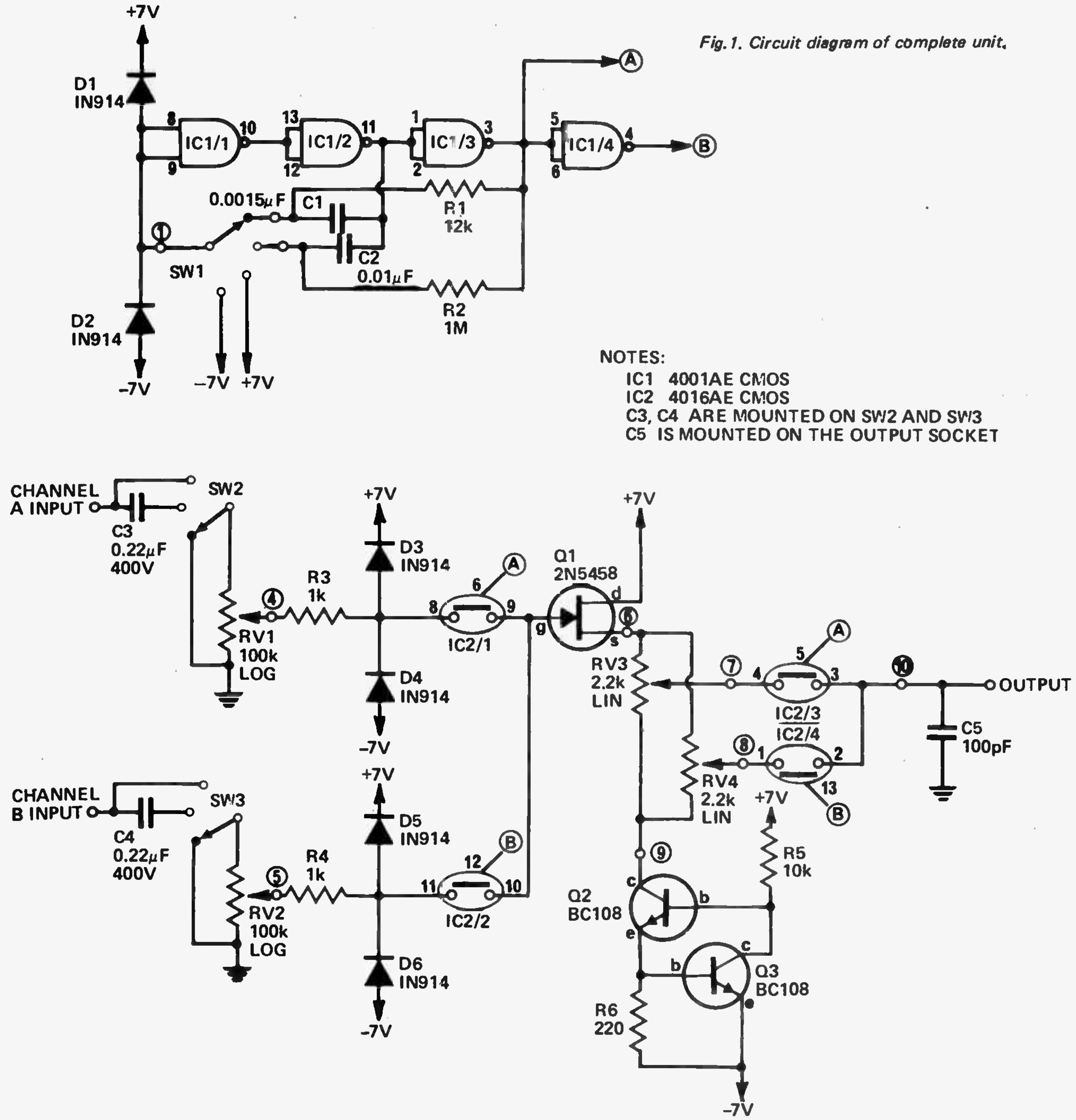
Our prototype was mounted in a small aluminium minibox as illustrated. As individual requirements will vary, details of front panel layout and metalwork only are supplied.

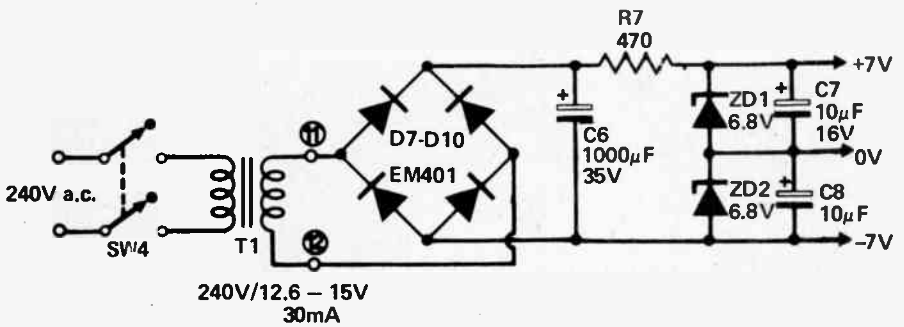
USING THE ADAPTOR

Connect the output of the adaptor to the input of the CRO. The two adaptor inputs now become A and B trace inputs to the CRO. A triggering signal should be applied direct to the trigger input of the CRO as otherwise the CRO will tend to synchronize to the chop frequency and not to either input signal.

It is preferable that the two input signals have approximately the same amplitude as there is no input amplifier or range selection provided

Fig. 1. Circuit diagram of complete unit.





on the adaptor. However there is an attenuator provided on each input so that some adjustment may be made.

If only one input is to be applied it is best to switch to that input only thus eliminating the second trace and any cross talk which may occur due to the high input impedances.

Two chopping frequencies are used, having widely different frequencies, so that if the input signal is a harmonic of the chopping frequency, (see Fig. 4) choosing the other chop mode will prevent the chop frequency being visible.

Normally CHOP 1 would be used for high frequency inputs, and CHOP 2 for low frequency inputs. An ALTERNATE mode has not been included (entails obtaining an output

SPECIFICATION

Input Level	
dc	± 4 volts max
ac	2 volts RMS max
dc insulation on ac	±400 volts max
dc level shift	± 1.5 volts
Frequency Response	
– 3dB point	> 1 MHz
Chopping Frequencies	
A	60 Hz
B	35 kHz
Input Impedance	100 k ohm

DUAL BEAM ADAPTOR

from the CRO of unknown level and availability) as the CHOP 1 mode is similar and almost as effective.

By means of the two shift controls traces A and B may be separated by up to ±1.5 volts.

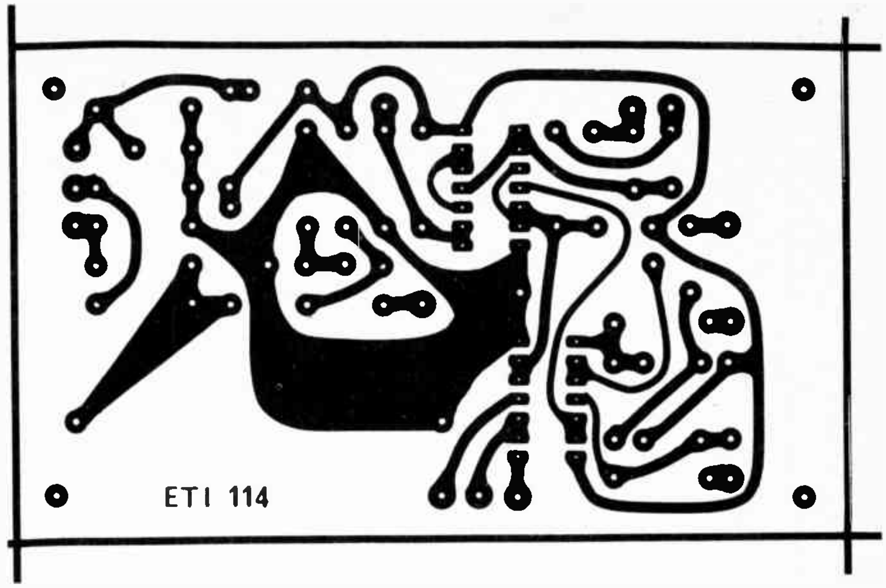


Fig.2. Printed circuit board pattern for the adaptor. (Shown fullsize).

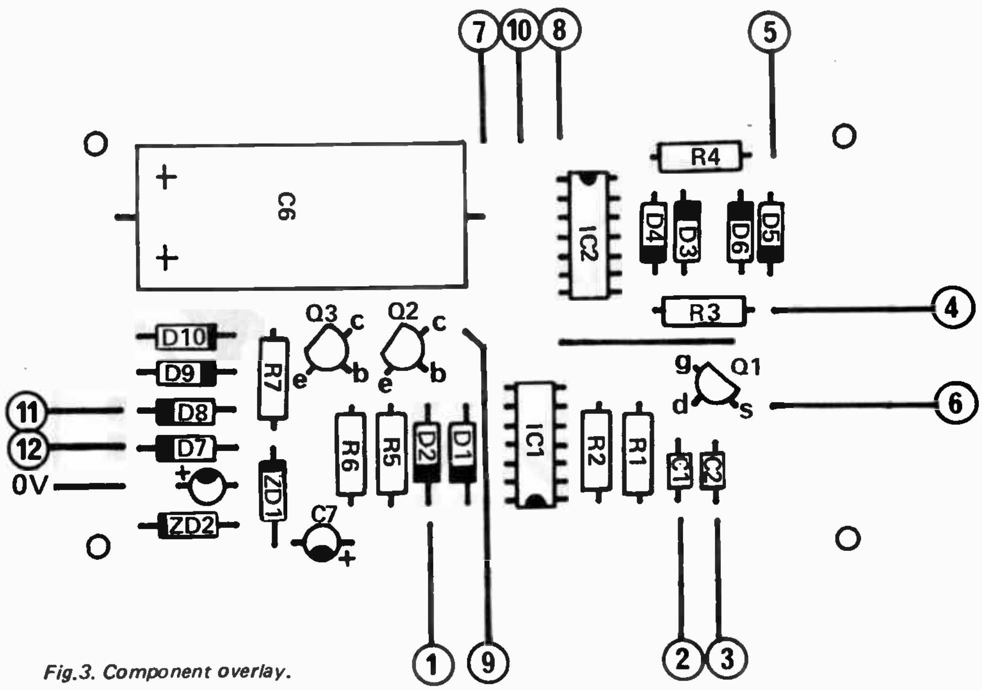


Fig.3. Component overlay.

HOW IT WORKS – ETI 114

Switches SW2 and SW3 select dc or ac coupling, or input shorted, for channel A and channel B inputs respectively. The signals are applied to the sensitivity potentiometers RV1 and RV2 and then passed to IC2/1 and IC2/2 which select one of the signals as an input to source follower Q1.

Transistor Q1 is supplied with a constant current (approximately 2.7 mA) by transistors Q2 and Q3. Hence, there is about 3 volts across RV3 and RV4, and this is unaffected by changes in input signal level. These potentiometers therefore provide a level-shift facility. When channel A is selected by IC2/1, IC2/3 selects RV3, and when channel B is selected by IC2/2, IC2/4 selects RV4. Thus as each signal has an independant level shift the two traces may be separated when chopped.

The CMOS gates of IC2 are driven by the outputs, A and B, the circuitry associated with IC1. The drive circuit mode of operation is selected by SW1, a four position switch, such that channel A only, channel B only, A and B chopped at 60 Hz or, A and B chopped at 35 kHz may be selected. The operation is as follows.

Integrated circuit IC1 forms a multivibrator which can run at 60 Hz or 35 kHz, or be locked in A-high B-low, or A-low B-high output states. For example, if SW1 selects –7 volts, IC1 pin 10 will be at +7, IC1 pin 11 will be at –7, IC1 pin 3 will be at +7 and IC1 pin 4 will be at –7 volts. The CMOS switches of IC2 will be "on" if the control voltage is at +7 volts and "off" if the control voltage is at –7 volts. Thus when –7 volts is selected by SW1, "A" will be at +7 volts, and IC2/1 and IC2/3 will select channel A. Similarly if +7 volts is selected by SW1, IC2/2 and IC2/4 will select channel B.

If C2 and R2 are selected by SW1 the multivibrator will be free to run at 60 Hz and channels A and B will be alternately selected at this frequency. Similarly if C1 and R1 are selected, channels A and B will be alternately selected at 35 kHz.

The power supply is a simple full-wave bridge type which uses two Zeners to provide the +7 and –7 volt supplies required.

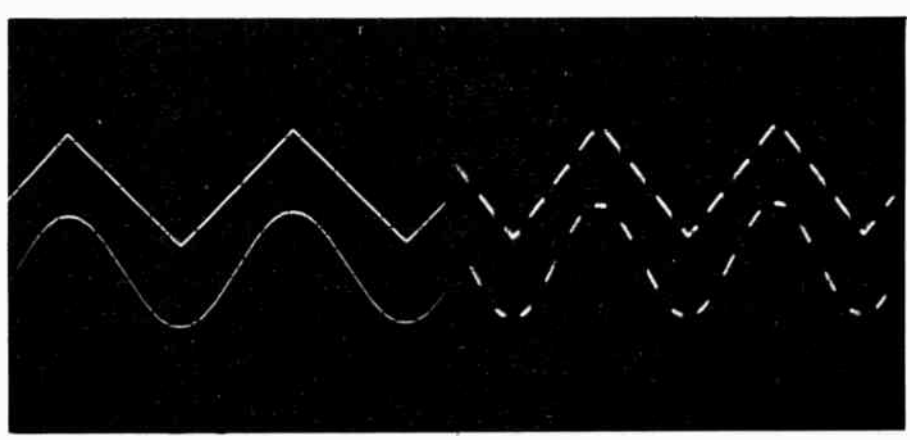
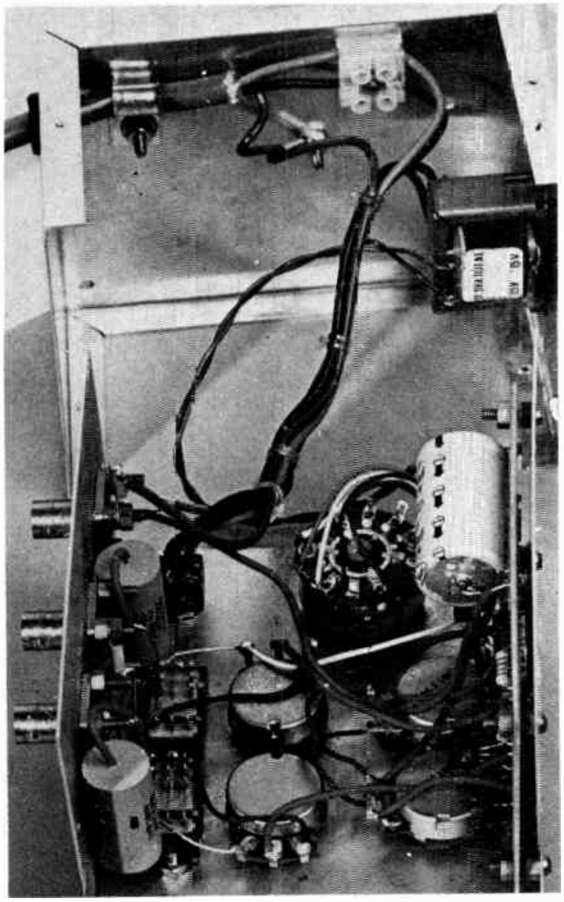


Fig.4a. Two signals, correctly displayed using the dual beam adaptor.

Fig.4b. Use of incorrect chopping frequency for a particular input signal (chop frequency a harmonic of signal) results in above effect. To cure use other chop frequency.



Layouts of components within the unit can be see from this and accompanying photographs.

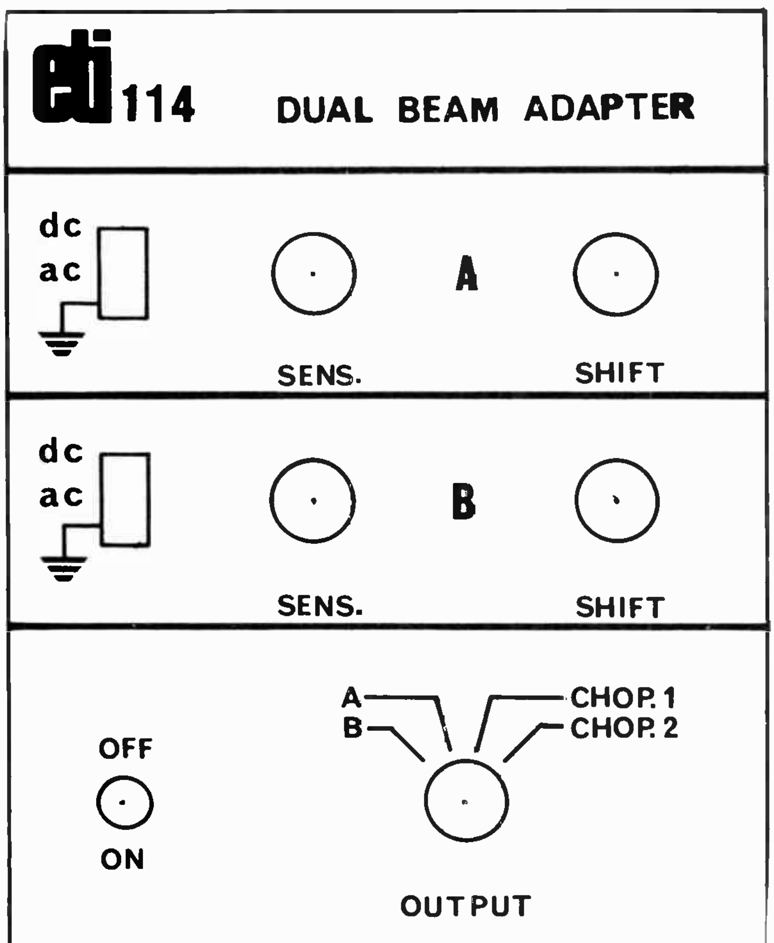


Fig.5. Artwork for front panel of the edaptor.

PARTS LIST — ETI 114

R6 Resistor 220 ½W 5%
R7 " 470 ½W 5%
R3,4 " 1k ½W 5%
R5 " 10k ½W 5%
R1 " 12k ½W 5%
R2 " 1M ½W 5%

RV1,2 Potentiometer 100k log rotary
RV3,4 Potentiometer 2.2k lin rotary

C5 Capacitor 100pF ceramic
C1 " 0.0015μF polyester
C2 " 0.01μF polyester
C3,4 " 0.22μF 400V poly.
C7,8 " 10μF 16V electrolytic
C6 " 1000μF 35V "

D1-D6 Diode IN914 or similar
D7-D10 " EM401 or similar
ZD1,ZD2 Zener Diode BZY88C6V8 or similar

Q1 Transistor 2N 5458
Q2,Q3 " BC108, BC548 or similar

IC1 Integrated circuit 4001AE CMOS
IC2 Integrated circuit 4016AE CMOS

T1 transformer 12.6V — 15V @ 300ma PF2851, PF3786, A&R7577 etc.
PC Board ETI 114
SW1 switch one pole 4 position rotary
SW2,3 switch 3-position slide switch
SW4 switch 2-pole on-off toggle 240V rated.
Metal Box 130mm x 105 mm x 80 mm
3 sockets to suit CRO leads
Knobs for front panel.

SILENT A-B SWITCH

Speakers may be A-B tested using this simple modification to our tone-burst generator

WHEN evaluating speaker systems in A-B listening tests, the first few seconds of listening convey the truest impression of sound quality. Listening for longer than a few seconds not only fails to give further information, but may well give a false indication. For this reason it is usual to switch rapidly between the reference speaker and the speaker under test. This is generally done by using the amplifier's A/B speaker selector switch, or by wiring a change-over relay in the speaker wiring.

Whilst such switching methods are simple and reliable they have one major drawback. That is that switching may take place at any point in the waveform and as a consequence switching transients may be introduced which tend to mask the subtle differences for which one is listening. Hence a method of switching at zero-crossing points would be of great value.

When the ETI Tone-Burst Generator was constructed it was realised that it contained all the circuitry needed to performance this switching task and that it could be modified to do so very simply.

The switching must be done at low level and hence the unit is used at the input of a stereo power amplifier. The reference speaker and the speaker under test are each connected to one channel of the amplifier and the silent switch switches the input to the amplifiers as required. Thus the arrangement is mono only but this is all that is required to assess the transient response and performance of a speaker in comparison to a reference speaker.

HOW IT WORKS – ETI 124 AB

As this unit is based on the operation of the tone-burst generator ETI 124 described on pages 47-51, that article should be thoroughly read first. Only the changes necessary to that unit are detailed in this article. A-B switch would be a little simpler if designed specifically for that purpose, the modifications required to the tone-burst generator are so simple that we thought it not worth while to design a special circuit.

To make the generator act as an A-B switch it is necessary to disable the existing mode switch. We do this by plugging in an external control switch, SW6, via a stereo phone socket. The phone socket has two change-over contacts fitted which are used to disconnect the plus and minus six volts supplies from SW3 when the jack is inserted. One of the phono contacts also disconnects the plus six volts from the common of the socket when the jack is removed. As the common of the socket is required to be at plus six volts the phono socket must be insulated from the front panel which is at 0 volts.

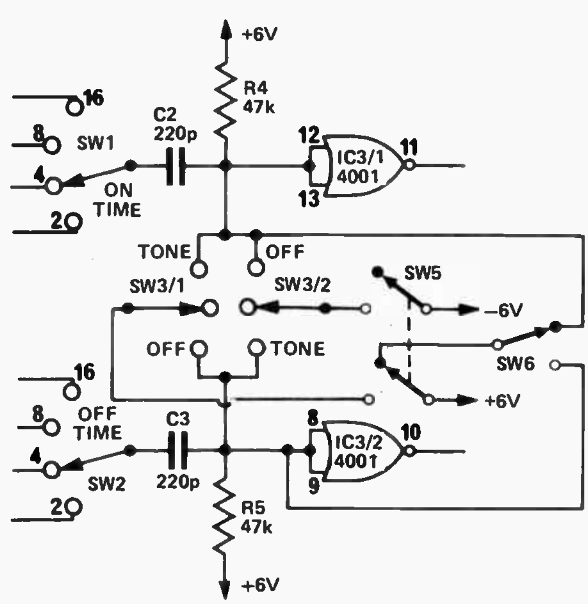
The control switch, SW6, effectively shorts either R4 or R5 thus stopping the pulses from C2 or C3 triggering the flip-flop. When the switch is actuated there is a delay until the number of cycles as set by the front panel switch have occurred and then, at the next zero crossing, the change-over occurs. The delay is necessary to ensure that any contact bounce of the SW6 contacts does not cause unwanted switching of the circuit.

CONSTRUCTION

The ETI 124 Tone-burst Generator should first be constructed as detailed on page 47 except that the wiring to SW3 is changed as detailed in Fig. 1 and 2 of this article. The dual-RCA socket and the phono socket are then mounted on one side of the box. If a metal box is used make sure that the phono socket is insulated from the case of the box as it is at a potential of six volts. The switch, SW6, should be mounted in a small pill container or similar housing and fitted with a three-core cable that is terminated at the other end by a stereo phone jack. Note that the common of the switch should be connected to the common of the jack but that the other wires may be wired to either of the remaining contacts.

USING THE SWITCH.

The audio switch requires a reasonably high level of signal to ensure correct zero-crossing switching. There are two suitable points in a conventional amplifier. The first position is between the tape-in and tape-out sockets but the second and preferable position is between the pre and main amplifiers provided that the main amplifier has a volume control that is independent of the preamplifier.



PROJECT 124 AB

Fig. 1. Partial circuit diagram of the tone-burst generator modified to perform AB switching.

To connect the unit for AB testing apply a single input, from the preamplifier (switched to mono), to the normal input socket of the generator. The normal output socket of the generator is not used but the two RCA output sockets are connected back to the left and right channel inputs of the main amplifier. When SW6 is operated the mono input will be silently switched between right and left channel speakers.

If using the tape sockets the monitor switch should be in the 'monitor' position and the balance control should be adjusted so that the levels from the two speakers are apparently the same. Make sure that the tone controls are in the flat position, as they can cause phase shifts which prevent the switching occuring at the zero-crossing point.

If the pre and main amplifier terminals are used the preamplifier volume should be adjusted to about half way and separate volume controls used to balance for the difference in efficiencies of the two speakers. If the main amplifier does not have separate volume controls then external ones must be added if balance is to be achieved. In this case the tone controls may be used if required without upsetting the crossover point.

Change over may be effected by using either a toggle switch or a push button. The tone-burst generator controls should be set for eight cycles on and off as this position will effectively remove any contact bounce.

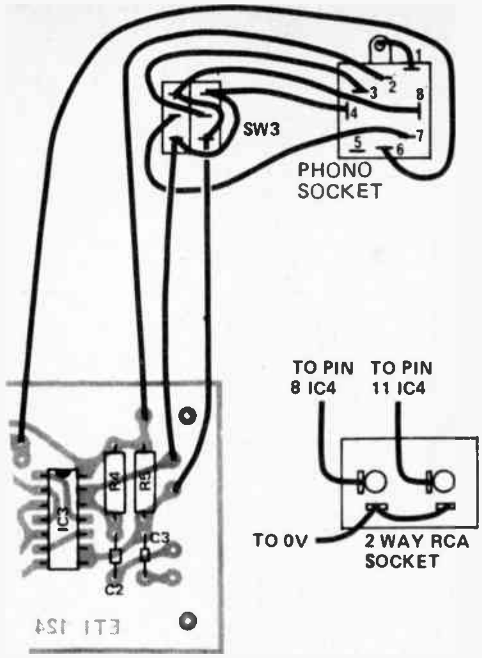


Fig. 2. Interconnection diagram to phono socket and RCA output sockets of AB switch.

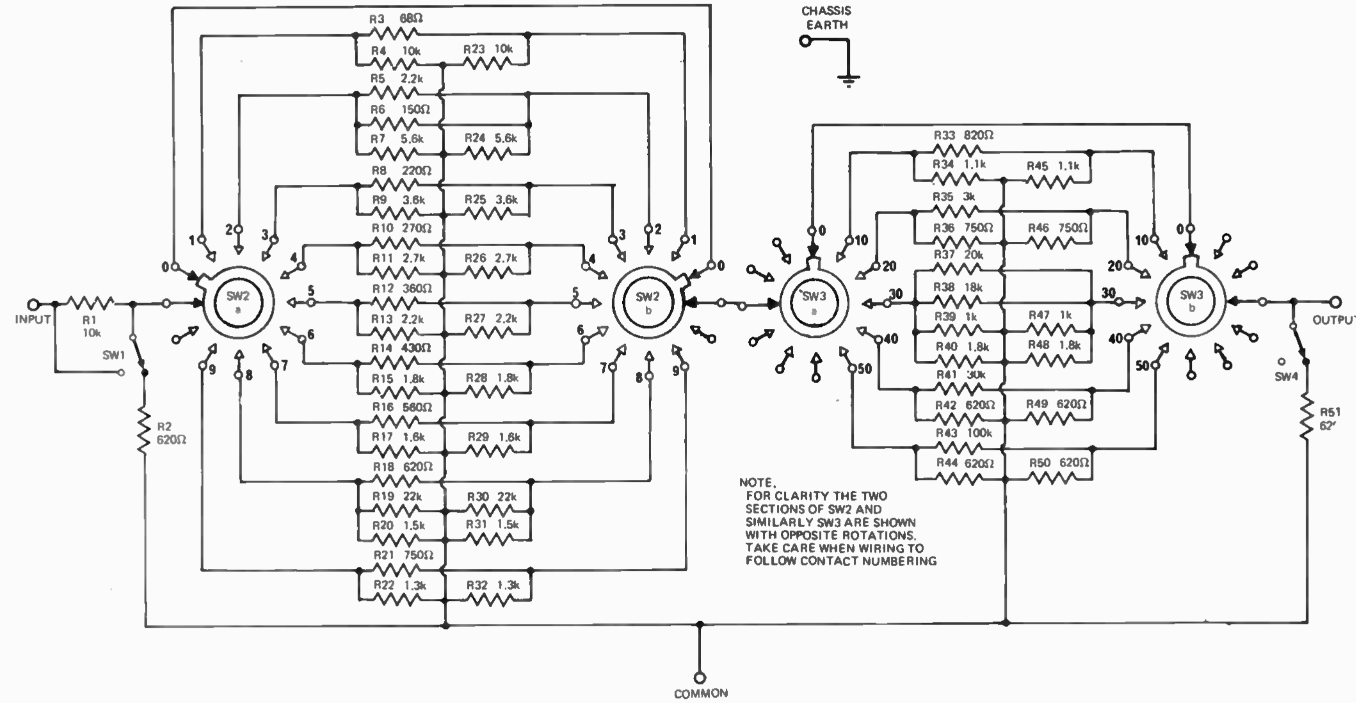
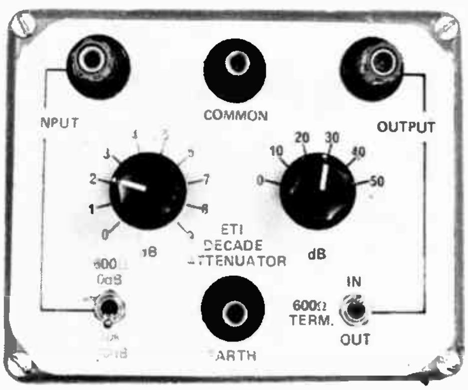


Fig. 1. Circuit diagram of the attenuator.

This useful audio attenuator project for the experimenter provides 0-59dB attenuation in one dB steps.

AUDIO ATTENUATOR

PROJECT 112



ACCURATE attenuators are required in a multitude of design, service, testing and measuring situations. These units are designed with varying degrees of accuracy and as many steps of attenuation as the designer feels necessary. They may be balanced or unbalanced and have whatever input and output impedances the designer requires.

There are three common types of attenuator configuration, Pi, T or L. The latter is mainly employed where the output impedance is not required to be constant.

We have chosen Pi type sections for our unit. We could have connected the various sections in tandem to form a ladder attenuator, but this would have made more complex rotary switches necessary. Instead, we chose to employ a separate section for each step of attenuation, making only simple rotary switches necessary.

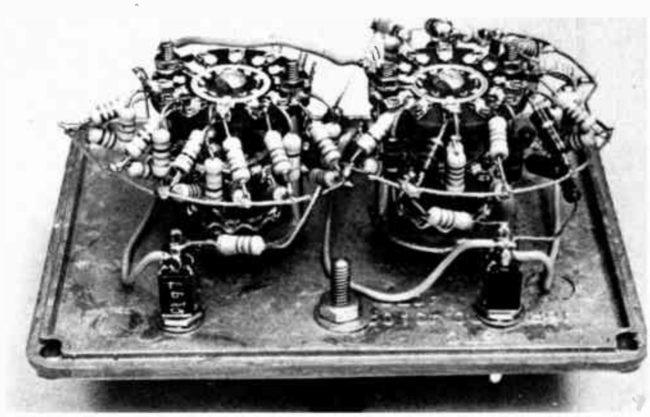
The input and output resistances of the unit remain relatively constant at 600 ohms over the full attenuation range. The input impedance can be changed to 10k by SW1 but an additional 30dB of attenuation is added. The output can also be terminated internally by SW4 when using a high impedance load such as a meter.

The maximum attenuation when the input and output resistances are set at 600 ohms is 59dB. There are ten 1dB steps from 0dB to 9dB, via a 10 position rotary switch, and a further six 10dB steps from 0dB to 50dB via a six position rotary switch, giving a

SPECIFICATION

Max attenuation	59dB
Resolution	1dB
Accuracy	±0.3dB
Frequency range	dc to 100kHz
Input impedance	600 Ω nominal
	10k switched (+30dB attenuation)
Output impedance	600 Ω nominal
Max input voltage	15 volt

Internal switched termination resistor for use with high impedance loads.



PARTS LIST ETI 112

R1	Resistor	10k	2% ½W
R2	"	620Ω	" "
R3	"	68Ω	" "
R4	"	10k	" "
R5	"	2.2k	" "
R6	"	150Ω	" "
R7	"	5.6k	" "
R8	"	220Ω	" "
R9	"	3.6k	" "
R10	"	270Ω	" "
R11	"	2.7k	" "
R12	"	360Ω	" "
R13	"	2.2k	" "
R14	"	430Ω	" "
R15	"	1.8k	" "
R16	"	560Ω	" "
R17	"	1.6k	" "
R18	"	620Ω	" "
R19	"	22k	" "
R20	"	1.5k	" "
R21	"	750Ω	" "
R22	"	1.3k	" "
R23	"	10k	" "
R24	"	5.6k	" "
R25	"	3.6k	" "
R26	"	2.7k	" "
R27	"	2.2k	" "
R28	"	1.8k	" "
R29	"	1.6k	" "
R30	"	22k	" "
R31	"	1.5k	" "
R32	"	1.3k	" "
R33	"	820Ω	" "
R34	"	1.1k	" "
R35	"	3k	" "
R36	"	750Ω	" "
R37	"	20k	" "
R38	"	18k	" "
R39	"	1k	" "
R40	"	1.8k	" "
R41	"	30k	" "
R42	"	620Ω	" "
R43	"	100k	" "
R44	"	620Ω	" "
R45	"	1.1k	" "
R46	"	750Ω	" "
R47	"	1k	" "
R48	"	1.8k	" "
R49	"	620Ω	" "
R50	"	620Ω	" "
R51	"	620Ω	" "

SW1 Single pole change over miniature toggle switch
SW2 2 pole 11 position rotary switch
SW3 2 pole 11 position rotary switch
SW4 Single pole change over miniature toggle switch
Diecast box 4¾ x 3¾ x 2
4 Terminals type L1568/1S or similar
2 Knobs

total of 60 steps from 0dB to 59dB. This range of attenuation is adequate for most purposes. Although further sections could be added, noise becomes a limiting factor in a simple attenuator such as this.

CONSTRUCTION

It is advisable to employ separate wafers for each switch pole. If the type of switch that has two poles on one wafer is employed, there may be problems at the high frequency end due to stray capacitance. This would be evident as spikes on the leading edges of high frequency square waves.

The common rail for each switch is a length of 18 gauge tinned copper wire formed into a ring to allow termination of the shunt resistors (R4, R23, R7 and so on). The series resistors are connected directly between the relevant switch contacts. Layout of the unit may be seen by the accompanying photographs. ●

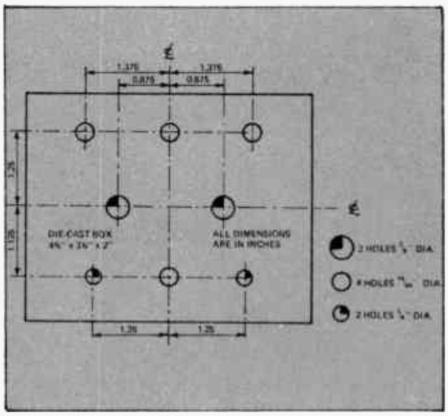


Fig. 2. Drilling details for the die cast box.

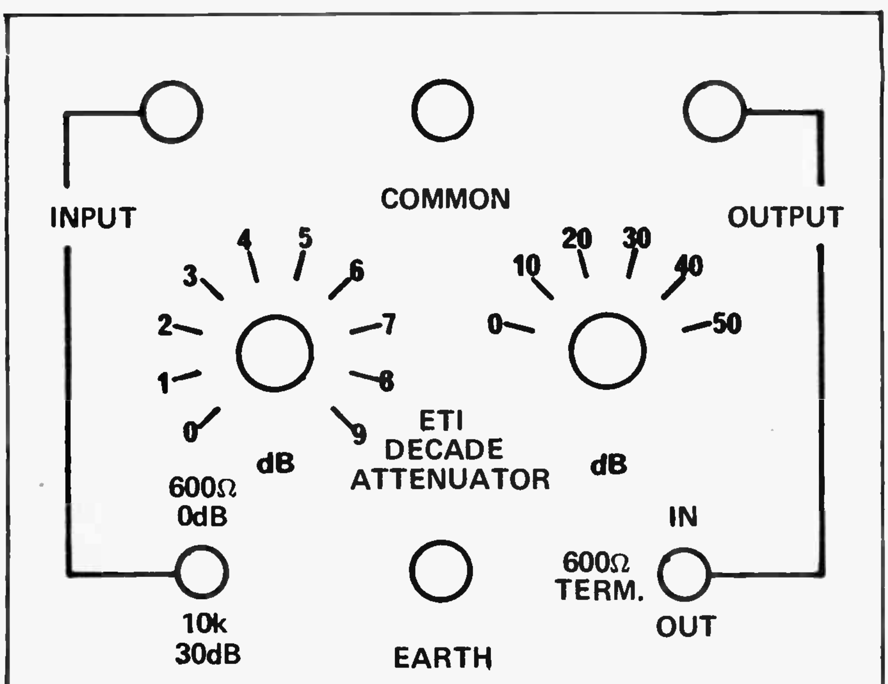
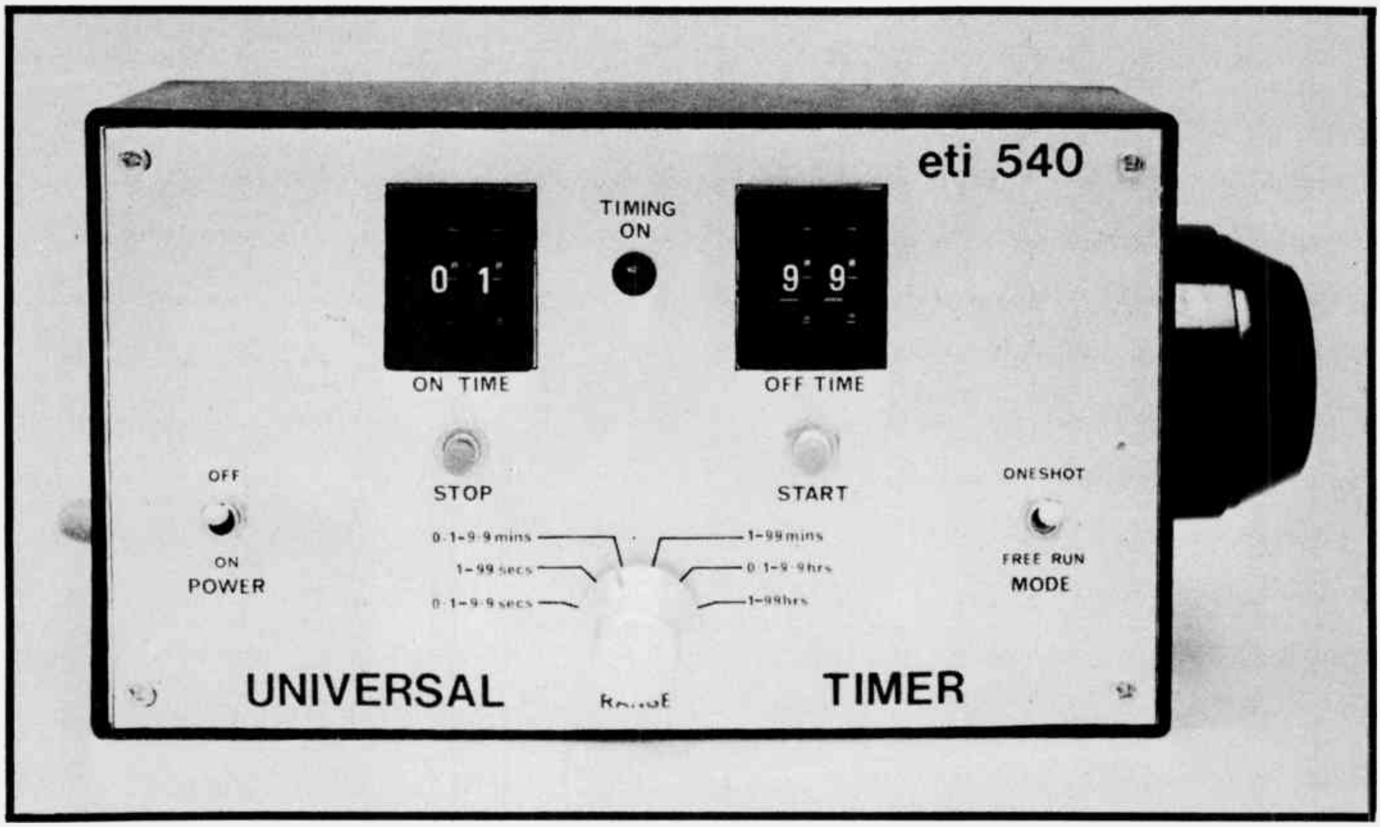


Fig. 3. Lettering and front panel artwork — full size.

Universal timer

One tenth of a second to 99 hours. Both on and off times programmable. Manual or automatic operation resettable at any time.



THE TIMING OF EVENTS and processes is becoming an ever-increasing necessity particularly in applications involving automation.

Unfortunately most timers are either specifically made for a particular application – and difficult to adapt to others – or have restricted timing range, accuracy and facilities.

The ETI Universal Timer described in this project is free of most such constraints. It is extremely flexible, accurate and versatile. Its timing range is from 0.1 seconds to 99 hours. Both 'on' and 'off' times can be programmed (for example 12 hours on and 47 hours off). It can be manually started, stopped, or reset at any time, can be set for automatic cycling or for single cycle operation. It may be triggered by an external source (light, sound or pressure transducer etc). Finally, as the unit is digital – the 50 Hz mains is used as the reference – timing accuracy is very high indeed, and a manual reset facility enables the timer to be synchronized with local time if so desired.

Clearly not all users will need all the facilities provided – so if the unit is required for a specific permanent use it is a simple matter just to leave out those ICs not required – several variations are described at the end of this project.

CONSTRUCTION

We strongly recommend that this unit be assembled using the printed circuit board shown.

Begin construction by fitting the links to the board as shown on the component overlay. Note that there are two points labelled 'a' and two points labelled 'b'. Link 'a' to 'a' and 'b' to 'b' using insulated hook-up wire routed on the copper side of the board.

Mount the resistors to the board followed by the diodes, transistors, capacitors and finally the ICs. Take particular care to ensure that all the polarized components are orientated correctly — especially the integrated circuits.

Wires should now be attached to the board for later connection to the front panel switches. We used rainbow cable for the connections to the thumb-wheel switches as this makes the wiring easier and also helps to keep the wiring tidy. Mount the printed-circuit board into the case and mount the power outlet socket. Assemble the switches to the front panel and then interconnect the printed-circuit board, front panel and power socket in accordance with the interconnection diagram.

Finally after wiring the 240 Vac power circuitry insulate all 240 V terminals with tape to ensure that there is no risk of personal contact when fault finding is required at any later date.

CUSTOMIZING

The unit need not necessarily be built in its complete form and many different modifications are possible to lessen the cost of the unit when it is to be used for one particular application only. The modifications required for a number of specific applications are described below.

Specific fixed time — delete selector switches SW3 to SW6, and replace by wiring links from the appropriate outputs of IC4 and IC5 to the inputs of IC6/1 and IC6/2 respectively. The range switch may also be omitted by installing a link between the appropriate output of IC1 to IC3 and pin 13 of IC4.

Single shot operation — connect both inputs of IC6/2 to ground and omit switches SW5 and SW6.

Timing 99 hours or less — omit IC3 and connect inputs of IC7/3 and IC7/4 to ground.

Timing 99 seconds or less — omit IC2, IC3 and IC7.

External triggering — simplest way is a relay contact in parallel with start or stop button.

SPECIFICATION ETI 540

MODES

Freerun
On/off (note 1)
One shot
Manual override (note 2)

TIMING RANGE

0.1 seconds to 99 hours (note 3)

ACCURACY

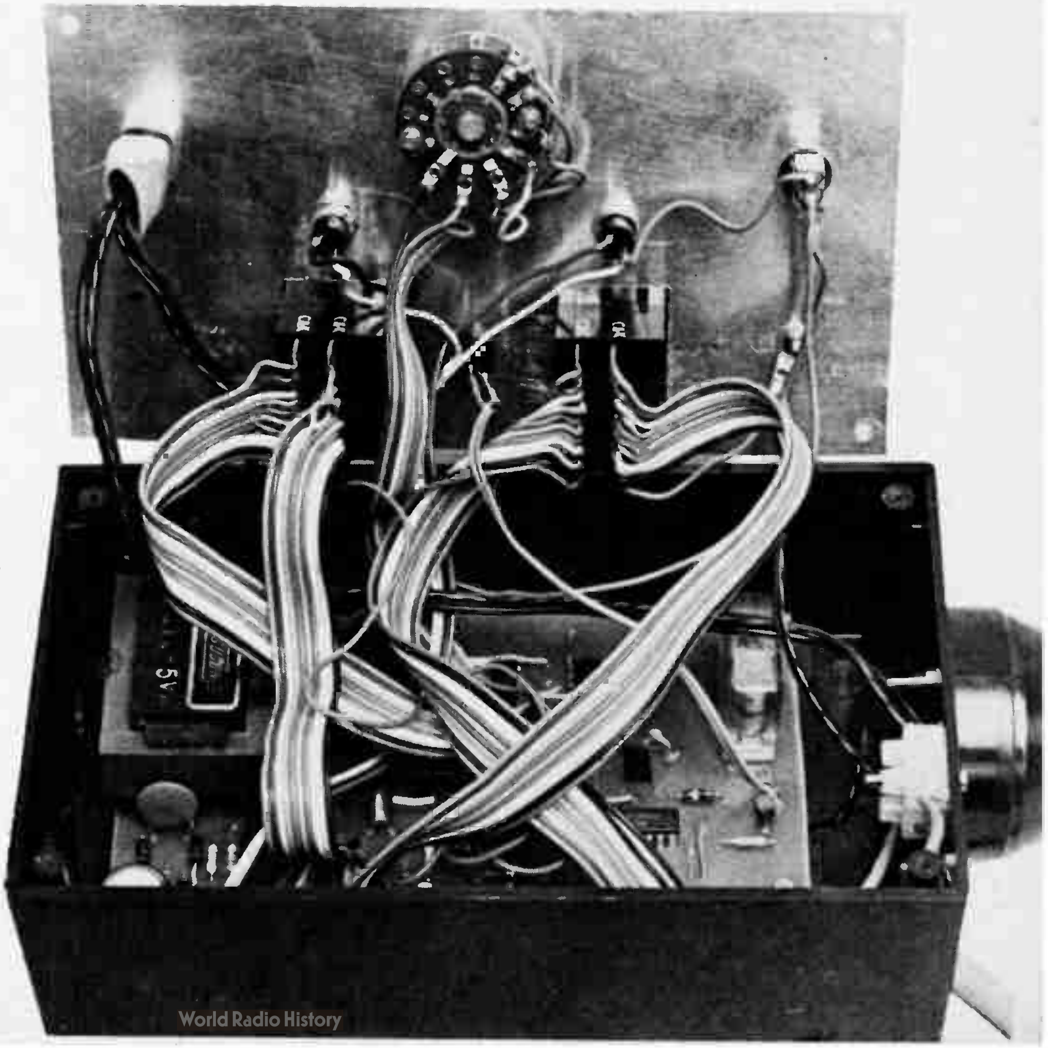
Mains synchronized

OUTPUT

240 volts ac relay switched

Note 1. Both on and off times are variable independently.
Note 2. Unit may be stopped or started at any time. If the appropriate button is pressed whilst in the same mode the timing is recommenced.
Note 3. Timing is adjustable by a common coarse control which gives ranges having a full scale of 9.9 seconds, 9.9 minutes, 99 minutes, 9.9 hours and 99 hours. Each range is adjustable from 1 to 99 that is one second on and 99 seconds off is possible whereas one second on and two minutes off is not (different coarse range is required).

The main consideration when making any changes is that the logic is CMOS and any unused inputs must be connected to ground or to +12 volts to prevent damage to the IC (which may overheat with unconnected inputs).



Universal timer

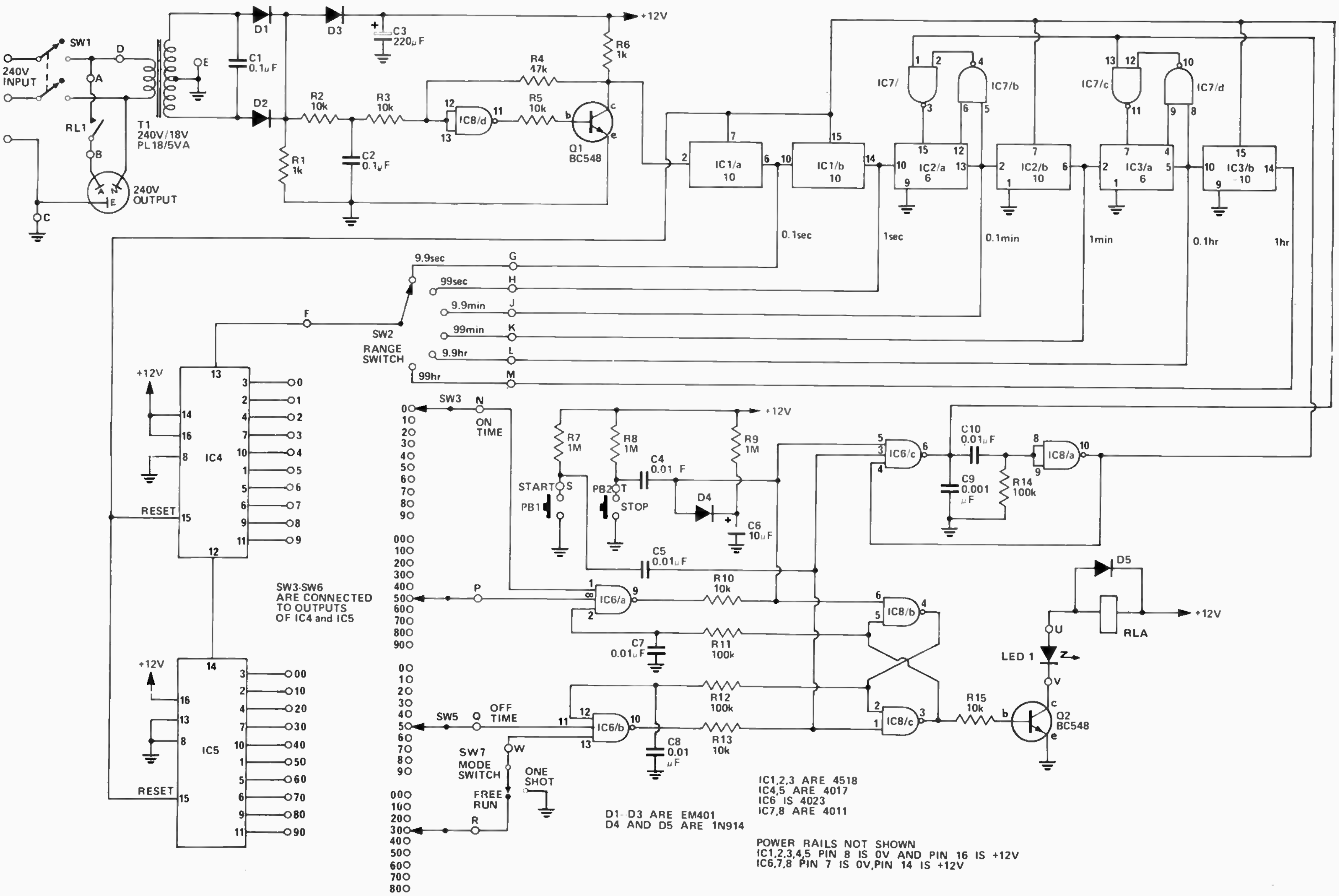


Fig. 1. Circuit diagram of the complete timer.

HOW IT WORKS – ETI 540

THE 240 Vac is reduced to 12 Vdc by transformer T1 and diodes D1 to D3. Diode. D3 isolates the smoothing capacitor C3 from the rectifiers and therefore 100 Hz ripple appears across R1. This waveform is used for the basic timing reference for the timer. To operate the counting ICs reliably a very fast rise-time waveform is required at the clock input. This is obtained by feeding the 100 Hz to a Schmitt formed by IC8/1 and Q1. Capacitor C2 is included to prevent the control tones superimposed on the mains for the control of hot-water services from upsetting the timing accuracy.

The 100 Hz from the Schmitt trigger is divided by 10 by IC1/1 to give a 10 Hz or 0.1 second output – the first required. Note that due to the low frequencies involved from now on the outputs will be referred to as time periods not as frequencies. A second divide by ten stage is used to give a one second output. A division by six is then performed by IC2/1 with IC7/1 and IC7/2 being used to decode the six count and reset the counter. This gives the one minute (or sixty second) period required. Further divisions of 10,6 and 10 are used to provide the six outputs required to select periods from 0.1 seconds to one hour.

One of these six outputs is selected by the range switch SW2 and is fed to a 4017 IC – the first of a pair of decade counters which have ten decoded outputs. The ten outputs of each IC go high in turn for one clock period each. As the two 4017 ICs are in series, a total division of 100 is obtainable. We have labelled the outputs of IC4 and IC5 as 0 to 9 and 00 to 90 respectively. IC4 is triggered by the clock enable as negative edge triggering is required. The second IC is clocked normally by the carry output from IC4.

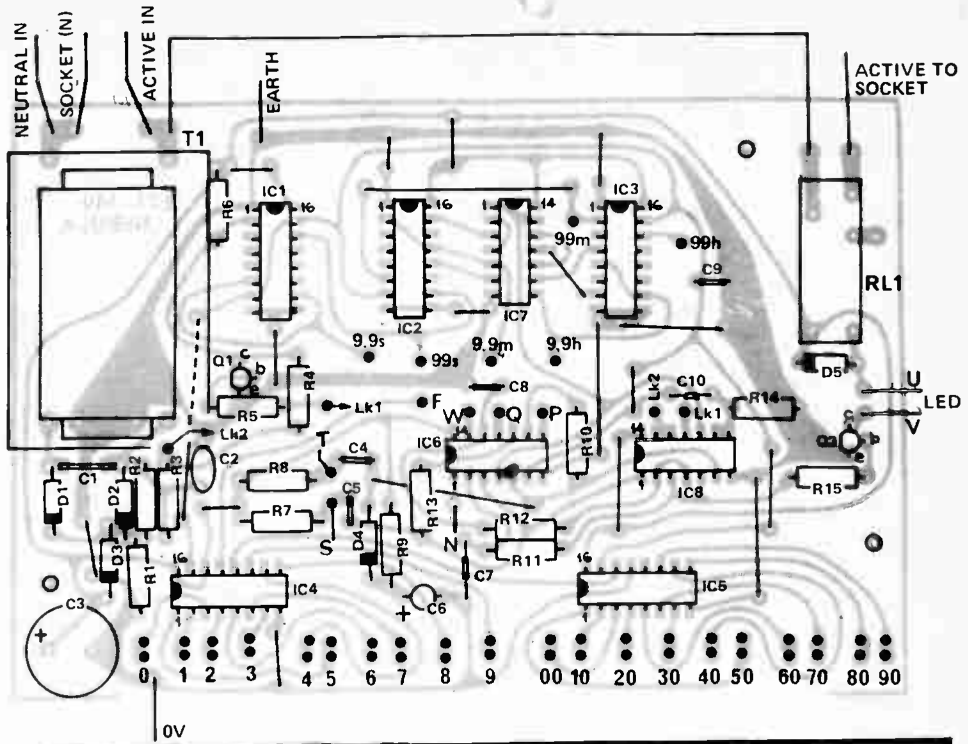
We pause at this point to go straight to the control output which is via a relay RL1, this in turn being controlled by the flip-flop made up of IC8/2 and IC8/3. This flip-flop can be controlled either manually by PB1 (manual on) and PB2 (manual off) or automatically by IC6/1 and IC6/2. To toggle the flip-flop automatically the output of either IC6/1 or IC6/2 must be low and for the output to be low the three inputs must all be high. This occurs only when the number selected by SW3 and SW4 (for IC6/1) and SW5 and SW6 (for IC6/2) is held by the counters IC4 and IC5 and the third input from the flip-flop is used to ensure that the off-time of the relay is controlled only by the off-time selector switches. A small time delay is incorporated in the signal back from the flip-flop to avoid the ambiguity that could arise with equal times.

If the output of either IC6/1 or IC6/2 goes low the monostable formed by IC6/3 and IC6/4 is triggered and its resultant output is used to reset all the counters to zero. This reset also occurs if either of the manual push buttons is pressed. The push buttons are coupled into the logic by capacitors so that only the initial part of the press actuates the logic and there is therefore no dependency on the length of time for which the button is pressed.

The sequence of events is as follows assuming that initially the switches are set for 25 seconds on and 14 seconds off.

On first switch-on C6 ensures that the flip flop is toggled into the off state and also that the counters are all reset to zero. The control inputs from the flip-flop to IC6/1 and IC6/2 are low and high respectively. Therefore until the flip-flop changes state only IC6/2 can have the three high inputs necessary to provide a low at the output. Meanwhile the counters IC4 and IC5 are counting up at the rate of one count per second. After 14 seconds all three inputs to IC6/2 are high and the output goes low toggling the flip flop. The monostable is then triggered and all counters are reset to zero. This removes the three high inputs to IC6/2 and the output goes high again. The pulse output of IC6/2 is very narrow and is about a microsecond long. As the flip-flop has now changed state the relay has been closed and IC6/1 has been enabled (control input to pin 2 now high). After 25 seconds all the inputs to IC6/1 are high and the same procedure as before resets the counters and changes the state of the flip-flop.

In the one-shot mode of operation one input of the off timer is grounded and the off time procedure is effectively disabled. The only way that the timer can now start is for the manual start button to be pressed.



PARTS LIST – ETI 540

Resistors

R1	–	1 k	½ W	5%
R2,3	–	10 k	"	"
R4	–	47 k	"	"
R5	–	10 k	"	"
R6	–	1 k	"	"
R7-R9	–	1 M	"	"
R10	–	10 k	"	"
R11.12	–	100 k	"	"
R13	–	10 k	"	"
R14	–	100 k	"	"
R15	–	10 k	"	"

Capacitors

C1	–	0.1 µF 50 V disc ceramic
C2	–	0.1 µF polyester
C3	–	220 µF 16 V electro
C4,5	–	0.01 µF polyester
C6	–	10 µF 16 V electro
C7,8	–	0.01 µF polyester
C9	–	0.001 µF "
C10	–	0.01 µF "

Diodes

D1-D3	–	EM401 or similar
D4, 5	–	IN914 or similar
LED 1	–	RL4850 or similar

Transistors

Q1,Q2 – BC548 or similar

Integrated Circuits

IC1-IC3	–	4518
IC4,5	–	4017
IC6	–	4023
IC7,8	–	4011

Transformer 240 V/18 V CT PL18/5 VA
pc Board ETI 540
Relay, single pole 280 Ω coil 240 V 5A contact

Switches

SW1	double pole toggle switch
SW2	single pole 6 position rotary
SW3-6	single pole 10 position *
SW7	single pole toggle
PB1,2	single pole "make" push buttons

* C&K 321100000 is a 2 section Thumbwheel switch forming SW3 + 4 and SW5 + 6 (2 required)
Case plastic 196 x 113 x 60 mm
power cord, plug and clamp
3 pin power outlet socket

Universal timer

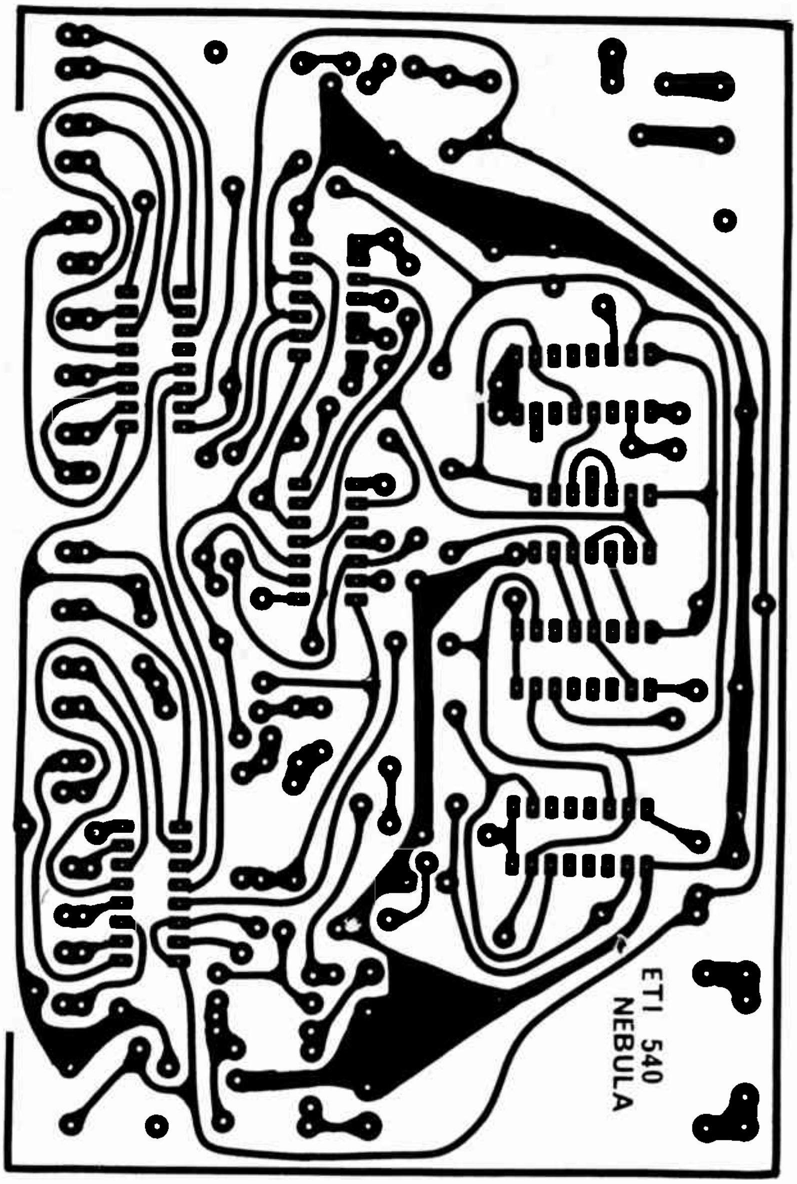


Fig. 4. Printed-Circuit board layout for the timer. Full size 153 x 100 mm.

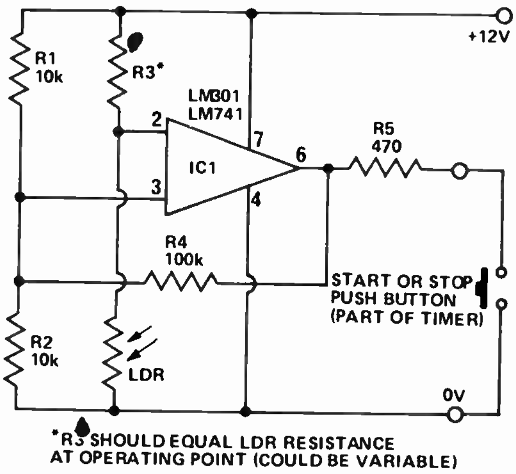
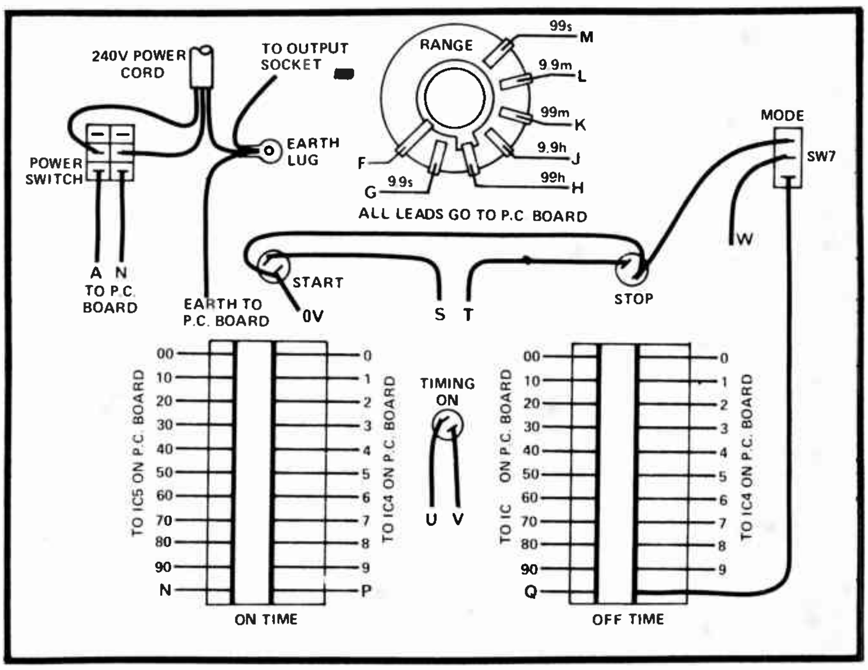


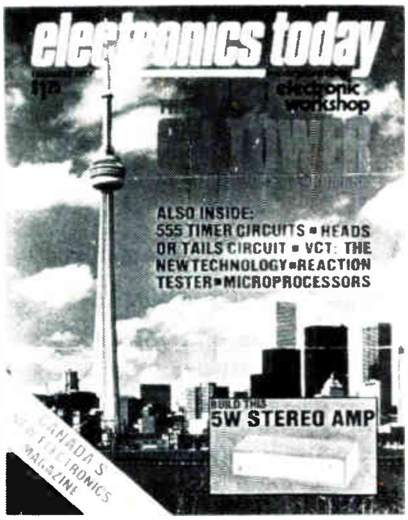
Fig. 6. For triggering timer from a change in light level this circuit will be found suitable.

Fig. 3. Interconnection diagram. ▶

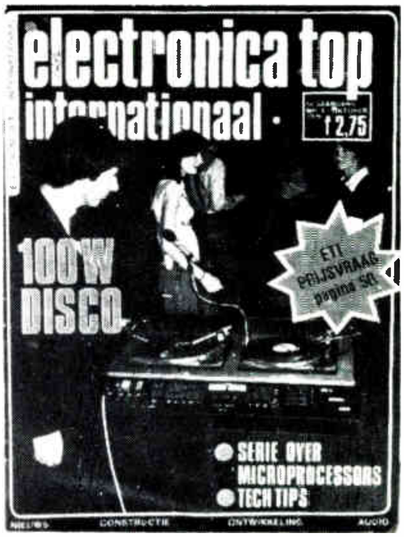


Scotchcal front panels for the Universal Timer are available from Electronics Today for $4.00 – plus a self-addressed stamped envelope at least 120 mm x 200 mm. The panels are 109 mm x 190 mm. Please make cheques payable to 'Scotchcal Offer', Electronics Today International, 15 Boundary Street, Rushcutters Bay, NSW, 2011.

subscribe!



Just a sample selection from the international editions of E.T.I. Australia, United Kingdom, Holland, France and Canada. Each is edited locally to give national coverage as well as international resources. We must be doing something right to become the fastest growing electronics magazine publishers in the world! Because of our tremendous growth in circulation, a lot of readers have difficulty getting hold of copies — why not take out a subscription and make sure!



Subscriptions to:

AUSTRALIA	ETI Magazine, Ryrie House, 15 Boundary Street, Rushcutters Bay, NSW, 2011.	*Payment in Aust. dollars*	**$14.00**
CANADA	ETI Magazine, Unit 6, 25 Overlea Boulevard, Toronto, Ontario, M4H 1B1.	*Payment in Canadian dollars*	**$12.00**
HOLLAND	Electronica Top Internationaal, P.O. Box 260, Emmen, Holland.	*Payment in Dutch currency*	**f27,50**
UNITED KINGDOM	ETI Magazine, 25-27 Oxford Street, London, W1R 2NT.	Overseas £7.00 *Payment in Sterling*	**£6.00**