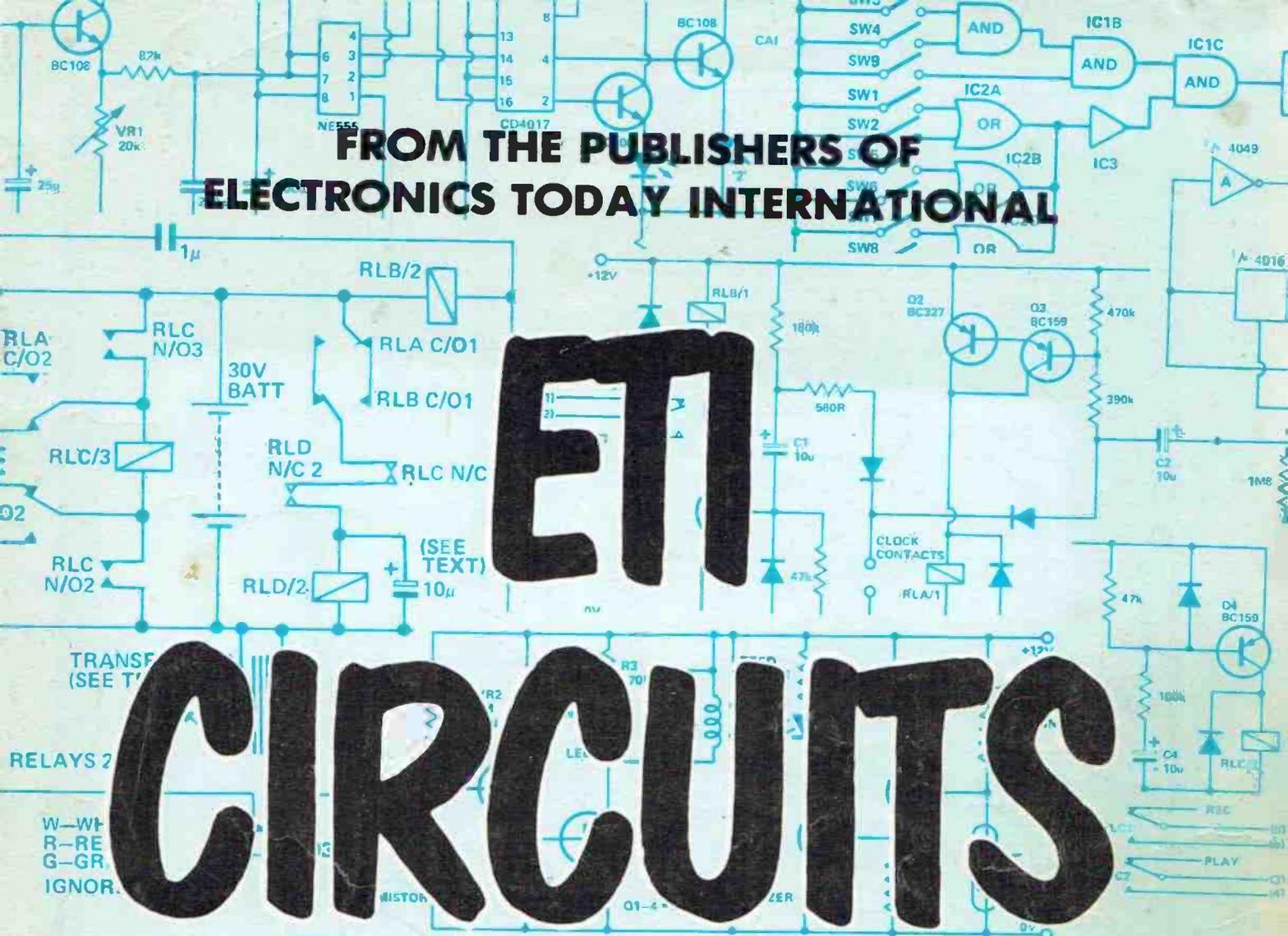
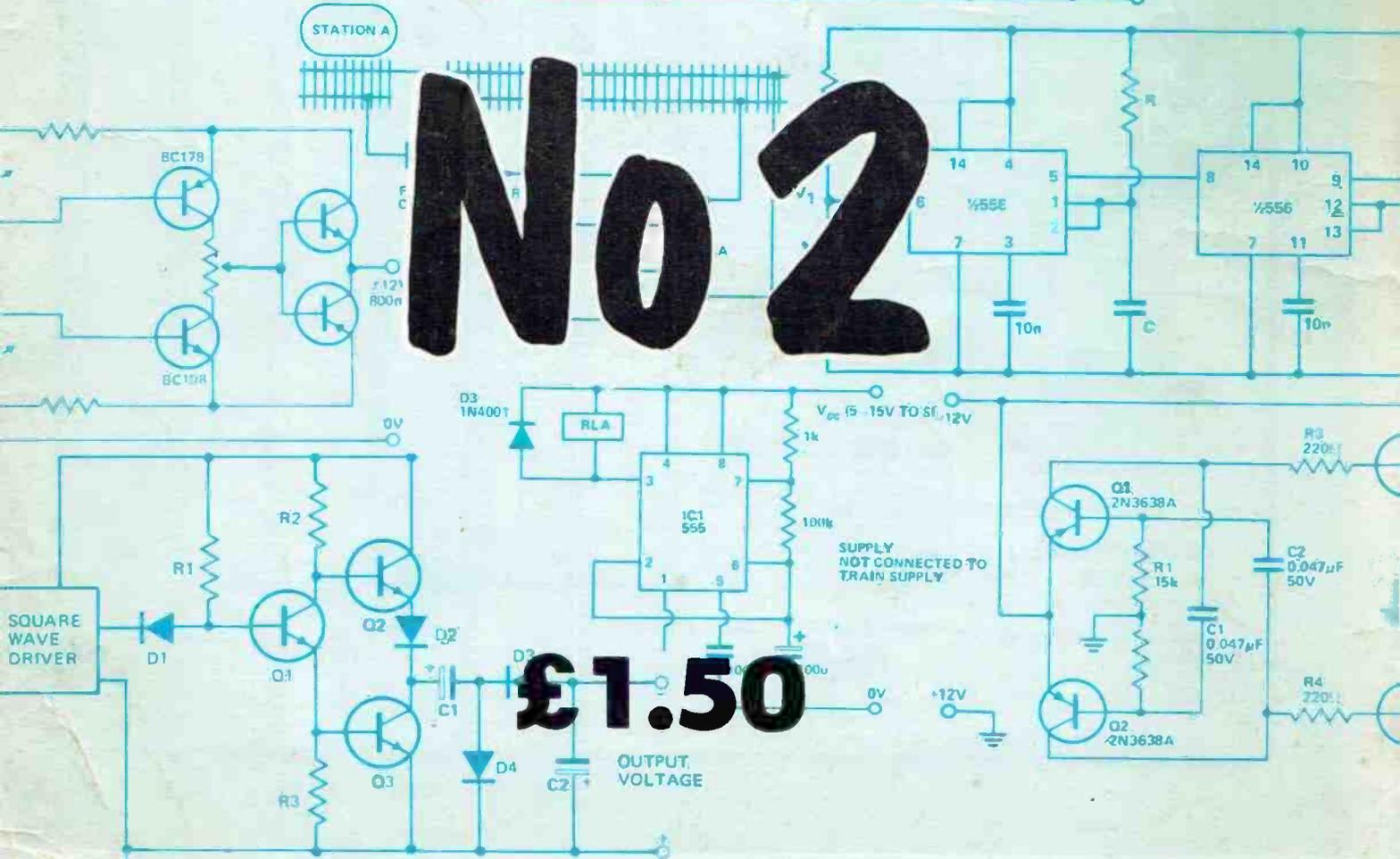


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# ETI CIRCUITS BOOK: No. 2

IDEAS AND DATA FOR EXPERIMENTERS FULL CONTENTS OVERLEAF

EDITED BY JIM PERRY

PUBLISHED MARCH 1978

For a long time we have published a section in Electronics Today International specifically for the experimenter-Tech Tips. This has always been one of the most popular sections of the magazine, due mainly to the tremendous variety of circuits that find their way into it.

But by the very nature of being a monthly feature, it becomes impossible (without some form of filing system) to remember particular circuits — or compare them with similar ones.

Our answer to this problem is this series of Circuits Books, this issue includes an enlarged data section and uses other sources besides Tech Tips — to produce an invaluable experimental source book. (We never were modest at ETI!).

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<b>AUSTRALIA</b> Collyn Rivers Publisher Les Bell Acting Editor	25-27 Oxford Street, London W1R 1RF. Telephone 01-434 1781/2. Telex 8811896
<b>HOLLAND</b> Anton Kriepsman Editor-in-chief	Halvor W. Moorshead Editor
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	Tim Salmon, Brenda Goodwin Advertising
	Mark Strathern (Manager), Tom Moloney Advertising
 <b>ABC</b>	<b>PUBLISHED BY</b> Modmags Ltd., 25-27 Oxford Street, London W1R 1RF
	<b>DISTRIBUTED BY</b> Argus Distribution Ltd (British Isles)
	Gordon & Gotch Ltd. (Overseas)
	<b>PRINTED BY</b> QB Limited, Colchester

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## Comprehensive Burglar Alarm

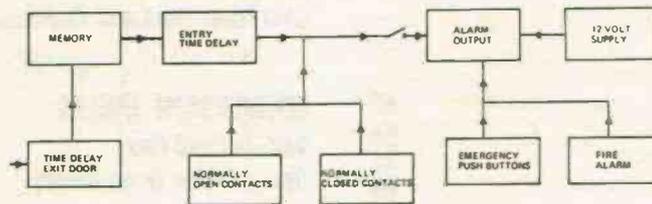
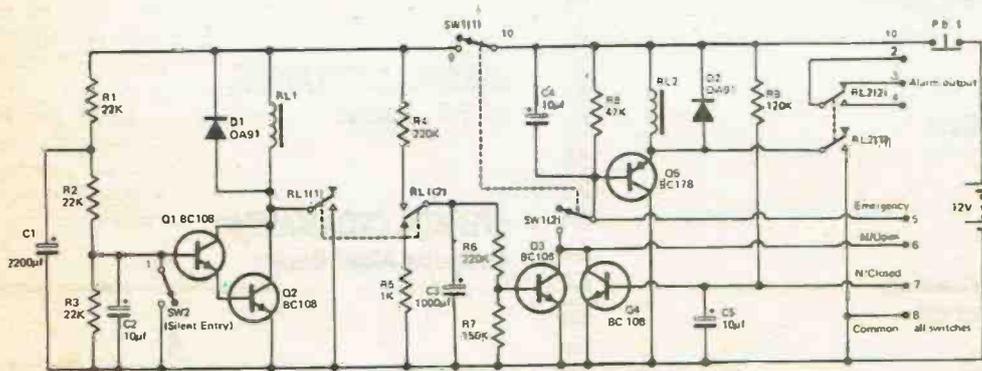
Examination of the block diagram will show the overall operation of this alarm circuit. SW1 is the main on/off switch and should be key operated. PBI is the reset control, and should be either hidden or wired in parallel with a third set of contacts on SW1 — this

is to prevent the alarm being disabled without a key.

SW1 is a normally closed contact on the exit door. When leaving the delay section disables RL1 for about 2 minutes and also on entry. If this exit is used and the alarm is not switched

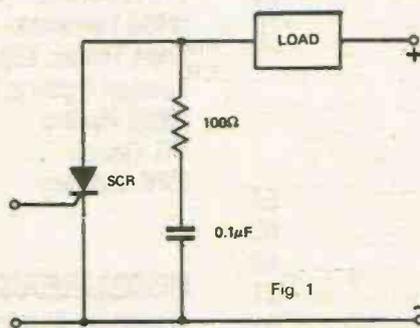
off before the 2 minutes are up — the alarm will sound.

The rest of the circuitry is quite straightforward. Q3 and Q4 are used to ground the base of Q5 if an alarm condition exists. Q5 operates RL2 which is used to sound a bell or siren externally.



## SCR Alarms

In some SCR circuits the SCR will tend to 'switch on' the moment the dc supply voltage is applied, at the SCR's gate. This is because the SCR is sensitive to the rate at which the supply voltage is applied, and if this rate of rise exceeds a certain level, switch-on will occur. The effect may be eliminated by connecting a series resistor/capacitor combination across the SCR. This is known as dv/dt suppression and its effect is to slow down the rate of voltage rise. A diode, connected in the same effective polarity as the SCR, may be paralleled across the resistor for maximum effectiveness. In most applications the values shown in Fig. 1 will prove effective.



False triggering can also be caused by transients induced into the gate circuits. This is a very common problem with a number of burglar alarms — even commercially made ones from manufacturers who should know better.

The most commonly used SCR burglar alarm circuit is basically that shown in Fig. 2. In this configuration, the gate of the SCR is connected to the positive rail via a 10k resistor, but an external loop interconnecting a number of normally closed trip switches, effectively clamps the gate at zero potential. However if any switch is opened, or if the external loop is cut, the SCR will immediately be triggered into conduction, thus energizing a series connected bell.

The problem with this circuit is that although the gate of the SCR appears to be held very firmly at zero potential by the external loop, transient energy induced into the external loop by electro-magnetic phenomena (caused by lightning, arc welders, fluorescent lighting starters etc) can reach quite

## Car Radio Protector

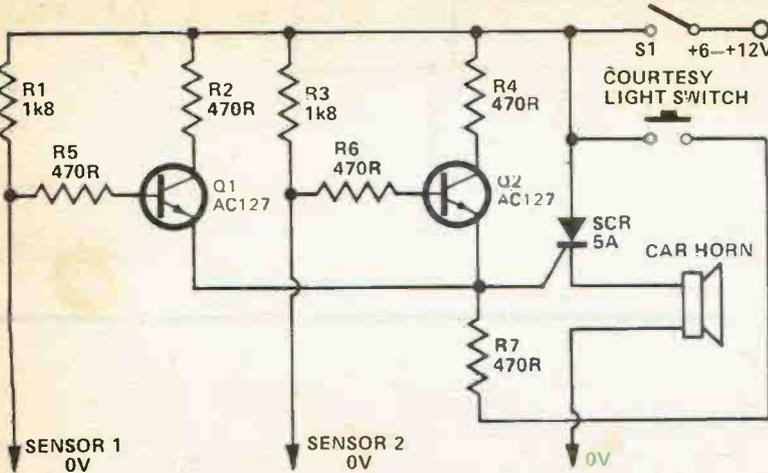
Many circuits have appeared for protecting radios and stereos in soft-top motor cars whose interiors are readily accessible to thieves. These circuits however, have the disadvantages of high parts count and expensive relays to switch on and latch the alarm.

The circuit operates as follows: sensor leads 1 and 2 are connected to the chassis of the equipment to be protected, therefore holding the bases of Q1 and Q2 at earth potential and thus switched off. If one of the sensor leads is broken, current flows to the base of the respective transistor and switches it on. This gates the SCR and

sounds the alarm. The self-latching characteristics of the SCR now make the transistor and its sensor lead inoperative and the alarm can only be stopped by switching off the concealed switch, S1.

To prevent thieves tampering with the wiring, a separate car horn and courtesy light switch (obtained from a breaker's yard) were fitted under the bonnet. If thieves cut the normal horn wires the alarm is unaffected; also any attempt to lift the bonnet to disconnect the battery will trigger it. The sensor leads are multistrand flexible cable with only one strand connected to the equipment, therefore easily broken while trying to remove it from the dashboard.

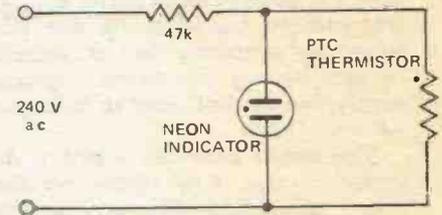
The transistors used are not critical and most NPN, general purpose transistors should suffice. The stand-by current is very low (typ. 13mA), and therefore is designed to be left switched on. The owner can never forget to switch on the alarm when leaving the vehicle. He must, however, switch off before lifting the bonnet.



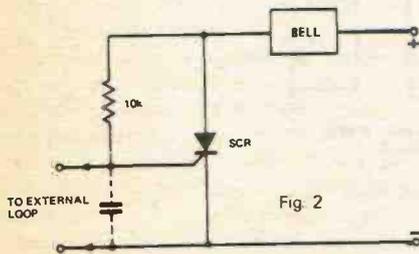
## Fire Alarm, Simple

A voltage divider made up of a resistor and positive temperature co-efficient thermistor has its incremental voltage fed to a neon indicator lamp. The thermistor is used as the temperature sensor. Its value is such that under normal ambient conditions the neon voltage across it is

below striking voltage for the neon. If a fire causes the thermistor to heat up, its resistance rises, and the neon ignites giving a visual alarm. The value of the PTC thermistor is selected to give the necessary resistance change to ignite the particular neon lamp used. An audible alarm can be activated by adding suitable electronic circuitry.



high voltage levels at the 'open' SCR end of the loop. And these levels are more than sufficient to trigger a sensitive SCR.



In some instances this type of false triggering can be overcome by connecting a 2uF capacitor between

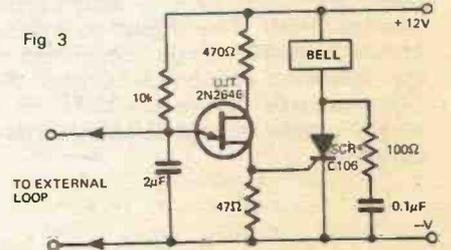
the SCR's gate and cathode but generally speaking it is bad practise to connect long 'aerials' directly to the gate circuit of an SCR.

A better solution is to use a UJT as a 'buffer stage' — as shown in Fig. 3. This will ensure that the gate circuit is totally immune from false triggering no matter how long the external circuit.

False triggering may also be caused by switching transients if long external leads are used in the anode or cathode circuit of the SCR. This sometimes occurs with burglar alarms and other control and warning systems if a bell (or other load) is located some distance away from the SCR.

This problem can almost invariably be overcome by using dv/dt

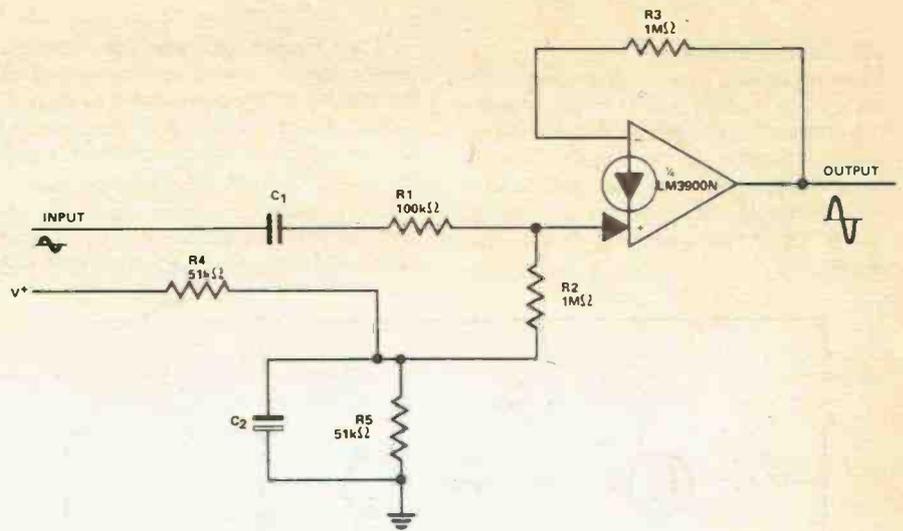
suppression (as shown in Fig. 1). In extreme cases it may be necessary to use a 5uF capacitor and a 5k series resistor, but values of 0.1 uF and 1k will generally suffice.



## Non-Inverting Amplifier

The circuit shows an amplifier which provides an output in phase with the input. The gain is equal to  $R_3/(R_1 + r_d)$  where  $r_d$  is the small signal impedance of the input diode. The value of  $r_d$  is equal to 0.026 divided by the current passing through  $R_2$  to the non-inverting input.

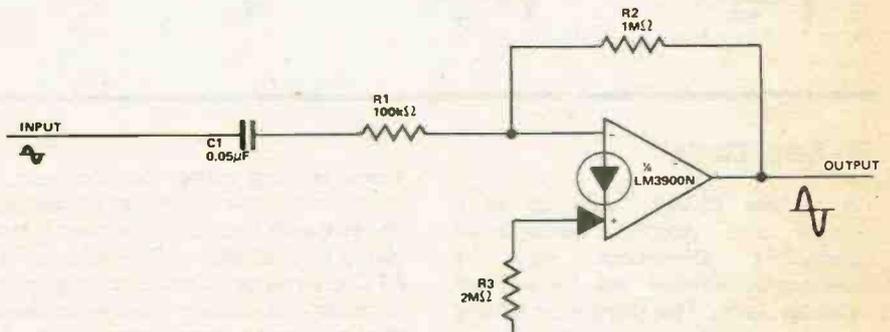
The capacitor values should be chosen so that the impedance of these components is considerably less than the circuit impedance at the points concerned.



## AC Amplifier, Simple

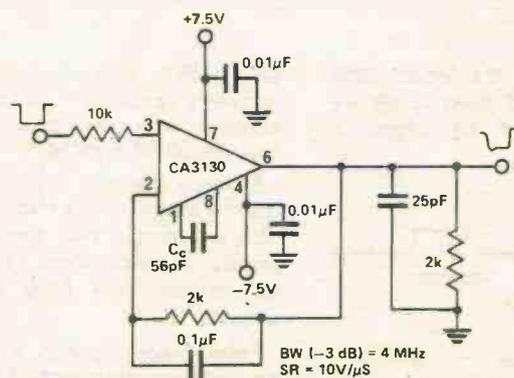
The gain is approximately equal to  $R_2/R_1$ , or 10 with the circuit values shown. The means potential at the output is half the supply voltage. The value of  $R_3$  should be twice that of  $R_2$ , since the current passing through either of these resistors is then the same. The positive supply and ground connections are not shown for simplicity, but  $R_3$  should be returned to the same positive supply line as that used to feed pin 14.

The circuit provides a phase inverted output. Any ripple on the power supply line will appear on the output at half amplitude.



## Voltage Follower

Operational amplifiers with very high input resistances, like the CA3130, are particularly suited to service as voltage followers. Figure shows the circuit of a classical voltage follower, using the CA3130 in a split-supply configuration. The digital-to-analog converter (DAC) circuit, described in the following section, illustrates the practical use of the CA3130 in a single-supply voltage-follower application.



## Flexible Response

Figure 1 gives the circuit of a high input-impedance amplifier with a nominal voltage gain of 48 and a bandwidth of from 10 Hz to at least 50 kHz. In the prototype the measured value of input impedance was  $10\text{M}\Omega$  at 1 kHz. This value will vary slightly with frequency and with the particular layout employed, but in any case is likely to be as high as will normally be required for most applications.

As an ac connection, via a capacitor, is provided at the non-inverting input there would be no dc return for bias current, at that input, if  $R_3$  were not present. The value of  $R_3$  is 47 k however, bootstrapping is used to raise the apparent value of  $R_3$  to the value of 10 megohm as quoted, in the following manner.

Due to the extremely high gain of the op. amp, and to the feedback between the output and the inverting input pin 2, there is very little difference in the signal levels at the +ve and -ve inputs, and, since  $C_1$  has a negligible reactance, there is similarly very little difference in signal voltage at either end of  $R_3$ . Accordingly, very little signal current can flow into  $R_3$  from the signal input, thus  $R_3$  appears, to the input signal, to be many times its actual value.

With the op. amp. arranged in the non-inverting configuration, the voltage gain is:

$$A_v = \frac{R_1 + R_2}{R_1} = 48$$

This amplifier set-up is most likely to be used in the design of a pre-amplifier for an oscilloscope or millivoltmeter, where the high value of input impedance is necessary in order to load the circuit under test as little as possible.

In audio applications, a 'tailored' frequency response is often called for; for example, the output of a tape replay head should be fed to a stage with a gain rising at 6dB per octave below about 2.5 kHz, and a flat response above that frequency. (The actual value of the break frequency depends on the tape speed and the particular replay characteristic employed). Such a response is readily arrived at by replacing  $R_2$  of Fig. 1 with the network shown in Fig. 2a.

At high frequencies  $C_5$  has a reactance low compared to  $R_6$  and hence it can be ignored. Thus the gain

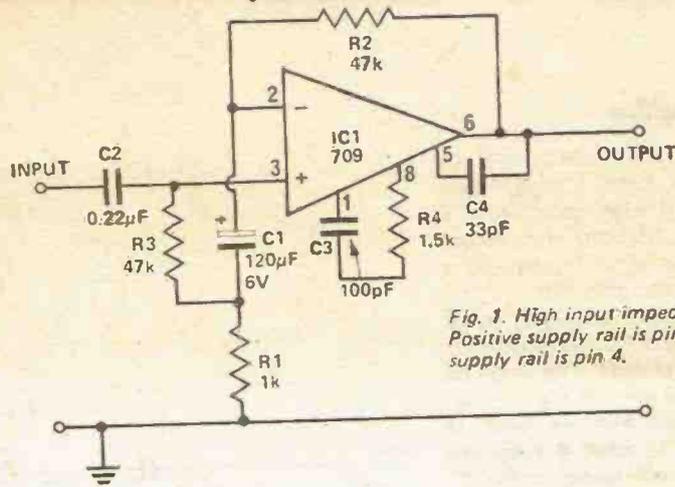


Fig. 1. High input impedance ac amplifier. Positive supply rail is pin 7 and negative supply rail is pin 4.

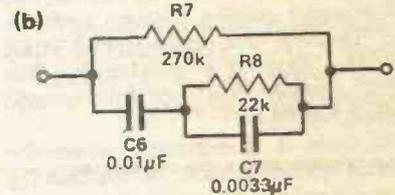
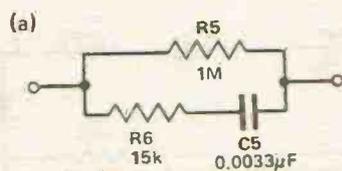
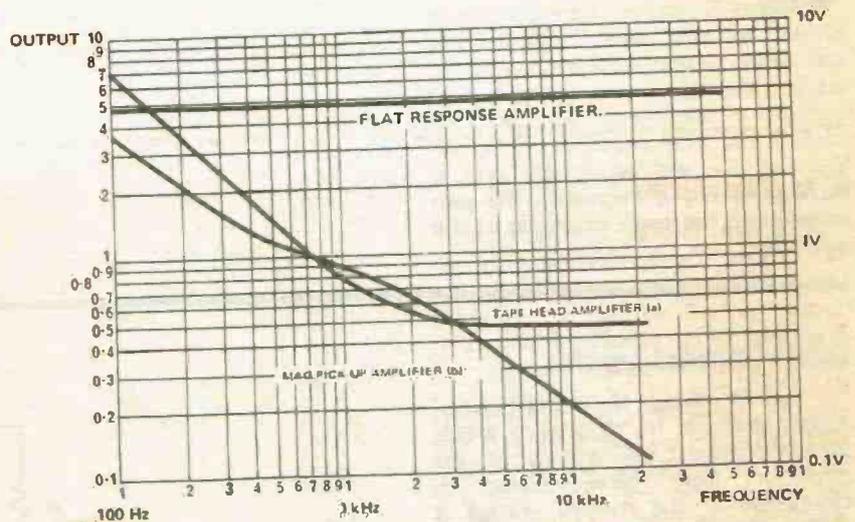


Fig. 2. Alternative feedback networks. (a) Tape head; (b) Magnetic pick-up.

is determined by  $R_6$  alone (although  $R_5$  is in parallel its value is large enough to be disregarded). As the frequency is lowered, the reactance of  $C_5$  rises and consequently the feedback is reduced, so giving the frequency response shown in Fig. 3a. Resistor  $R_5$  provides a dc connection for the negative input of the op. amp, and limits the gain at very low frequencies.

The voltage gain of this circuit at high frequencies is about 16 times; this will make the tape head output comparable to that from a magnetic pick-up. If more gain is called for, this is best done by increasing the value of

$R_6$  and reducing the value of  $C_5$  in proportion.

What if a response suitable for pre-amplification of the output of magnetic pick-up is required? In this case the network of Fig. 2b is a suitable replacement for  $R_2$  in the original circuit; the overall response of the stage is now as given in Fig. 3b.

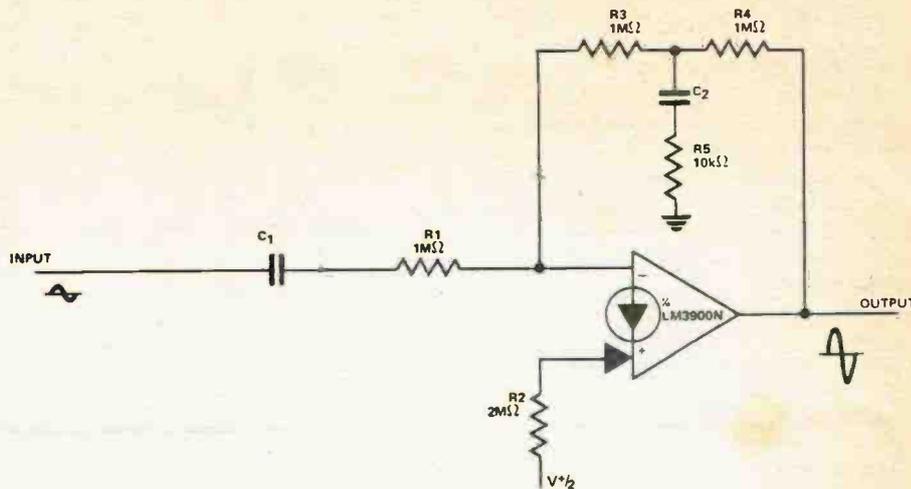
Similar reasoning to that given for the tape head amplifier applies here also — the gain rises at lower frequencies as  $C_6$  reactance becomes larger, falling at the higher frequencies as the reactance of  $C_6$  and  $C_7$  both fall. As before,  $R_7$  sets the low-frequency gain.

## Hi Z, Hi Gain Amplifier

This circuit has been designed so that it provides both a high input impedance and a high gain using a simple amplifier. With the component values shown, the input impedance is one megohm and the gain 100.

The voltage applied to  $R_2$  is made equal to the output voltage (which is half the supply voltage). The value of  $R_2$  is equal to the sum of  $R_3$  and  $R_4$ ; these resistors set the dc bias. If desired,  $R_2$  may be made 4 megohms and its lower end connected to the  $V^+$  supply.

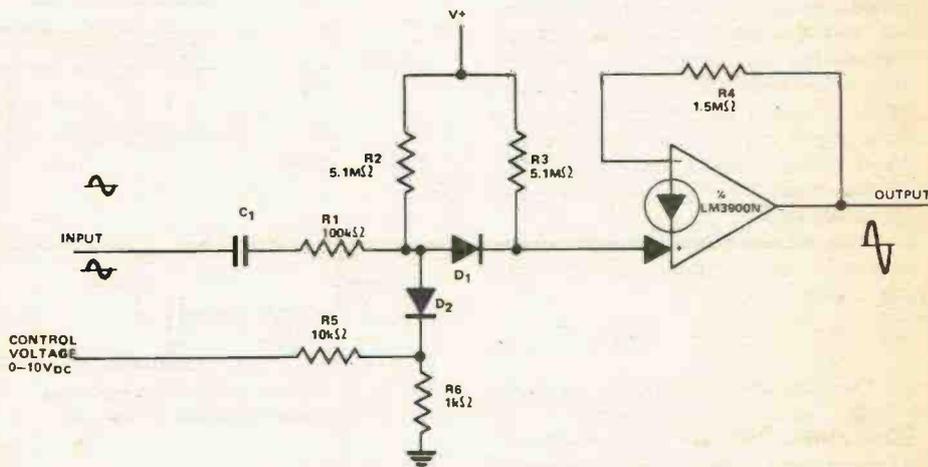
Resistors  $R_4$  and  $R_5$  form a potential divider so that only 1/100 of the alternating output voltage is developed across the  $C_2 - R_5$  circuit. This fraction of the output voltage is fed back to the inverting input via  $R_3$ . As  $R_3$  and  $R_1$  are equal, the gain is  $R_4/R_5$ . As  $R_5$  is decreased, the gain approaches the open loop gain of the amplifier.



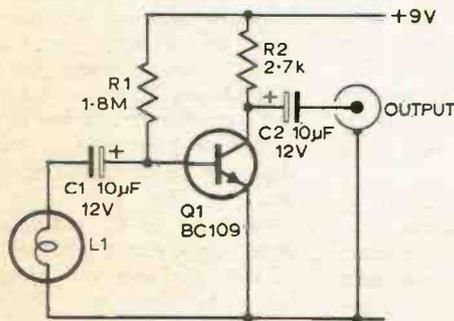
## Voltage Controlled Amplifier

A current flows from the positive supply through  $R_3$  to provide a bias which prevents the output of the amplifier from being driven to saturation as the control voltage is varied. When  $D_2$  is non-conducting, the currents passing through both  $R_2$  and  $R_3$  enter the non-inverting input and the gain is a maximum. This occurs when the control voltage approaches 10 V.

The gain is a minimum when the control voltage is zero. In this case  $D_2$  is conducting and only the current passing through  $R_3$  enters the non-inverting input of the amplifier.

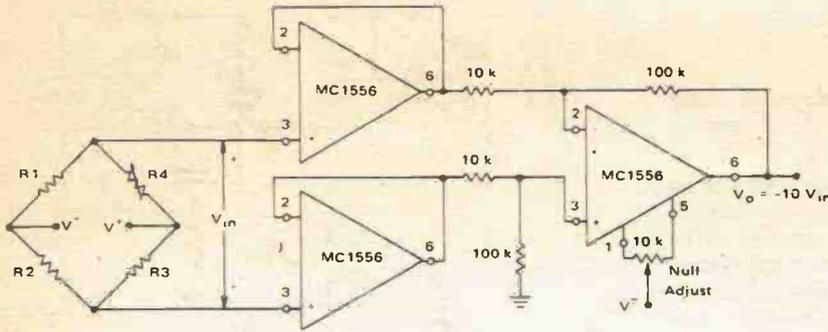


## Recording Pickup



It is often inconvenient to interfere with a circuit to take an audio tap off for recording etc. However by using a telephone pickup coil and placing this near the coil of almost any loudspeaker, excellent quality may be obtained with no direct electrical connection. The varying magnetic flux in the loudspeaker is induced directly into the coil. As the output may be low for some uses the very simple amplifier shown in the circuit will raise the level. This may not work well with some TV sets due to high frame pulses from the transformer which may cause a low frequency buzz though this depends on the proximity to the loudspeaker.

## High Impedance Bridge



The MC1556 operational amplifier may be used as a voltage follower in a bridge amplifier application. The high

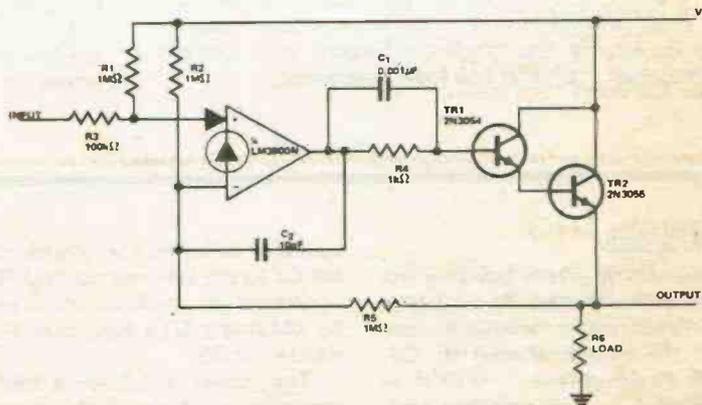
input impedance avoids loading effects on the bridge and transforms the impedance down to a level where a

third amplifier used in a differential mode can provide voltage gain, 10 in this case. The third amplifier employs the standard offset adjust circuit to provide nulling capability for the configuration.

Although the circuit is shown for complementary supply voltages, it lends itself well to operation from a single supply since the bridge can be operated just as well from the single supply. One must, however, provide for biasing the now-grounded 100 kΩ resistor to half the supply voltage using a simple resistive divider. Also, of course, the output is no longer referenced to ground, but to half the supply voltage.

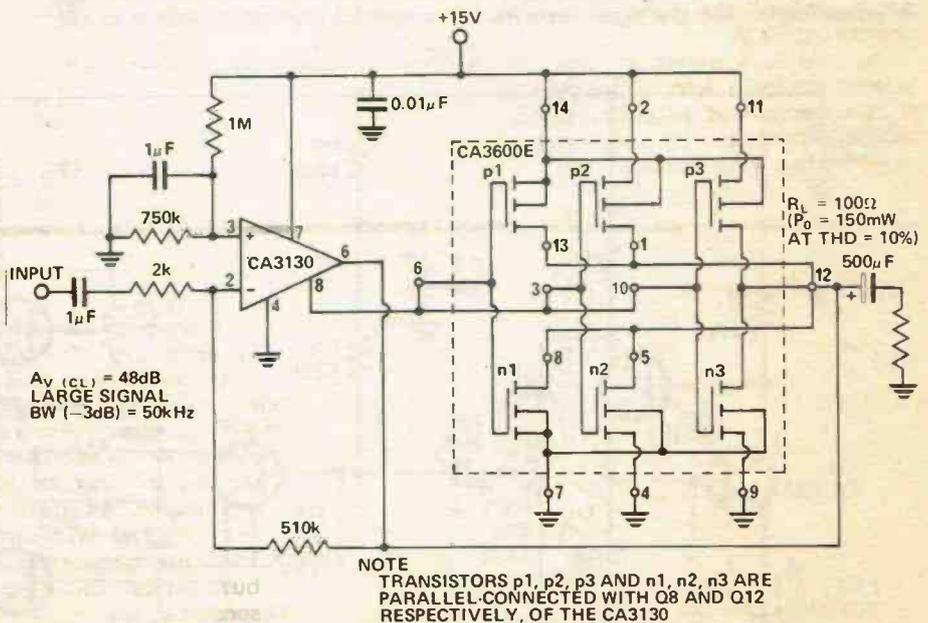
## Direct Coupled Power

In the circuit, the output from an LM3900N amplifier is fed to a Darlington pair of power transistors. This circuit can deliver over three amps into a suitable load when the transistors are correctly mounted on heat sinks.



## CMOS Power Booster

The current-sourcing and sinking capability of the CA3130 output stage is easily supplemented to provide power-boost capability. In the circuit three COS/MOS transistor pairs in a single CA3600E IC array are shown parallel connected with the output stage in the CA3130. In the Class A mode of CA3600E shown, a typical device consumes 20 mA of supply current at 15V operation. This arrangement boosts the current-handling capability of the CA3130 output stage by about 2.5.

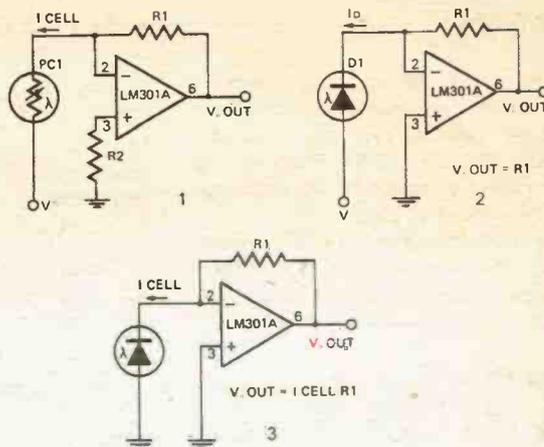


## Photocell Amplifiers

Amplifiers for photoconductive, photodiode and photovoltaic cells are shown in Figures 1, 2 and 3 respectively.

All photogenerators display some voltage dependence on both speed and linearity. It is obvious that the current through a photoconductive cell will not display strict proportionality to incident light if the cell terminal voltage is allowed to vary with cell conductance. Somewhat less obvious is the fact that photodiode leakage and photovoltaic cell internal losses are also functions of terminal voltage. The current-to-voltage converter neatly side-steps gross linearity problems by fixing a constant terminal voltage, zero in the case of photovoltaic cells and a fixed bias voltage in the case of photo-conductors or photodiodes.

Photodetector speed is optimized by operating into a fixed low load impedance. Currently available photovoltaic detectors show response times in the microsecond range at zero load impedance and photoconductors, even though slow, are materially faster at low load resistances.



The feedback resistance,  $R_1$ , is dependent on cell sensitivity and should be chosen for either maximum dynamic range or for a desired scale factor and should be chosen to minimize bias current error over the operating range.

## 12 Volt PA System

This circuit was originally built for use in a negative earth car. A miniature speaker, impedance immaterial, is connected in the emitter circuit of Q1, and acts as a microphone.

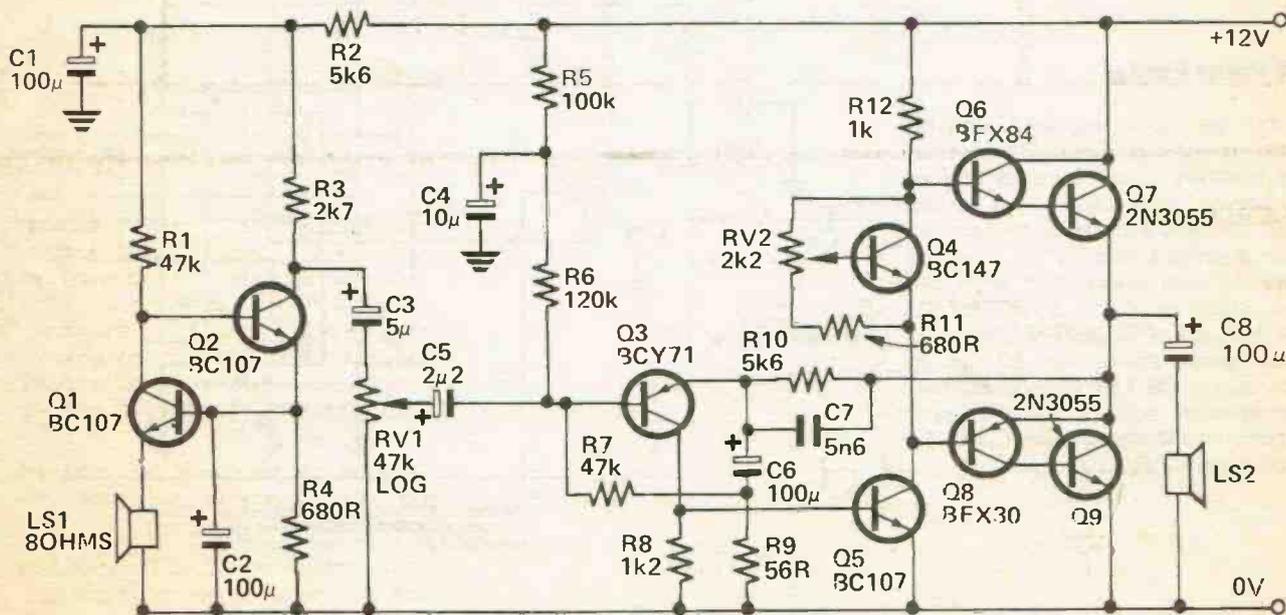
Q1 operates in the common base mode and a highly amplified signal appears at its collector. Q2, used in the common emitter mode, provides further amplification and the signal from its

collector is fed via the blocking capacitor C3 to the volume control RV1.

Overall de-stabilisation is provided by obtaining Q1's base bias from the emitter of Q2.

The power amplifier is fairly conventional and fitted with a heavy duty output stage to enable a pair of 3Ω P.A. type horns to be driven in parallel. Under these conditions 8W is available. A single 3Ω unit can be driven to 4W.

Since the unit is intended for the reproduction of speech a wide bandwidth is not required and C7 is incorporated to roll off the response above 5kHz. C6 also provides a rapid roll off in the bass region. Q7 and Q9 should be fitted to a 5" x 4" finned heatsink and the body of Q4 should be thermally in contact with this.



## Class A Amplifier

The main advantage of class A amplifiers is the absence of crossover distortion. Against this major advantage must be weighed the disadvantage of permanently hot heatsinks and large capacity power supplies.

The circuit shown here contains several novel features and will deliver 5W of pure class A sound into an 8 ohm load.

Q1 and Q2 form, with the associated components, a high quality voltage amplifier with overall ac and

dc feedback applied from the collector of Q2 via R6 to the emitter of Q1.

The output stage proper, consists of Q6 and Q7 connected as an emitter follower darlington pair. These transistors are driven by IC1, a 741 op amp, and are included in the latter's feedback loop.

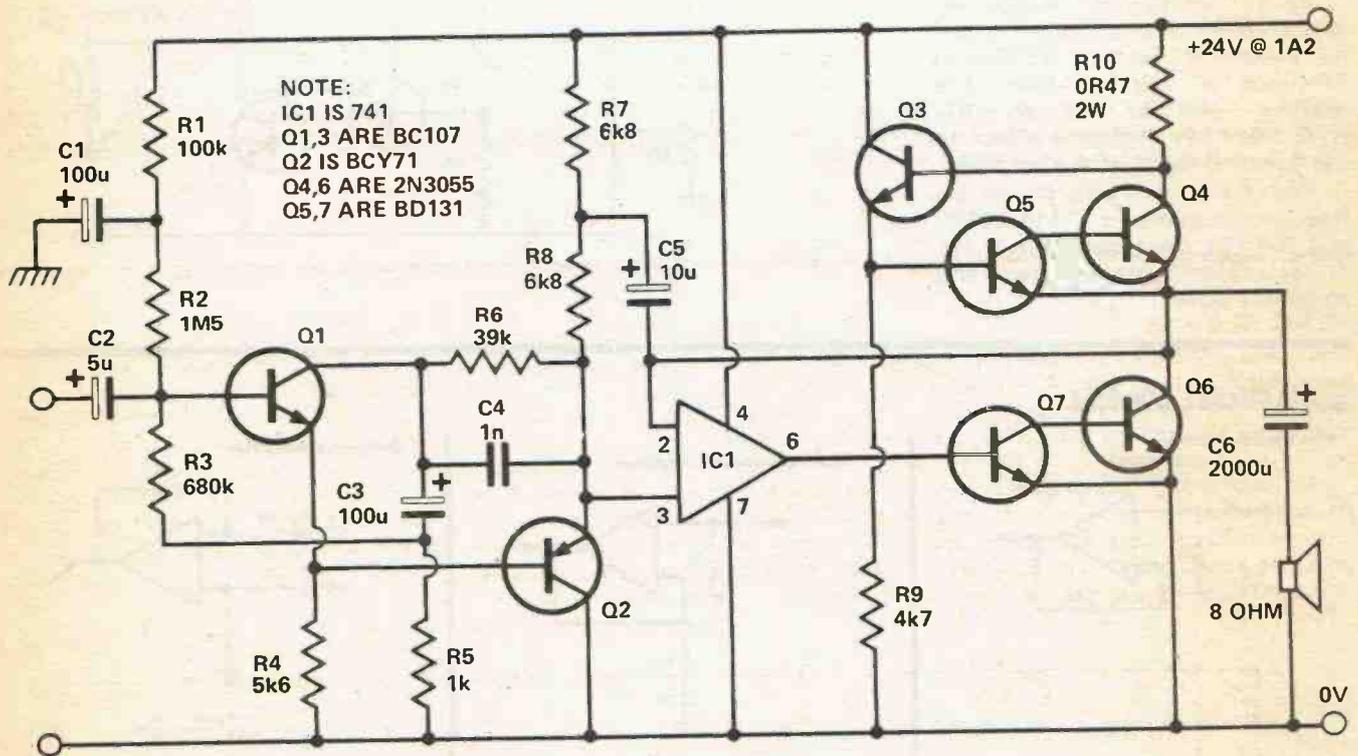
These three form a near perfect output stage with an input impedance of several megohms and a bandwidth extending from dc to over 100KHz.

Quiescent current is provided by the constant current source Q3, Q4, Q5, R9 and R10. The use of a

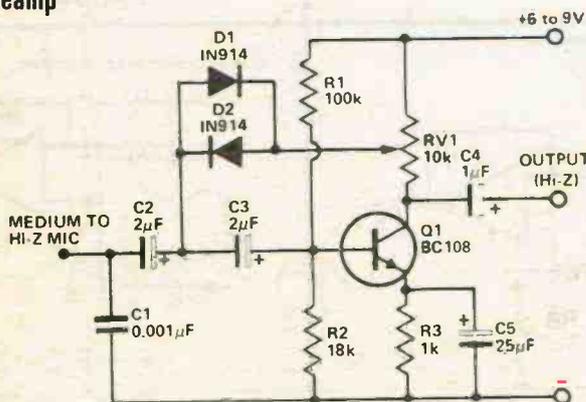
constant current source here effectively isolates the output from line variations and ripple.

With the components shown, the circuit has a bandwidth of 10Hz - 30KHz -3db, a distortion of less than 0.1% before the onset of clipping, an input impedance of 1.5MΩ and a sensitivity of 180mV for full output.

Transistors Q4 to Q7 must be mounted on an adequate heatsink, a 5" by 4" finned type is suitable, but must be mounted vertically and in such a position as to allow ample ventilation.



## Clipper Preamp



Maintaining a high average modulation level for mobile communications transmitters considerably improves the effectiveness of a transmitter, especially under difficult conditions.

This circuit provides a small amount of preamplification as well as variable clipping level (preset).

The two diodes should be a matched pair or clipping will not be symmetrical.

It is possible to mount the complete unit in many styles of hand-held microphone cases.

# AMPLIFIERS & PREAMPLIFIERS

## Headphone Amplifier

The circuit will deliver full 'orchestral' levels to four pairs of stereo headphones connected in parallel across the output.

Input signals are coupled to the non inverting input of a 741 op amp via the volume control RV1.

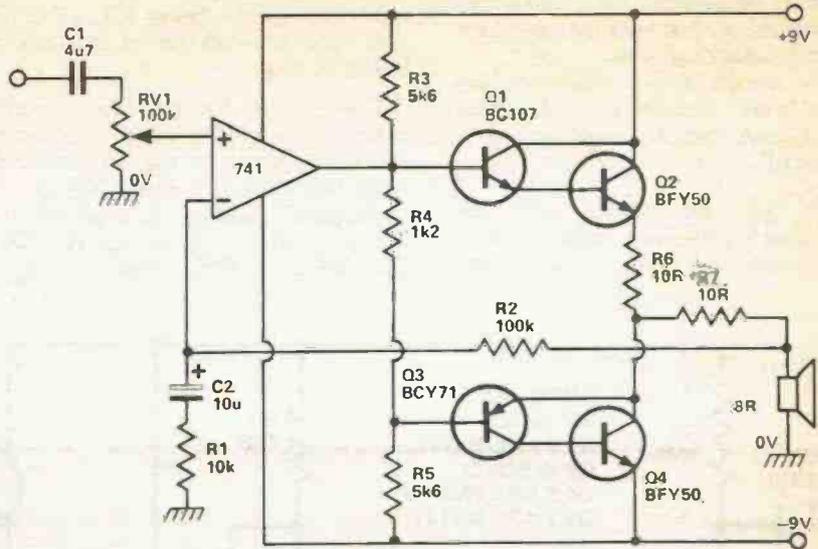
This IC is used to drive a quasi-complementary output stage consisting of Q1-4.

Quiescent current in the output transistors is provided by the voltage drop across R7 and local feedback provided by R6 in Q2's emitter circuit.

R6 is included to render the whole amplifier short circuit proof (to protect Q2 and Q4). Overall feedback is applied from the 'earthy' end of R6 so this component has negligible effect on the damping factor of the amplifier.

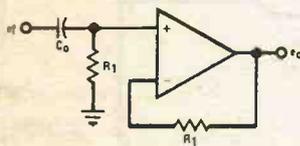
With the components shown the frequency response is -3dB at 4Hz and 100kHz, distortion below 0.1% at 1kHz (50mW out, 8Ω load), and sensitivity 60mV.

ONE CHANNEL ONLY SHOWN



## Op-Amp Circuits, Standard

Non-Inverting Buffer



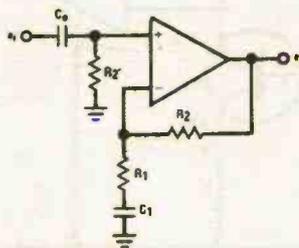
$$A_v = 1$$

$$R_{in} = R_1$$

$$f_o = \frac{1}{2\pi R_1 C_0}$$

Definitions  
 $A_v$  = Closed loop AC Gain  
 $f_o$  = Low frequency -3dB corner  
 $R_{in}$  = Input Impedance

Non-Inverting AC Amplifier

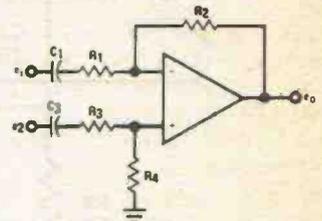


$$A_v = 1 + \frac{R_2}{R_1}$$

$$R_{in} = R_2$$

$$f_o = \frac{1}{2\pi R_2 C_0} = \frac{1}{2\pi R_1 C_1}$$

Difference Amplifier



$$e_o = \left( \frac{R_1 + R_2}{R_3 + R_4} \right) \frac{R_4}{R_1} e_2 - \frac{R_2}{R_1} e_1$$

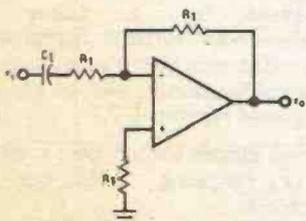
IF  $R_1 = R_3$  AND  $R_2 = R_4$  THEN

$$e_o = \frac{R_2}{R_1} (e_2 - e_1)$$

$$f_o = \frac{1}{2\pi R_1 C_1} = \frac{1}{2\pi (R_3 + R_4) C_3}$$

$R_2 = R_4$  FOR MINIMAL OFFSET ERROR

Inverting Buffer

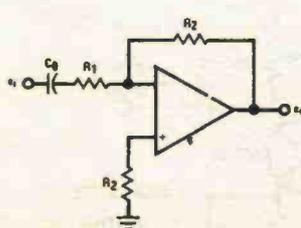


$$A_v = -1$$

$$R_{in} = R_1$$

$$f_o = \frac{1}{2\pi R_1 C_1}$$

Inverting AC Amplifier

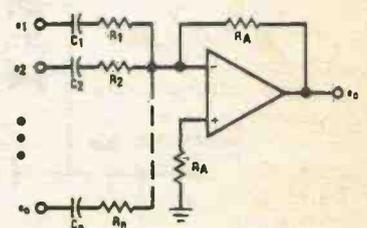


$$A_v = -\frac{R_2}{R_1}$$

$$R_{in} = R_1$$

$$f_o = \frac{1}{2\pi R_1 C_0}$$

Inverting Summing Amplifier



$$e_o = -R_A \left( \frac{e_1}{R_1} + \frac{e_2}{R_2} + \dots + \frac{e_n}{R_n} \right)$$

IF  $R_1 = R_2 = \dots = R_n$  THEN

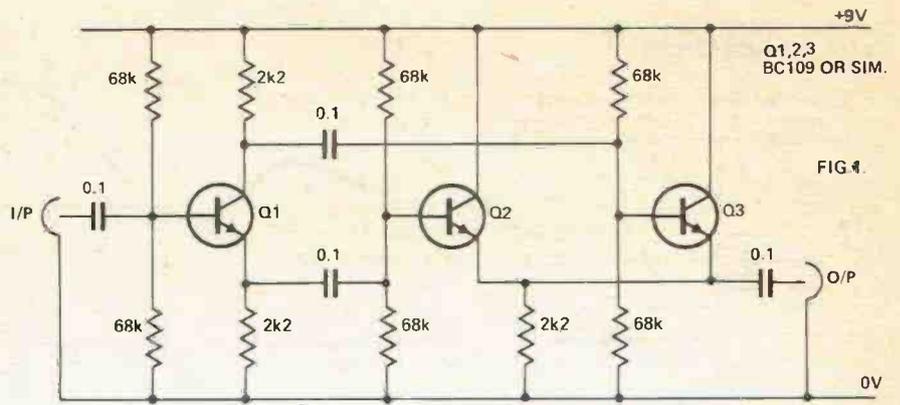
$$e_o = -\frac{R_A}{R_1} (e_1 + e_2 + \dots + e_n)$$



# SIGNAL PROCESSORS

## Frequency Doubler

This is a simple three transistor circuit to raise an audio frequency by a factor of two i.e., one octave. Q1 is connected as a phase splitter with anti-phase signals appearing at its collector and emitter. These signals are fed to two emitter followers Q2 and Q3, which have a common emitter resistor, and thus add the two anti-phase signals. A degree of distortion is inevitable as shown in Fig. 2, but is acceptable for speech and soloists and produces a sound similar to the Chipmunks or Pinky and Perky.



Q1,2,3  
BC109 OR SIM.

FIG. 1.

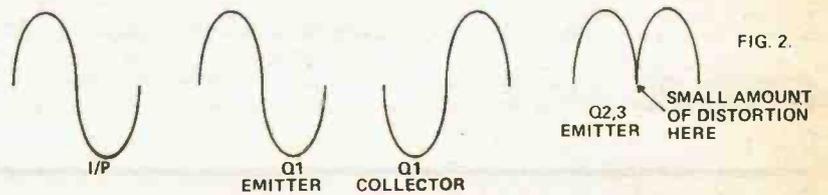
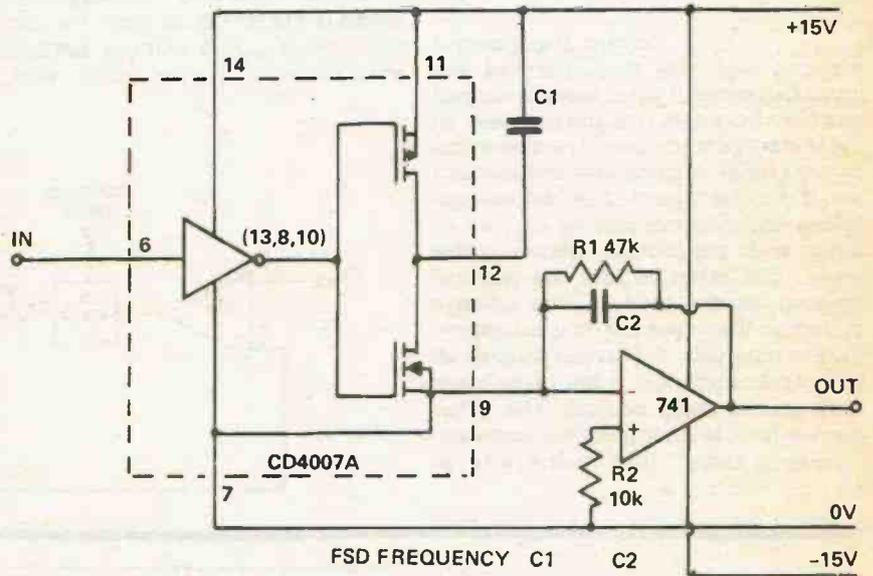


FIG. 2.

## Frequency to Voltage Converter

Figure shows a linear frequency to voltage converter which works by charging a capacitor up once for every input cycle, the charge to do so being passed by a MOSFET into a summing amplifier. The component values given are based on an approximate five volt output for the given frequency. The resistor R1 should be made a 100kΩ preset if it is required to set a range exactly. The capacitor C2 "smooths" the output and need not be changed from 10μF if fast-response on the upper ranges is not needed. The linearity achieved on the top range will depend on the particular "741" used and if reliable operation is required a higher speed op-amp should be used.



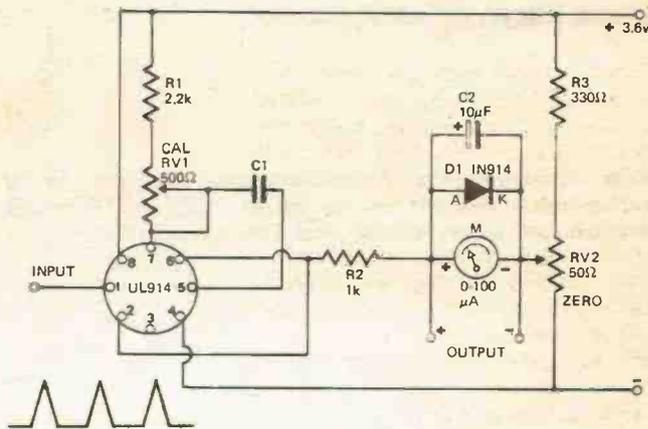
FSD FREQUENCY	C1	C2
10Hz	1μF	10μF
100 Hz	100nF	1μF
1kHz	10nF	100nF
10kHz	1nF	10nF
100kHz	100pF	1nF

## Frequency Meter, Analogue

This circuit may be used as a pulse counter, tachometer, or if preceded by a Schmitt trigger, an analogue frequency meter.

Output linearity better than 2% can be obtained with duty cycles of less than 30%. The meter is selected so that it reads full scale when the duty cycle of the uL 914 is 30%. The choice of C1 and the meter sensitivity determines the range of measurement.

Potentiometer RV1 is used to calibrate the meter to full scale deflection, and resistor R2 counteracts the slight zero offset of the saturated IC. Diode D1 protects the meter.

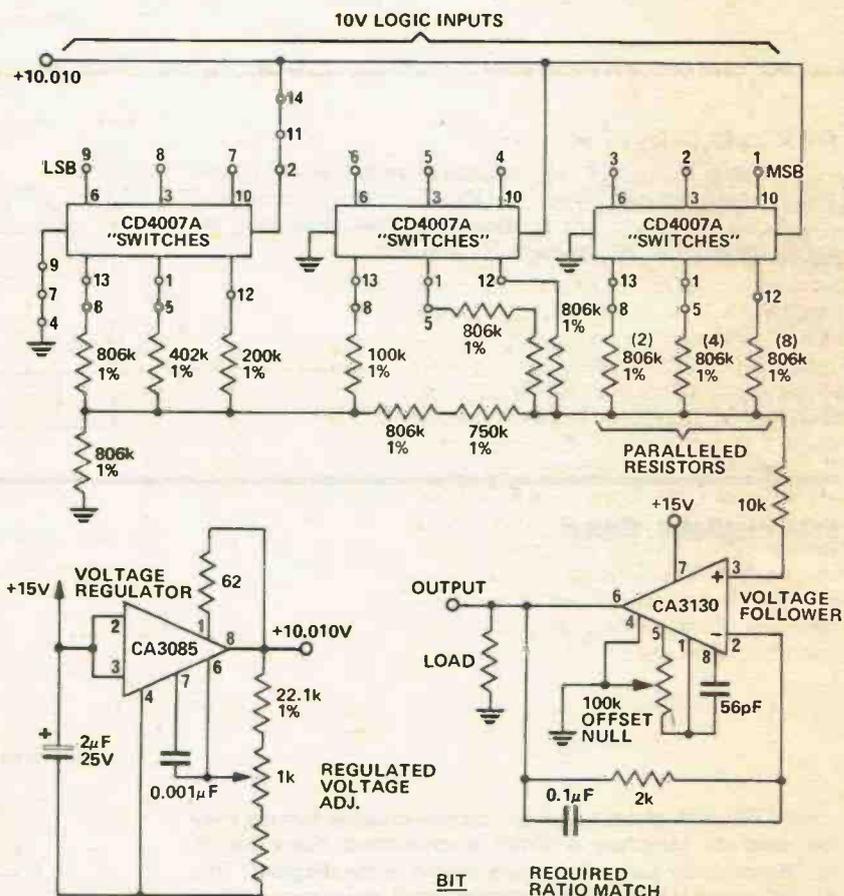


## Digital to Analogue Converter

The circuit of a 9-bit Digital to Analogue Converter (DAC) is shown. This system combines the concepts of multiple-switch COS/MOS IC's, a low-cost ladder network of discrete metal-oxide film resistors, a CA3130 op-amp connected as a follower, and an inexpensive monolithic regulator in a simple single power-supply arrangement. An additional feature of the DAC is that it is readily interfaced with COS/MOS input logic, e.g. 10-volt logic levels are used in the circuit.

The circuit uses an R/2R voltage-ladder network, with the output potential obtained directly by terminating the ladder arms at either the positive or the negative power-supply terminal. Each CD4007A contains three "inverters", each "inverter" functioning as a single-pole double-throw switch to terminate an arm of the R/2R network at either the positive or negative power-supply terminal. The resistor ladder is an assembly of one per cent tolerance metal-oxide film resistors. The five arms requiring the highest accuracy are assembled with series and parallel combinations of 806,000-ohm resistors from the same manufacturing lot.

A single 15-volt supply provides a positive bus for the CA3130 follower amplifier and feeds the CA3085 voltage regulator. A "scale-adjust" function is provided by the regulator output control, set to a nominal 10-volt level in this system. The line-voltage regulation (approximately 0.2%) permits a 9-bit accuracy to be maintained with variations of several volts in the supply. The flexibility afforded by the COS/MOS building blocks simplifies the design of DAC systems tailored to particular needs.



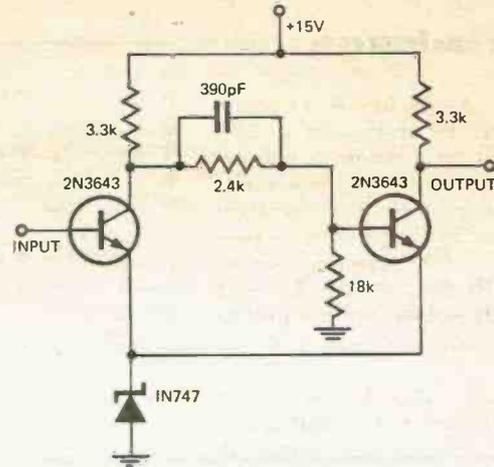
BIT	REQUIRED RATIO MATCH
1	STANDARD
2	±0.1%
3	±0.2%
4	±0.4%
5	±0.8%
6-9	±1% ABS

ALL RESISTORS IN OHMS.

# SIGNAL PROCESSORS

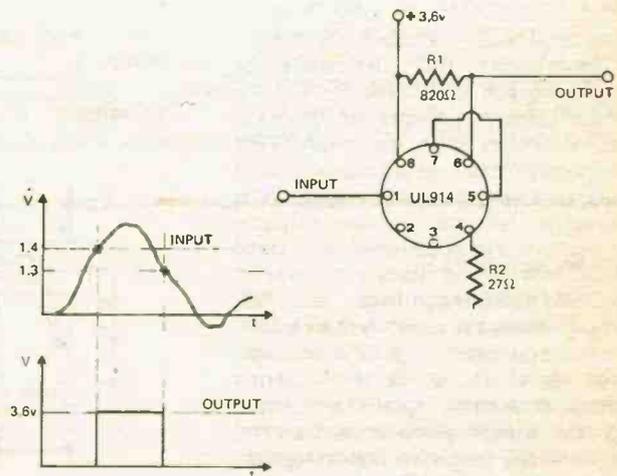
## Schmitt Trigger, Without Hysteresis

By replacing the common-emitter resistor in a conventional Schmitt by a zener diode, the hysteresis normally associated with these circuits is eliminated.



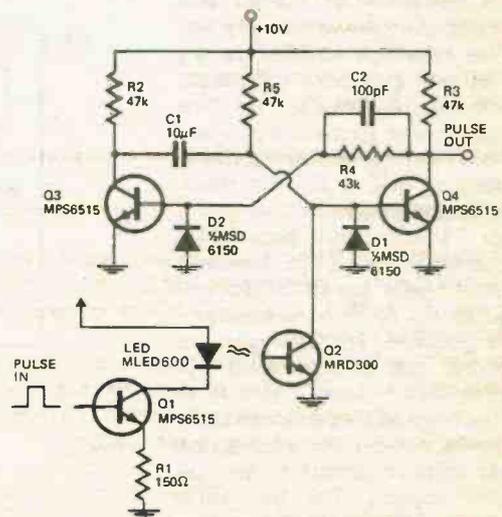
## Schmitt Trigger, Simple

One cheap IC, uL914, can be used as an extremely simple and effective Schmitt trigger suitable for many applications. Hysteresis of the circuit is about 0.1 Volt. This may be varied by altering the values of R1 and R2.



## Pulse Lengthener, Optical

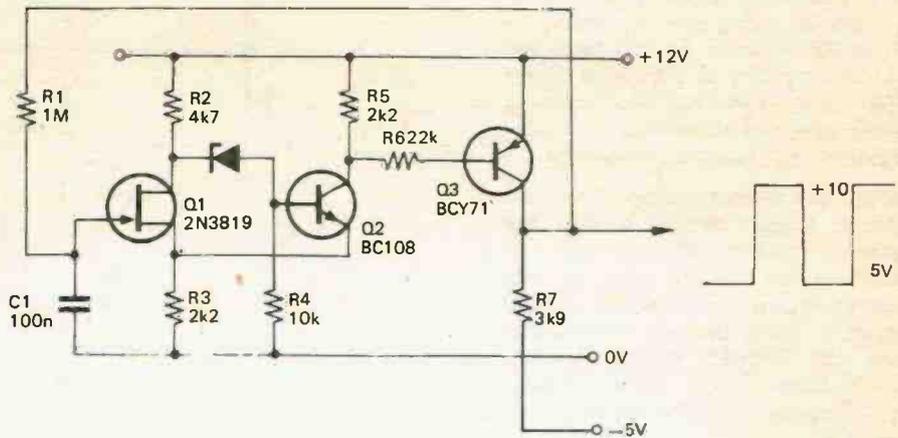
An LED and phototransistor optical-coupler system may be used to lengthen a three microsecond pulse to 55 milliseconds by using the values shown in the diagram. This allows complete isolation between input pulse circuits and the lengthener to be obtained.



## Square Wave, Low Frequency

A drawback of low frequency oscillators using bipolar transistors or TTL logic is that the timing capacitor usually has to be a high value electrolytic. Using a field effect transistor at the input of a schmitt trigger, means a low value capacitor can be employed. The trigger by Q1 and Q2 has a hysteresis of approximately 3V. This is controlled by the 3V zener.

With C1 uncharged Q1 is off and Q2 is forward biased. The voltage at the source of Q1 is approximately +4V. Q2 conducts, thus turning on TR3. The output is therefore at +10V. C1 then charges via R1 and the gate voltage of Q1 goes positive. When the gate voltage is sufficiently positive Q1 conducts, turning off Q2. The positive feedback from the emitter of Q2 to the source of Q1 ensures a rapid switch off. Q3 also



switches off and the output goes to -5V. Capacitor C1 now discharge towards -5V, but when the voltage across C1 falls by approximately 3V, Q1 ceases to conduct, turning on Q2.

The collector load of Q3 is connected to a negative supply giving a

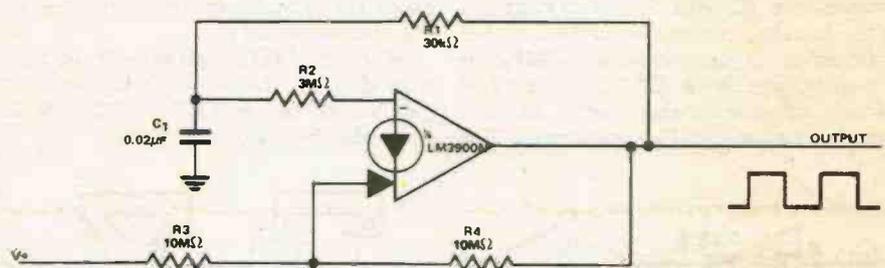
50% duty cycle. (The circuit still oscillates if R7 is connected to 0V but the duty cycle will change, the output remaining at 0V for a longer period than at +10V).

With the components as shown the frequency of the output is approximately 0.025Hz.

## Square Wave Generator

The multiple amplifiers in the LM3900N device are very suitable for use in waveform generators at frequencies of up to about 10 kHz. Voltage controlled oscillators (the frequency of which is dependent on an input voltage) can also be designed using the device.

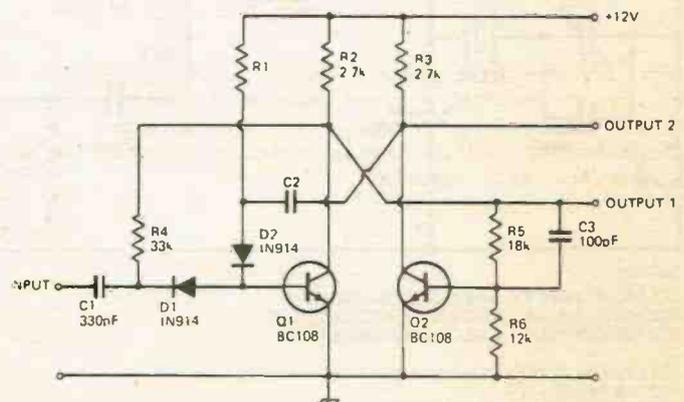
A simple square wave generator is shown. The capacitor C<sub>1</sub> alternately charges and discharges between voltage limits which are set by R<sub>2</sub>, R<sub>3</sub> and R<sub>4</sub>. The circuit is basically of the Schmitt trigger type, the voltages at which triggering occurs being approximately  $V^+/3$  and  $2V^+/3$ .



## Monostable Multivibrator

The time constant T of this circuit is equal to  $0.7 R_1 C_2$ . Where T is in seconds, R<sub>1</sub> in ohms and C<sub>2</sub> in farads. For example when R<sub>1</sub> = 10 k and C<sub>2</sub> = 100 microfarads the time constant will be one second.

Capacitor C<sub>2</sub> may be selected over wide a range and R<sub>1</sub> may be a potentiometer 100 k maximum. Outputs 1 and 2 provide pulses of opposite polarity but the rise time of output 2 is long due to the charging current of C<sub>2</sub>.



# SIGNAL GENERATORS

## Triangular Waveform Generator

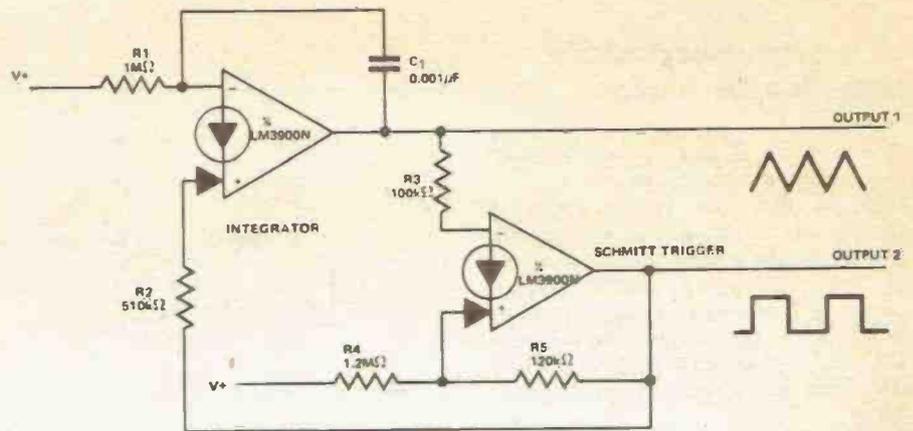
A triangular waveform generator can be made by using one amplifier of a LM3900N device as an integrator and another amplifier as a Schmitt trigger circuit. A suitable circuit is shown, it has the unusual advantage that only the one power supply is required.

When the output voltage from the Schmitt trigger circuit is low, the current flowing through  $R_2$  is integrated by  $C_1$  to produce the negative slope of the triangular wave at output 1. When the output 2 voltage from the Schmitt trigger is high, current flows through  $R_2$  to produce the rising part of the waveform at output 1.

The output waveform will have good symmetry if  $R_1 = 2R_2$ . The output frequency is given by the equation:

$$f = \frac{V^+ - V_{BE}}{2R_1C_1V}$$

where  $R_1 = 2R_2$ ,  $V_{BE}$  is the steady voltage at the inverting input (0.5 V) and  $V$  is the difference between the tripping points of the Schmitt trigger.



## Marker Generator

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.

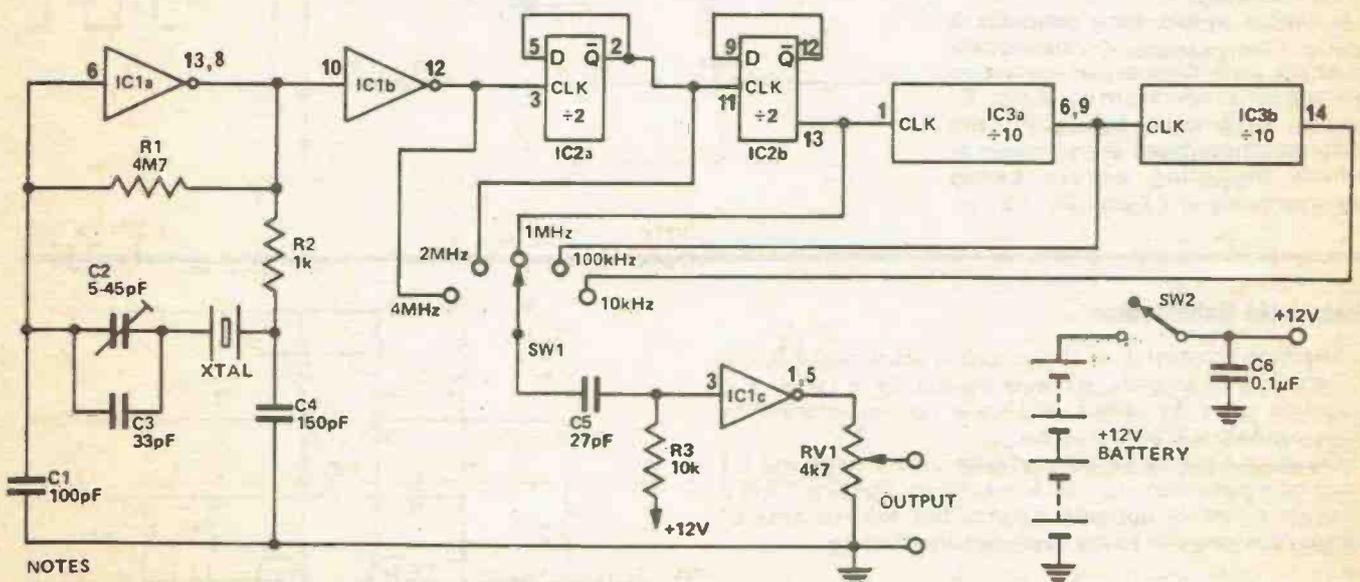
The oscillator is IC1a in which  $R_1$  biases the IC into linear operation. The crystal determines the basic frequency of operation at 4 MHz in conjunction with  $C_1$ , 2, 3 and 4 which appear to the crystal as one parallel capacitor. The capacitor  $C_2$  is

used to tune the oscillator exactly to frequency as explained in the text. The resistor  $R_2$  adds extra phase shift but also reduces the gain. Thus if the oscillator is slow in starting reducing  $R_2$  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  $C_5$  and  $R_3$  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.



### 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

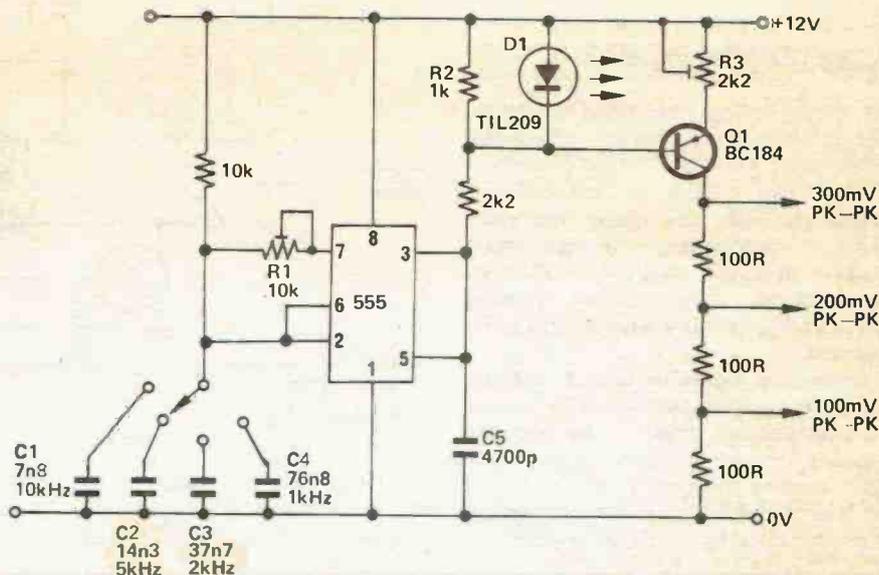
IC1 PINS 14, 2 AND 11 ARE +12V  
IC2 PIN 14, IS +12V  
IC3 PINS 2,10 AND 16 ARE +12V

## Voltage and Frequency Calibrator

This circuit provides simultaneous voltage and frequency calibrations by generation of a precision squarewave.

The 555 timer IC is used in a slightly unusual configuration, having the advantage that an exact 50:50 mark/space ratio may be attained by trimming R1. The frequency of oscillation may be set between 10 kHz and 1 kHz by switching timing capacitors C1-4. C5 decouples the internal reference potential-divider of the 555 from supply transients.

The squarewave output from pin 3 of the IC, while stable in frequency, is not stable in peak-to-peak voltage as this depends on the supply voltage. This is used to switch on and off a temperature-compensated constant-current source Q1. R2 ensures that the current-source turns off completely when pin 3 goes high. The current-source output, trimmed by R3 to be exactly 1 mA, drives a resistor ladder network so that a series of precise squarewave voltages are generated. The advantage of current drive rather



than voltage drive for this sort of network is that calibration is much easier. A simple ladder network is shown by way of example, and more complex ones may simply be constructed to give a wider variety of output voltages.

The non-standard component values used were obtained by paralleling standard values. For the timing capacitors several in parallel had to be used, and only the resultant value is shown on the diagram.

## VCO, Simple

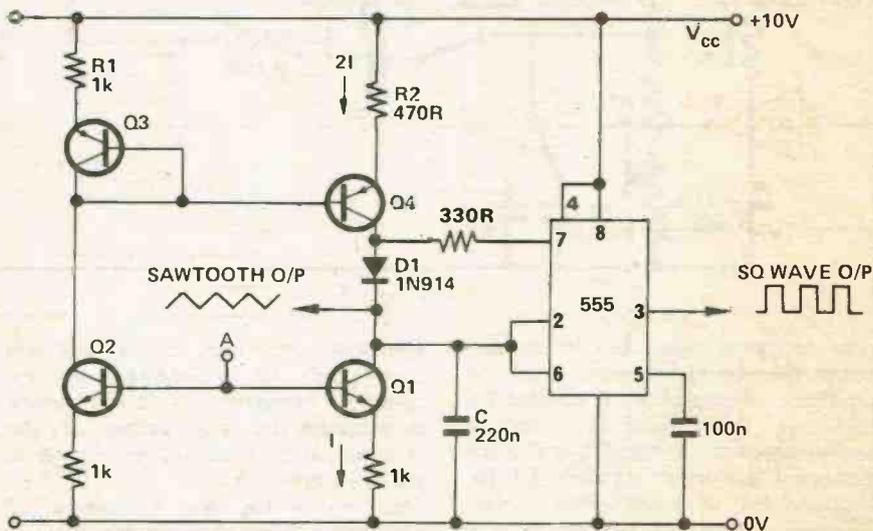
This circuit generates sawtooth and triangle waveforms at a frequency set by an external control voltage.

Current source Q1 draws a current  $I$  from timing capacitor C. Simultaneously current source Q2 draws the same current from current mirror Q3, Q4; this is set up (by R1 and R2) to deliver (from the collector of Q4) twice the current leaving Q2.

Hence C receives a current  $2I$  from the top rail, at the same time delivering  $I$  to the bottom rail, the net effect being that the capacitor is charged by a constant current  $I$ , its voltage rising linearly until the 555's upper trigger point (at  $2/3V_{cc}$ ) is reached.

The output (pin 3) then goes low, as does the open-collector output at pin 7. The latter shunts the output of the current mirror to earth, D1 becoming reverse-biased and isolating C.

Now only current source Q1 is connected to the timing capacitor which is now linearly discharged by current  $I$ . In this way C is alternately charged and discharged. When the voltage on C falls to the 555's lower trigger point at  $1/3V_{cc}$ , the output and discharge pins go high, and the



cycle recommences; the repetition frequency is determined by the magnitude of  $I$ , which is set by the voltage applied at the input point A.

With the component values shown, the frequency range is from approx.

2.5 kHz to less than 10 Hz, as the control voltage varies from +10 V to zero; the frequency is directly proportional to the control voltage. Other ranges may be obtained by altering the value of C.

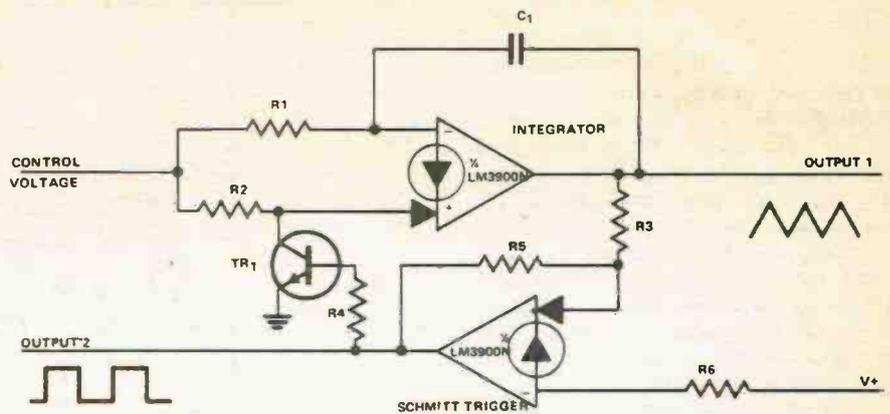
# SIGNAL GENERATORS

## Voltage Controlled Oscillator

A simple voltage controlled oscillator circuit which produces both triangular and square wave outputs is shown.

Then the output of the Schmitt trigger is high, the clamp transistor TR<sub>1</sub> is conducting and the input current passing through R<sub>2</sub> is shunted to ground. The current passing through R<sub>1</sub> causes a falling ramp to be formed.

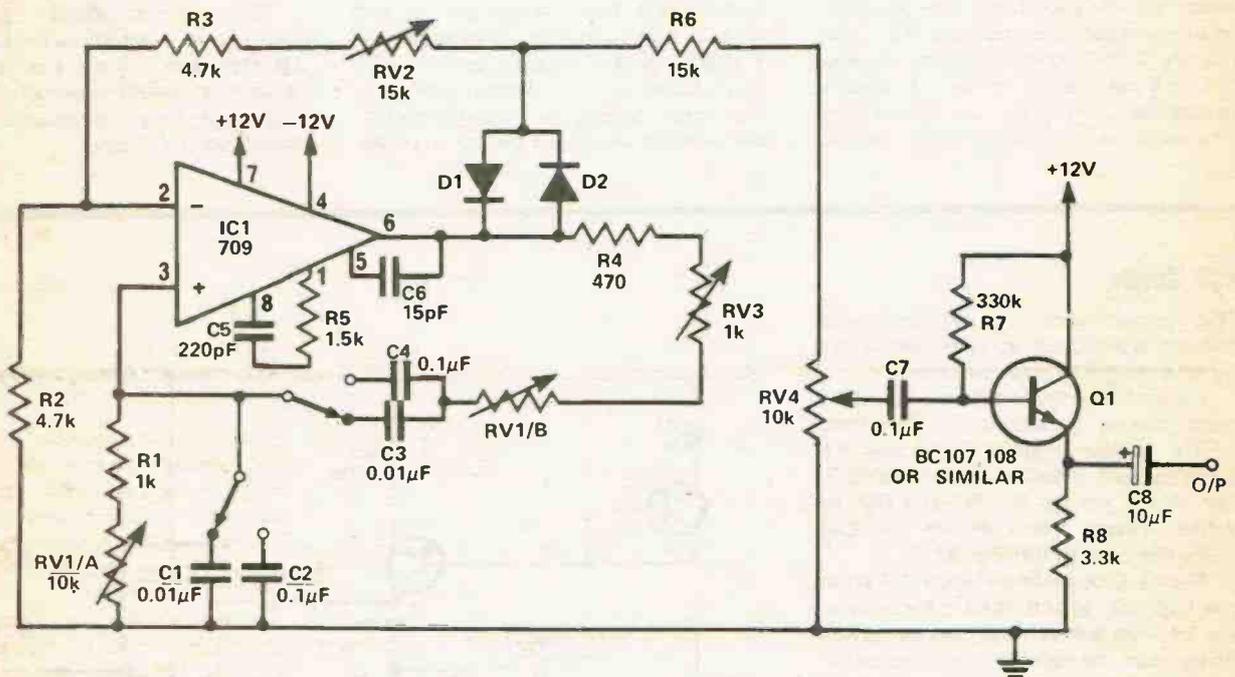
When the Schmitt circuit changes state, its output switches TR<sub>1</sub> to the non-conducting state. The current flowing through R<sub>2</sub> can be made twice that flowing through R<sub>1</sub> ( $R_2 = R_1/2$ ) so that the rising part of the ramp has a similar slope to the negative part.



The greater the value of the control voltage, the greater the frequency of oscillation. However,

the voltage must exceed the constant input voltage ( $V_{BE}$ ) or the circuit will fail to oscillate.

## Sine Wave Oscillator



The oscillator makes use of the well known Wien-bridge network to set the frequency of operation. A resistor (in this case RV1a and R1) and a parallel capacitor (either C1 or C2) are connected to further resistors (RV1b, RV3 and R4) in series with a further capacitor (either C3 or C4). It is a property of the Wien network that the junction of the two RC arms, has, at a single frequency only, a voltage in phase with, but smaller than, that applied to the whole network. Since, in the oscillator, this in-phase voltage is fed to the non-inverting terminal of the op. amp. it constitutes positive

feedback, and thus oscillations will occur and be maintained at one specific frequency - a frequency determined by the values of the resistors and capacitors employed in the Wien network.

So much for the frequency of oscillation. What of its amplitude?

Consider for a moment what would happen if, with the oscillator already giving a sine wave output, the output amplitude should increase for some reason. If it continues to do so, eventually the voltage will become so large that it will be limited by the supply rails and a clipped sine wave

will result. Conversely, if the amplitude of oscillation should decrease, then oscillations will eventually die away to nothing.

Such variations in amplitude can easily arise due to temperature changes etc., and will in any case occur as the frequency is altered, due to tolerances in the capacitor values and tracking errors in the twin gang potentiometer.

Thus, some means of automatic gain control is essential in order to maintain a constant output amplitude.

It will be recalled that the signal voltage applied to the op. amp. non-inverting input was smaller than

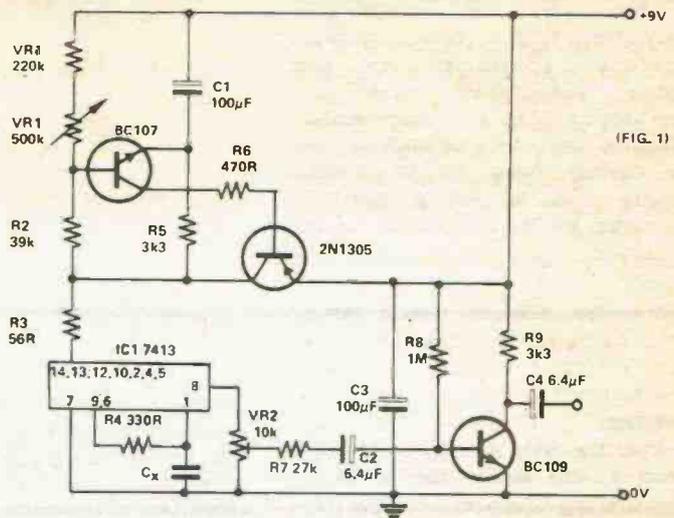
## Tone Burst Generator

The circuit in Fig. 1 generates the waveform shown in Fig. 2. The output is basically oscillations at a certain frequency outputted in small pulses. This type of waveform has varied uses ranging from a beat for an organ or synthesizer to audio or radio frequency testing.

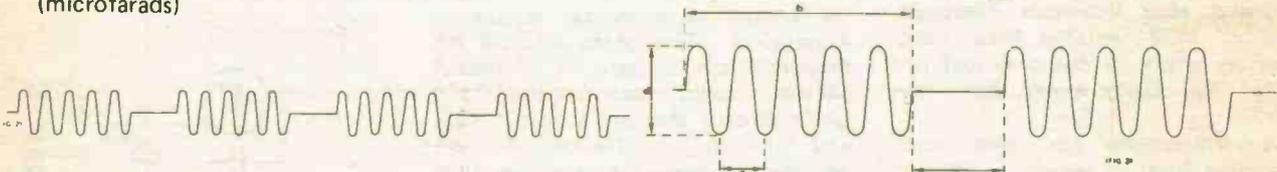
The variable parameters of the waveform are shown in Fig. 3:-

VR1 alters the time between pulses.  
C1 alters the length of the pulse.  
VR2 alters the amplitude of the waveform.

Cx alters the frequency of the waveform within a pulse. This ranges from .0005 giving RF, to 5 giving AF. (microfarads)



(FIG. 1)



the output voltage due to the attenuation in the Wien network. To maintain oscillation the op. amp. must have a gain equal or exceeding this attenuation — which is in fact x3. The desired gain is obtained by selecting the ratio of feedback resistance to input resistance of the inverting input  $(RV2 + R3)/R2$ .

If the overall gain, including feedback, exceeds unity the circuit will produce sine wave oscillation at a frequency set by the Wien network.

Stabilisation of the gain is brought about by the action of diodes D1 and D2.

When the instantaneous output voltage is close to zero, neither diode conducts, since even a germanium diode requires 0.4 volts or so forward voltage to bias it on. Consequently, the negative feedback loop is open (giving maximum gain) and, under the action of the positive feedback via the Wien network, oscillations build up rapidly. As soon as their amplitude is

sufficient to bias on either D1 or D2 (depending on the polarity of the output voltage swing), then R2, R3 and RV2 provide negative feedback, so limiting oscillations to a convenient level.

Re-inforcement of such oscillations takes place close to each zero crossing when D1 and D2 are open i.e. non-conducting; the setting of RV2 determines the final amplitude.

This method of stabilisation does give rise to a very small amount of crossover distortion, but the effect of this can be minimised by setting VR2 for the largest possible sine wave without clipping. In any event, some distortion is a small price to pay for such a simple, easy-to-get-working sine wave oscillator and, further, it is a low level of distortion — some class B audio amplifiers are worse!

Range switching is confined to a choice of two ranges, in the interest of simplicity and cheapness, but more ranges could easily be provided if the

constructor is so inclined.

A simple emitter-follower output stage completes the unit, with a logarithmic potentiometer as a level control, enabling the output to be set from 1 V rms down to 10 mV rms or so.

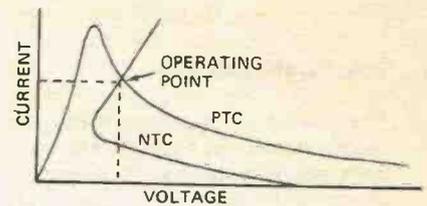
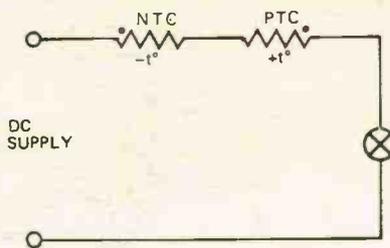
With wiring up completed and thoroughly checked, switch on and, if possible, monitor the output on an oscilloscope. No 'scope? Then a pair of headphones, of reasonably high impedance, can be used instead. Set RV4 about half way, S1 to "low" and RV1 about half way. If there is no output, adjust RV2; clockwise rotation should give increased output.

With an ac meter, measure the signal level at the junction of D1, D2 and RV2. Adjust RV2 for 3 volts rms. This will ensure the highest output level (thus reducing the effect of crossover distortion) consistent with sine wave operation (no clipping). This should provide about one volt rms at the output.

# SIGNAL GENERATORS

## Thermistor Oscillator

A simple very low frequency oscillator can be made by interconnecting one positive temperature co-efficient and one negative temperature co-efficient thermistor in series. For conditions of oscillation the characteristics of the two devices have to be chosen carefully. The operating point is determined by the intersection of the two curves.



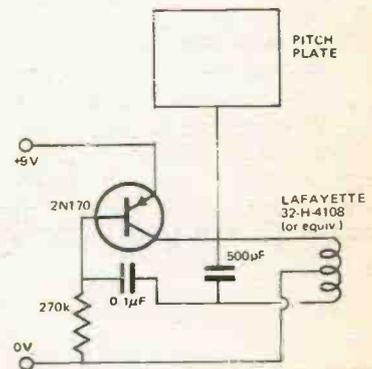
## Theremin

This is the only musical instrument known to the writer that is played without being touched! The Theremin is named after Professor Theremin who, in 1928, amazed New York audiences when he demonstrated his ability "to obtain music from the ether".

The instrument has two rods protruding from its housing, each one forming one 'plate' of a capacitor. The performer's hands become the second

'plates' when held near the rods. The capacitive changes engendered in one rod controls the pitch of the tone, the other rod, responding by varying the amplitude of the output.

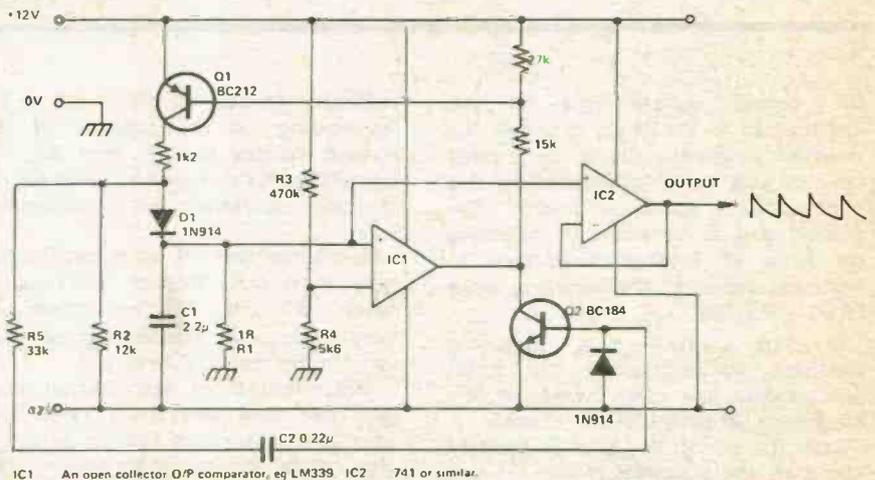
A simple experimental circuit is illustrated. The plate should be around 30cm. square. It is placed next to a radio receiver tuned to a fairly strong station around 900 kHz. The slug of the coil is then adjusted to obtain the most pleasing tone. When the hand is moved near the plate, the pitch of the tone will change.



## Exponential Waveform Generator

This circuit produces a waveform that decays exponentially from a set voltage to near-zero, and then rapidly resets to re-start the cycle.

Initially C1 is charged to +12V, and Q1, Q2 are both off. The timing capacitor there discharges slowly through R1, the exponentially decaying voltage appearing at low impedance at the output of unity gain buffer IC2. R2 prevents the leakage current from Q1 affecting the discharge as D1 is reverse-biased. When the voltage on C1 reaches a value just above zero that is set by R3, R4, the open-collector O/P of IC1 goes low, turning on Q1 and rapidly recharging C1. IC1 of course reverts to its original state almost at once, but the recharge mode is prolonged for several milli-



IC1 An open collector O/P comparator, eg LM339 IC2 741 or similar.

seconds by the positive feedback loop through R5, C2 and Q2, to ensure C1 charges fully. After this time C2 is also fully charged, and Q2 turns off, turn-

ing off Q1, and allowing the slow discharge of C1 to begin again.

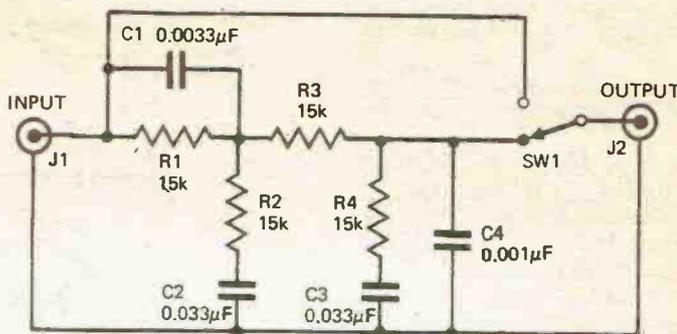
With the component values shown, each cycle lasts about ten seconds.

**Loudness Control**

Many modern high quality amplifiers have loudness controls built in. In most instances they are manually switched into circuit when required — in a few amplifiers the circuit is switched in at all times.

Nevertheless there are innumerable older or present-day low-priced amplifiers that are not fitted with loudness compensation — and it is for units such as these that this simple project has been designed.

The device shown is for a mono amplifier — two are required for stereo amplifiers. It can be very simply assembled on tag strips or matrix board, and, when completed connected between your pre-amplifier and main amplifier. If yours is an integrated unit it should be readily possible to break into the volume control circuit — just connect the unit in series with the slider terminal of the potentiometer. Screened leads may be necessary of long lengths are required.



We would like to emphasize that this is a "compromise" circuit. Ideally a loudness control must be designed specifically to suit the amplifier for which it is intended. Also the degree of loudness compensation should be related to the volume control setting. This latter requirement involves replacing the existing volume control by a suitably tapped potentiometer — a device that is not readily available

"off the shelf" — so the circuit shown here introduces a fixed amount of compensation that is adequate for moderate listening levels.

This circuit will suit most amplifiers quite well — and in any case can be adjusted by minor variation of component values if required.

Switch SW1 should be a double-pole double-throw type if stereo operation is required.

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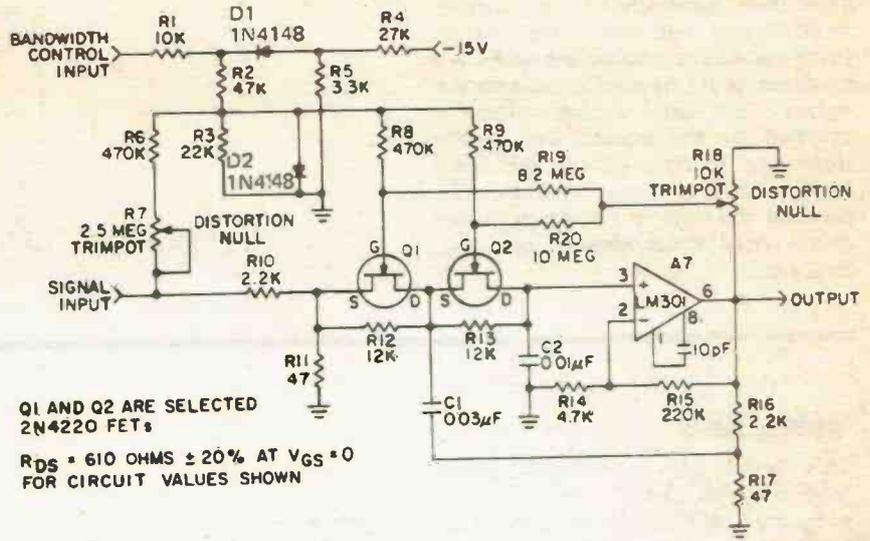
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## Voltage Controlled Filter

An input attenuator (R10 and R11) limits the signal amplitude presented to the FETs to about 0.1 volt p-p at 0 VU to ensure low distortion. Output amplifier A7 makes up exactly for this loss. An op amp having external frequency compensation was used here so that this relatively high-gain stage could be tailored for flat response to 15 kHz (a 741 could be used, but would roll off slightly above 10 kHz). Resistors R16 and R17 attenuate the output signal by an amount equal to the gain, so that this amplifier doubles as the unity-gain buffer required for filter operation. The highest cutoff frequency is dictated by minimum FET resistance and capacitors C1 and C2. The latter should have values in a ratio of about 3:1 to produce the desired Butterworth response.



Q1 AND Q2 ARE SELECTED 2N4220 FETs  
 $R_{DS} = 610 \text{ OHMS} \pm 20\% \text{ AT } V_{GS} = 0$   
 FOR CIRCUIT VALUES SHOWN

## Tone Control. Active

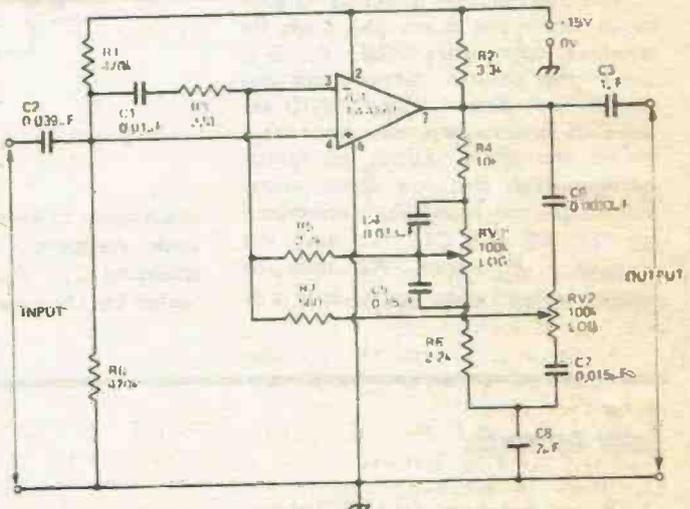
The input signal is applied to the non-inverting input of the IC which is a Siemens TAA861 operational amplifier. Bass and treble boost and cut are controlled by the potentiometers RV1 and RV2 respectively.

Control range is 20 dB of boost or cut at 50 Hz and 15 dB boost or 20 dB of cut at 12 kHz.

The overall gain of the circuit at 1 kHz is 15 dB and the input impedance is greater than 80 k ohm. Total harmonic distortion for 2.4 volts output is less than 0.5% and remains below 4% for up to 3.5 volts output.

Correct law for the potentiometer is antilog. This may be obtained by using slide potentiometers which are mounted in reverse (end-for-end) to normal.

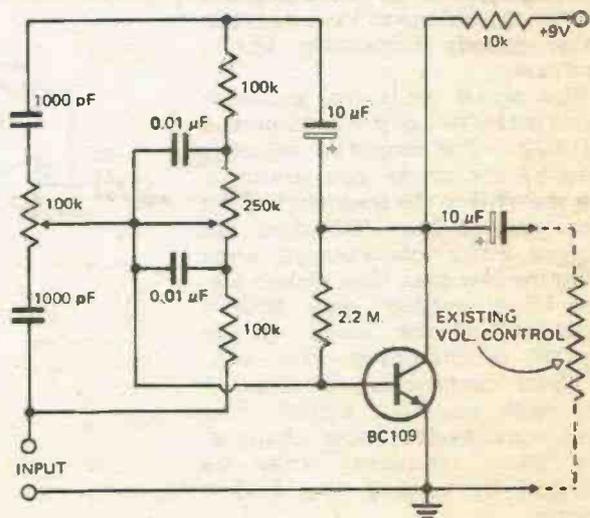
Note that equalization is not incorporated in this preamplifier.



## Tone Control Circuit

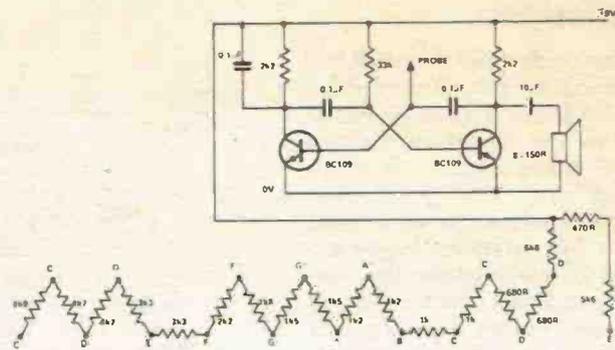
This simple single-transistor circuit will give approximately 15 dB boost or cut at 100 Hz and 15 kHz respectively. A low noise audio type transistor is used, and the output can be fed directly into any existing amplifier volume control to which the tone control is to be fitted.

The gain of the circuit is near unity when controls are set in the "flat" position.



## Organ, Simple

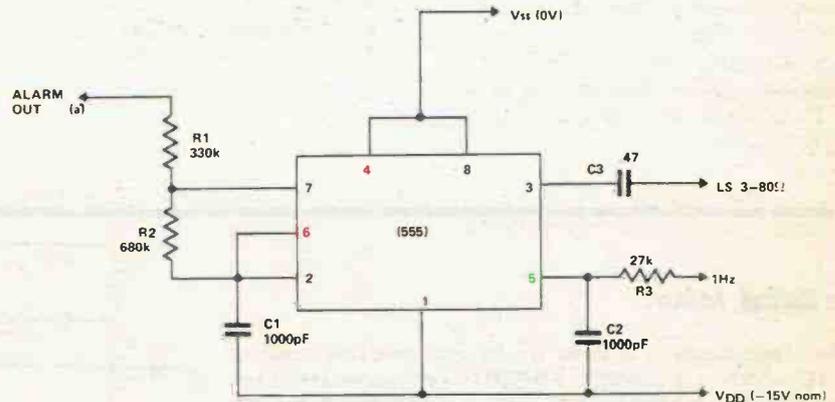
The tone generator is an astable multivibrator with one of the resistors being variable to change the notes. An amplifier could be used to increase the volume, but quite a high volume is attained by the astable. Due to the simplicity of the circuit the wave form is rather irregular in shape. (To produce the note, the probe is moved across metal strips wired to points A, B, C etc.)



## Warbling Alarm

This device gives a two-tone alarm from a digital clock. It may be used with any CMOS alarm clock chip having an active high alarm output and 1Hz (optional) output. It was built to work with a CT7001 chip and requires no interface components.

The 555 operates in normal astable mode when the alarm goes high (ie point (a) approaches VSS). Pin 5 is the normal control voltage input and swings from almost VSS to VDD via the 27K resistor at a 1Hz rate. This causes the audio output to switch between high and low tones, above and below the frequency determined by R1, R2 and C1. To vary the frequency difference, R3 may be altered within wide limits, but it is



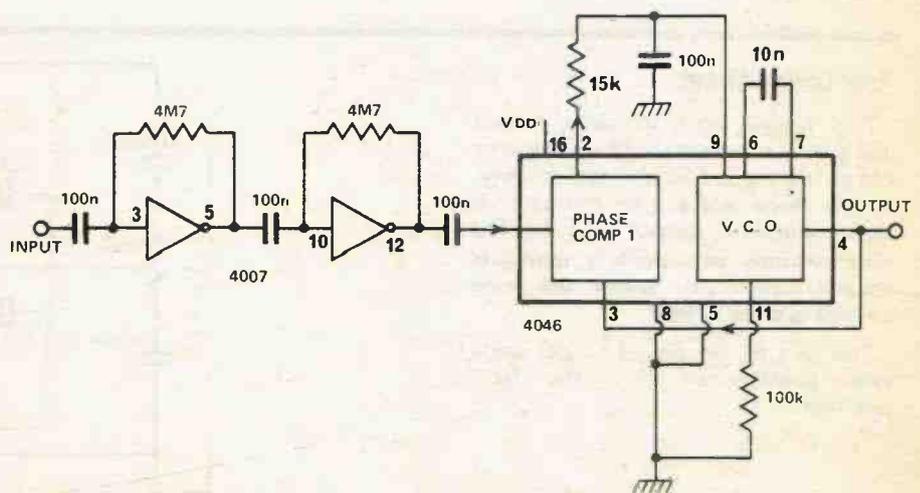
inadvisable to keep it below 15K. The basic frequency is best varied by changing C1. Audio output may be varied by changing C3 (depending on

LS impedance). In the original, a 35Ω speaker was used with -12V VDD and was sufficient to rouse an expert heavy sleeper.

## Guitar Synthesiser

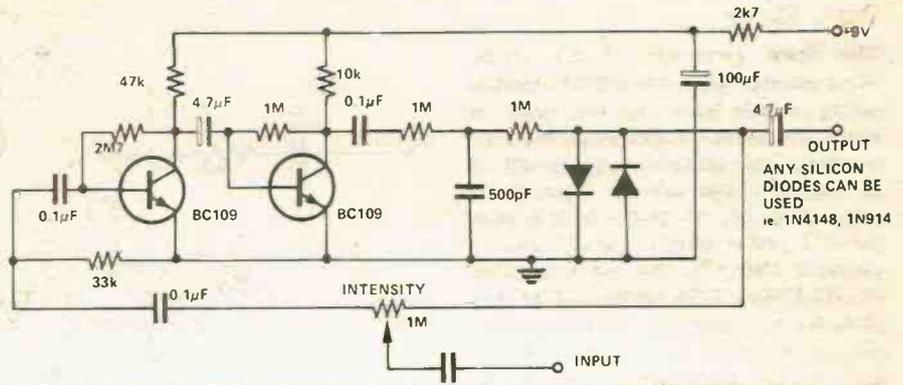
This circuit uses a CMOS Phase Locked Loop, the 4046, to produce a very unusual sound from a guitar, which sounds something like a synthesiser.

The signal from the guitar is amplified by two of the amplifiers in the 4007. The amplified signal is used by the phase comparator to lock the VCO to the frequency of the note played. The VCO does not oscillate until a note is played, when using the low pass filter shown (i.e. the 15 k resistor and 100 n capacitor). If the value of the resistor is increased, the VCO oscillates continuously at about 1 kHz (with no input signal). This gives very smooth note changes. The basic frequency may be changed by varying the 100 k resistor.



## Guitar Fuzz

The input signal is amplified by the two transistors. The distorted output is then clipped by the two diodes and the high frequency noise is filtered from the circuit via the 500pF capacitor. The 1M pot adjusts the intensity of the fuzz, but this tends to make the unit oscillate, so a 33k resistor is put between the input and ground to stop this. When the pot is at minimum intensity the unit may be switched off to allow normal playing.

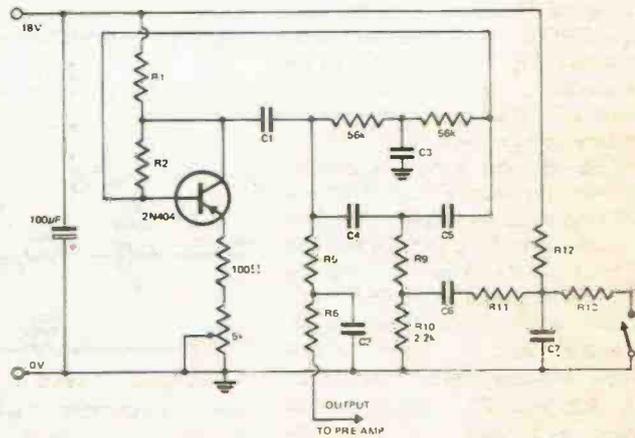


## Drum Simulating

A variety of percussive sounds may be obtained with variations on a simple twin-T oscillator of the type illustrated. The table is given as a guide to the frequency determining and envelope shaping components.

	DRUM	TOM-TOM	BONGO	BLOCKS
R5	22K	82K	82K	330K
R6	10K	82K	82K	not used
R9	2.7K	6.8K	6.8K	6.8K
R11	82K	22K	27K	not used
R12	1M	0.56M	1M	1M
R13	2.7K	2.7K	2.7K	6.8K
C1	0.1	0.047	0.047	0.047
C2	0.1	0.01	0.01	not used
C3	0.1	0.047	0.033	0.01
C4	0.1	0.027	0.015	0.0033
C5	0.1	0.027	0.015	0.0033
C7	0.1	0.1	0.01	0.1

all 1/2 watt 10%

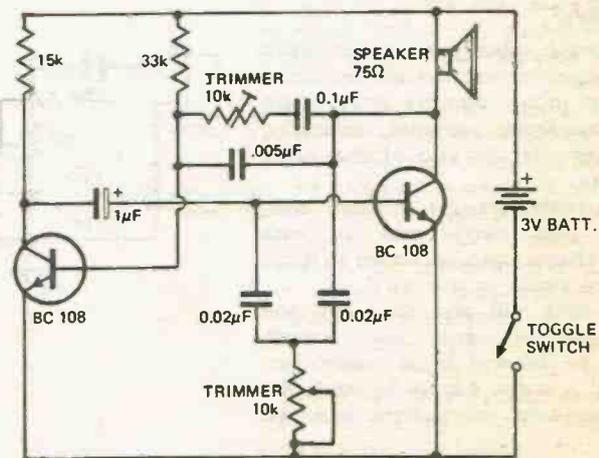


## Fishcaller, Transistorised

A lot of controversy exists among amateur fishermen as to the effectiveness of "fish-callers". Some swear by them, others just shake their heads.

Here's an inexpensive way of finding out. The two-transistor circuit drives the speaker. Varying the two potentiometers produces a wide variety of sounds. You may be lucky and hit on one that will bring in the big ones.

An inexpensive waterproof housing is a *thick-walled* polythene bag with a few lead sinkers inside. An on-off toggle switch can be manipulated without opening the bag when switching power on and off. The bag opening is sealed with good quality electrical tape to make system waterproof. Tape seal should be renewed after each use.



## Audio Mixer

The amplifiers of a LM3900N device can be conveniently used to make a mixer unit for audio purposes; the unit enables three separate audio signals to be mixed together to produce a composite output. The circuit shown provides this facility using only a single LM3900N device and also enables any one channel to be selected by switches. The currents passing through the resistors  $R_4$ ,  $R_8$  and  $R_{12}$  are summed in the input circuit of the fourth amplifier.

If  $S_1$  is open, amplifier 1 will be driven to saturation by the current passing through  $R_2$ . It will therefore be inactive.

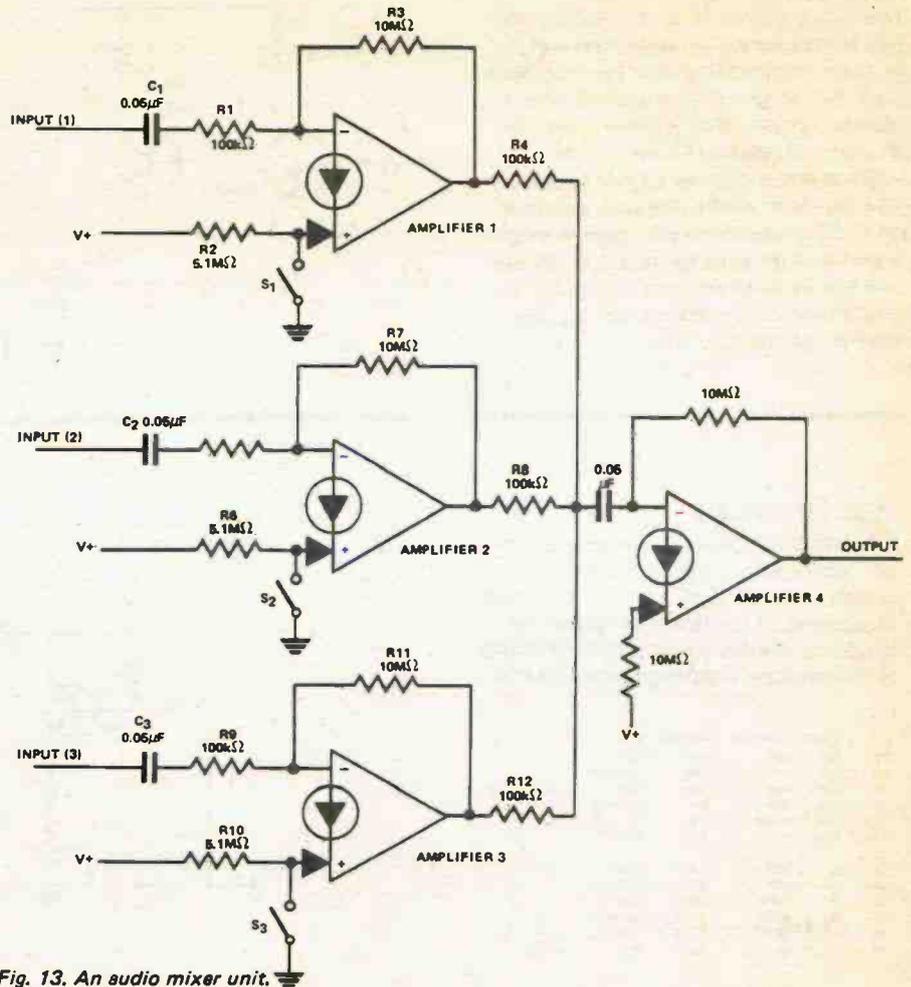


Fig. 13. An audio mixer unit.

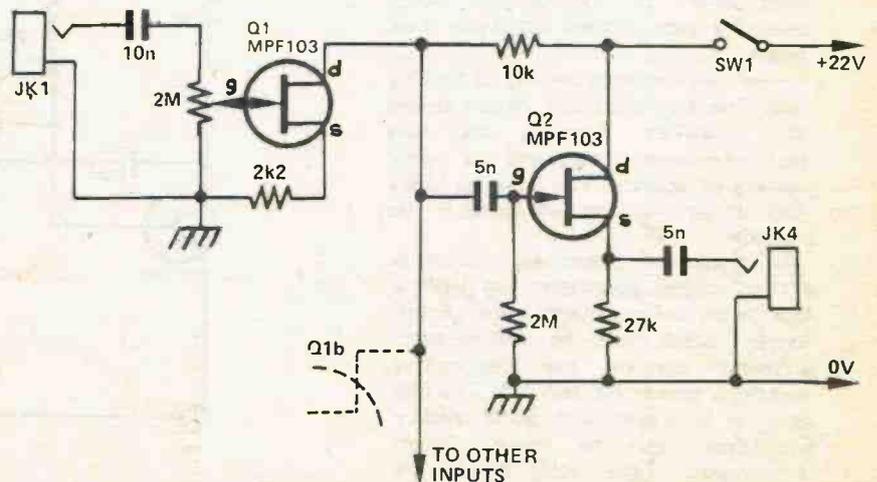
## Basic Mixer

This simple mixer circuit will work with two or three channels, providing excellent input isolation and exceptional frequency response, extending well over the top end of the audio spectrum.

It is usable by one or more instruments plus microphone, or with special effects, such as mixing an input with pink noise, to give 'surf'.

The unit will give 8db gain, and since low-level signals are involved, should be housed in an aluminium box. If a mains supply is used, the usual anti-hum precautions must be taken.

It is useful to use scaled slider potentiometers, so that effects may be re-created.



## Switched Mixer

The circuits illustrated first appeared in "dB", March 1969. It is, in essence, a twelve-channel ring sequencing device, with provision for external trigger. When the switch is at 'interrupt' a single 12-event sequence occurs. The start button is pressed to insert a pulse into the ring generator and, by pressing this button more than once, it is possible to generate highly complex sound patterns, as the multiple pulses follow each other round the ring.

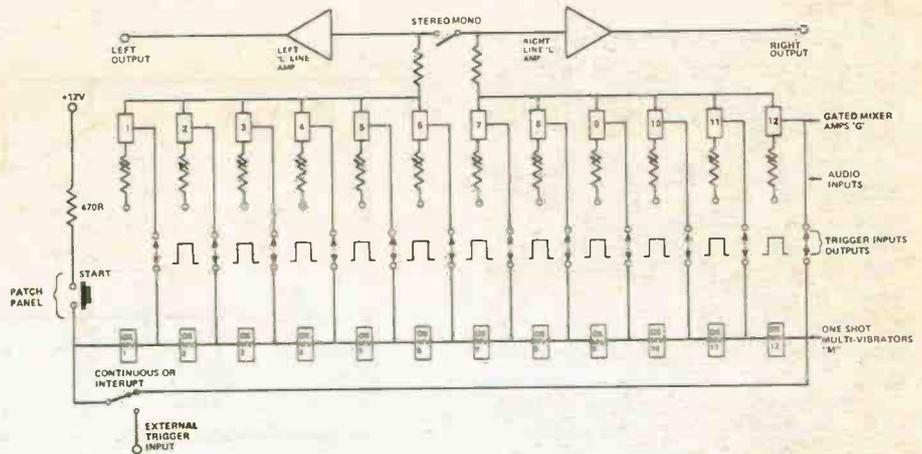


Fig. 1. Twelve channel switching mixer.

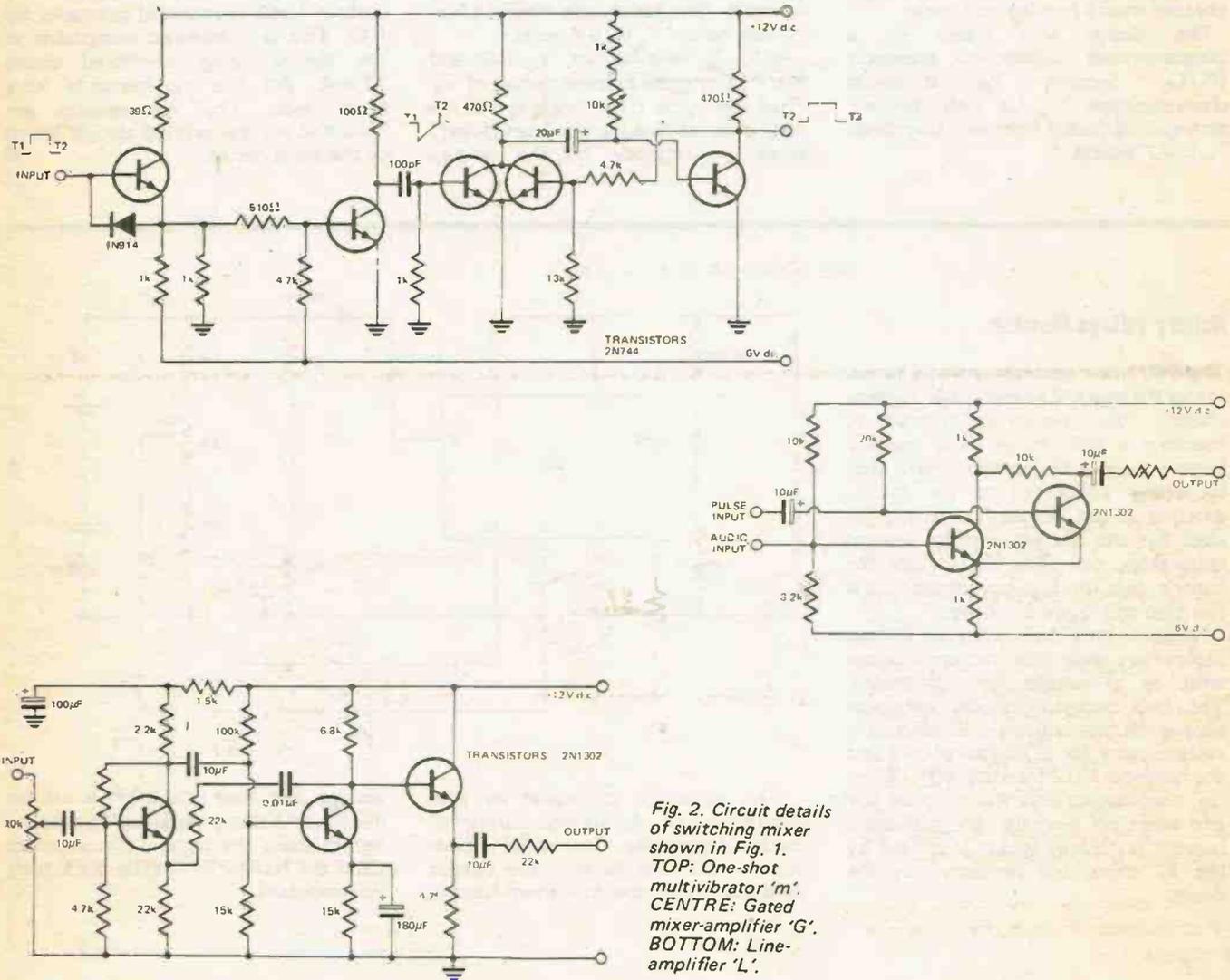


Fig. 2. Circuit details of switching mixer shown in Fig. 1. TOP: One-shot multivibrator 'm'. CENTRE: Gated mixer-amplifier 'G'. BOTTOM: Line-amplifier 'L'.

# DETECTORS & COMPARATORS

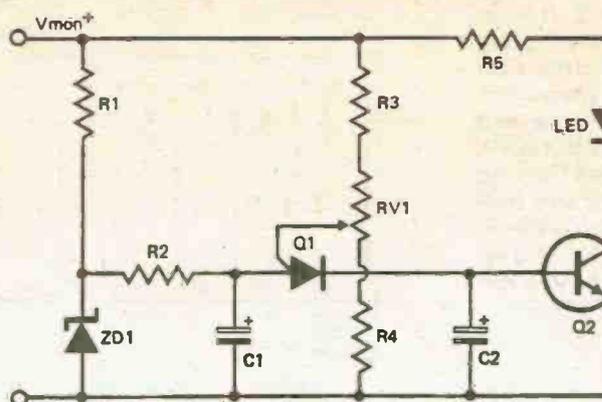
## Low Battery Warning

The prototype of this device will be used in a hospital operating theatre in conjunction with battery operated medical equipment (powered by four 'pen-light' cells).

A moving coil voltmeter was not appropriate as, in the designers' experience, medical staff have difficulty in interpreting a voltmeter — and sometimes find themselves half way through an operation with exhausted batteries. Therefore, the requirements for the indicator were that:

- 1) the display be eye catching, easily understandable and provide a sense of urgency as the battery approaches exhaustion;
- 2) provide adequate warning of battery failure (at least 1 hour);
- 3) current consumption of the indicator be low in relation to the main equipment;
- 4) preferably, be more rugged and cheaper than a moving coil meter.

The design was based on a programmable unijunction transistor (PUT), because its threshold characteristics can be well defined, arranged to flash a light emitting diode (L.E.D.) indicator.



R1	2k2	¼ W
R2	100k	..
R3	18k	..
R4	18k	..
R5	330R	..
RV1	2u2	..
ZD1	4V3 400mwW	
Q1	2N6027 PUT	
Q2	BC107	

The circuit is shown in Fig. 1. The PUT (Q1) is used in a relaxation oscillator circuit. As the voltage being monitored ( $V_{mon}$ ) falls, the voltage on the gate ( $V_g$ ) falls whilst the anode voltage ( $V_a$ ) remains essentially constant. Oscillation commences when  $V_g$  falls below  $V_a$  by 0.6 volts.

As  $V_{mon}$  falls further,  $V_g$  falls and the PUT triggers at lower values of  $V_a$ . Thus the cycle time shortens and the frequency of flashing increases giving a sense of urgency as the battery

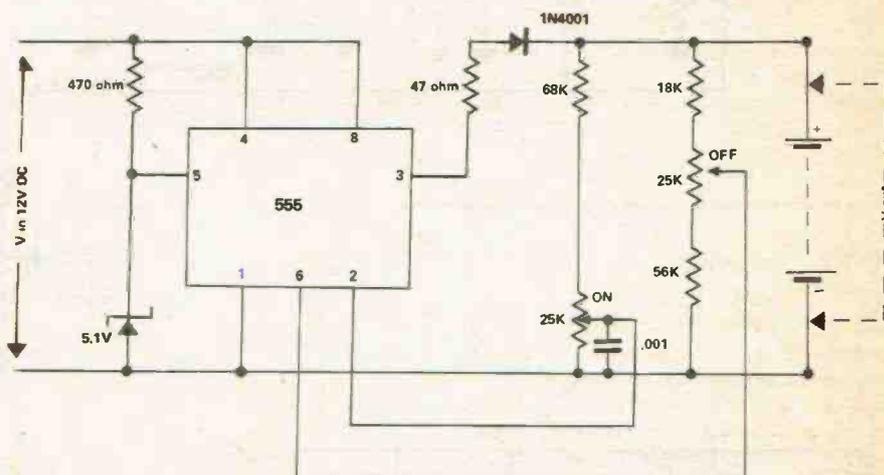
approached exhaustion. Transistor Q2 and C2 act as a pulse stretcher and amplifier to drive the L.E.D. display.

In the prototype the trigger point can be adjusted from 4.5-5.5 volts and the current drain when  $V_{mon}$  is 6 volts is 1 mA (controlled primarily by R1). This is considered acceptable as the device being monitored draws 17 mA. All the requirements have been met. The components are mounted on the printed circuit board of the main device.

## Battery Voltage Monitor

The 555 timer can conveniently function as the heart of an automatic battery charger, the circuit is intended to maintain a full charge on a standby battery supply for an instrument that is always connected to the mains, whether in use or not. It can also be used for the charger unit for pocket calculators, etc. The circuit uses the timer's two on-chip comparators, the flip flop and driver amplifier.

A zener provides a reference voltage somewhere near the battery voltage with an allowance for adjustment. The two potential divider networks supply the comparators with adjustable voltages, one for LOW (switch on) and the other for HIGH (switch off). When on, the output gives a maximum of 10V and when off gives 0V, the maximum current is 150mA which is limited by the 47 ohms and protected by the diode.

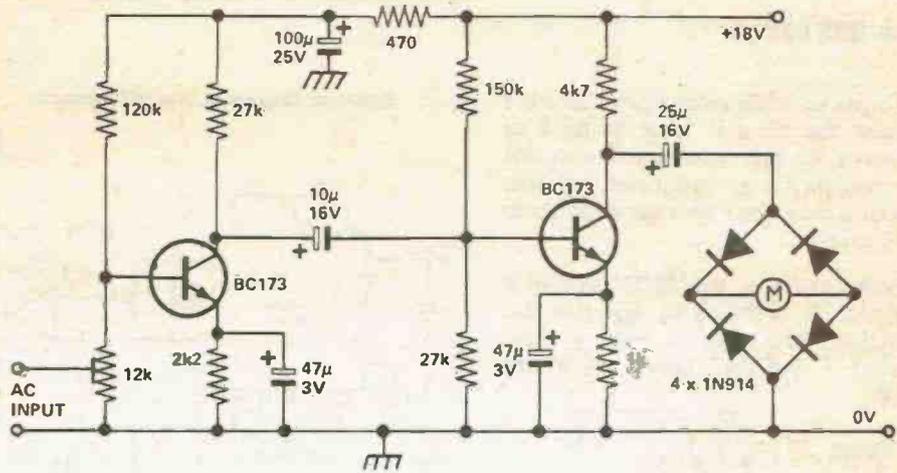


The circuit is calibrated by substituting a variable voltage supply for the NiCd batteries. The HIGH adjustment is set first so that the output switches off at the maximum battery

voltage and then the LOW is set for minimum battery voltage. It is simplest to leave the output disconnected from the resistor until after the setting up procedure.

## Recording Level Meter

The circuit shows a two-stage voltage amplifier driving a recording level meter. The AC signal input is amplified, rectified, and the resultant DC voltage shown on the meter. The circuit can be used with a tape-recorder or audio mixer and should be fed from a point early in the pre-amp. Current consumption in a no-signal state is 2.8mA. The 12K preset gives a variation in sensitivity. The meter can be any general purpose type.



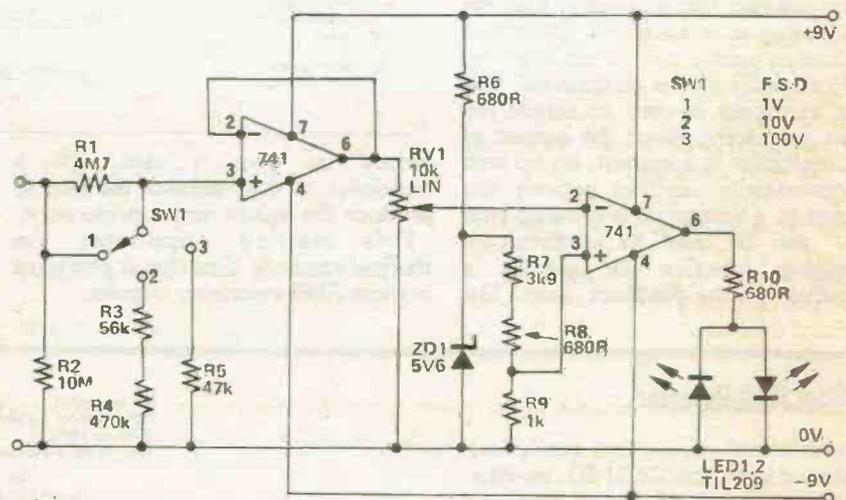
## Comparator Voltmeter

This circuit, although simple, is capable of accurate voltage measurement. The input is applied to the high impedance input of IC1 via the attenuator comprising of R1 to R5 inclusive.

Since this IC is used as a unity gain buffer, the output at pin 6 is equal to the input voltage at pin 3, but at a low impedance. IC2 is connected as a comparator driving a pair of LEDs, D1 and D2.

The inverting input samples a portion of the unknown input voltage, whilst the non-inverting input is connected to a 1V reference obtained from the stable voltage across ZD1.

In use RV1 is adjusted till D2 just illuminates. At this point, if the control knob is of the 0 - 10 calibrated type, the pointer will indicate the input voltage.

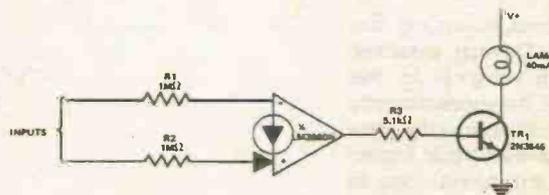


For example, with SW1 in position 2, and with a reading of 2 on RV1, the input voltage will be 2V. With a little practice, the voltage can be read to  $\pm 2\%$ , comparable to a moving coil instrument. The input impedance on all ranges is 3.2M $\Omega$ .

## Voltage Comparator

The circuit shows how an LM3900N amplifier may be employed to compare two input voltages and to indicate the result by means of a small lamp. If the input voltage connected to the non-inverting input is appreciably more positive than the other input, the output of the amplifier will provide a positive voltage which renders the TR<sub>1</sub> conducting. The lamp will then be illuminated.

One of the inputs may be a reference voltage so that one can then compare a single input voltage against this constant reference.



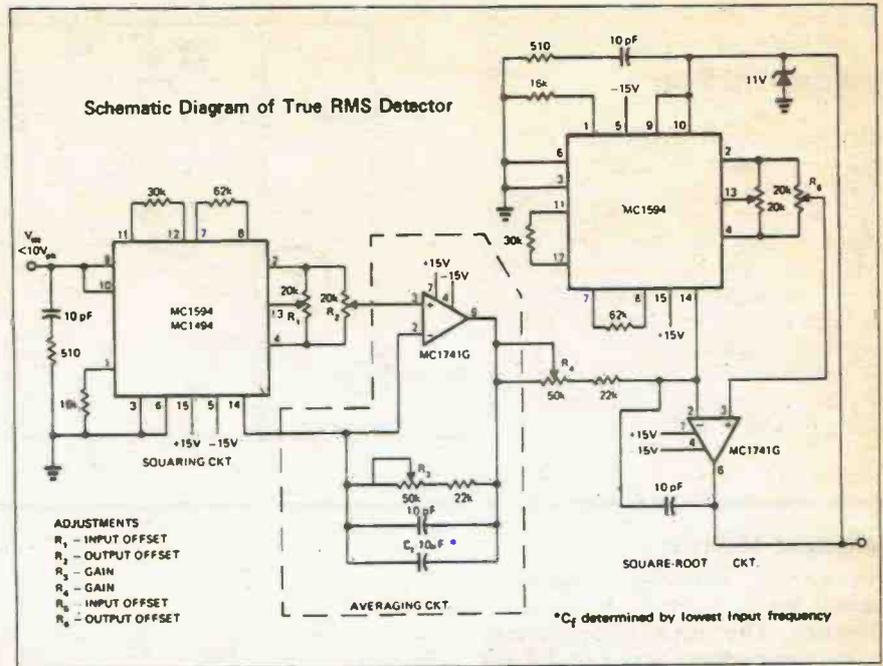
## True RMS Detector

To get an RMS value when you can't afford the time it takes to heat an element, try this technique. It may not be feasible for a multimeter but how about a sampling voltmeter good up to 600 kHz?

Mathematically, the RMS value of a function is obtained by squaring the function, averaging it over a time period  $T$  and then taking the square root:

$$V_{RMS} = \sqrt{\frac{1}{T} \int_0^T V^2 dt}$$

In a practical sense this same technique can also be used to find the RMS value of a waveform. Using two multipliers and a pair of op amps, an RMS detector can be constructed. The first multiplier is used to square the input waveform. Since the output of the multiplier is a current, an op amp is customarily used to convert this output to a voltage. The same op amp may also be used to perform the averaging function by placing a capacitor in the feedback path. The

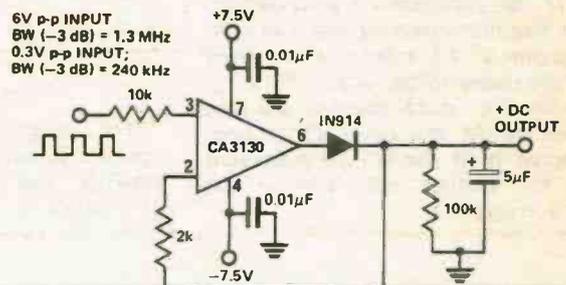


second op amp is used with a multiplier as the feedback element to produce the square root configuration. This method eliminates the thermal-response time that is prevalent in most RMS measuring circuits.

The input-voltage range for this circuit is from 2 to 10 Vpk. For other ranges, input scaling can be used. Since the input is dc coupled, the output voltage includes the dc components of the input waveform.

## Positive Peak Detector

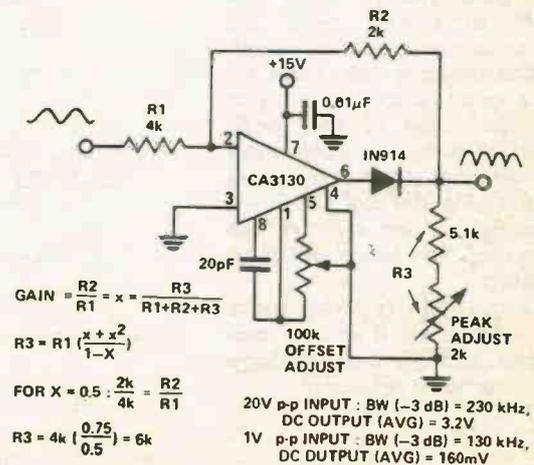
Peak-detector circuits are easily implemented with the CA3130, as illustrated. It should be noted that with large-signal inputs, the bandwidth of the peak-negative circuit is much less than that of the peak-positive circuit. The second stage of the CA3130 limits the bandwidth in this case.



(a) PEAK POSITIVE DETECTOR CIRCUIT

## True RMS Converter

An absolute-value circuit, using the CA3130 is shown. During positive excursions, the input signal is fed through the feedback network directly to the output. Simultaneously, the positive excursion of the input signal also drives the output terminal (No. 6) of the inverting amplifier negative such that the 1N914 diode effectively disconnects the amplifier from the signal path. During the negative-going excursion of the input signal, the CA3130 functions as a normal inverting amplifier with a gain equal to  $-R2/R1$ . When the equality of the two equations shown is satisfied, the full-wave output is symmetrical.



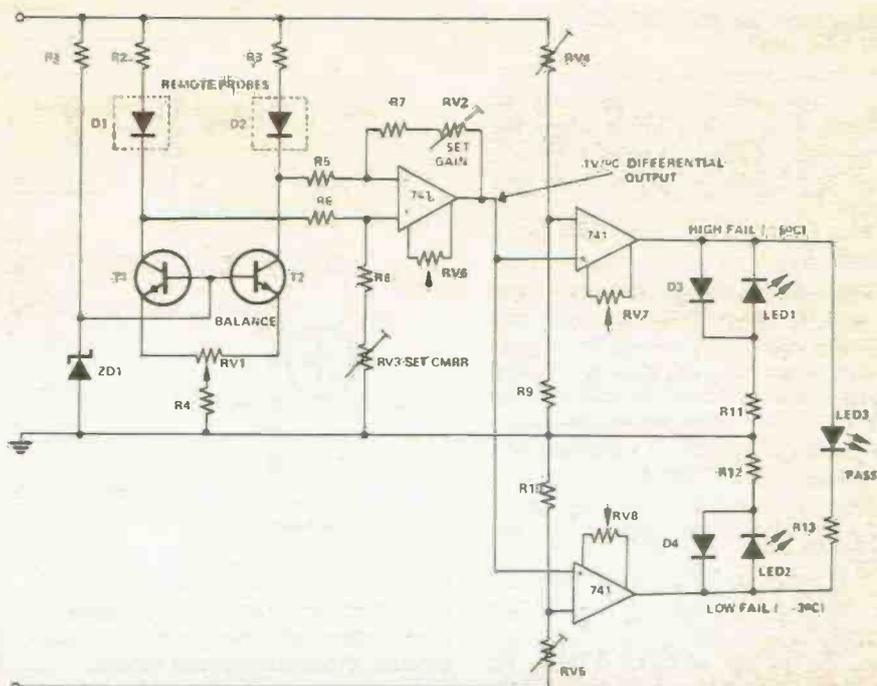
## Temperature Sensor, Differential

The circuit is comprised of three parts (i) the differential temperature sensor (ii) a differential amplifier to provide gain (iii) a switching circuit to monitor the output from the differential amplifier.

Two diodes D1 and D2 are used as probes for the sensor. A small preset, RV1 provides fine adjustment of the current through each branch so as to give zero differential output between D1 and D2 when they are at the same temperature.

A gain of 500 must be provided at the differential output to provide a useful voltage to switch the LED's (...ie 1V corresponding to 1°C.) RV2 provides fine adjustment of the gain and RV3 adjusts the CMRR.

A potential divider network is set up by RV4, R9, R10, RV5 to provide the necessary switching voltages for the voltage comparators, thus enabling LED1 or LED2 or LED3 for voltages set up by RV4 and RV5...ie.. -3V and +5V.



### SETTING UP

1. Adjust offset-null on all Op. Amps for zero output by connecting input terminals together and taking to ground and adjusting either RV6, RV7 and RV8.
2. Adjust CMRR for differential amplifier by shorting input terminals and connecting to +15V line, then adjusting RV3.
3. Apply probes D1 and D2 to a liquid, say at room temperature, and adjust RV1 until there is zero output across collectors of T1 and T2.

### RESISTORS

R1	2.2k
R2,R3	51k
R4,R10,R13	1k
R5,R6	2k
R7,R8	910k
R9	390Ω
R11,R12	1.2k
PRESETS	
RV1	100Ω
RV2,RV3	100k

### RV4,RV5

RV4,RV5 2.2k

RV6,RV7,RV8 10k

### TRANSISTORS

T1,T2 BC108

### DIODES

D1,D2 1N4004

D3,D4 1N914

### LED1,LED2

miniature RED

LED3 miniature GREEN

ZD1 400mW,3V3

### 3 Operational Amplifiers

741

4. Apply probe D1 to a liquid at a temperature 10°C different from above, then adjust gain control RV2 until there is 10V at the diff. amplifier output. The CMRR

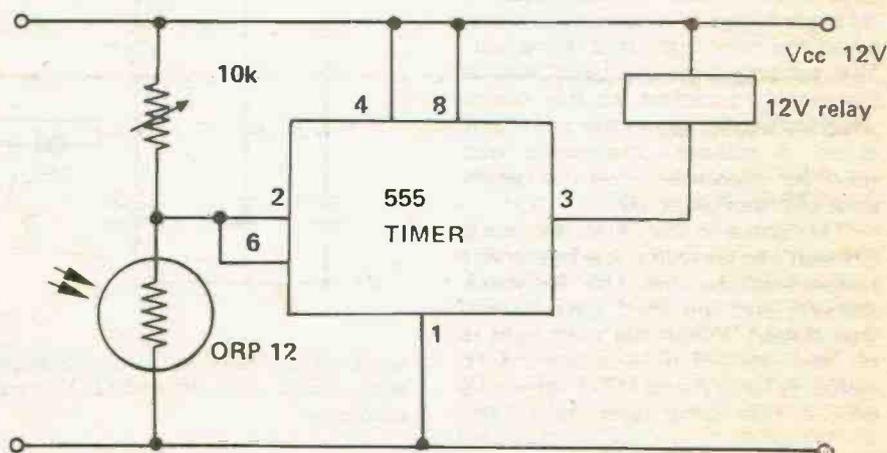
should again be set.

5. Adjust RV4 and RV5 so that the comparators switch at -3V and +5V corresponding to -3°C and +5°C.

## Schmitt, 555

A very useful schmitt trigger can be made by utilising a single 555 timer with its trigger and threshold inputs connected together. The schmitt has a very low input current (1.5uA) and can directly drive a relay taking up to 200mA of current.

The circuit shows a 555 schmitt being used to energise a relay when the light level on a photoconductive cell falls below a preset value; the relay energises when the voltage on pins 2 and 6 is greater than  $2/3V_{cc}$  and de-energises when the voltage falls



below  $1/3V_{cc}$ . This gives a hysteresis of  $1/3V_{cc}$ . The circuit can be used in many other similar applications where

a high input impedance and low output impedance are required with the minimum component count.

## Temperature Sensor, Remote

The circuit shows a temperature sensing device which can be used to indicate at a remote point when the temperature passes through a certain value or to give an alarm when this occurs.

The sensing unit itself contains a 2N930 transistor. The base-emitter voltage of this device appears across R1 and (as the base current is far less than the collector current) the voltage at the upper end of R2 will be the emitter-base voltage multiplied by  $(R2 + R1)/R1$ . The base-emitter voltage changes with a temperature coefficient of  $-2.2\text{mV}/^\circ\text{C}$  and this change is multiplied by the same factor before being applied to the LM339 circuit.

The potential at point A is set by the resistors R3 and R4. As the temperature of the sensor rises, the voltage at point B falls. At the time this voltage falls below that at point A, the output of the LM339 voltage comparator will go 'high'. If, however, the input connections to the LM339 are reversed, the output will go 'low' when the temperature of the sensor falls below the preset point.

The LM339 contains four separate voltage comparators in one package; only one of these comparators is used in the circuit shown. The other three comparators could be used with

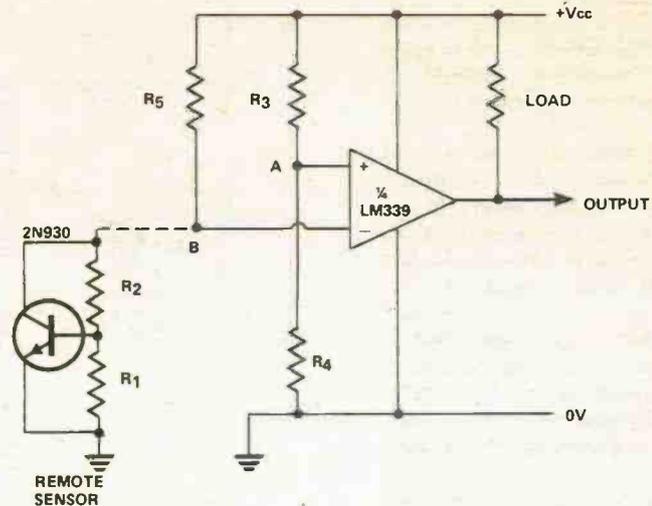
another three temperature sensing transistors so that an indication is given when the temperature passes through three other preset values.

The value of R5 should be chosen so that the current passing through the remote sensor unit is about  $10\mu\text{A}$ . If the temperature range over which operation is required is narrow, the ratio R2/R1 may be large so that the system is very sensitive to small temperature variations. A potentiometer may be substituted for R3 and R4 so that the temperature at which the comparator switches is variable. The voltage at point B is highly linear

over a very wide temperature range (about  $-65^\circ\text{C}$  to  $+150^\circ\text{C}$ ) and therefore the potentiometer which replaces R3 and R4 can be given a linear calibration.

A feedback resistor may be connected from the output to the non-inverting input to provide a small amount of hysteresis (so that the temperature at which the output changes when the temperature is rising is different from that when it is falling); one then has the basis of a thermostat.

The output current has a maximum value of about  $15\text{mA}$ .



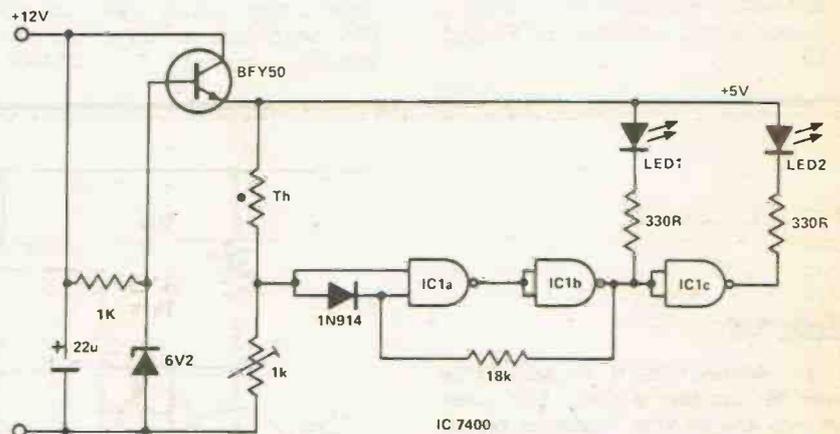
## Warmth Indicator

A simple indicator was required for a gas fridge in a caravan to show when the pilot light had gone out. The sensing element used was a thermistor, attached to the outlet which is 'warm' when the pilot light is on. A rod-type thermistor was used for cheapness, with a resistance of about  $3\text{k}$  at  $20^\circ\text{C}$ .

Two gates of the 7400 provide a Schmitt trigger with a low hysteresis (determined by the  $18\text{k}$  feedback resistor) and the third gate inverts that output. When the pilot light is on, the input of IC1a is high, IC1c output is logic 0 and LED2 (green) is on. If the pilot light fails, the

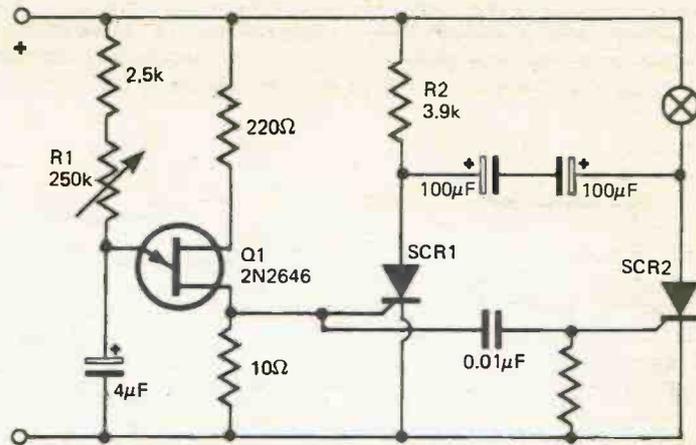
temperature falls, all gates change state, LED2 goes off and LED1 (red) comes on.

The temperature at which the changeover takes place is set by the  $1\text{k}$  preset.



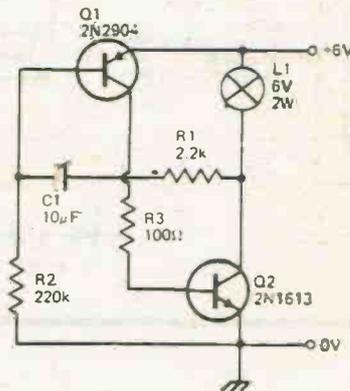
## Warning Flasher

This circuit will operate reliably from noisy or fluctuating power supplies — and unlike many multivibrator circuits — is inherently self-starting when power is applied. In this circuit unijunction transistor Q1 is used as a relaxation oscillator supplying a continuous train of pulses to the gates of the SCRs. Assume that SCR2 has been triggered into conduction and that lamp 2 is energized. The next trigger pulse from Q1 triggers SCR1, this discharges C2 and the resultant commutation pulse turns off SCR2. The resistor R2 in the anode of SCR1 is of a value high enough to prevent SCR1 from latching on. SCR2 is retriggered by the next triggering pulse from Q1. Using the component values shown, the flash rate of this circuit is adjustable — by R2 — from 35 to 150 flashes a minute.



## Transistorised Flasher

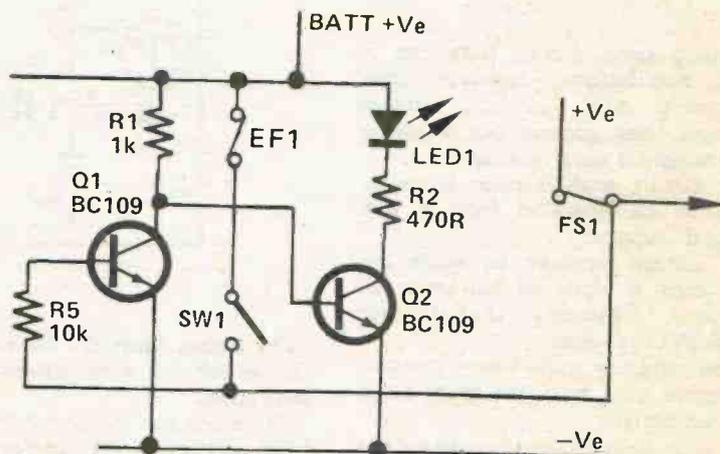
This simple circuit will flash a 6 volt lamp at a rate determined by the size of capacitor C1. It is most economical on power as it only draws current when the lamp is ON. When the lamp is OFF both transistors are biased OFF.



## Blown Fuse Indicator

Base current for Q1 is taken from the 'earthy' side of FS1. Q1 will conduct its collector voltage falling to zero. Q2 base will also be zero, switching LED 1 off.

If FS1 were to 'blow' or cease to exist, depart for its maker, have a rest, go to sleep, peg out, become inoperative, deceased, out of order, or duff, kick the bucket, bite the dust, pass away, self destruct, become no longer intact, or cease to conduct in any way, due to war, flood, corrosion or act of God etc., Q1 would switch off, causing its collector to rise to 12 V, switching Q2 and LED 1 on. R2 is the current limit resistor for LED 1. SW1 will bypass FS1 via emergency fuse 1, until FS1 can be replaced.



# INDICATORS

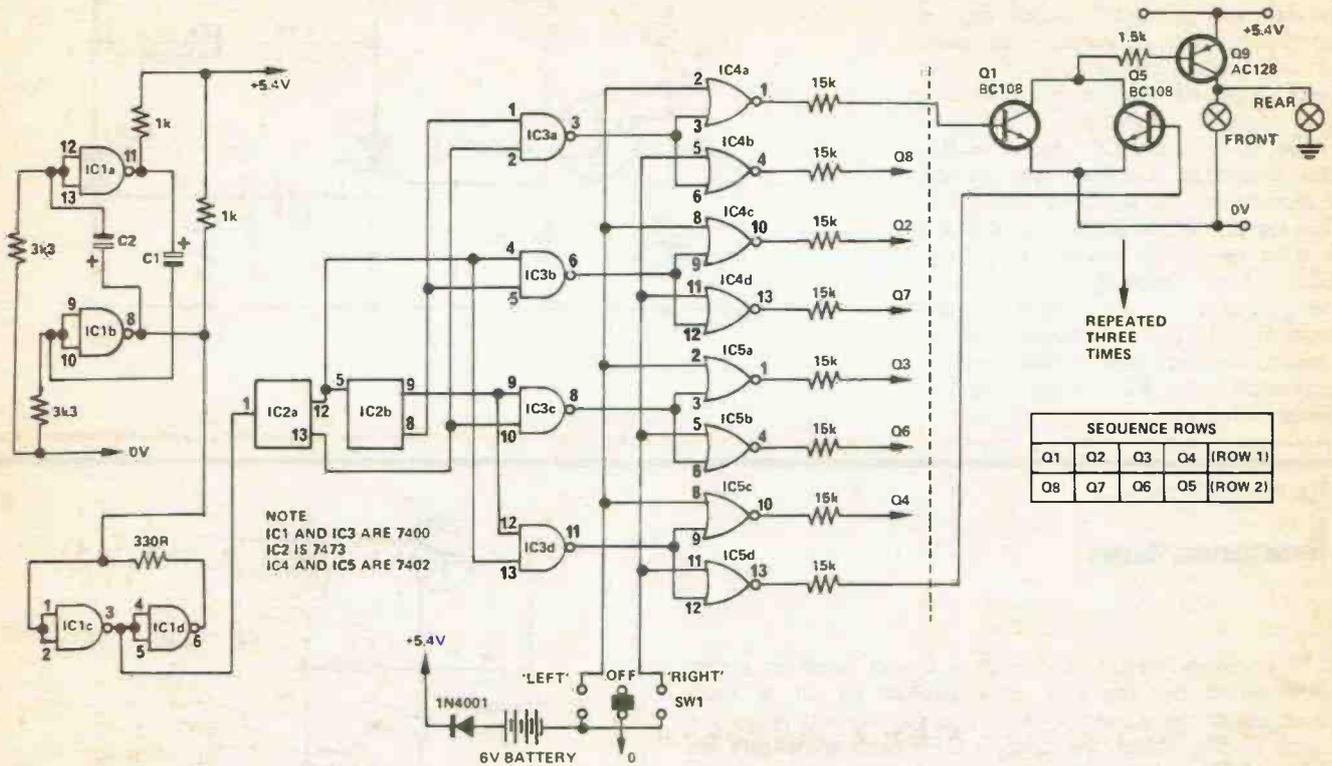
## Novel Indicators

Since a bicycle has no effective width, normal indicator lamps placed on each side do not give a clear indication of direction when seen from a distance,

especially at night.

The circuit shown is a four stage ring counter which sequentially drives four yellow lamps giving an impression of movement i.e. towards the left or right. Lamp sequencing rate can be altered by changing C1 and C2. (50uF

was found to be about right). Oscillator pulses are shaped by schmitt trigger IC1b. The decoding and output gating are performed by ICs 3, 4, and 5. Driver transistors Q1 to Q8 can be any low current, medium gain NPN silicon.



## Neon Tube Flasher

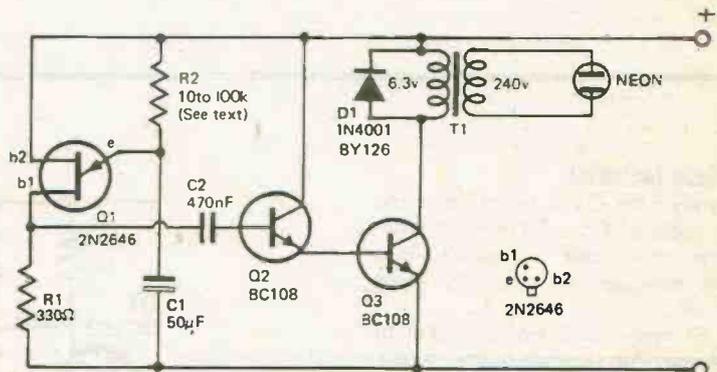
Flashing neon globes have use in many applications, however their relatively high working voltage precludes their general use where a mains supply is not available.

This circuit enables neon tubes or bulbs to be operated from a low voltage dc supply.

The voltage required to ignite the neon tube is obtained by using an ordinary filament transformer (240-6.3V) in reverse.

Battery drain is quite low — being in the region of 1 to 2 milliamps for a nine volt battery.

Q1 is a unijunction transistor and operates as a relaxation oscillator. Its frequency of operation is determined by R2-C1.



The pulses from Q1 are directed to Q2 which in turn drives Q3 into saturation.

The sharp rise in current through the 6.3V winding of the transformer as Q3 goes into saturation induces a high voltage in the secondary winding causing the neon to flash.

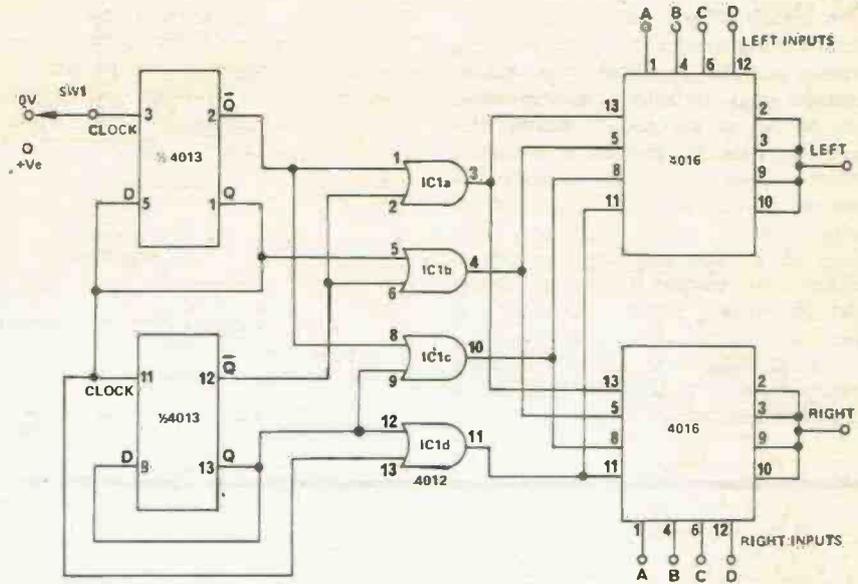
The diode D1 protects the transistor from high voltage spikes generated when switching currents in the transformer.

## Stereo Input Selector

Four different inputs can be switched through by the continual pressing of SW1.

IC1 is a dual 'D' type flip flop. The Q outputs are connected to the D inputs so that the clock inputs are divided by two. The two flip flops are connected in series, giving a two-stage binary counter.

IC2 is a quad AND gate. This is used to decode the four states of the counter. The outputs are used to control the quad switches at IC3 and IC4 (4016AE).

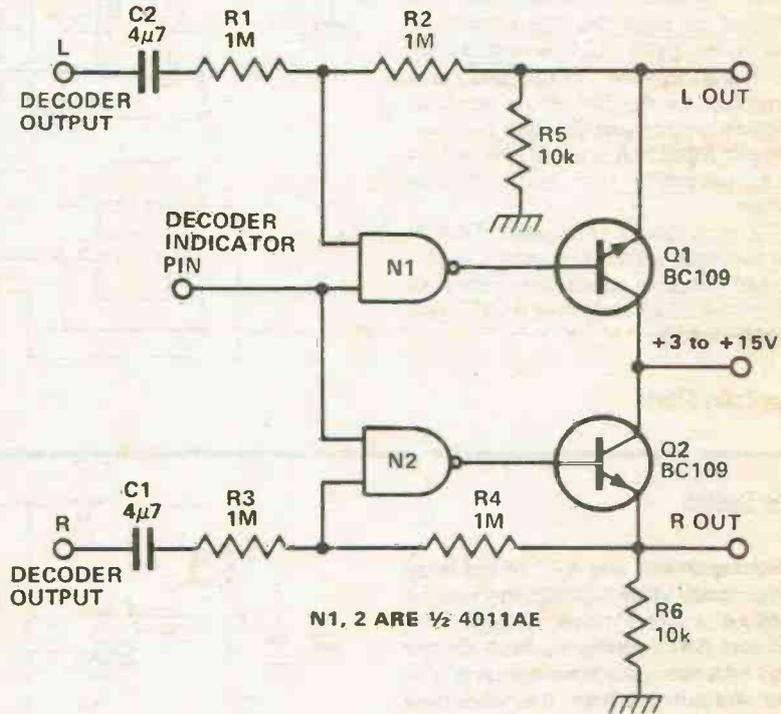


## Stereo Switch, Simple

A device to switch the audio from a stereo tuner only when a stereo signal is being received.

Two CMOS NAND gates and two transistors are employed. One of the inputs from each gate is connected together and to the indicator output of the decoder IC.

The other gate inputs are connected to the emitter's of Q1 and Q2 respectively, by means of the feedback resistors R2 and R4. On reception of a stereo signal the indicator output of the decoder goes high and the feedback resistors bias the gates into the linear region passing the signal. On reception of a mono signal or interstation noise, no signals pass through the gates, the circuit providing a mute function.

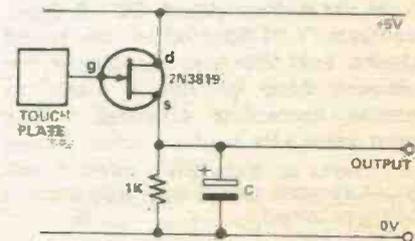


## Logic Touch Switch

An n-channel field-effect transistor is the basis of this simple trigger. In its quiescent state the voltage at the output is about 3V. When the plate is briefly touched with a finger, the minute currents between the body and the plate alter the electric field at the gate of the transistor. The effect is to cause a drop in output voltage. It falls almost to zero and can be used to

trigger a TTL flip-flop. This can be constructed in the usual way, using two NAND gates from a 7400 IC. If several triggering circuits are required, it is more convenient to use the 74118 sextuple bistable latch.

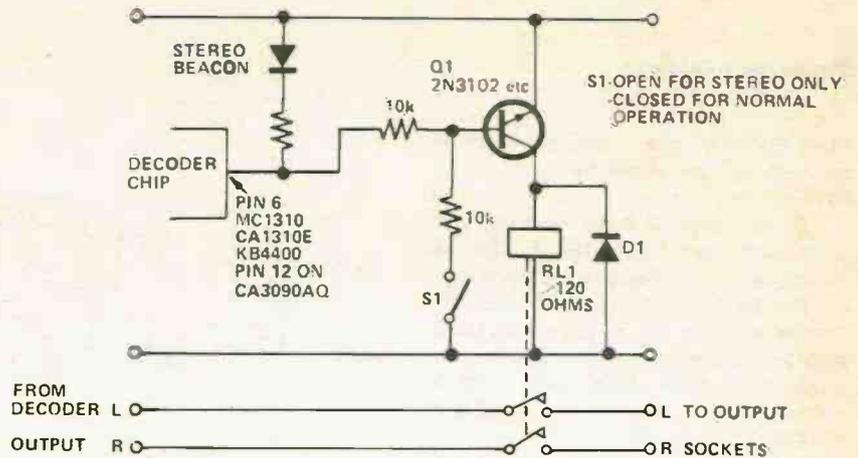
The value of the capacitor is not critical, but 10uF is convenient. The touch-plate can be an area of copper etched on a circuit-board, a square of aluminium foil, or simply a drawing-pin pressed into an insulating support.



# SWITCHING

## Stereo Only

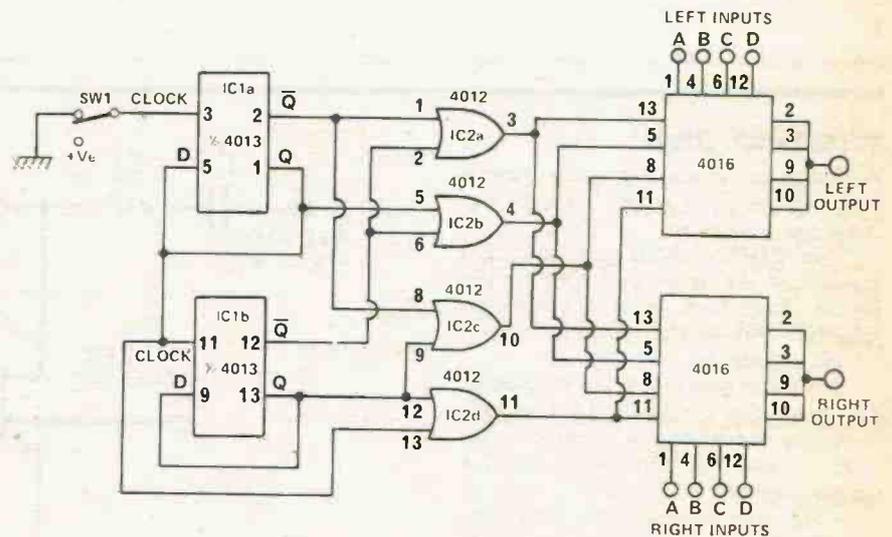
This circuit allows only stereo broadcasts to be outputted by a tuner using either a 1310 or 3090 type stereo decoder chip. In both cases the stereo beacon driver is used to switch the audio output of the tuner. When a stereo signal is being received the beacon driver output is low which turns the Q1 and energises reed relay RL1. The two contacts which switch the output lines are closed and the stereo signal is available at the tuner output sockets. RL1 can be any reed relay with a coil resistance greater than 120 ohms and two normally open contacts.



## Input Selector, Sequencing

Four different inputs can be switched through by the continual pressing of SW1. IC1 is a dual 'D' type flip flop. The Q outputs are connected to the D inputs so that the clock inputs are divided by two. The two flip-flops are connected in series, giving a two stage binary counter.

IC2 is a quad OR gate. This is used to decode the four states of the counter. The outputs are used to control the quad switches of IC3 and IC4 (4016AE).

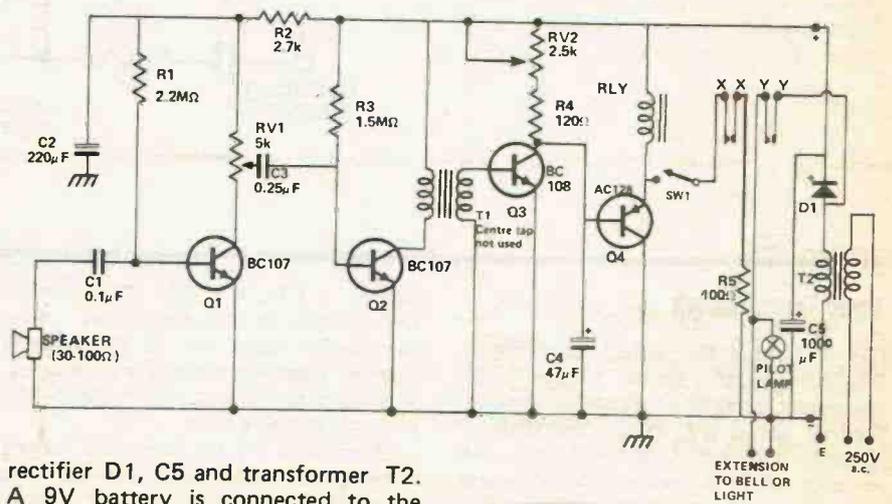


## Audio Switch

Contacts X-X and Y-Y of the relay are normally open. When the relay is energised, current flows through SW1, X-X and R5, thereby locking on the relay. At the same time contacts Y-Y close and current from the secondary of T2 is available at the extension lead sockets, to operate the remote bell or lamp.

If the lock-on switch SW1 is open, contacts Y-Y open when the sound ceases, and this is satisfactory if the warning lamp or bell is likely to receive immediate attention. The pilot lamp LP1 is of aid when setting the unit, as with SW1 open it will show at what sound level operation is being obtained.

For occasional use battery operation is possible by omitting the silicon



rectifier D1, C5 and transformer T2. A 9V battery is connected to the positive and negative points shown, with an on-off switch in circuit. Quiescent current is about 6mA,

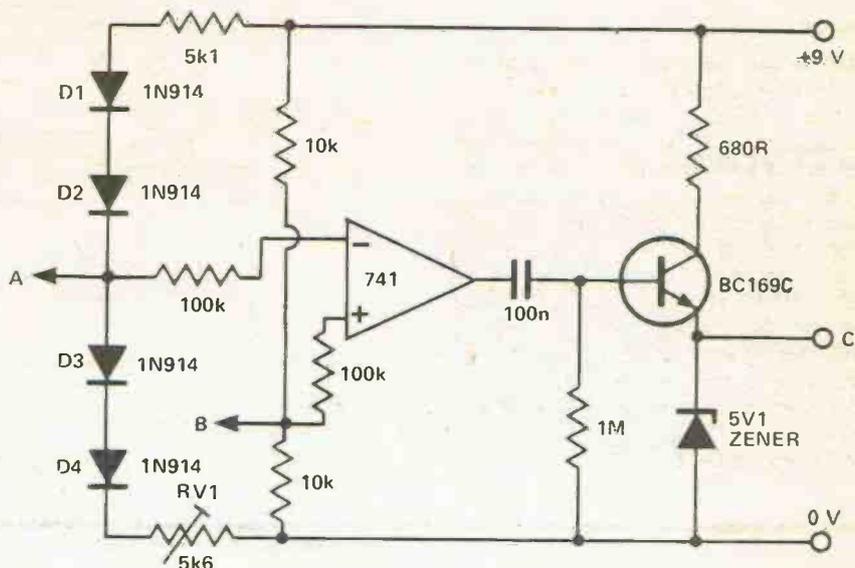
which is not very heavy and will allow a long period of working from a PP9 or similar battery.

## Touch Switch, Thermo

The following touch switch works on the temperature dependence of the forward voltage of silicon diodes. At 0°C this is about 650mV, but drops by 2mV per °C increase in temperature.

When a finger is placed on D3 and D4 the voltage at A will drop below that at B and the O/P of the Op-Amp will go high, causing a TTL compatible pulse to appear at C. D1 and D2 provide compensation against ambient temperature changes. VR1 is initially set so that VA is greater than VB by about 10mV.

The system has the intrinsic advantage that it may be used in moisture-prone conditions in which ordinary touch switches would be most unsatisfactory due to their principle of operation.

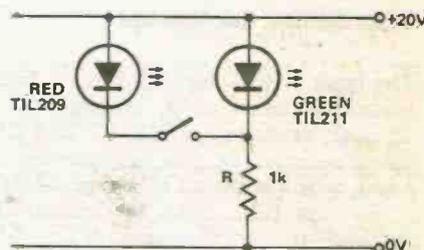


## LED Changeover Circuit

This configuration allows a green LED to be turned off and a red LED turned on by the operation of one 'make' contact only, thus simplifying the design of circuitry to indicate, for example, safe/unsafe or standby/on states.

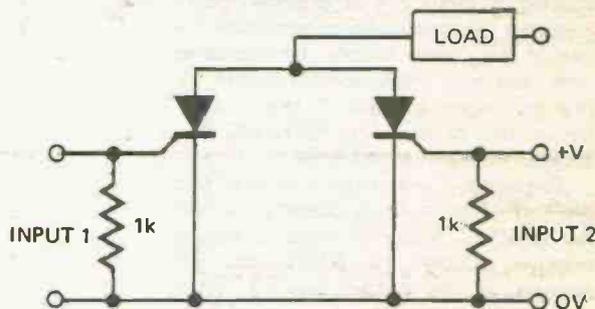
The circuit relies on the fact that a green LED has a slightly higher "on" voltage than a red LED of the same size, and hence is turned off when the red LED is paralleled with it.

For the diode types shown, R should be chosen to give a current drain of about 20mA from the chosen supply rail voltage.



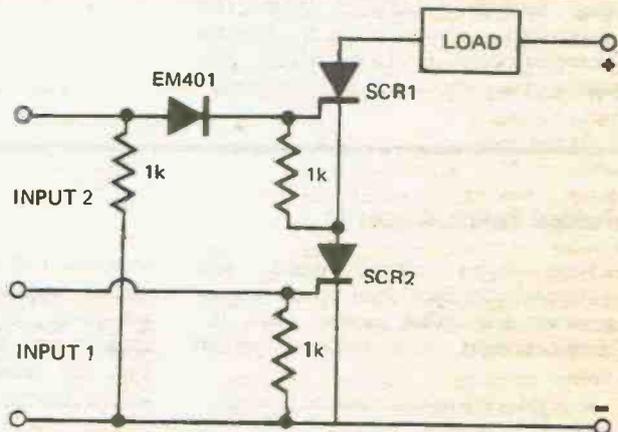
## OR Gate, SCR

An 'OR' gate, again using C106s, is shown. Here, an input to either 1 or 2 will energize the load.



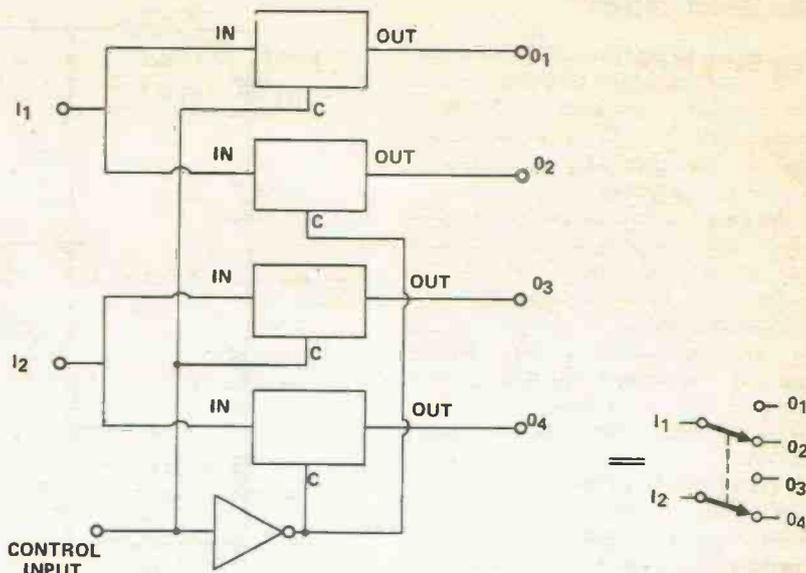
## AND Gate, SCR

Figure shows how a pair of C106s may be used as an 'AND' circuit capable of switching up to four amps. In this circuit, unless inputs 1 and 2 occur simultaneously, no voltage can exist across the load.



## 4016 DPDT Switch

It should be appreciated that the output impedance of the switch is fairly high and so for low signal distortion, a load greater than  $10k\Omega$  is necessary. Using a high supply voltage (10-15V) also helps to achieve this end. The gates will pass signals above the 10MHz mark but as the frequency becomes higher, crosstalk between the switches and distortion will inevitably increase.

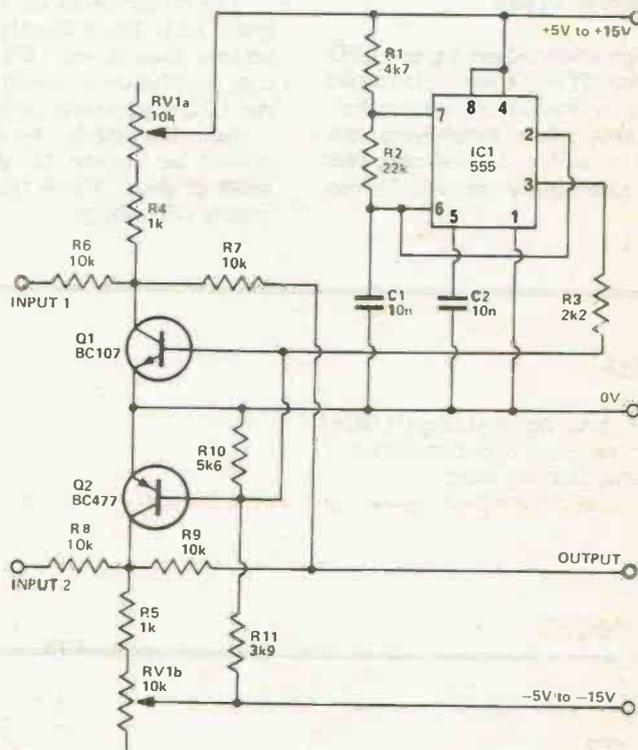


## Beam Splitter, Oscilloscope

The basis of the beam splitter is a 555 timer connected as an astable multivibrator, components R1, R2 and C1 being selected to give approximately equal high/low pulses of about 3 kHz.

Resistor R3 couples the output of the oscillator to the npn/pnp pair Q1 and Q2. When the output of the oscillator is low, resistors R10 and R11 allow Q2 to be on so that any signal applied to input 2 is effectively short-circuited via resistor R8 to the common line of the power supply. At the same time, the npn transistor Q1 is off, so that any signal at input 1, plus a positive voltage provided by RV1a and R4, appears at the output via R7.

Conversely, when the output of the oscillator is high, Q1 is biased on whilst Q2 is off. A signal at input 2 plus a negative voltage via RV1b and R5 appears at the output via R9. Thus signals at the two inputs are alternately displayed on the oscilloscope with a clear separation between them. The separation is controlled by the tandem potentiometer RV1a/b which also varies the amplitude of the traces.



## Twilight Switch, Automatic

Here is a circuit which will automatically light your porch light or activate any other device when the ambient light drops below a certain level.

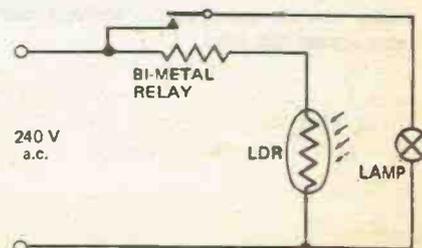
A light dependent resistor is used in series with a relay.

The resistor has a value in excess of 1 megohm when illuminated, this drops

to below 110 ohms when dark.

It is important that the LDR be positioned in such a place as not to receive any spurious illumination as this will cause the relay to drop out intermittently.

A bimetallic strip type relay will give sufficient delay to ensure that incident light flashes have no influence.



## Time Delay Switch

IC1a is provided with resistive and capacitive feedback to form an integrator with initial conditions. IC1b is in an "open loop" mode so that its output is either high or low depending on its inputs, and changes state when the output of IC1a goes more negative than the voltage set at ZD2. When the output of IC1b goes positive the transistor Q1 biases hard on switching the SCR on. Diodes D1-D4 are to make the SCR conduct on both halves of the mains wave form.

The delay period is set by the components ZD1, ZD2, C, RV1, and

R. If ZD1 is chosen to be 0V5 and ZD2 at 5V, then the maximum delay period is given by  $T = 10 \cdot C \cdot R$ .

$$RV1 = \frac{ZD2}{ZD1} \times R \leq 10 R$$

The meter is a voltmeter with a fsd equal to the value of ZD2. The switch then operates when the meter reaches fsd. The meter can therefore be calibrated to show remaining delay with 0V equal to T and fsd equal to zero.

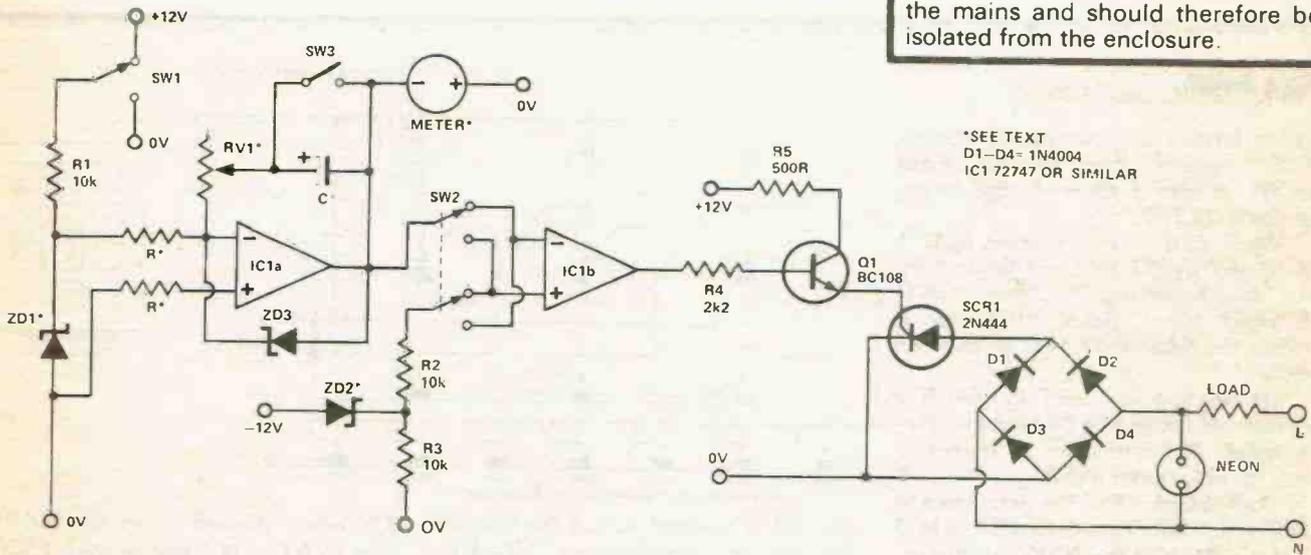
SW2 changes round the inputs of the op-amp so that the output either swings from high to low, or, low to high. SW3 is to reset the time delay which it does by discharging the capacitor. ZD3 should be chosen

to be a value slightly higher than ZD2, this is to stop the capacitor charging beyond a set limit and therefore overloading the meter. SW1 is the run-hold switch. When the switch is at +12 volts the integrator charges the capacitor. When the switch is set to 0V the charging of the capacitor is stopped until the switch is set back to 12 volts.

Q1 is a buffer to avoid loading on the IC and to trigger the SCR. The supply voltage should be 12-0-12 and does not need to be well smoothed as the zener diodes set the timing function.

### Warning

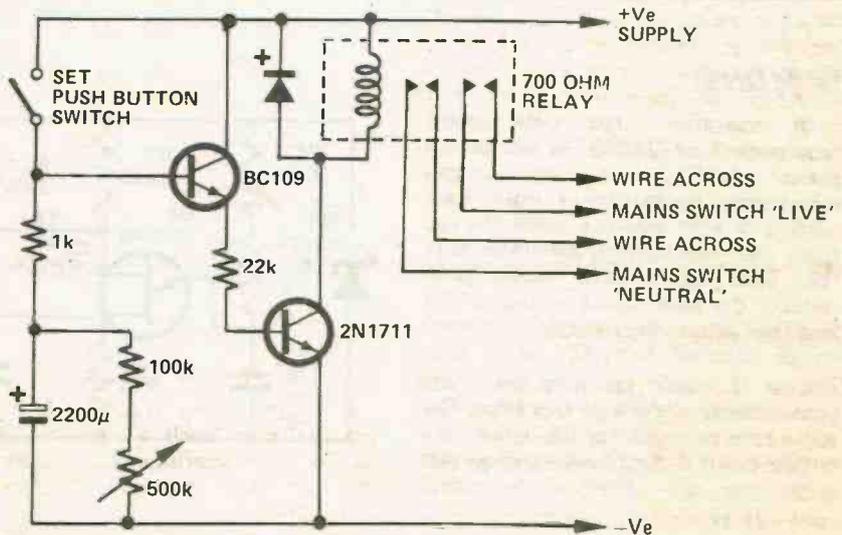
The circuitry is not isolated from the mains and should therefore be isolated from the enclosure.



## Snooze Delay Unit

When the Set switch is depressed the large electrolytic capacitor is charged via the limiting resistor (1k). This charge causes the BC109 to conduct which supplies enough base current to switch on the 2N1711 space and operate the relay. The relay contacts are wired in parallel with the mains switch so that if the mains switch is now turned off, the equipment will continue.

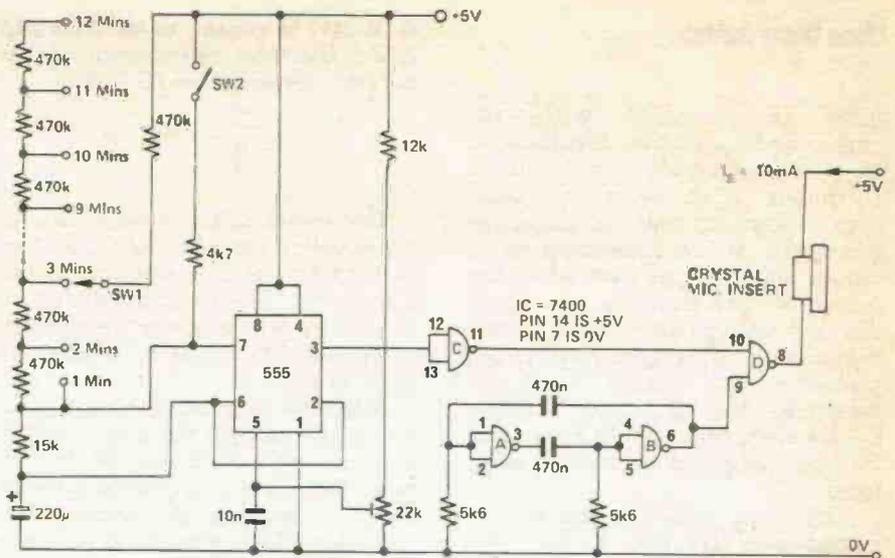
The supply voltage is taken from the equipment in which the unit is fitted and will determine the choice of relay. The maximum delay being 1.75 hours.



## Timer, 1-12 Minutes

The circuit is centred around the NE555V timer which provides a logic 0 level at pin 3 every 1 - 12 minutes, depending upon the position of the 12 way switch (SW1). SW2 is a push-button switch which synchronises the first pulse (originally a switch circuit was fitted to the reset pin, 4, but the first timing period was never the same as the subsequent periods). The variable 22k resistor allows a degree of fine adjustment for the timing period.

Gates A and B form a tone generator. Gate C inverts the output of the NE555V so that a logic '1' is fed to gate D at the end of each timing period. Thus a tone burst of a few seconds is produced by the transducer (any surplus crystal microphone insert should be suitable).



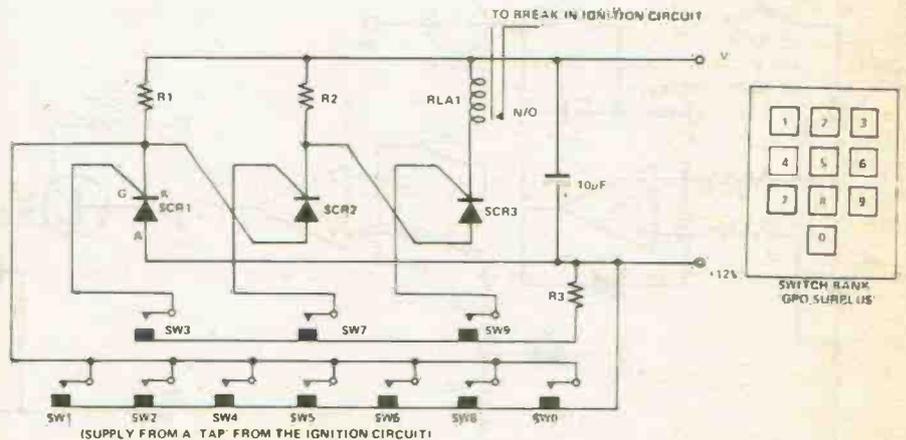
## Code Switch

When button 3 is pressed R3 'Gates' SCR 1, which remains on with a load of R1. It also supplies voltage to the anode of SCR 2.

When button 7 is pressed SCR 2 is 'Gated' by R3 also, and held on by R2, thus supplying the anode of SCR 3, which when 'Gated' by button 9 closes the relay and makes an external circuit.

It can also be used to switch a circuit off depending on how the relay is wired. This would be an advantage in a home intruder alarm.

Components: The Thyristors can be any type and values for R1 and 2 selected to hold the SCR's in conduct-

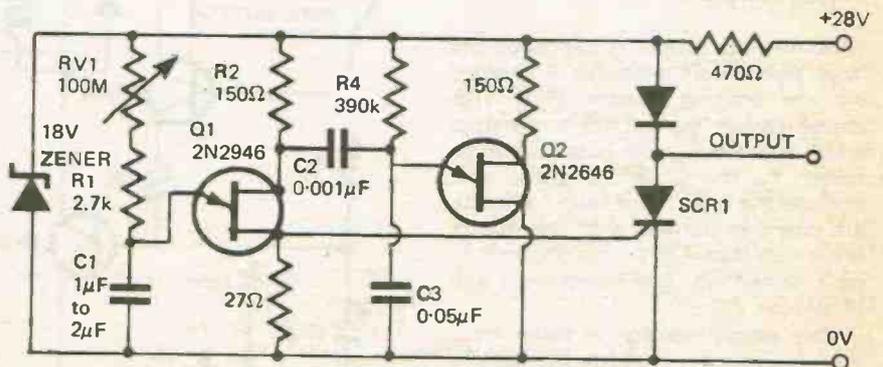


ion. R3 is selected to suit the thyristors. The remaining buttons, 1,2,4,5,6,

8,0 when pressed short out SCR1, thus switching off any following SCR.

## Timing Circuit

In operation, the peak point requirement of UJT Q1 is reduced to about 1/1000 of its normal requirement by pulsing its upper base with a ¼ volt negative pulse derived from the free running oscillator UJT Q2. This regular pulse momentarily reduces the peak point voltage of Q1 and thus allows the peak point current to be supplied from C1 rather than R1, as it would be with the more conventional circuits of this type. The pulse rate of oscillator Q2 is not very critical but it should have a period that is less than one fiftieth of the overall time delay.



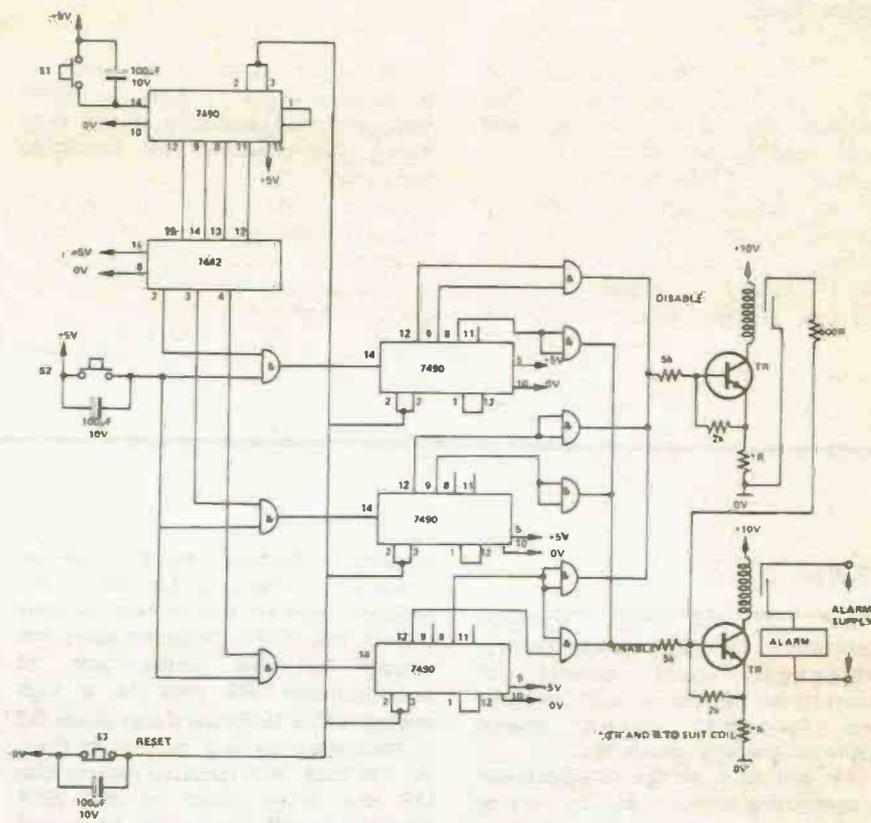
## Combination Lock

The circuit and switching system is simplified by the use of a multiplex system. S1 inputs pulses to the decade counter 7490. The resulting BCD is decoded by the 7442. It is the decimal output of this which carries out the multiplexing via the AND-gates.

S2 inputs pulses which are transferred to the other 7490 decade counters by the AND-gate multiplex system. The BCD output from the 7490's is taken to the AND-gates whose outputs control the Alarm 'Disable' and 'Enable' switch system.

The 'Disable' function effectively prevents TR2 from being biased on and hence prevents the 'Enable' Reed relay from working.

This circuit has several advantages over conventional electronic combination locks as only two switches need be installed on the object to be guarded, regardless of the number of figures in the combination. The value of the example combination is 314. The alarm is triggered if any of these digits is exceeded in value. While the circuit is capable of directly driving an actuator it is recommended that it is only used to disable an alarm system —



conventional locks doing the actual locking. (To operate the example the switch sequence would be: S1, S2, S2, S2, S1, S2, S1, S2, S2, S2, S2.)

## Flexible Timer

The precision time delay circuit shown will provide accurate and repeatable time delays adjustable from a few milliseconds to a minute or two. This is a very flexible circuit in which the operating current and voltage depends only on the choice of SCR.

The timing sequence may be initiated either by applying power to the circuit — or by opening a shorting switch wired across C1. Timing capacitor C1 is charged via R1 and R2 until the voltage across C1 reaches the peak point voltage of the UJT Q1. When this occurs, Q1 fires, generating a pulse across R4, triggering the SCR, and applying power to the load. Holding current for the SCR is provided via R5 and D1.

The circuit is reset by momentarily removing the supply voltage.

If the circuit is to be used in an application where both rapid cycling

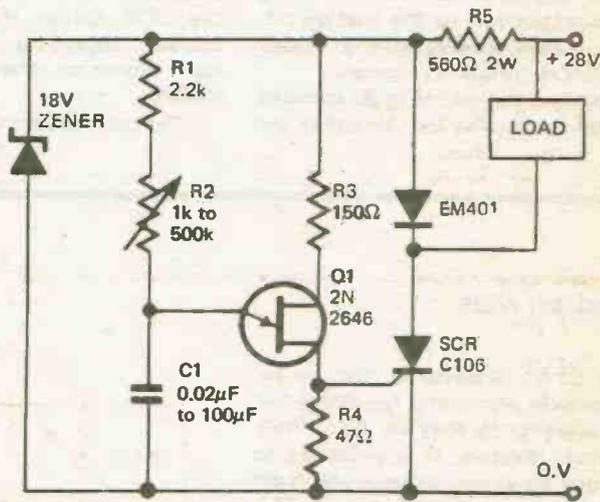


FIG. 12

and accurate, repeatable timing is required — some provision must be made to ensure that C1 is discharged to zero before each timing sequence. This can most easily be done by interconnecting a pair of switch contacts with the reset system so as to momentarily short out C1 whenever the circuit is reset.

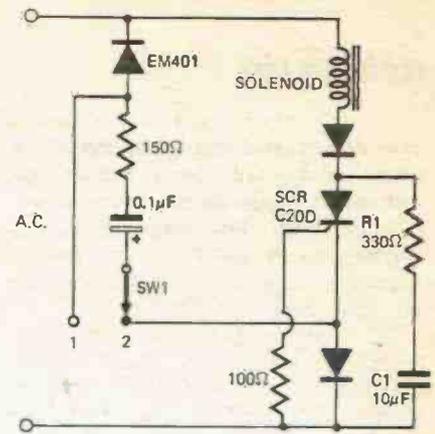
Temperature compensation for this circuit is provided by R3. Increasing the value of this resistor causes the circuit to have a positive temperature coefficient. It is possible to obtain zero coefficient over a small range of ambient temperatures by optimizing R3.

## Impulse Power

This circuit is often used in electrically powered stapling machines, impulse hammers etc, and causes load current to flow through the load for one complete half-cycle of the ac supply whenever SW1 is actuated (i.e. moved from its normal position (1) to energise-load position (2)). The circuit is arranged so that the SCR is always triggered at the

beginning of a positive half-cycle of the ac supply, even though the switch may be closed randomly at any time during the previous two preceding half-cycles.

Resistor R1 and capacitor C1 should be chosen so that their series combination supplies just sufficient holding current for the SCR for one complete half-cycle.



## Half-Wave Control

One of the most common applications for SCR phase control systems is speed control of commutator motors — such as those used for food mixers, sewing machines, pottery wheels etc.

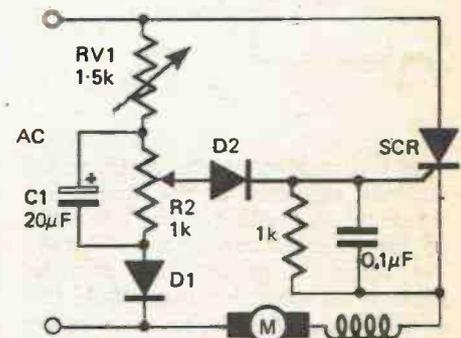
However one of the disadvantages of controlling motor speed by varying input power is that as the effective power input is reduced to slow down the motor — the torque available is reduced as well.

This may be overcome by using a feedback signal to advance the firing angle in proportion to the load on the motor — thus increasing the power input if more torque is required.

The circuit shown in Fig 20 achieves this load compensating function by

deriving a feedback signal from the armature back-emf (produced by the residual field of the motor). In this circuit, the SCR is triggered when the voltage on the wiper arm of potentiometer R2 rises to a high enough value to forward bias diode D2 — thus allowing gate current to flow. As the back emf tends to reverse bias D2, the firing point of the SCR depends largely upon the back emf and this in turn is a function of speed. If the motor is loaded, the speed reduces, thus also reducing the back emf — hence D2 becomes forward biased earlier in the cycle (triggering the SCR earlier in the cycle), and thereby supplying the motor with more power to offset the effect of the loading.

The component values shown are

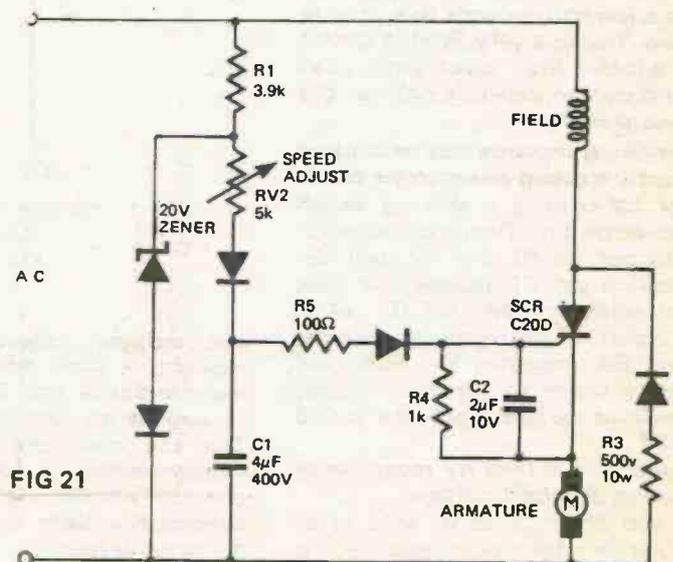


suitable for most fractional horsepower motors — for optimum results it will be necessary to adjust component values to suit the motor used.

The circuit described above will provide stepless speed control over a wide range of motor speed — but tends to cause jerky operation at low speeds.

## Improved Half-Wave

The above drawbacks can be almost entirely overcome by using the circuit shown. As may be seen from the circuit diagram, it is necessary to bring out separate connections from the armature and field windings. This is generally a simple operation and providing it can be done the circuit will provide stepless speed control down to virtual standstill. In this circuit the 20V zener diode provides a constant voltage for the discharge of C1. Capacitor C2 and resistor R4 are connected from gate to cathode of the SCR to stabilize the circuit by preventing the SCR from being triggered by extraneous signals.

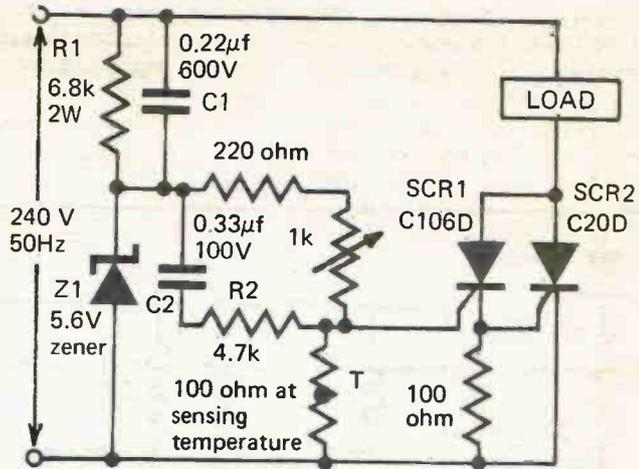


## Zero Switching

A very simple yet effective zero voltage switching provides half-wave control only, but is satisfactory for commercial applications where the heating elements can be designed to suit.

The circuit is extremely stable and unaffected by quite large variations in line voltage and ambient temperature. The response time depends upon the characteristics of the thermistor which is used — times of one to two seconds are typical. The sensing differential is around  $\frac{1}{4}^{\circ}\text{F}$  at normal ambient temperature.

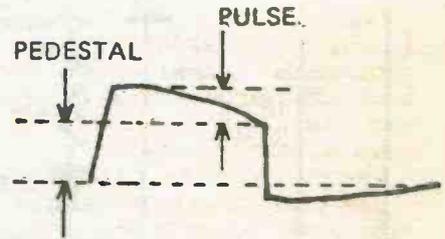
The Zener diode Z1 forms a voltage pedestal of 5.6 volts nominal amplitude by clipping the incoming positive half-cycle of mains voltage. This pedestal is differentiated by R2, C2 and associated resistors to form a pedestal of reduced amplitude with a pulse superimposed on top of the pedestal. This waveform is applied to the gate of SCR1. The capacitor C1, which is connected in parallel with R1, provides a leading phase shift to the pedestal so that SCR1 is triggered into conduction by the peak of the positive decaying pulse which is superimposed on the pedestal. It does this at the beginning of the positive going half-cycle



of line voltage appearing at the anodes of both SCRs.

The thermistor controls the amplitude of the pedestal and thus provides a semi-proportional control with a small temperature differential.

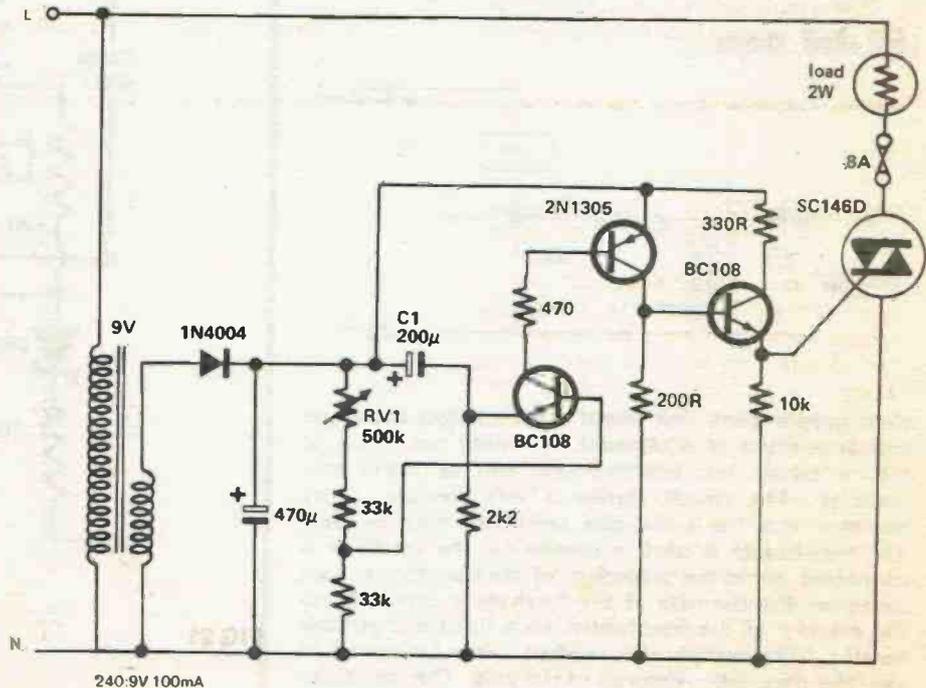
The lock-in configuration of SCR1 and SCR2 reduces the effects of ambient temperature variations. The cost of this circuit is very low compared to a phase control circuit of the same power handling capacity as no rfi components are required.



## Triac Lamp Flasher

The circuit is a relatively simple triac lamp flasher, probably of most interest to those in the disco business. The flasher will handle a load of up to 2kW with a variable flash rate of about 20/200 flashes per minute, achieved by altering the value of RV1.

C1, the timing capacitor, can be experimented with to obtain the most satisfactory results. Even though little power is dissipated in the triac (15W on full load), it should be mounted on a heatsink.



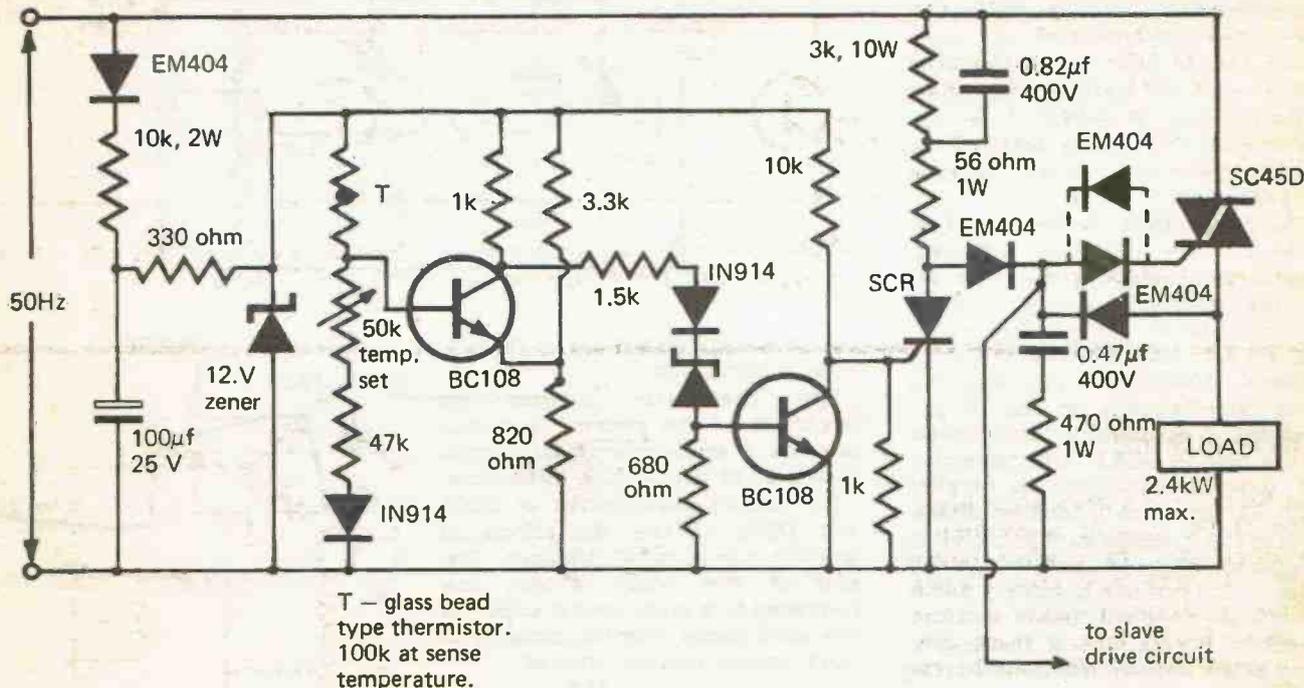
240-9V 100mA

## Triac Slave Controller

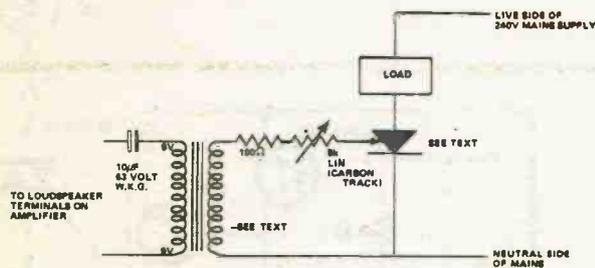
The circuit shown will provide full-wave control of heating loads of almost any size. The triggering

circuit will drive Triacs of any size from 1 amp up to 125 amps. In addition almost any number of additional Triacs can be slave driven by the main triggering circuit.

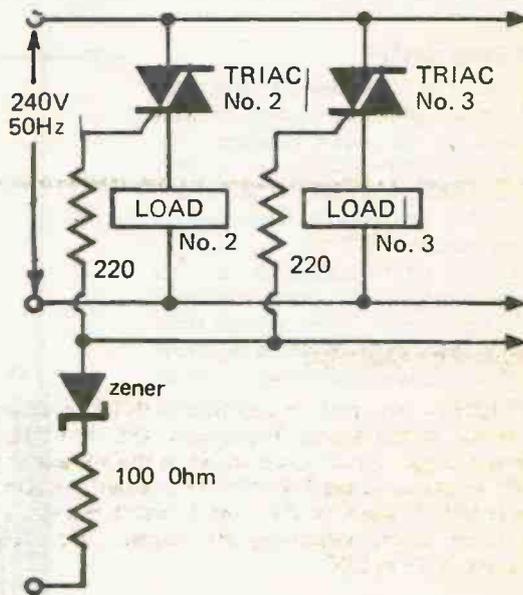
The differential of this circuit is approximately  $\pm 1/6^{\circ}\text{C}$ . This circuit has a semi-proportional action, and is suitable for applications where large amounts of power have to be controlled accurately and at low cost.



## Light Show, Simple



Most people think that sound activated light shows are expensive items of equipment, this need not be so, in fact a simple but effective unit can be made very cheaply. The circuit shown is very flexible. Any thyristor that has a low gate sensitivity may be used. The transformer is used in reverse i.e. the amplifier is connected across the secondary of the transformer via a capacitor and the gate of the thyristor is connected to the primary of the transformer via a fixed and variable resistor. The components marked \* can be altered to vary the frequency response of the unit. The sensitivity of the unit is not as high as more expensive units.

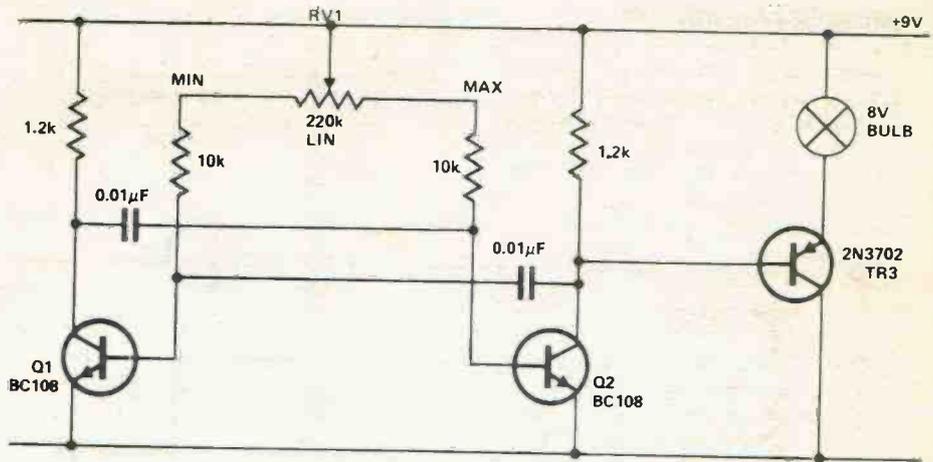


## DC Lamp Intensity

A simple and inexpensive unit for reducing the brilliance of a lamp and at the same time reducing the current drain, thus increasing battery life considerably. A power saving does not happen by merely inserting a resistor in series with the bulb, hence the reason for the above circuit.

Q1 and Q2 form an astable multivibrator, the RV1 varying the mark/space ratio. The output from Q2 collector is fed to Q3 base, either saturating Q3 or turning it off. Varying the mark/space of the lamp.

A notable point is that as Q3 is either fully 'on' or 'off' it need not be a high wattage type. As an

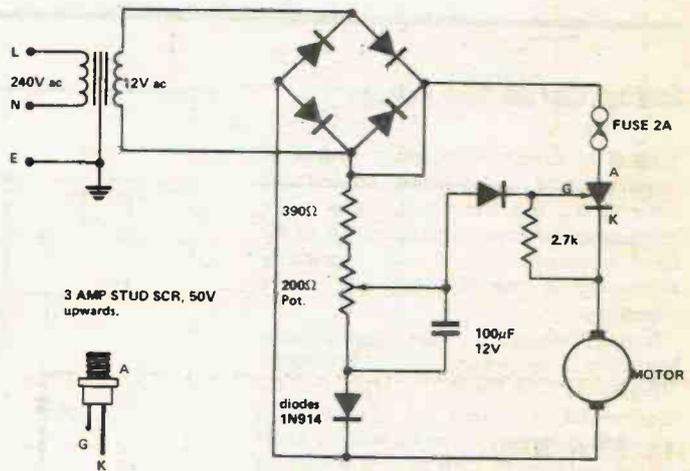


example, when Q3 is saturated,  $V_{ce} = 0.3V$  approx, and the lamp current is

$$0.2A \text{ the power across } Q3 \text{ is } W = V \times I = 0.3 \times 0.2 = 0.06.$$

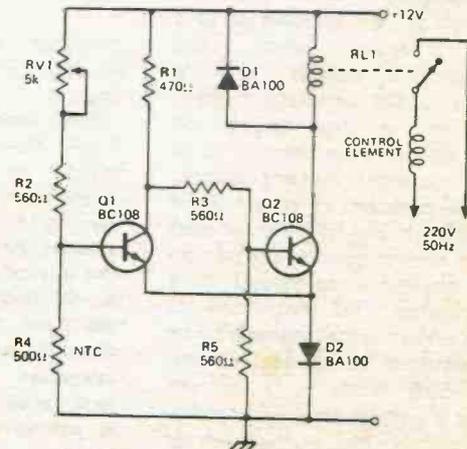
## Train Speed Control

The following is a low voltage adaptation of the type of speed control popularly used to regulate power drills. It gives very good starting torque and excellent speed regulation of the model. A reversing switch may be incorporated in the leads to the motor.



## Temperature Controller

A negative temperature coefficient (NTC) resistor is used to sense temperature. Transistors Q1 and Q2 form a Schmitt trigger which switches when the voltage at the base of Q1 increases above 1.4 volts. Thus when the temperature falls below that set by RV1 the Schmitt changes state and the relay opens switching the heater 'ON'. Regulation accuracy is 1° to 2°C.



# POWER SUPPLIES

## Current Source, Drift Free

The conventional type of constant-current source, as shown in Fig. 1, will drift in output current immediately after switch-on. This is because of the voltage drop across Q1, causing a significant amount of power to be dissipated in the transistor, heating it and its Vbe. Hence the output current slowly increases after switch-on, typically reaching a stable value about two minutes later. In tests the current increased by about 4% for a small signal transistor dissipating 100mW.

This effect is greatly reduced by the configuration shown in Fig. 2, which fixes the voltage across Q1 at a very low level by virtue of the common-base transistor Q2. The main voltage drop occurs across Q2, leaving about 600mV across Q1, this being set up by the two extra diodes in the bias chain, (D1, D2) which fix the emitter potential of Q2.

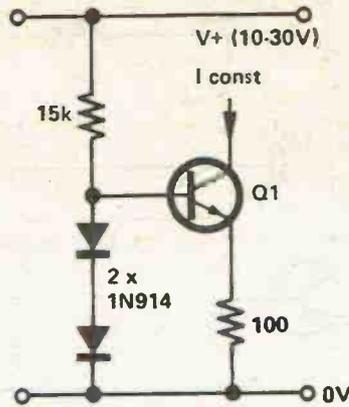


Fig. 1

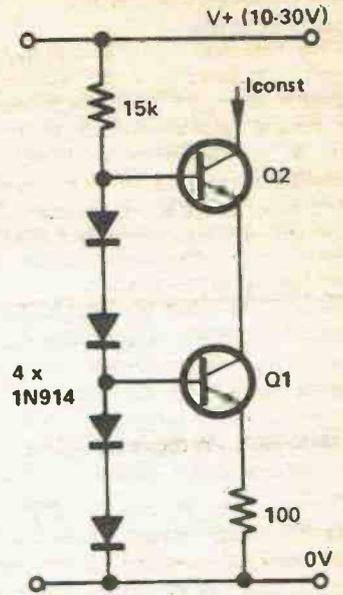


Fig. 2

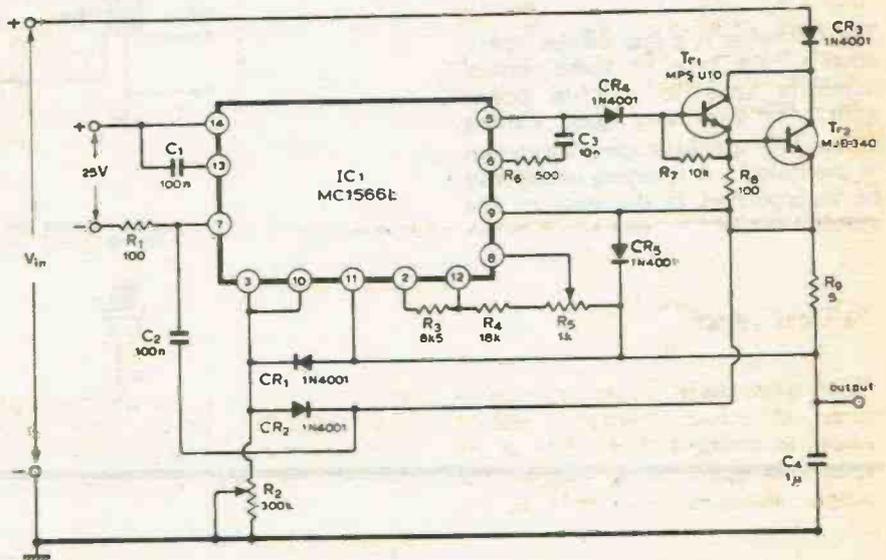
## Constant Current, High Voltage

WHEN a constant current source is required and the various advantages offered by the use of IC's are to be exploited, an input voltage limit of 40 or, possibly, 50 V is normally necessary if the IC's are not to be damaged.

Neil Wellenstein, an applications engineer working in Motorola's Phoenix, Arizona, laboratories, discovered a means of obtaining a variable constant current supply with input voltages as high as 750 V using a standard regulator IC. In fact, the input voltage is limited only by the breakdown voltage of the series pass transistors employed.

The IC used by Wellenstein was the Motorola MC1566L which has the ability to "float" on its own output voltage. However, when used conventionally, a voltage sensitive error occurs in the constant current mode and this is large enough to prevent the device from being used as a precision constant current source. Normally the constant current feature of the MC1566L would only be used to provide short circuit protection when the device is employed as a voltage regulator. The magnitude of the current error is small enough to be of no consequence in this application.

The MC1566 contains a current sensing and a voltage sensing amplifier which "float" on the output voltage and which are supplied from an on-chip regulator. The on-chip



regulator receives its input from an auxiliary 25 V supply external to the chip.

When used conventionally a constant 1 mA flows from pin 3 through a resistor to ground to establish the reference voltage for the voltage sensing amplifier. The error voltage appears between pins 8 and 9. When the device goes into the current limit mode (short circuit conditions) part of the 1 mA output from pin 6 can flow through a diode to pin 9 thereby upsetting the error voltage and producing a voltage sensitive output current error.

Wellenstein discovered by reversing the roles of the voltage and the current

sensitive amplifiers, he could eliminate this problem altogether. The net effect is that any portion of the reference current that appears in the load must pass through the current sensing resistor (R9) which cannot be bypassed as was previously the case.

The maximum input voltage to the circuit is limited by the series-pass transistor. In the case of the MJE340 shown, the maximum input voltage is 300 V. The circuit provides a constant current output which is adjustable from 200  $\mu$ A to 100 mA; above 10 mA take care not to exceed the ratings of the MJE340. At both the 200  $\mu$ A and the 1 mA settings, output impedance exceeds 20 M ohms.

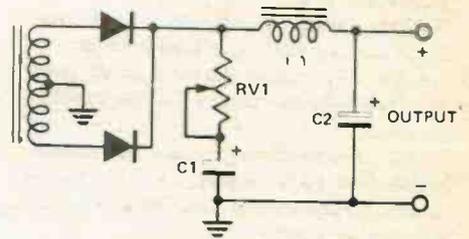
## Output Voltage Adjustment

When the output voltage of a power supply comes out a little higher than expected it can be adjusted by making the simple addition illustrated above. R1 is a 500 ohm or 1000 ohm potentiometer (10 watt rating) inserted in series with the input filter capacitor. Adjust it to give the correct

voltage under load

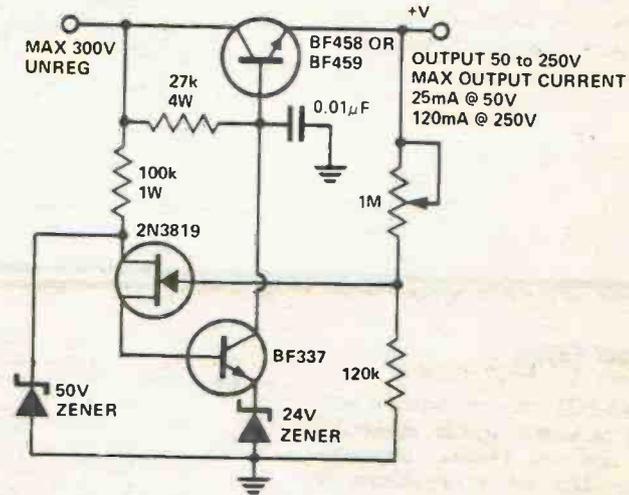
For low voltage supplies (i.e. up to 50 V or so) a 50 ohm or 100 ohm pot might be more suitable.

The pot could be connected between the negative lead of C1 and ground and would not then need to be insulated.



## High Voltage, Variable Regulator

This regulator is ideal for SSB linear amplifier tube screens. It would also have application in the repeller supply for a reflex-klystron microwave oscillator. CRO deflection amplifier supply is another possible application. Regulation is about 0.5%. The output transistor will need to be mounted on a small, insulated heatsink. A BF459 is preferred (30V V<sub>ceo</sub>) as the BF458 is sailing a bit close to the wind when the output is down to 50 V.

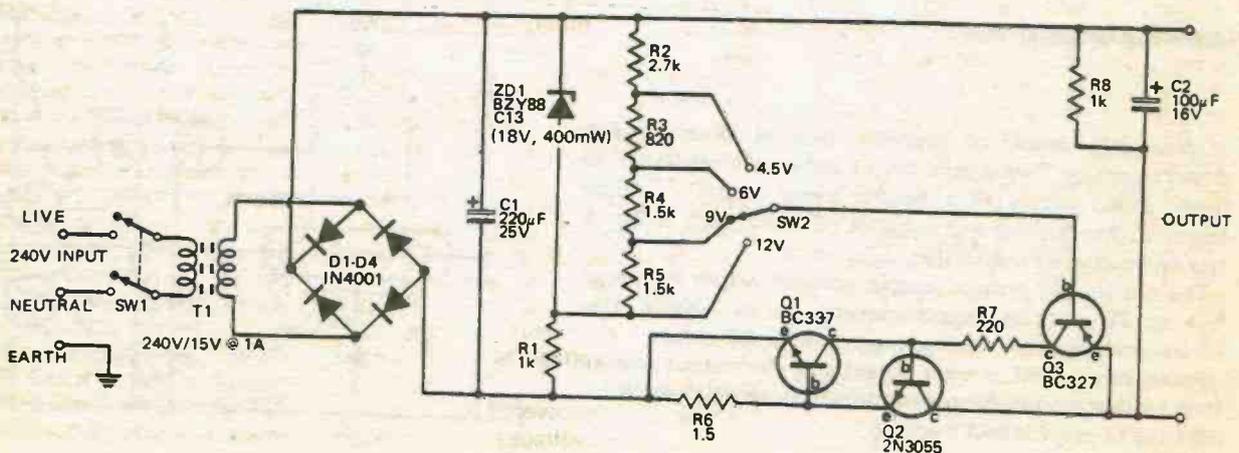


## Switched Output

THIS little power supply provides a range of switch selectable output regulated voltages from 4.5 to 12 volts, selectable by a switch. 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.

Nominal output voltage 12V, 9V, 6V and 4.5V  
Output current 0 – 400mA  
Current limit approx. 500mA



# POWER SUPPLIES

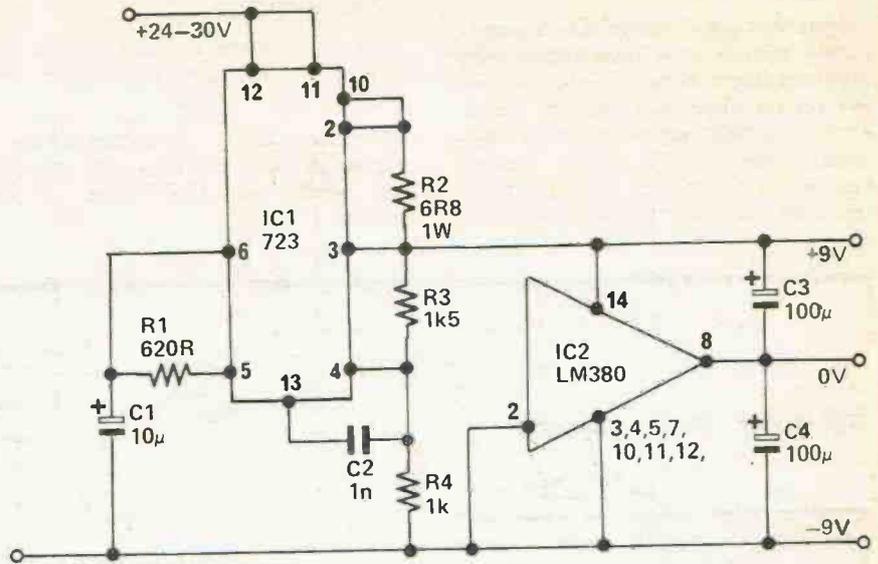
## Dual PSU

Anyone who experiments with op-amps will need a dual stabilised power supply. The circuit shown was designed to power up to twenty 741s simultaneously.

IC 1 is used to provide an output voltage of 18 V at pin 10. R2 sets the short circuit current at 100 mA and should be rated at 1 W.

IC 2 is used as a precision short circuit proof voltage divider.

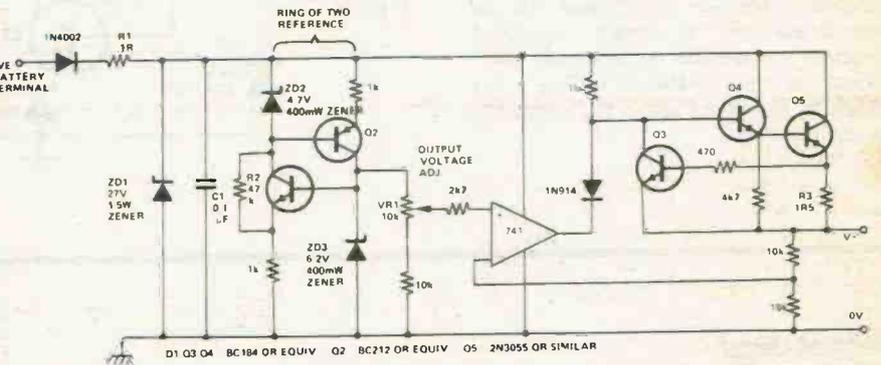
A 9.0-9 V supply has been found in practise to cater for most op-amps with the notable exception of the CA3130 which has a maximum supply rating of 16 V.



## Mobile Power Supply

R1, C1 and ZD1 provide clipping and smoothing of supply spikes, while D1 protects against reverse polarity connection. The reference voltage is provided by the ring-of-two, since in this configuration the zeners bias constant-current sources for each other, the output across ZD3 is almost totally independent of supply variations. R2 ensures the ring starts reliably.

A set fraction of the reference voltage is applied, via VR1, to the 741, which in conjunction with current amplifiers Q4, Q5 forms a negative feedback loop to maintain the output voltage constant. It may be set



between 10 and 6 volts, so, for instance, most battery cassette equipment may be driven. Short-circuit protection is provided by Q3; when the output current exceeds 400mA

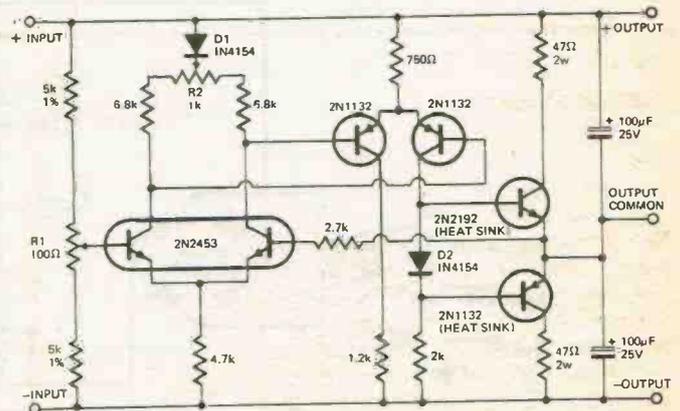
sufficient voltage is dropped across R3 to turn on Q3, which shunts drive away from the base of Q4 and hence prevents the output current from rising further.

## Converting Single to Dual

Operational amplifier circuitry requires double-ended power supplies. This simple circuit converts a conventional single ended supply to a double ended operation. Once adjusted, the positive and negative rails will track within a few millivolts without further adjustment.

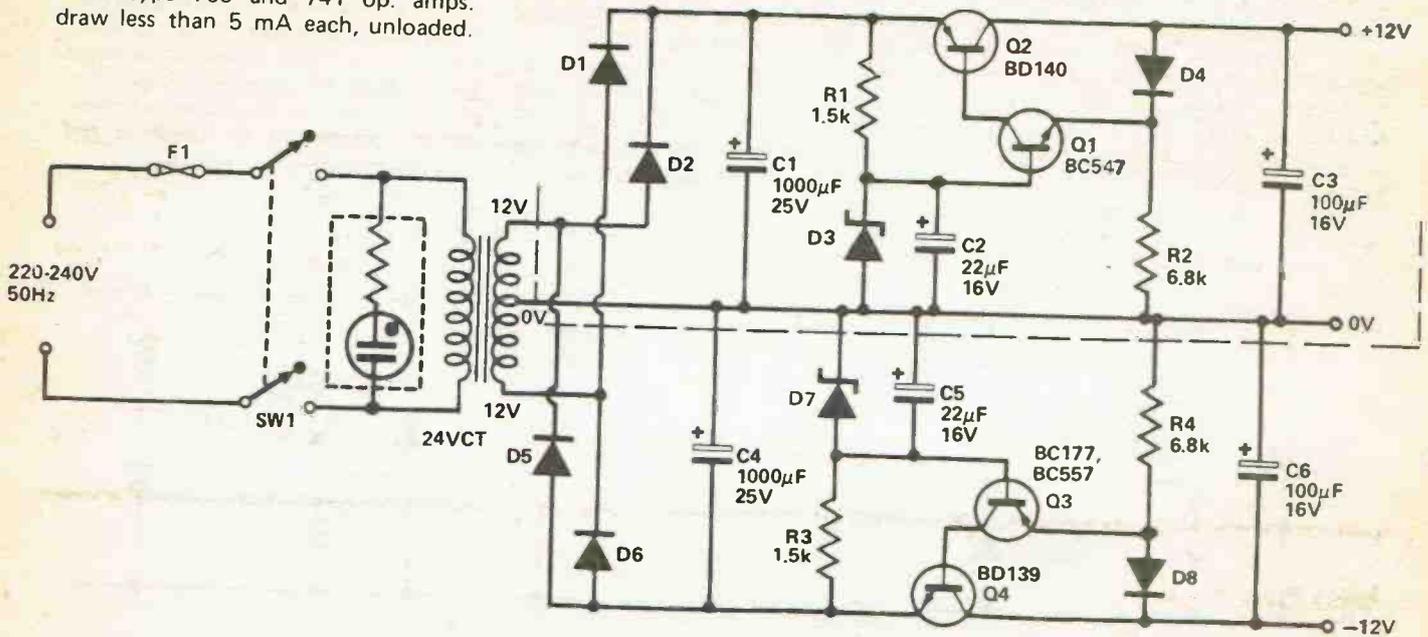
The circuit will provide output voltages within the range five to 25 volts at output currents up to 100mA. The corresponding supply voltage range is 10 to 50 volts.

Potentiometer R1 is used to balance the output voltage (test by precision divider network) and potentiometer R2 is adjusted to provide best tracking.



## Op-Amp Supply

The rated output, of 100 mA from either side, will be found to be more than adequate for the intended use, since type 709 and 741 op. amps. draw less than 5 mA each, unloaded.

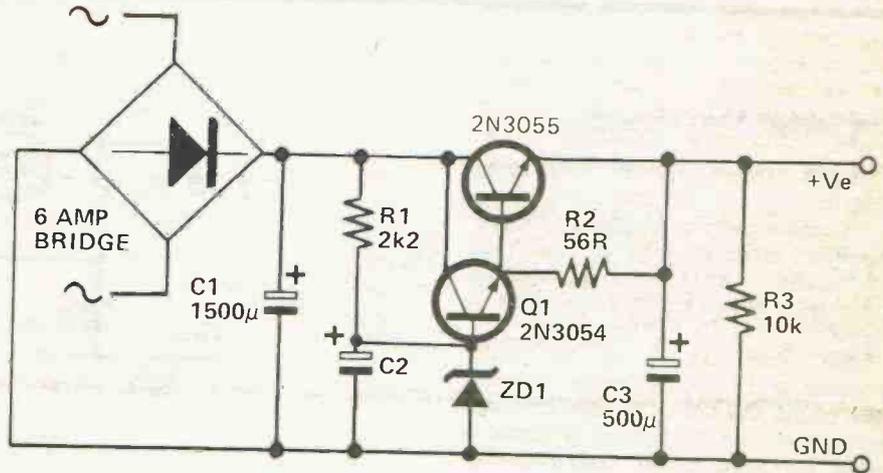


## Low Ripple PSU

The power supply circuit shown may be used where a high current is required with a low ripple voltage (such as in a high powered class AB amplifier when high quality reproduction is necessary).

Q1, Q2 and R2 may be regarded as a power darlington transistor. ZD1 and R1 provide a reference voltage at the base of Q1. ZD1 should be chosen thus:  $ZD1 = V_{out} - 1.2$ .

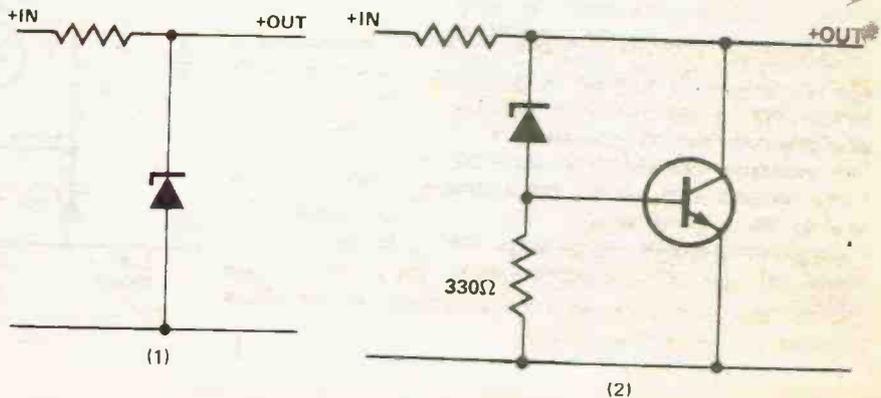
C2 can be chosen for the degree of 'smoothness' as its value is effectively multiplied by the combined gains of Q1/Q2, if 100µF is chosen for C2, assuming minimum  $h_{fe}$  for Q1 and Q2,  $C = 100 \times 15(Q1) \times 25(Q2) = 37,000 \mu F$ .



## Zener Assistance

The simple zener shunt of diagram (1) may not handle sufficient current if the zener available is of low wattage. A power transistor will do most of the work for the zener in circuit (2).

The output voltage is increased by 0.7V but it is stabilisation rather than exact voltage which is often required.



# POWER SUPPLIES

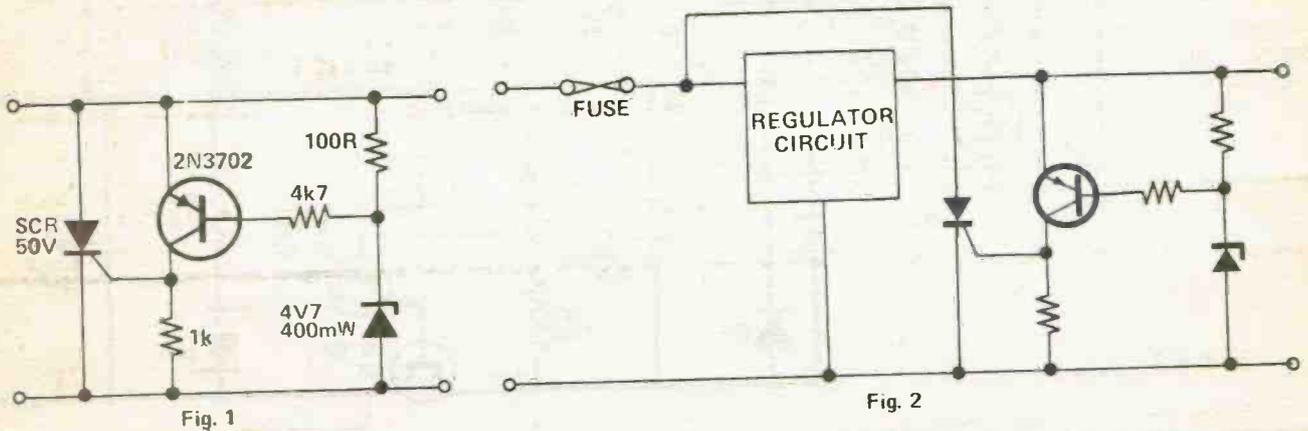
## Crowbar, Simple

This circuit provides overvoltage protection in case of voltage regulator failure or application of an external voltage. It is intended to be used with a supply offering some form of short circuit protection, either foldback, current limiting or simple fuse. The circuit is less effective in the latter case however, as a good deal of damage can be done in the time taken to blow a fuse.

The most likely application is a 5V logic supply, since TTL is easily damaged by excess voltage. The values chosen in Fig.1 are for a 5V supply, although any supply up to about 25V can be protected by simply choosing the appropriate zener diode. When the supply voltage exceeds the zener voltage +0.7V, the transistor turns on and fires the thyristor. This shorts out the supply, and prevents the voltage rising any further. In the case of a supply

with only fuse protection, it is better to connect the thyristor across the unregulated supply as shown in Fig.2 to prevent damage to the regulator circuit when the crowbar operates.

The thyristor should have a current rating about twice the expected short circuit current and a maximum voltage greater than the supply voltage. The circuit can be reset by either switching off the supply, or by breaking the thyristor circuit with a switch.



## Low Voltage, Short Protection

These short-circuit protected stabilisers give 6, 7.5 and 9 V from an automobile battery supply of 13.5 V nominal, however, they will function just as well if connected to a smoothed dc output from a transformer/rectifier circuit.

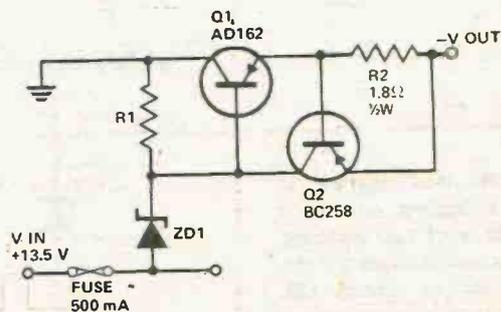
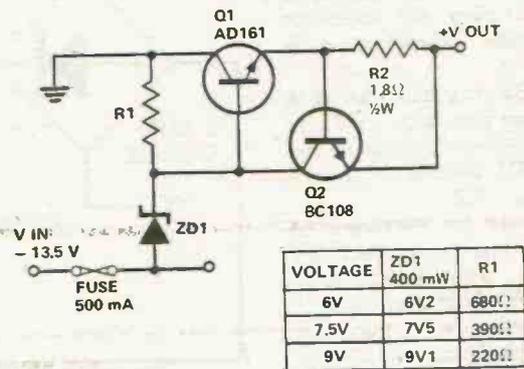
Two types are shown for both positive and negative earth systems. The power transistors in each case can be mounted on the heatsink without a mica insulating spacer thus allowing for greater cooling efficiency.

Both circuits are protected against overload or short-circuits.

The current cannot exceed a value of 330 mA. Under normal operating conditions the voltage across R2 does not rise above the 500 mV necessary to turn Q2 on and the circuit behaves as if there was only Q1 present.

If excessive current is drawn, Q2 turns on and cuts off Q1 protecting the regulating transistor.

The accompanying table gives the values of R1 for different zener voltages.



## Low Ripple at Low Current

In the normal circuit (Fig. 1) the ripple at 1 amp is at least 2 volts. Cheap power amps use this circuit (with low supply ripple rejection) and produce annoying amounts of hum at low signal levels.

In the circuit in Fig. 2 the ripple is considerably reduced at low levels and at high currents the supply voltage is only minimally affected.

Maximum low ripple current  $(I_m) = V_z/R$  where  $P_{tot}$  must be more than  $V_z^2/R = I_m V_z$ .  $I_m$  = maximum total current so  $P_{tot} = I_m \cdot I_m V_z$ . A typical set of values for  $I_m = 1/2$  Amp is  $V_z = 3V$ ,  $R = 1 1/2$  ohms.

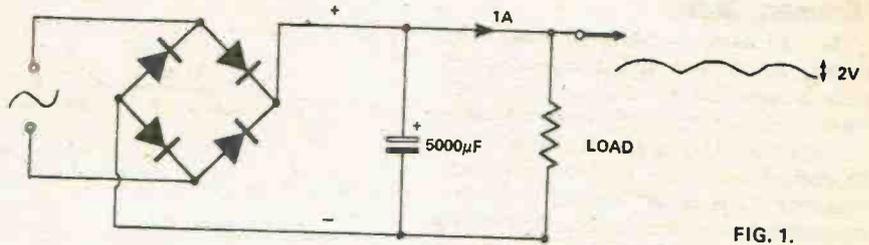


FIG. 1.

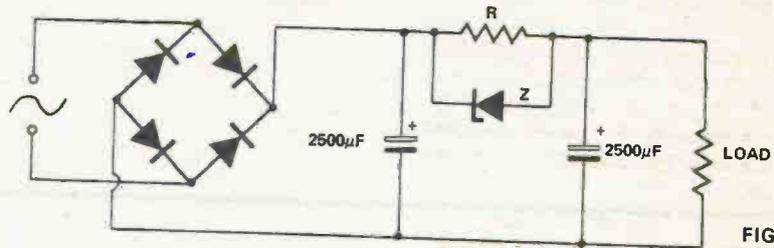
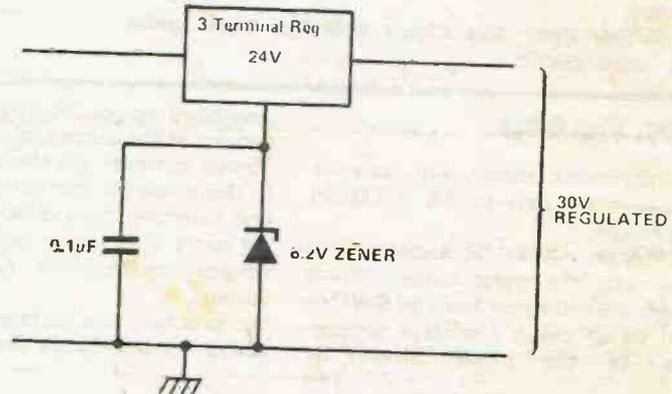


FIG. 2.

## 30 Volt Regulators

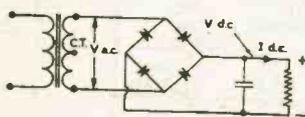
Three-terminal voltage regulators are available in 5,9,12,15,18 and 24V types. If you require a 30V supply use a 24V regulator with a 6.2V zener diode in the earth lead as shown. This increases the voltage to 30V. A 0.1µF capacitor should be connected across the zener diode as shown.

The zener should be of suitable wattage rating. In a similar manner for 27V use a 3.3V zener or for 33V a 9.1V zener.



## Standard Configurations

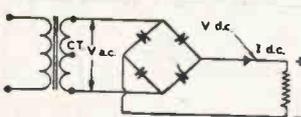
### FULL-WAVE BRIDGE - CAPACITIVE INPUT FILTER



$$V_{dc} = 1.41 \times V_{ac}$$

$$I_{dc} = 0.62 \times I_{ac}$$

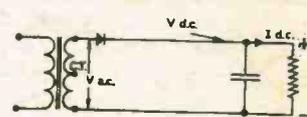
### FULL-WAVE BRIDGE - RESISTIVE LOAD



$$V_{dc} = 0.90 \times V_{ac}$$

$$I_{dc} = 0.90 \times I_{ac}$$

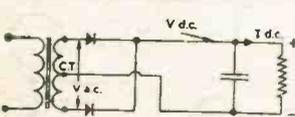
### HALF-WAVE - CAPACITIVE INPUT FILTER



$$V_{dc} = 1.41 \times V_{ac}$$

$$I_{dc} = 0.28 \times I_{ac}$$

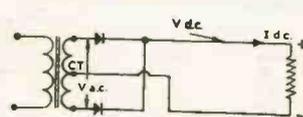
### FULL-WAVE - CAPACITIVE INPUT FILTER



$$V_{dc} = 0.71 \times V_{ac}$$

$$I_{dc} = 1.0 \times I_{ac}$$

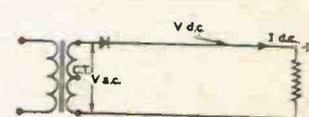
### FULL-WAVE - RESISTIVE LOAD



$$V_{dc} = 0.45 \times V_{ac}$$

$$I_{dc} = 1.27 \times I_{ac}$$

### HALF-WAVE - RESISTIVE LOAD

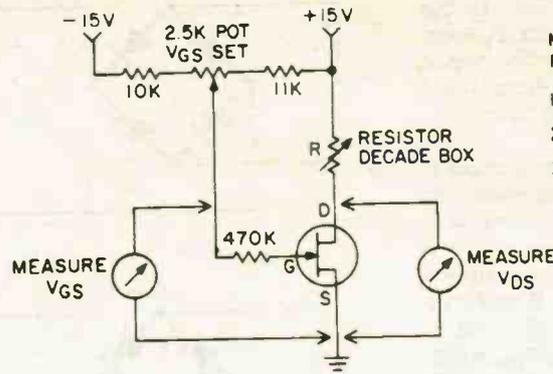


$$V_{dc} = 0.45 \times V_{ac}$$

$$I_{dc} = 0.64 \times I_{ac}$$

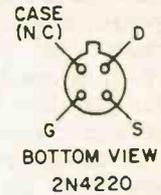
## FET Testing, Static

A transistor socket will facilitate changing FETs. A good procedure is to first measure  $R_{DS}$  at  $V_{GS} = 0$ . Then increase  $V_{GS}$  (negatively for N-channel FETs) until  $R_{DS}$  is about three times the zero-bias value; this corresponds to a mid-range cutoff frequency where matching is the most critical. With this  $V_{GS}$  setting try different FETs until a 10 percent or better match is found. If  $R_{DS}$  values seem to cluster higher or lower, try another unit as a reference and try matching to it. When matched units are found, check the match at minimum  $R_{DS} = +0.5V$  and at 10 times this value of  $R_{DS}$ . A 20 percent mismatch can be tolerated at these extremes.



### MEASURING PROCEDURE $R_{DS}$ VERSUS $V_{GS}$ :

1. SET DESIRED  $V_{GS}$
2. SET R FOR  $V_{DS} = 0.050 V$
3.  $R_{DS} = \frac{R}{300}$  (APPROX)

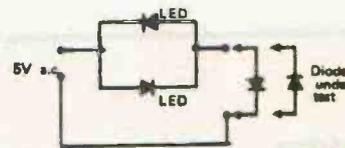


## Diode Tester

This is a diode tester with light emitting diodes. If we change the polarity of the diode under test the appropriate LED will light.

If both of the LEDs go on, this means that the diode is shorted.

If neither light, this means that the diode under test is open circuit.



## Ammeter, Wide Range

The instrument shown will measure currents from  $1\mu A$  to 1A F.S.D. in seven ranges.

IC1 is connected as a unity gain buffer and the input current flows through the resistor selected by SW1 to earth. In so doing a voltage proportional to the input current is

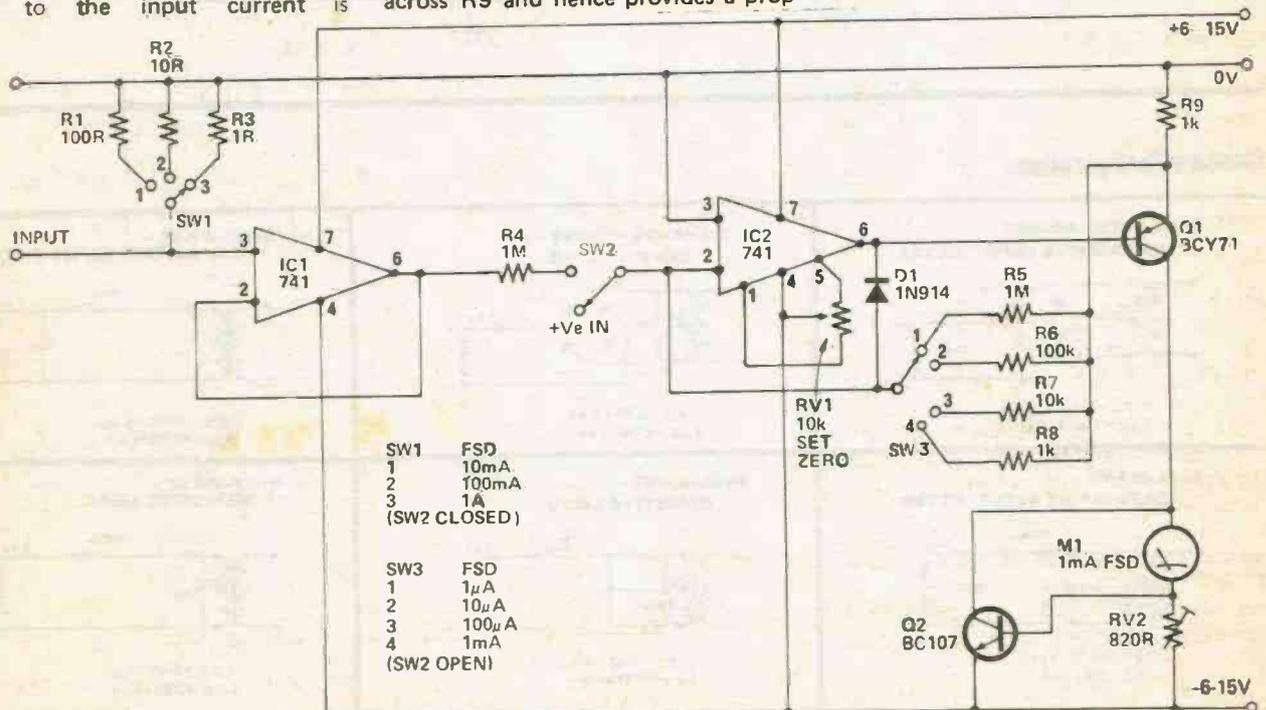
developed across the resistor and this appears at the output, pin 6.

Small currents are measured by IC2. In this mode the current flows into the non inverting input. Since this is a virtual earth, the output will generate a voltage proportional to the input current.

In practice, this voltage is developed across  $R_9$  and hence provides a prop-

portional current through Q1 and M1.

Q2 and RV1 form a meter protection circuit and the latter component should be adjusted so that Q2 starts to conduct at F.S.D. D1 is included to prevent damage to the base emitter junction of Q1 in the event of an input of wrong polarity.



**Millivoltmeter, Audio**

This circuit was conceived as an easily built instrument offering a  $\pm 2\%$  accuracy.

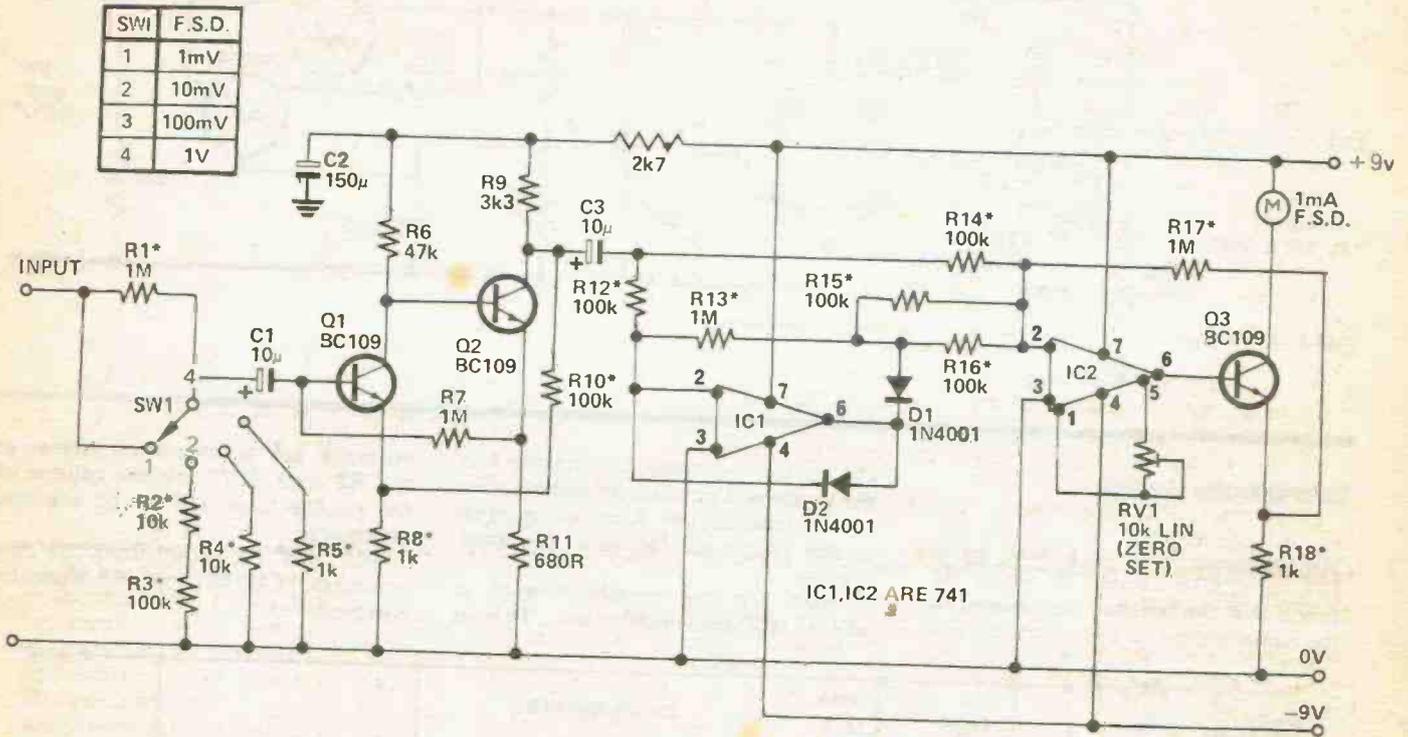
Q1 and Q2 are connected as a high gain feedback pair with a fixed gain of 100.

IC 1 and IC 2 form a precision-full wave rectifier with a gain of 10 over a bandwidth extending from below 20 Hz to above 50 kHz.

The gain of the whole circuit therefore is 1000 and input signals above 1 mV must be attenuated.

No setting up is required except for the zero adjustment of RV1. Input impedance is 1 M $\Omega$  on all ranges.

Resistors marked with an asterisk are 1% types.



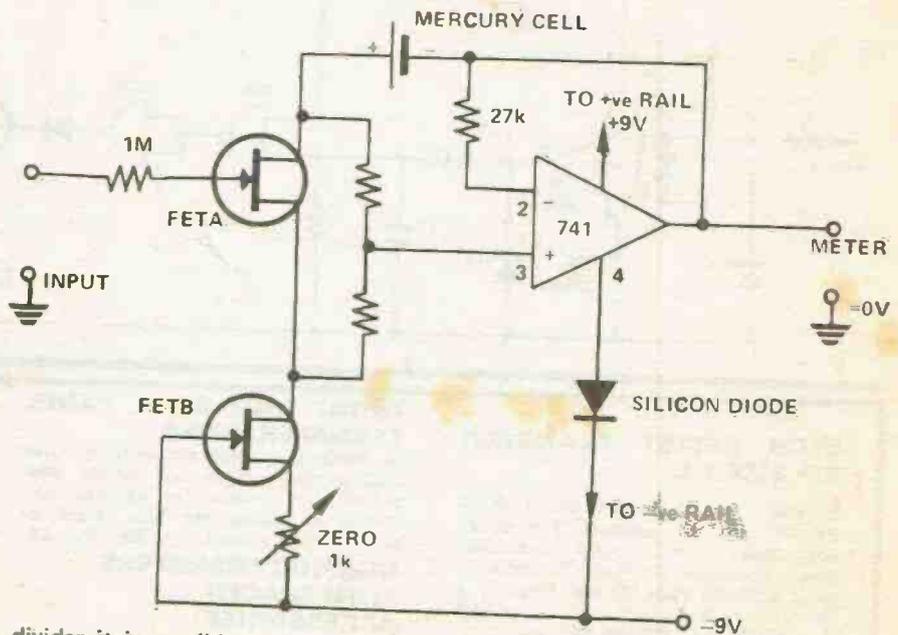
**DC Probe, 100 000 Megohm**

The input current of a junction FET, usually less than 1nA, flows out of the gate, and is constant at a particular temperature, provided the voltage across the device is constant. By making the gate positive to the source this leakage current can be made to flow back into the device, reducing the input current almost to zero.

FET A should be a low loss, low  $V_p$  device (ideally  $V_p$  should be about 0.5V). FET B must be somewhat higher but is less critical, the bargain pack is usually a good source of such devices. Forward bias should be about 150mV and current through the FETs about 400 $\mu$ A.

The mercury cell holds the voltage across the input FET constant at 1.5V (1.35V plus 150mV) and the silicon diode in the op amp's negative lead prevents the cell from discharging when the power is off.

By adjusting values in the potential



divider it is possible to achieve input currents within a few picoamps either

way and to measure the voltage on a small capacitor without changing it.

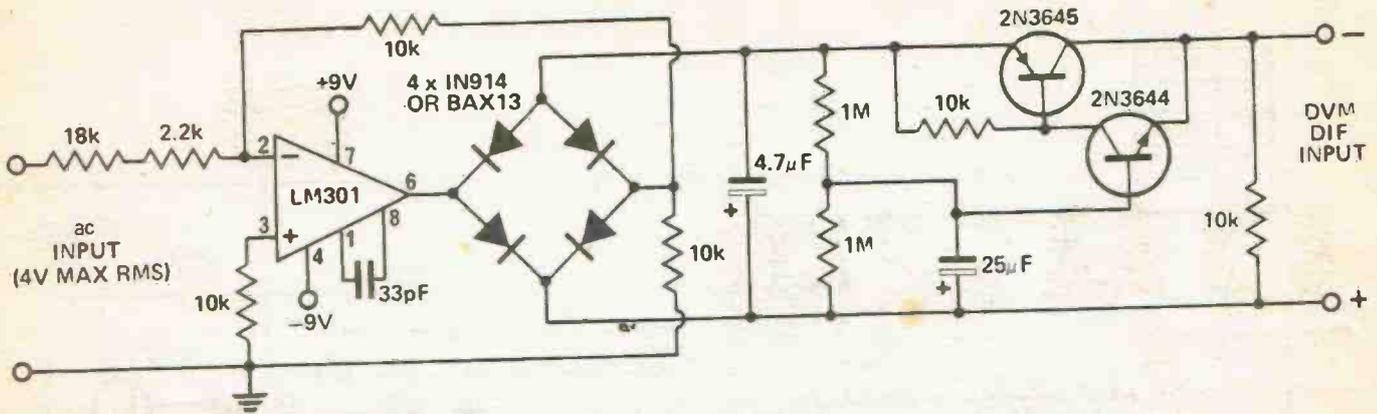
## Measuring RMS with a DVM

The above circuit may be used for measuring the RMS value of ac (sinewave) with a dc digital voltmeter. It has a frequency response to beyond 10 kHz and will measure signals as low

as 400 mV. The error rises at low frequencies, somewhat below 50 Hz, to about 4% mean. The LM301 supply may be as low as  $\pm 4V$ , or up to  $\pm 15V$ , if desired, with reduced

sensitivity at the lower voltage.

The DVM input must be floating and a differential input is required. To increase the input range a step attenuator may be used.



## Logic Indicator, Audible

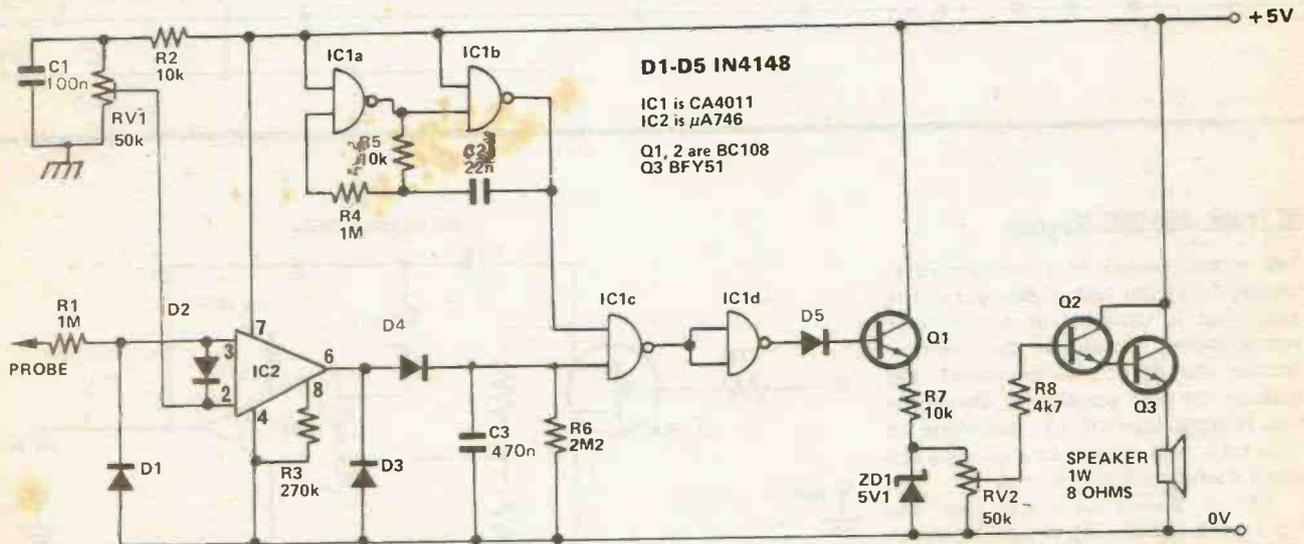
The indicator will work with either TTL or CMOS circuits. A useful feature is that the unit can be powered by the same supply as the one supplying

the circuit under test. Logic state 1 at the probe will produce an audible tone on the loudspeaker. A switching signal at the probe also activates the loudspeaker.

RV1 sets the threshold level at which IC2 will switch on. This is

normally set at maximum (wiper at the R2 end). RV2 sets the volume of the audible tone, and can be adjusted as required.

IC2 can be substituted by the equivalent LM748, but R3 must be removed first.



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## Hex to 7 Segment

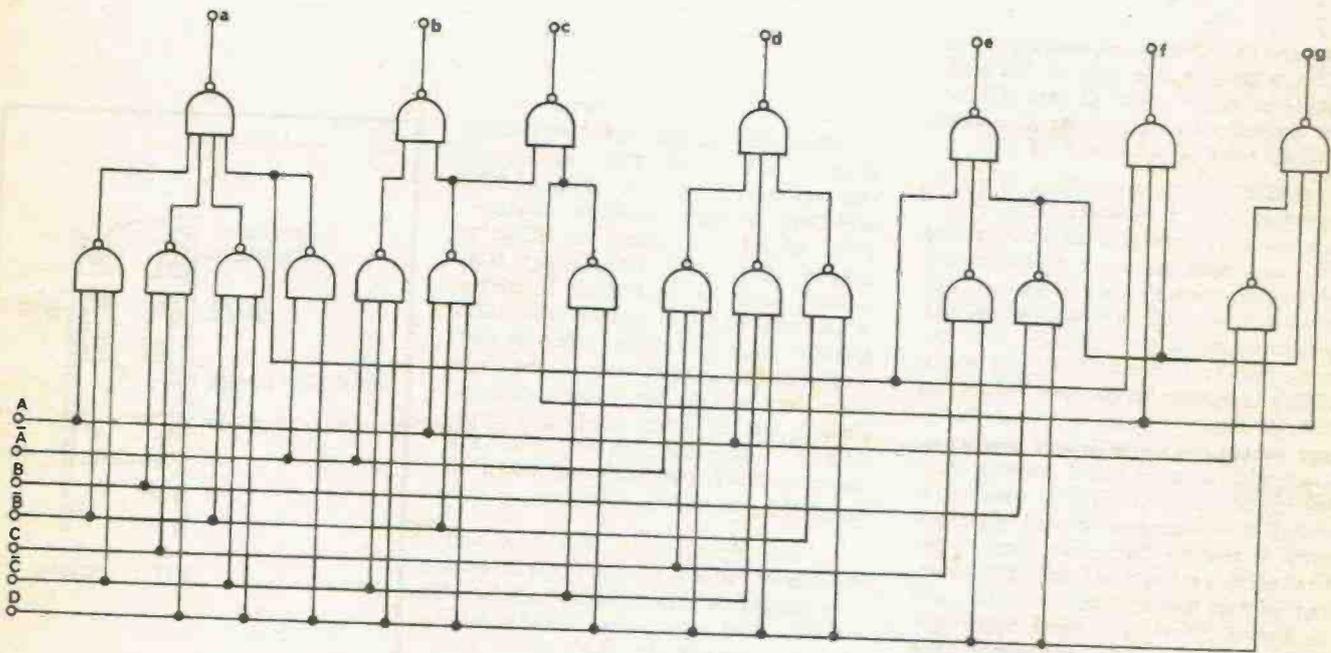
The circuit described below provides an extension to the 7448 BCD to seven-segment decoder, converting it into a hexadecimal to seven-segment decoder which will give the numerals 0-9 and the characters A,B,C,D,E, and F as output for a four bit binary input. (Inputs of A,  $\bar{A}$ , B,  $\bar{B}$ , C,  $\bar{C}$ , D,  $\bar{D}$  are needed with an inverting buffer - fan out 30 - on the  $\bar{D}$  input.)

The 7448 is disabled by bringing the blanking input low when the input is greater than 0111<sub>2</sub> (i.e.  $\bar{D}$  is connected to B1/RB0 on the 7448.) Outputs from the 7448 and the add-on decoder are OR-ed together creating a single seven-segment output.

	a	b	c	d	e	f	g
0	0	0	0	0	0	0	0
1	0	0	0	0	0	1	1
2	0	0	0	0	0	1	1
3	0	0	0	0	0	1	1
4	0	0	0	0	0	1	0
5	0	0	0	0	0	1	1
6	0	0	0	0	0	1	1
7	0	0	0	0	0	0	1
8	0	0	0	0	0	1	0
9	0	0	0	0	0	1	0
A	1	1	1	1	1	1	1
B	1	1	1	1	1	1	1
C	1	1	1	1	1	1	1
D	1	1	1	1	1	1	1

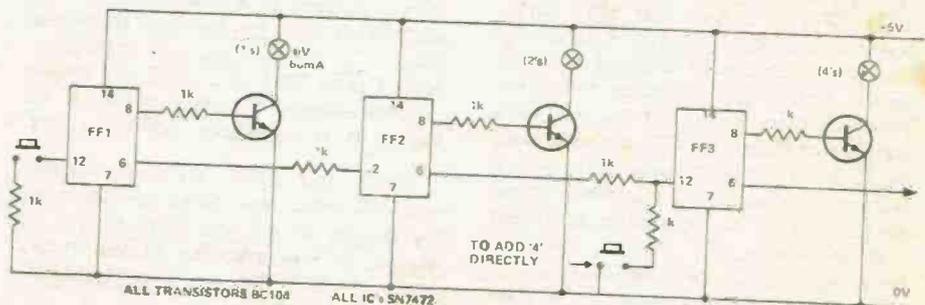
TRUTH TABLE for the 'add-on' decoder. Note that when the input is 0110<sub>2</sub> (6<sub>10</sub>) a logical one is inserted in

the 'a' column to provide the resulting seven-segment '6' with a cap, thus differentiating it from a 'B'.



## Binary Calculator

This simple circuit allows infinite addition in binary (base 2). The circuit can be split into many identical stages, each consisting of a flip flop and lamp driver. An input of 'state 1' initiates the first flip flop. Hence the 1's lamp is on. A second pulse alters the first F.F to switch off the lamp and send a pulse to the second flip flop which illuminates the 2's lamp. The third pulse causes F.F1 to light its lamp without altering the second. This means that the 1's and 2's lamps are on (1+2=3) a total count of three. This on/off process continues for all



the stages.

There is no limit to the total count of the circuit. Each additional stage doubles the count. i.e: 9 stages, total

count is 511. To enter large numbers a press button shorts the input of the intermediate stage to 0V via a 1k resistor.

## TTL Keyer

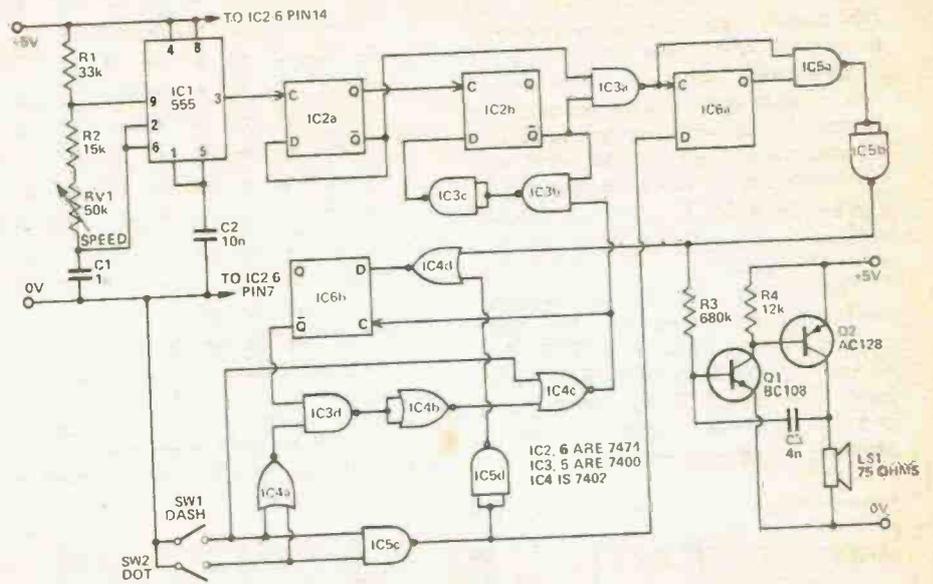
This device can be used to send perfectly spaced Morse at very high speeds - up to twice as fast as with an ordinary Morse key. It uses six integrated circuits, and also requires two special switches, SW1 and SW2, which are described later.

To describe the operation of the circuit fully would take up over a page of ETI, and so a simplified explanation is given here. IC1 is a 555 timer connected as an astable multivibrator, whose frequency is varied by RV1. The output is fed to IC2a, a D flip-flop, which divides the input frequency by 2, producing a square-wave with a 1:1 mark-space ratio (dots).

If SW1 and SW2 are both open, the D inputs of IC2b and IC6a are both at logic 0, so that the dots from IC2a are inverted by IC3a, but blocked by IC5a. IC5b output is a 0, and so the audio oscillator made up of Q1 and Q2 and the associated components is disabled and no tone is fed to the speaker.

If SW2 is closed, IC6a's D input becomes logic 1. However IC6a's output can only change state on the rising edge of a clock pulse (i.e. the beginning of a dot). Hence if a dot has already started when SW2 is closed, it will not get through to the speaker, but the next dot will, because it will make IC6a's Q output to go to 1. Hence the dots now get through to the oscillator and successively enable and disable it, causing dots to be heard coming from the speaker. When SW2 is opened, if a dot is in progress it will continue until it has finished, and then at the beginning of the next dot, IC5a output will go low and no more dots will be heard. There is a short delay between the beginning of the dot and the Q output going low, which does cause a short 'blip' at IC5b output, but the blip is too fast to be heard.

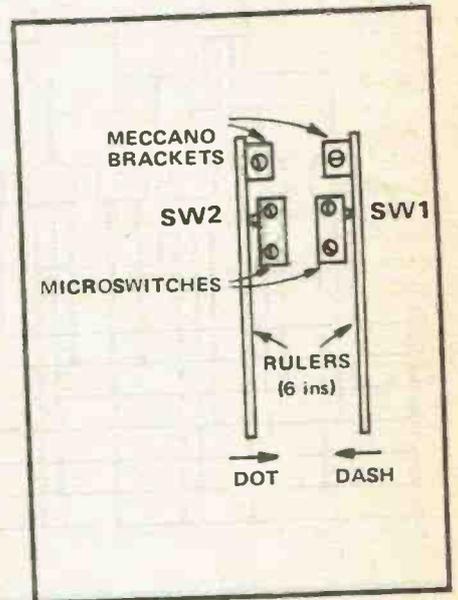
If SW2 is closed, but SW1 open, IC4c output goes to 1 and IC2b's Q output is effectively shorted to its D input. This causes IC2b to divide the string of dots from IC2a by two. The outputs of IC2a and IC2b are combined by IC3a to produce a waveform with a 3:1 mark-space ratio (dashes). These are passed on to the audio oscillator just as before. The dashes, like the dots, are self-completing. Notice that IC4c output determines whether dots or dashes are produced.



While SW1 or SW2 is closed, IC6b's D input is fed from IC4c output and clock pulses come from IC5b. If SW1 and SW2 are both operated together, IC4a allows the output of IC6b to pass to IC4c, and IC4d inverts IC4c output again, so that IC6b's Q output is shorted to its D input. Thus IC6b changes state every time a dot or dash begins at the output, and causes alternate dots and dashes to be produced. This is useful when sending a letter like C (dash-dot-dash-dot), as the switches SW1 and SW2 each need to be closed and opened just once.

It was found after the unit had been built, that it was difficult to send a letter like A (dot-dash) at high speed because SW1 had to be closed a fraction of a second after SW2, which was difficult to achieve at the first attempt. Hence IC5d and IC4d were added. When both switches are released, IC6b input becomes 0. A clock pulse is then applied to IC6b by the 'blip' described earlier. This makes the output go high, and if now SW1 and SW2 are closed simultaneously, the first thing to be heard in the speaker will be a dot.

SW1 and SW2 are push-button microswitches, and these are operated by means of a lever arrangement as shown in the diagram. Plastic rulers were used on the unit built because they are flexible.

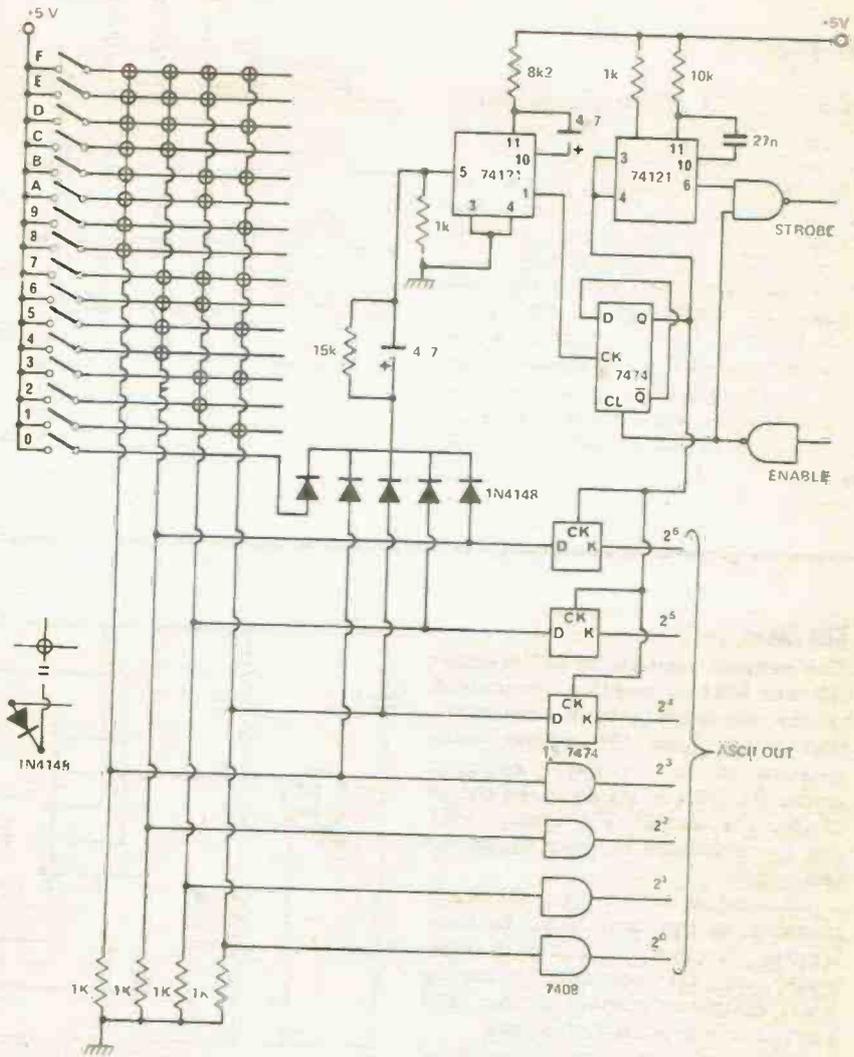


The component values shown around IC1 give a speed range of 11-30 words per minute. The upper limit can be raised by decreasing R2. I have so far reached a speed of 20 wpm on the unit, after only a week or so of using it. As it stands it is a Morse practise unit, but if IC5b output is taken to a transistor driving a relay, the relay contacts could be used in place of an ordinary Morse key in a C.W. transmitter.

## ASCII Keyboard

This circuit uses a 16 key calculator keyboard to generate the 7 bit ASCII code, using two hex numbers to define ASCII character.

If, for example, the code for A (41 hex) is required, '4' is pressed first. After 10mS (to avoid switch bounce) the binary code from the diode matrix is latched into three D-type flip-flops. '1' is now entered. This time, after the 10mS delay, a 200uS pulse is produced by the second 74121. If the ENABLE input is low, a negative pulse appears on the STROBE output, while the ASCII code for A appears on the other outputs. If the enable input is high, the circuit remains in its initial state with the strobe pulse disabled.

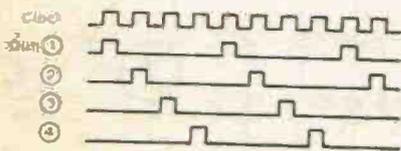
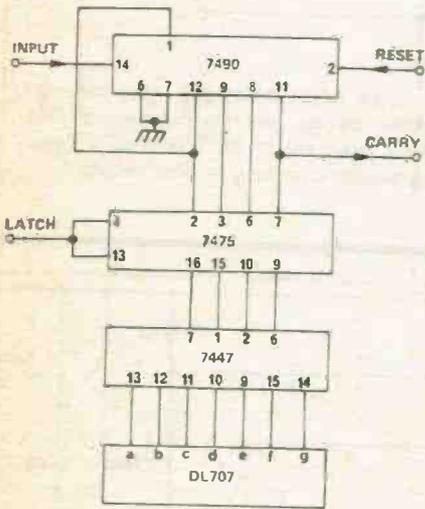


## Counter-Display Module

The signal is connected to the input line of the 7490 decade counter. When the 'latch' is high, the display will follow the count. When the 'latch' goes low, the display holds.

The module may be constructed on two pieces of veroboard, the boards being held together by wire soldered at each corner.

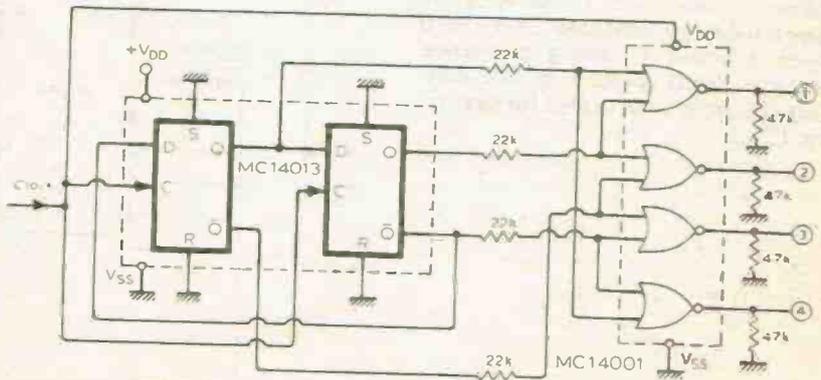
In order to cascade, common up all the latch hold and reset lines, and connect the carry line to the input line of the next module.



## Multiphase Clock Generator

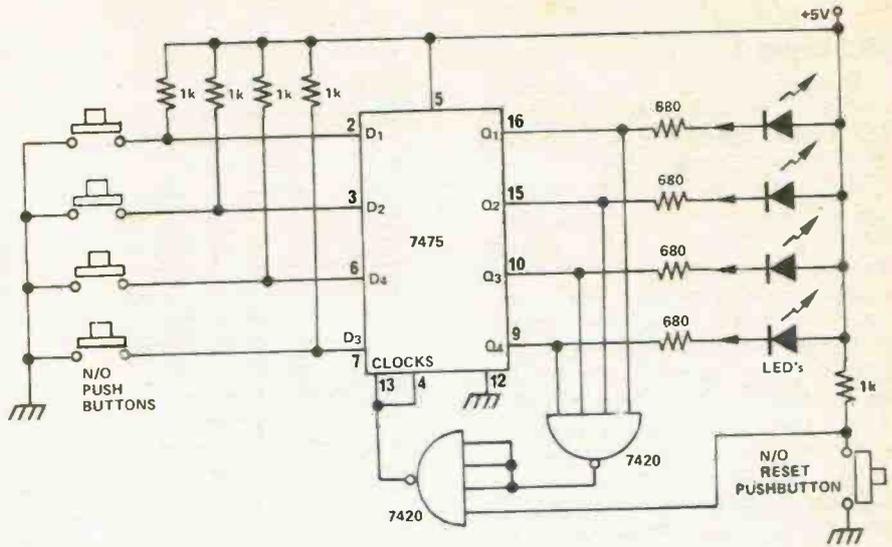
The circuit shown, uses only two CMOS ICs, was designed by Michel Burri of Motorola's Geneva applications laboratories. It will produce a pulse on each of the four output lines in turn. These pulses do not overlap one another.

Operation of the circuit is self-evident from an examination of the schematic; however, it is interesting to note that the power supply of the MC14001 is derived from the clock input. The maximum operating speed of this circuit is about 1 MHz.



## Windicator

With two TTL ICs and a handful of other components, a circuit can be constructed that will indicate which of four buttons was pressed first, as well as lock out all other entries. It is thus suitable for quizzes, games of Snap and the like. The appearance of a logic 0 at one of the Q outputs, lights the appropriate LED and locks out other entries by taking the clock input low. The TTL outputs are capable of sinking 10 TTL loads or 16mA. Running the LEDs at 5mA leaves adequate margin to sink the 1 load of the 7420 gate.



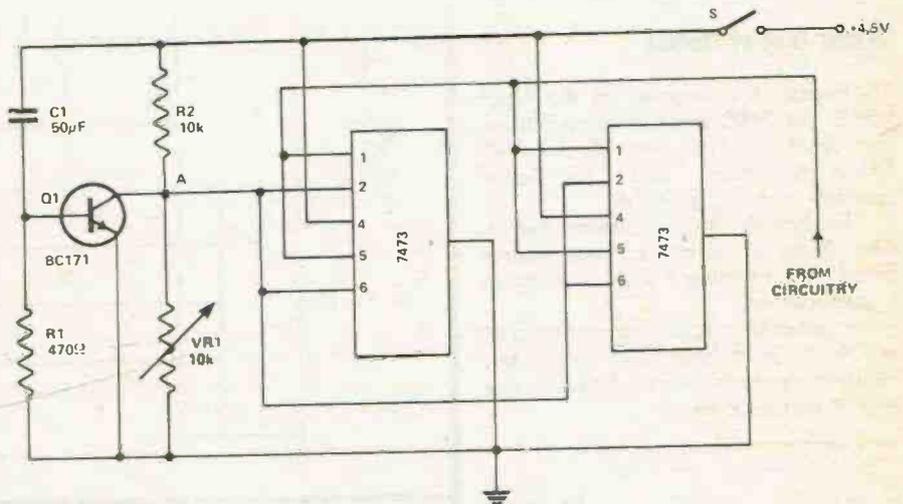
## Self-Clear

The network consists of two resistors R2 and VR1 arranged as a potential divider, the latter being shunted with a non linear load Q1 whose value depends on the voltage developed across R1. This is related to the charge of the capacitor C1. The resistor VR1 was made variable to make the design less critical.

As soon as S is closed, C1 starts charging; at the same time the base-emitter junction is being forward biased and Q1 conducts, bypassing VR1. Voltage at point A is "low" and a set pulse is produced, therefore.

When the charge on C1 reaches a given value, Q1 stops conducting and voltage at point A rises to a stabilized value which is approx.  $4.5 VR1 / (R2 + VR1)$ .

Component values are not critical although R1 and R2 must be close to



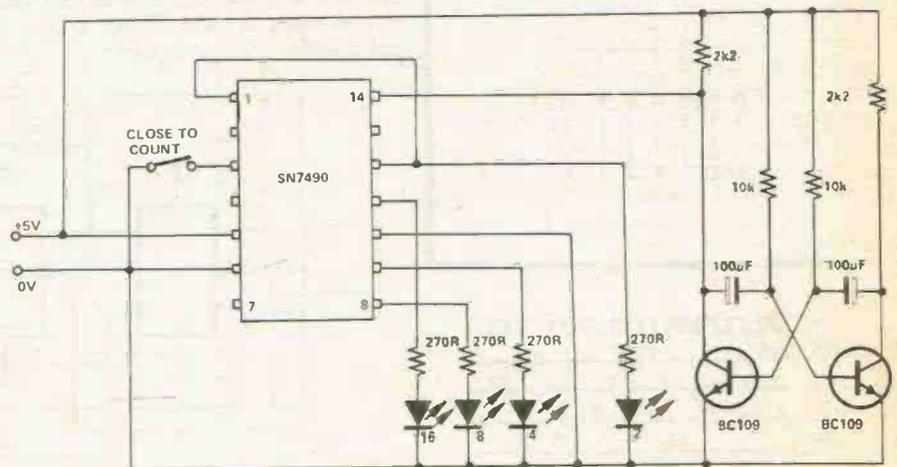
the indicated values.

Any NPN silicon transistor will work the prototype being assembled with the BC171.

VR1 adjustment depends, amongst other things, on the number of flip-flops and must be adjusted in each particular case to give best results.

## LED Counter

The astable multivibrator is used to generate pulses which operates the four integrated bistables. The 7490 gives a binary counting sequence and the 7490 gives a BCD count. This circuit is very useful for testing the IC's.





## Emergency Lights

Basically it's a 240 volt mains operated device that provides low voltage dc power, and switches instantaneously and automatically to battery operation in the event of power failure. When power is restored the unit automatically reverts to mains operation and recharges the battery.

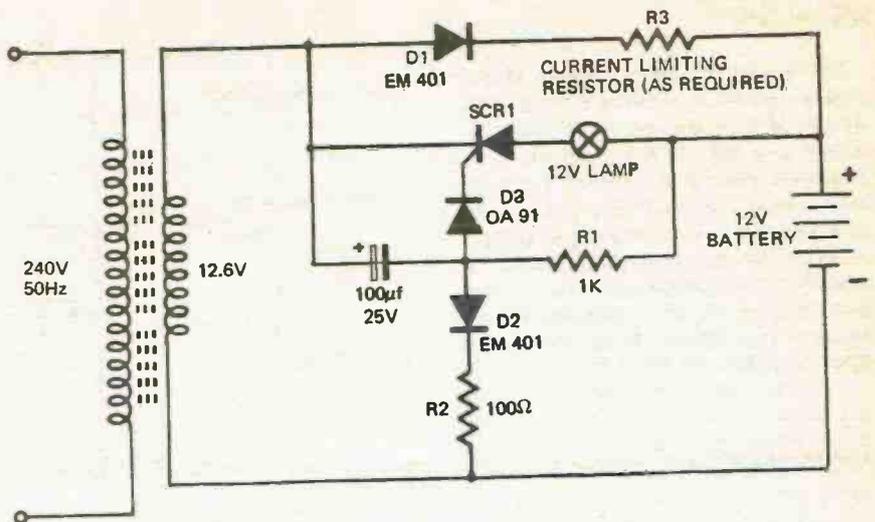
The unit may be used to provide emergency lighting in hospitals, or dark corridors, as an automatic battery change-over supply for intruder alarm systems, or as a power failure alarm for heaters or deep freeze systems.

The circuit may be used in many different forms with circuit component values and battery ampere/hour capacity chosen to suit individual applications.

Figure 1 shows a circuit designed to supply 12.0 volts at 1.0 Amp, this may be increased to at least 2.0 Amps if the SCR is mounted on a heat-sink.

Transformer T1 is a standard filament transformer with a secondary winding capable of supplying 12.6 Volts rms. Any SCR capable of handling a couple of Amps may be used for SCR1. We suggest a C106 series, primarily because they are generally available from most parts suppliers.

In theory, capacitor C1 should be non-polarised because, during emergency operation, it is reverse-charged to the SCR gate triggering voltage.



The 12 volt lamp shown in this circuit should be rated at 12 watts or less — this may be increased to 24 watts if the SCR is mounted on an adequate heat sink.

This rarely exceeds half a volt and in practise standard electrolytics may be used without fear of breakdown.

Resistor R3 must be chosen to limit the charging current of the battery to a safe level. This level varies from one type of battery to another — most nickel cadmium batteries, may be trickle charged continuously at 1/100th of their Ampere/hour rating — i.e., a one A/h battery may be continuously charged at 1/100th of an Amp = 10 mA.

The value of R3 may be calculated as:-

$$R3 = \frac{V_s - V_B}{I_c} \times 1000$$

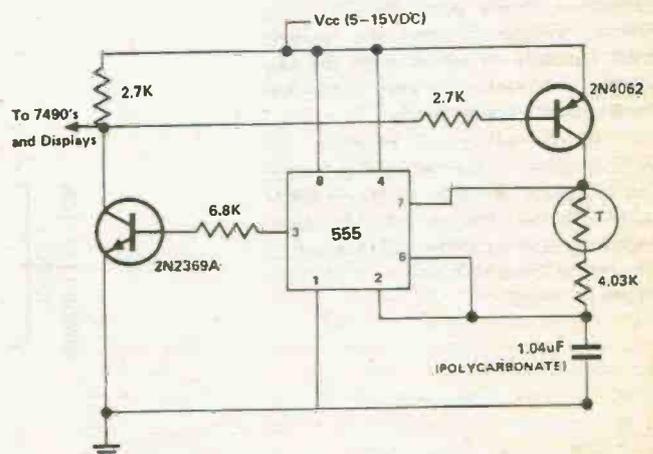
Where  $V_s$  = Supply voltage from transformer  
 $V_B$  = Battery voltage  
 $I_c$  = Charging current in milliamps

If in doubt, the maximum safe continuous charging current can be ascertained from the supplier or manufacturer of the battery that you have chosen.

To modify the circuit for other applications it is necessary to choose a transformer having an output about 5 or 10% higher than the standby battery, and rated to provide the load current required. The SCR must also be chosen with voltage and current ratings suited to the application.

## Digital Thermometer

This circuit we haven't tried yet but it looks very good, anybody who tries it, let us know how you get on. The circuit's output frequency varies in a nearly linear manner from 38 to 114Hz as the temperature changes from 37°F to 115°F. The 555 is set up in the normal astable configuration with one resistor replaced by a thermistor/resistor network and the other replaced by a transistor. The transistor's near zero on-resistance and very high off-resistance results in equal charge and discharge intervals that depend only on the thermistor/resistor network. The thermistor is one with a value of 5000 ohms at 25°C and a resistance ratio of

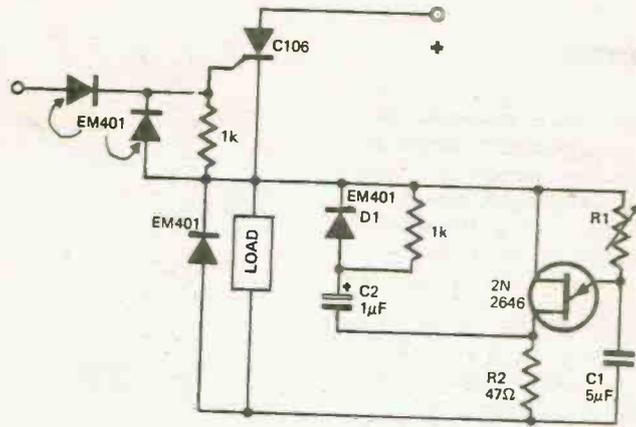


9.06:1 over the temperature range 0°C to 50°C. The capacitors need to be temperature stable and may need to be hand selected and added to give the

best results. It would seem that a similar circuit for Centigrade might also be possible — any ideas?

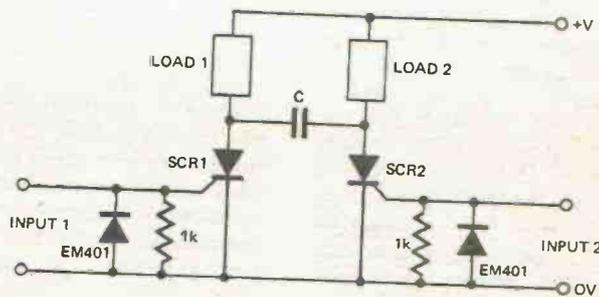
### SCR One Shot

The circuit shown is a 'one shot' or pulse generator. Here an incoming signal triggers the SCR and energizes the load. The load voltage energizes the UJT timing circuit. After a time determined by  $R1/C1$ , the UJT fires, and a pulse generated across  $R2$  is coupled to the cathode of the SCR through  $D1$  and  $C2$ . The SCR's cathode is momentarily lifted above the anode voltage and the SCR turns off.



### SCR Multivibrator

A triggered multivibrator — an input to 1 energizes load 1. A subsequent input to 2 energizes load 2, thus turning off SCR1 and de-energizing load 1.

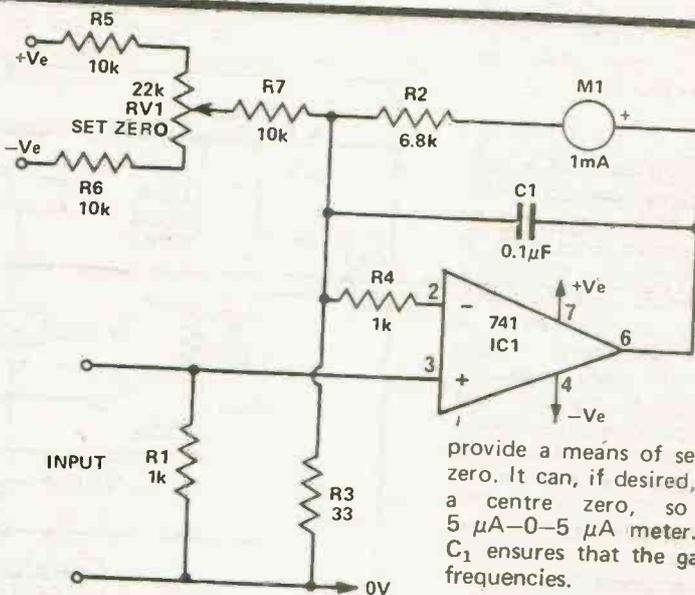


### Meter Amplifier

Now for the dc meter amplifier which uses a 741 type IC. The values shown give full scale deflection on a 1 mA meter with only  $10\mu A$  flowing into the input.

Circuit function depends on there being negligible difference between the voltages at the two inputs of an op. amp. when arranged in a negative feedback configuration. Accordingly, whatever voltage is applied to the non-inverting terminal, that is, across  $R_1$ , will appear at the inverting terminal, that is, across  $R_3$ . However,  $R_3$  is only 1/100th of the value of  $R_1$ , so that the current through  $R_3$  must be 100 times larger than that through  $R_1$ . It is, of course, the current through  $R_3$  that flows through the meter, and it is worth noting that the value of this current is not affected by resistor  $R_2$  in series with the meter — provided of course that  $R_2$  is not too large to allow the required meter current to flow. The value of  $R_2$  is chosen here to limit meter current to about twice the FSD current, so providing a useful safety device should an unexpectedly high voltage be applied to the non-inverting terminal.

Thus we have a circuit incorporating a meter of 1 mA basic



sensitivity but which appears to be a meter of 100 times that sensitivity.

Resistor  $R_4$  is included to improve the performance with regard to drift of the meter reading, as temperature changes cause changes in the op. amp. bias currents. It is best selected by experiment, although the value given was found to be satisfactory with three individual 741's.

The voltage at the slider of  $RV_1$  is fed via  $R_7$  to the inverting input to

provide a means of setting the meter zero. It can, if desired, be used to give a centre zero, so producing a  $5\mu A-0-5\mu A$  meter. The capacitor  $C_1$  ensures that the gain falls at high frequencies.

With a basic sensitivity of  $10\mu A$ , this amplifier enables a dc voltmeter of 100 kohm per volt to be constructed, by connecting the appropriate resistor in series with the input. The value of the resistor is given by:

$$R = 100 \text{ V kilohms}$$

where V is the input voltage required to give FSD

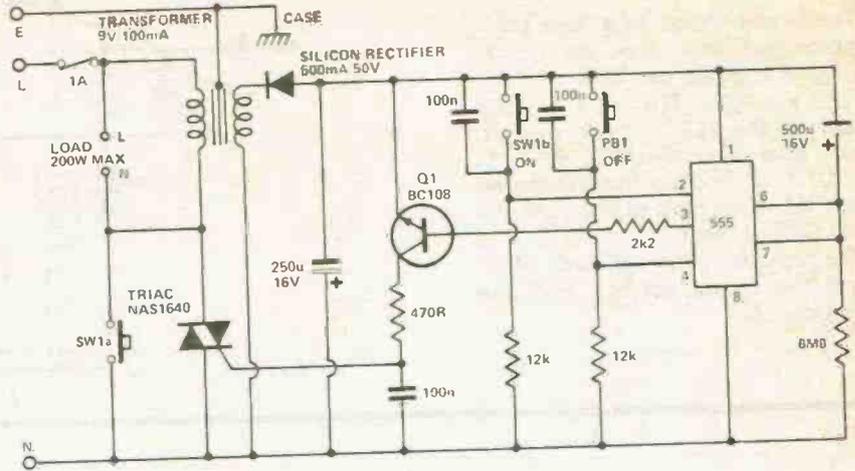
Note that the basic meter of 1mA is a type of movement that is much more robust, and yet cheaper, than others of greater sensitivity.

## Night Light, Automatic

This circuit was devised to turn off a bedroom light after a period of an hour. It could, however, be used to control any load up to a maximum of 200W. At the end of the period the unit switches off both itself and the load.

The timing period is generated by a standard 555 timer in monostable mode controlled by SW1b and PB1. For reliable operation timing capacitor C should be selected for low leakage. The output of the timer switches Q1 which in turn controls the gate current for the triac. During the timing period the triac is fully turned on so there is no degradation of the waveform across the load or RFI due to switching transients.

To initiate the timing period mains must be applied to the transformer to provide a DC supply for the timing circuitry. This is achieved by momentarily bypassing the triac with one pole of the ON switch, SW1a. Because this switch must also provide power to the load



it must be rated accordingly. SW1b is used to trigger the 555 and start the timing period. Q1 will then be turned on, providing gate current to turn on the triac. When SW1 is released the supply and the load is maintained until the end of the timing period. PB1 is provided so

that the load can be switched off at any time. It may be omitted if this feature is not required.

Great care must be exercised with this circuit as all components are connected to mains neutral even when inactive.

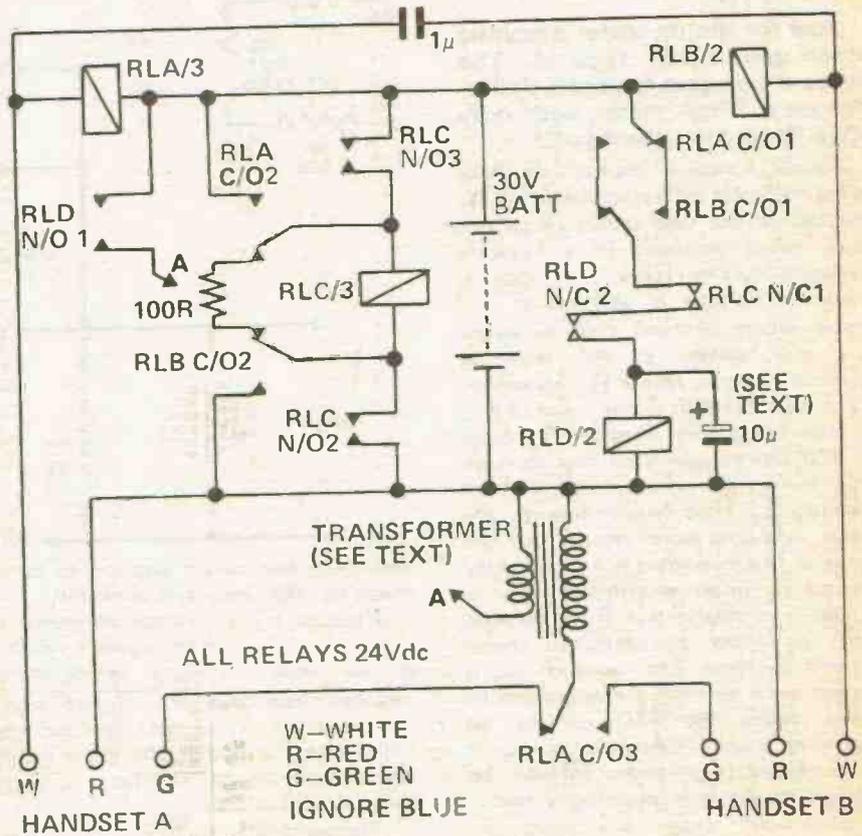
## Telephone Circuit

If handset A is lifted, RLA is activated changing over contacts RLA energising RLD. RLD has a pair of N/C contacts wired in series with its coil and this causes the relay to vibrate. The contacts would not however, change over fully, and a capacitor is wired in parallel and stores enough charge to allow the relay to change fully. The value of the capacitor will depend on the type of relay being used but the value selected should be chosen to cause the relay to vibrate at about 25 Hz.

A second set of contacts is wired in series with the transformer, which in the prototype was an old 250 to 125 V transformer with the 125 V winding being used as the primary. The output of the transformer is fed to the third set of contacts of RLA which selects which telephone is to be rung.

On lifting the other handset relay RLB will energise and the exclusive OR arrangement of contacts RLA and RLB will inhibit the bellringing circuit.

To prevent either bell ringing again if one of the handsets is replaced RLC is included. When both handsets are raised RLC energises and is self-latching, one set of its contacts being in series with RLD. When both handsets



are replaced, RLC is shorted through the 100 Ohm resistor and turns off, resetting the bellringing circuit. The

1 uF capacitor provides the required coupling for speech between the two handsets.

Headphone Adaptor

HEADPHONES have impedances which range from 8 ohms to 2 k ohms or more and handle a typical maximum power of 500 mW. To limit the power that may be delivered into the 8 ohm types, commercial amplifiers generally supply the headphones from the amplifier output via series resistors of around 220 ohms.

Although this technique allows the use of practically any type of headphones without fear of damage the series resistor drastically reduces the amount of damping the amplifier can apply to the phones.

A further problem with headphone listening is that the stereo separation is unnatural in that there is little right channel information fed to the left ear and vice versa.

This simple little adaptor is inserted between the amplifier and the leads to the speakers. It restores damping, by supplying the phones from a 10 ohm source, and has a blend control by which the separation between channels can be varied to obtain a more natural sound.

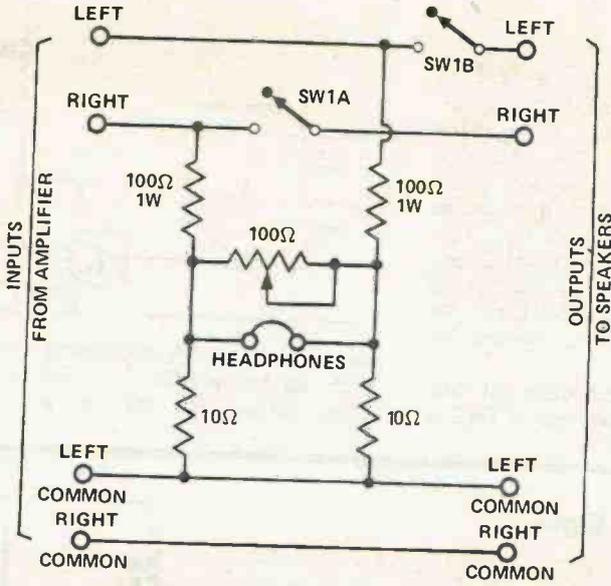


Fig. 1. Circuit diagram of the headphone adaptor.

NOTE  
SW1A AND SW1B ARE PART OF THE SOCKET USED FOR THE HEADPHONES

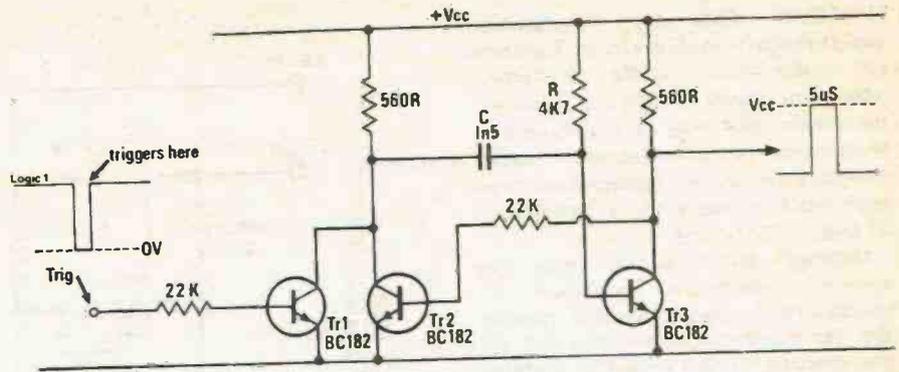
TTLs by TEXAS		C-MOS ICs		OP. AMPS		MEMORY I.C.s		MJE3055 80p		DIODES		BRIDGE RECTIFIERS	
7400	14p	74107	38p	4000	21p	NE531V	140p	MPF102/3	22p	2N2906/A	22p	8Y12	12p
7401	18p	74109	60p	4001	21p	CA3140	108p	2N2926B	9p	2N2907/A	25p	0A47	8p
7402	25p	74111	80p	4002	21p	CA3160	108p	MPF104/5	40p	2N2926B	9p	0A81	15p
7403	18p	74112	75p	4008	127p	LM301A	40p	2N2926G	9p	2N2926G	9p	0A85	15p
7404	24p	74116	21p	4009	180p	LM318N	175p	2N3000	300p	MPSA06	37p	0A91	8p
7405	25p	74118	160p	4010	67p	LM324N	130p	E18	18p	MPSA65	40p	0A91	8p
7406	43p	74119	225p	4011	21p	LM348N	130p	2N3005	65p	MPSU03	72p	0A95	8p
7407	43p	74120	225p	4012	23p	MC1458P	75p	2N3007	70p	MPSU05	78p	0A202	10p
7408	22p	74121	32p	4013	58p			2N3008	90p	MPSU55	90p	1N914	7p
7409	22p	74122	55p	4014	90p			2N3009	90p	OC28	432p	1N4001	28p
7410	18p	74123	75p	4015	90p			2N3010	90p	OC35	90p	1N4002	47p
7411	26p	74125	70p	4016	54p			2N3011	90p	OC71	35p	1N4003	47p
7412	25p	74126	85p	4017	100p			2N3012	90p	R20088	225p	1N4004	7p
7413	40p	74128	85p	4018	57p			2N3013	90p	R20106	225p	1N4005	78p
7414	55p	74132	82p	4019	110p			2N3014	90p	2N3015	90p	1N4006	108p
7415	40p	74136	81p	4020	140p			2N3016	90p	2N3016	90p	1N4007	108p
7417	40p	74141	85p	4021	140p			2N3017	90p	2N3017	90p	1N4008	108p
7420	18p	74142	300p	4022	140p			2N3018	90p	2N3018	90p	1N4009	108p
7421	43p	74145	95p	4023	140p			2N3019	90p	2N3019	90p	1N4010	108p
7422	26p	74147	205p	4024	23p			2N3020	90p	2N3020	90p	1N4011	108p
7423	36p	74148	160p	4025	23p			2N3021	90p	2N3021	90p	1N4012	108p
7425	33p	74150	130p	4026	23p			2N3022	90p	2N3022	90p	1N4013	108p
7426	40p	74151	81p	4027	84p			2N3023	90p	2N3023	90p	1N4014	108p
7427	40p	74152	81p	4028	110p			2N3024	90p	2N3024	90p	1N4015	108p
7428	40p	74153	81p	4029	84p			2N3025	90p	2N3025	90p	1N4016	108p
7430	18p	74155	97p	4030	87p			2N3026	90p	2N3026	90p	1N4017	108p
7432	37p	74158	97p	4031	87p			2N3027	90p	2N3027	90p	1N4018	108p
7433	37p	74157	85p	4040	150p			2N3028	90p	2N3028	90p	1N4019	108p
7437	37p	74159	250p	4042	97p			2N3029	90p	2N3029	90p	1N4020	108p
7438	37p	74160	130p	4043	100p			2N3030	90p	2N3030	90p	1N4021	108p
7440	18p	74161	130p	4046	150p			2N3031	90p	2N3031	90p	1N4022	108p
7441	85p	74162	130p	4047	150p			2N3032	90p	2N3032	90p	1N4023	108p
7442	75p	74163	130p	4049	84p			2N3033	90p	2N3033	90p	1N4024	108p
7443	120p	74165	150p	4054	120p			2N3034	90p	2N3034	90p	1N4025	108p
7444	105p	74166	160p	4055	140p			2N3035	90p	2N3035	90p	1N4026	108p
7446	108p	74167	320p	4056	140p			2N3036	90p	2N3036	90p	1N4027	108p
7447	80p	74170	260p	4060	130p			2N3037	90p	2N3037	90p	1N4028	108p
7448	85p	74172	750p	4068	30p			2N3038	90p	2N3038	90p	1N4029	108p
7449	18p	74173	190p	4069	30p			2N3039	90p	2N3039	90p	1N4030	108p
7451	18p	74174	130p	4071	30p			2N3040	90p	2N3040	90p	1N4031	108p
7453	18p	74175	97p	4072	30p			2N3041	90p	2N3041	90p	1N4032	108p
7454	18p	74176	130p	4073	45p			2N3042	90p	2N3042	90p	1N4033	108p
7460	18p	74177	100p	4078	30p			2N3043	90p	2N3043	90p	1N4034	108p
7470	38p	74180	160p	4081	30p			2N3044	90p	2N3044	90p	1N4035	108p
7472	32p	74181	324p	4082	30p			2N3045	90p	2N3045	90p	1N4036	108p
7473	36p	74182	150p	4083	104p			2N3046	90p	2N3046	90p	1N4037	108p
7474	37p	74184	260p	4510	140p			2N3047	90p	2N3047	90p	1N4038	108p
7474	70p	74185	90p	4511	140p			2N3048	90p	2N3048	90p	1N4039	108p
7475	48p	74186	90p	4516	130p			2N3049	90p	2N3049	90p	1N4040	108p
7476	37p	74189	120p	4518	110p			2N3050	90p	2N3050	90p	1N4041	108p
7480	54p	74191	120p	4518	110p			2N3051	90p	2N3051	90p	1N4042	108p
7481	108p	74192	120p	14433	110p			2N3052	90p	2N3052	90p	1N4043	108p
7482	90p	74193	120p	14533	540p			2N3053	90p	2N3053	90p	1N4044	108p
7483	88p	74194	160p	14583	160p			2N3054	90p	2N3054	90p	1N4045	108p
7484	18p	74195	110p					2N3055	90p	2N3055	90p	1N4046	108p
7485	120p	74198	130p					2N3056	90p	2N3056	90p	1N4047	108p
7486	38p	74197	130p					2N3057	90p	2N3057	90p	1N4048	108p
7489	340p	74198	270p					2N3058	90p	2N3058	90p	1N4049	108p
7490	90p	74199	218p					2N3059	90p	2N3059	90p	1N4050	108p
7491	90p	75107	875p					2N3060	90p	2N3060	90p	1N4051	108p
7492	58p	75182	250p					2N3061	90p	2N3061	90p	1N4052	108p
7493	38p	75324	400p					2N3062	90p	2N3062	90p	1N4053	108p
7494	80p	75325	400p					2N3063	90p	2N3063	90p	1N4054	108p
7495	75p	9601	175p					2N3064	90p	2N3064	90p	1N4055	108p
7496	90p	9602	175p					2N3065	90p	2N3065	90p	1N4056	108p
7497	290p							2N3066	90p	2N3066	90p	1N4057	108p
7498	140p							2N3067	90p	2N3067	90p	1N4058	108p
7499	140p							2N3068	90p	2N3068	90p	1N4059	108p
7500	140p							2N3069	90p	2N3069	90p	1N4060	108p

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## Rising Edge Trigger

The diagram shows a method of triggering a conventional monostable on the rising edge of a short negative-going pulse. The additional transistor, TR1, provides good isolation between the output pulse and the triggering circuitry. The circuit shown gives a pulse of 5µsec duration, but of course the usual design formula  $\tau = 0.65 RC$  can be used to determine circuit values for other pulse widths.

One slight disadvantage of this circuit is that the collector of TR2 is



held down by the triggering waveform, so the switch-on of TR3 is not regenerative. For this reason the

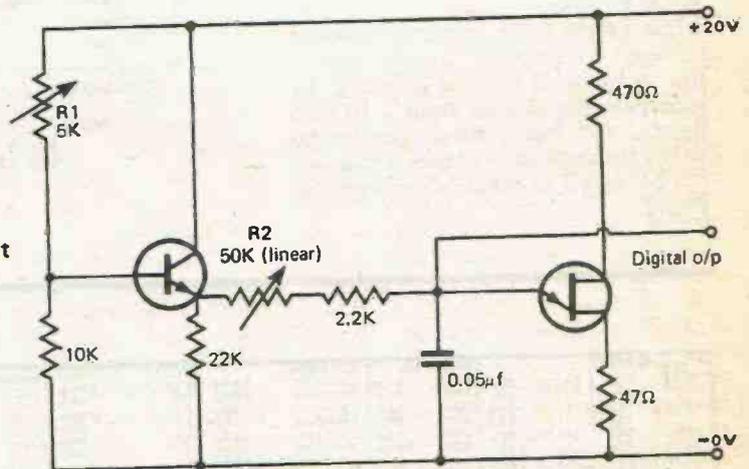
falling edge of the output pulse is not as fast as it might be, but is sufficient for most purposes.

## Position Transducer, Digital

A shaft angular-position to digital output transducer is shown in this circuit.

Rotating the angular position potentiometer R2 will provide a digital output varying from approximately 200 Hz to 2000 Hz.

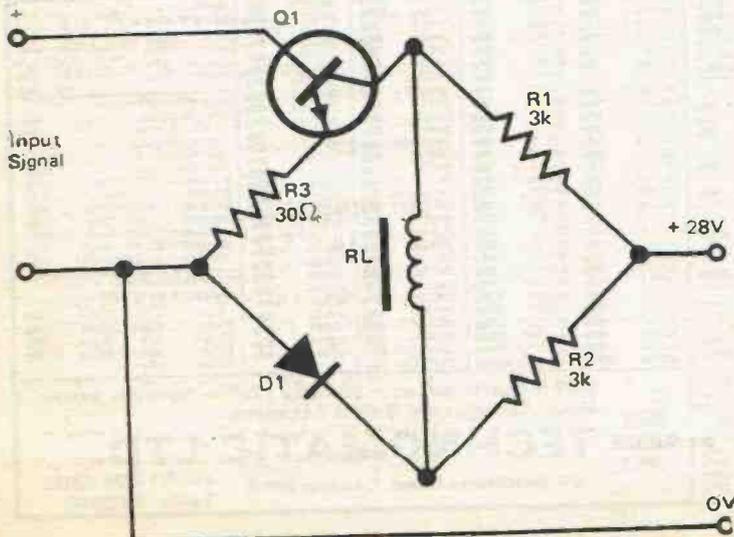
The 5k trim potentiometer R1 provides a rate adjustment of a further 50%. This trim adjustment is independent of the main timing potentiometer.



## Temperature Stabilized Relay

Accurate relay trip-point operation can be obtained over an ambient temperature range from -50°C to +90°C using this simple circuit.

The temperature sensitivity of the silicon transistor Q1 is balanced out by the silicon diode D1. Gain/temperature stabilization may be obtained if required by using a positive temperature co-efficient resistor for R3.



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Basic solid state crystal oscillator circuit techniques are by now well established, most circuits being adaptations of the well-known vacuum tube technology such as the Pierce, Hartley, Clapp and Butler oscillator and use both bipolar and FET devices. Whilst these circuits basically fulfil their intended purpose, there are many applications which require something different or where performance needs to be reliably characterised.

Presented here are a variety of circuits, for a range of applications from LF through the VHF range, that are not commonly found in current amateur use or literature.

## MODES OF OPERATION

A point not often appreciated, or just forgotten, is that quartz crystals can oscillate in a *parallel* resonant mode and a *series* resonant mode. The two frequencies are separated by a small amount, typically 2-15 kHz over the frequency range. The series resonant frequency is *lower* in frequency than the parallel. A crystal specified and calibrated for use in the parallel mode may be satisfactorily used in a series resonant circuit if a capacitor equal in value to its specified load capacitance (usually 20, 30, 50 or 100 pF) is connected in series with the crystal. Sadly, you can't invert the process for series resonant crystal in parallel mode circuits. The series mode crystal will oscillate higher than its calibrated frequency in this case and it may not be possible to capacitively load it down sufficiently.

Overtone crystals operate in the *series* mode usually on the third, fifth or seventh overtone, and the manufacturer normally calibrates the crystal at the *overtone frequency*. Operating a crystal in the parallel mode and multiplying the frequency three or five times produces quite a different result from operating the same crystal in the series mode on its third or fifth overtone. When ordering overtone crystals avoid confusion and specify the frequency you want, *not* the apparent fundamental frequency.

Fundamental crystals in the range 500 kHz to 20 MHz are usually specified for parallel mode operation but series mode operation can be requested. For low frequency crystals, up to 1 MHz, either mode can be specified. Overtone crystals generally cover the range 15 MHz to 150 MHz.

## WIDE RANGE or APERIODIC OSCILLATORS

Oscillators that do not employ tuned circuits can be very useful, whether they are simply used as 'crystal checkers' or some other purpose. Particularly for LF crystals, tuned circuits can be bulky. However, they aren't without their traps. Some crystals are prone to oscillation on unwanted modes, particularly the DT and CT cut crystals used for LF quartz oscillators. It is wise to check that the output is on the correct frequency and no mode instability is evident. Reducing feedback at the higher frequencies usually cures this. In extreme cases, the idea has to be

abandoned and an oscillator having a tuned circuit used instead, (LF crystal oscillators are discussed later).

The first circuit is an emitter-coupled oscillator, a version of the Butler circuit.

The output of the circuit in Fig. 1 is essentially sine wave; reducing the emitter resistor of Q2 increases the harmonic output. By doing this, a 100 kHz crystal produces good harmonics through 30 MHz. It is a series mode circuit.

A variety of transistors may be used. For crystals above 3 MHz, transistors with a high gain-bandwidth product are recommended. For crystals in the 50 kHz to 500 kHz range, transistors with high LF gain, such as the 2N3565 are recommended. Also, for crystals in this range, permissible dissipation is usually less than 100 microwatts and amplitude limiting may be necessary. Low supply voltage, consistent with reliable starting, is recommended. Modifying the circuit by the addition of diodes — as shown in Fig. 3 — is a better method, and starting performance is improved. The circuit will oscillate up to at least 10 MHz with appropriate transistors and emitter resistor values. An emitter follower or source follower buffer is recommended. Similar comments to the above apply to Fig. 2. An emitter follower buffer is included in this circuit. Both circuits are slightly frequency sensitive to power supply voltage changes and load variations. A load of 1 k or greater is recommended.

TTL IC can be used in crystal oscillator circuits but many published circuits have poor starting performance or suffer from non-repeatability owing to wide parameter spreads in IC's. The circuit in Fig. 4. has been tried by the writer over the range 1 MHz to 18 MHz and can be recommended. It is a series mode oscillator and suits AT-cut crystals. The output is about 3 volts peak to peak, square wave up to about 5 MHz beyond which it becomes more like half-sine pulses. Starting performance is excellent, often a critical factor with TTL oscillators.

## LOW FREQUENCY CRYSTAL OSCILLATORS

Crystals in the range 50 kHz to 500 kHz require special considerations not encountered with the more common AT or BT cut HF crystals. The equivalent series resistance (which determines 'activity' — that figure of merit of days of old) is much greater and their permissible dissipation is

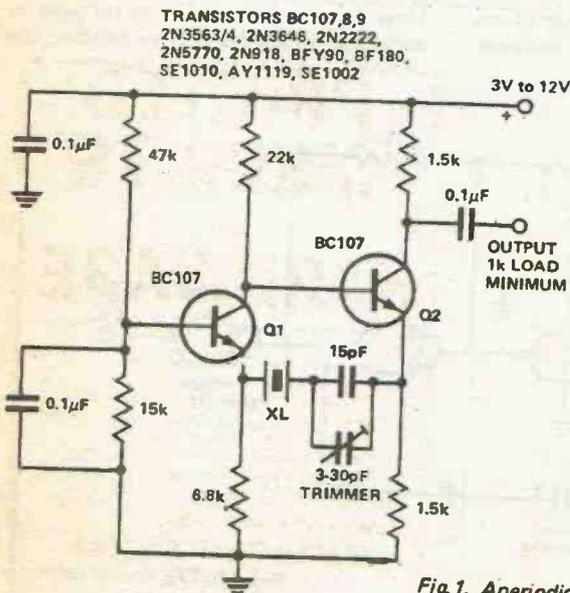


Fig. 1. Aperiodic Butler oscillator (series mode)

# CRYSTAL OSCILLATORS

Limited to less than 100 microwatts, preferably 50 microwatts or less.

The circuit in Fig. 5 is a series mode oscillator. It has the advantage of not requiring a tuned circuit, and has a choice of sine or square wave output. For crystals in the range 50-150 kHz, 2N3565 transistors are recommended although the author has found BC107's satisfactory. Either type will suffice for crystals in the range 150 kHz to 500kHz. If you find the crystal has a very high equivalent series resistance, in which case increase R1 to 270 ohms and R2 to 3.3 k. For square wave operation, C1 is 1 uF (or a value close to, or above it). For sine wave output, C1 is not in circuit. Amplitude limiting is unnecessary. Sine wave output is about 1 V rms, square wave output about 4 V peak to peak.

The circuit in Fig. 6 is a modified form of the Colpitts oscillator, with the addition of resistor Rf to control feedback. Capacitors C1 and C2 should be reduced by preferred values as the frequency is increased. At 500 kHz, values for C1 and C2 should be around 100 pF and 1500 pF respectively.

The circuit as shown gives sine wave output with the second harmonic about 40 dB down (or greater). This can be reduced by careful trimming of Rf and C1. Note that, at the reduced level of feedback necessary to achieve this, it takes some 20 seconds for the oscillator to reach full output. Output is about 2 to 3 volts peak to peak.

If you need an output rich in harmonics, the simple addition of a 0.1 uF capacitor across the emitter

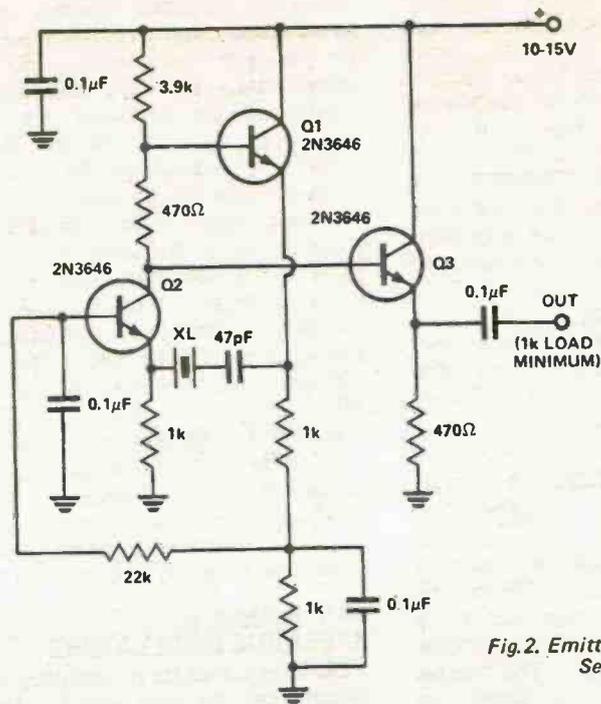


Fig.2. Emitter-coupled oscillator Series mode.

resistor will achieve this. Output then rises to about 5 V peak to peak. Power supply voltage can be reduced in this case to lower crystal dissipation.

Other transistors can be used, but bias and feedback may have to be adjusted. For cantankerous crystals determined to oscillate in modes other than those you wish, the circuit of Fig.7 is recommended. Feedback is controlled by tapping down the collector load of Q1. Amplitude limiting is necessary to keep the crystal dissipation within limits. For 50 kHz crystals the coil should be 2 mH and its resonating capacitor 0.01 uF. Output is about 0.5 V rms, essentially sine wave. The use of an emitter follower or source follower

buffer is recommended. If a parallel mode crystal is used the 1000 pF capacitor shown in series with the crystal should be changed to the crystal's specified load capacitance (usually 30, 50 to 100 pF for these crystals).

## HF CRYSTAL OSCILLATOR CIRCUITS

Solid state circuits for the popular AT-cut HF crystals are legion. However, results aren't always what one would expect. Most fundamental crystals up to 20 MHz are usually specified for parallel mode operation. However, such crystals can be used in series mode oscillators by putting the

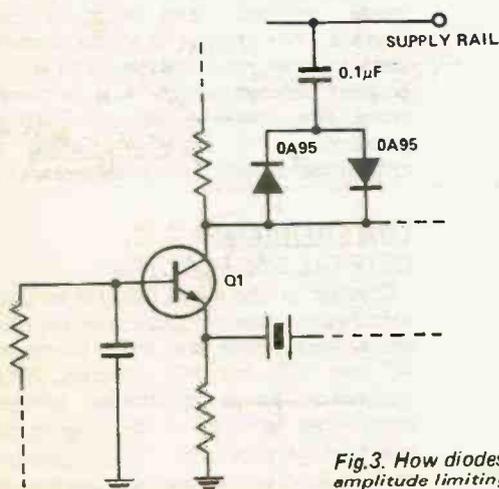


Fig.3. How diodes are used for amplitude limiting.

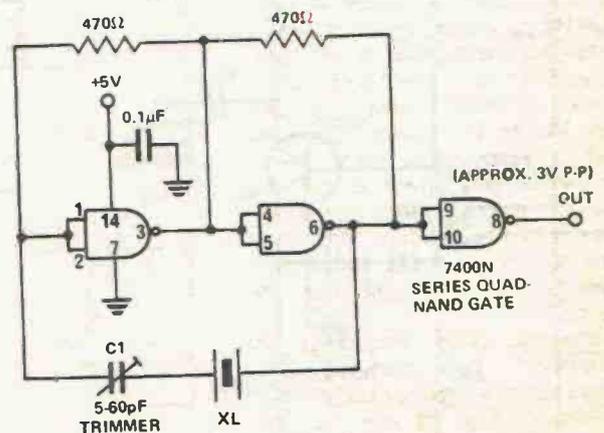


Fig.4. Reliable TTL crystal oscillator.

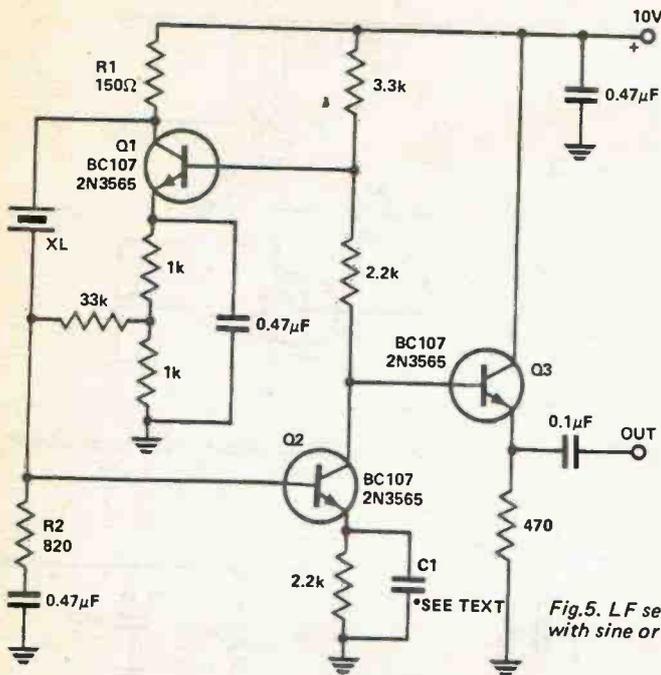


Fig. 5. L.F. series mode oscillator with sine or squarewave output.

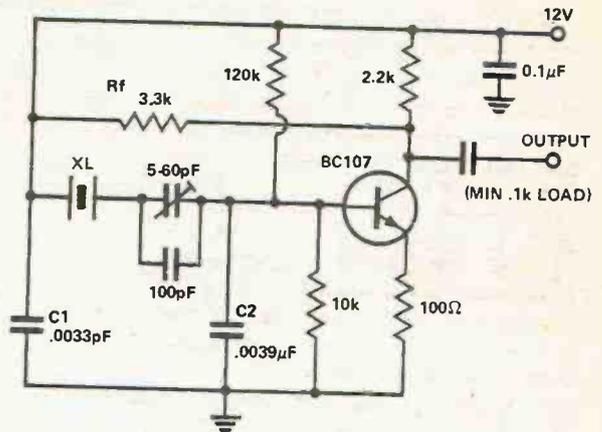


Fig. 6. Parallel mode L.F. oscillator.

specified load capacitance in series with the crystal as mentioned previously. Both types of circuit are detailed here.

A useful oscillator for the range 3 to 10 MHz that does not require a tuned circuit is given in Fig. 8 (a). It is, of course, the same circuit as Fig. 6. The circuit can be used down to 1 MHz if C1 and C2 are increased to 470 pF and 820 pF respectively. It can be used up to 15 MHz if C1 and C2 are reduced to 120 pF and 330 pF. Respectively. This circuit is recommended for non-critical applications where high harmonic output is wanted, or not a consideration.

The addition of a tuned circuit as in 8(b) reduces harmonic output considerably. A tuned circuit with as high a Q as possible is recommended. In a 6 MHz oscillator, I have obtained the following results. With a coil Q of 50 the 2nd harmonic was 35 dB down. With a Q of 160, it was -50 dB! Resistor Rf can be adjusted (increase slightly) to improve this. The output is also increased with a high Q coil. As previously noted, with reduced feedback it takes some tens of seconds to reach full output from switch on, however, frequency stability is excellent.

Operation at other frequencies is accomplished by changing the capacitors and coil appropriately.

This circuit (Fig. 8) can also be turned into a very effective VXO. A small inductance is placed in series

with the crystal and one of the capacitors in the feedback circuit is made variable. An ordinary two-gang 10-415 pF (or thereabouts) broadcast tuning capacitor will do the job nicely. Both gangs are paralleled. The tuning range depends on the crystal used, the inductance of L1 and the frequency. A greater range is usually obtainable with the higher frequency crystals. Stability is excellent, approaching that of the crystal.

## A VHF OSCILLATOR-MULTIPLIER

The circuit in Fig. 10 is a modification of the 'Impedance Inverting' overtone oscillator. Normally, with the impedance inverting circuit the collector is either untuned or grounded for RF. The collector can be tuned to twice or three times the crystal frequency. To reduce the output at the crystal frequency, a double tuned circuit is recommended. DO NOT tune the collector to the crystal frequency, otherwise the circuit will oscillate at a frequency not

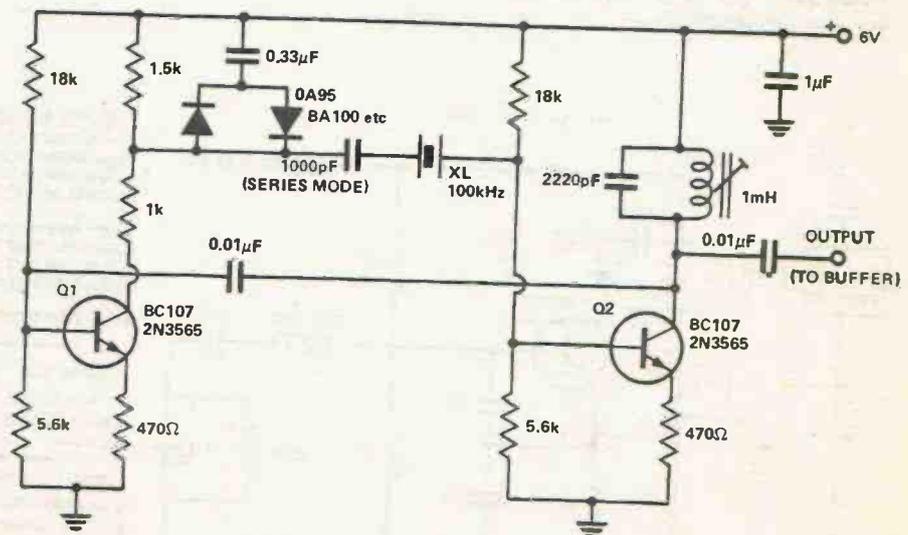


Fig. 7. 100 kHz crystal oscillator (with tuned circuit).



Speaker crossover networks are quite simple to design — if you have a computer to calculate the component values. So we presented a Honeywell computer time-sharing terminal with the task. Here are the results — crossover design made easy.

**I**N MODERN high-fidelity systems the loudspeaker must cover a range of frequencies from 30 Hz to at least 15 kHz.

Generally this requires the use of two or more speakers in each enclosure, each speaker operating within a controlled frequency range.

The extent of that part of the sound frequency spectrum handled by each speaker is controlled by 'crossover' or 'frequency dividing' networks consisting of two or more filters.

A two speaker system usually has a two-way network consisting of a high-pass filter and a low-pass filter. The high-pass filter limits the low frequency response of the high frequency speaker, and the low pass filter limits the high frequency response of the low frequency speaker.

A three speaker system will usually have a three-way network. This will consist of a high-pass filter, a band-pass filter, and a low pass filter. The high and low-pass filters act in a similar manner as those in a two-way system. The band-pass filter controls the frequency range of the mid-range speaker.

The effect, in theory at least, is a smooth transition from one speaker to another over the total frequency range of the system.

For a multi-speaker system to have a substantially flat response, it is essential that each speaker in the system has a usable frequency range overlapping the next. For example the bass speaker in a two speaker system may have a response that is substantially flat from say, 70 Hz to 3 kHz. The high frequency speaker chosen for this system would probably have a usable frequency range from 500 Hz to 15 kHz. Thus the overlap is 2.5 kHz.

## DETERMINING THE CROSS-OVER POINT

A crossover network for the system outlined above would be designed to operate somewhere between 800 Hz and 1 kHz.

A three-way system would probably be designed to crossover at 400 to 500 Hz and again at 5 kHz.

The optimum crossover frequency may be easily determined by studying the frequency response curves of the speakers to be used and arranging for the crossover to take place before the response of a given speaker unit falls off, or the movement of the diaphragm becomes non-linear. Few bass speakers, for example, have any really usable response beyond 2 to 3 kHz. The range of the frequency spectrum covered by the mid-range unit must be restricted to those frequencies at which the displacement of the diaphragm does not exceed the manufacturer's rating.

It should of course be quite clearly understood that the sole purpose of a crossover network is to control the operating range of each speaker. It is to prevent a tweeter with a cone travel of a few thousands of an inch from being driven by a 50 Watt amplifier at 35 Hz — and to ensure that a bass speaker does not have hysteresis trying to emulate Victoria de los Angeles.

A crossover network cannot be used to correct for deficiencies in the record player, amplifier, speaker drive units or enclosure design.

## DIFFERENT TYPES OF FILTER

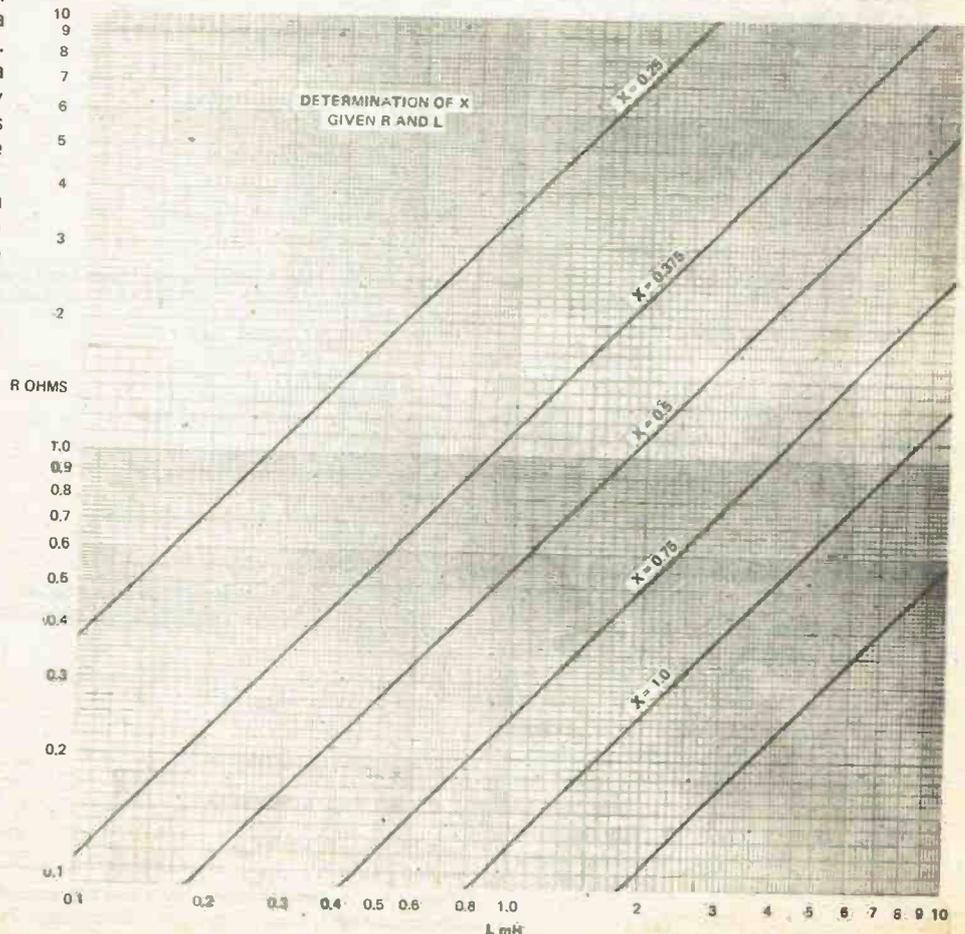
Figure 1 shows a typical frequency response for a crossover network (operating point 1 kHz) consisting of a low-pass filter and a high-pass filter. The graph does in fact show three different pairs of filters, each having a different rate of attenuation.

In practise an attenuation of 6 dB per octave is generally inadequate. The rate of cut-off is not always sufficient to protect the mid-range and high frequency drive units from being overdriven.

An attenuation rate of 12 dB per octave is commonly used, although 18 dB per octave filters are sometimes chosen. For amateur design it is advisable to stick to the 12 dB per octave filter.

Both series and parallel filters are used. Series filters are used only in two-speaker systems. Most commercially built networks use the parallel configuration because component values are the same for each filter and this reduces inventory costs. Apart from cost the parallel network has slightly better electrical characteristics in the transmission and

GRAPH 1



# SPEAKER CROSSOVERS

attenuation bands. Nevertheless the design data given later in this article covers both series and parallel networks.

Apart from the series and parallel configurations, filters used for crossover networks are known as 'constant k' or 'm derived'. It is not essential for the amateur to understand the difference between the two types. Basically the 'constant k' networks are limited to a cut off rate of 12 dB per octave, whilst the 'm derived' networks can operate at cut off rates of 18 dB per octave. The 'm derived network' is often used by designers of top quality speaker systems as the design approach permits closer control of impedance and attenuation characteristics.

## INSERTION LOSS

One of the most important design considerations is that the filter does not introduce any appreciable loss (this is called 'insertion loss') between the amplifier and the speaker drive units. The insertion loss — which is usually quoted in dBs — is caused by the dc resistance of the coils, together with the shunt and series reactance of the circuit elements. The insertion loss of a well designed filter should not exceed 0.5 dB (preferably less than this for high power systems).

For speaker systems driven by amplifiers of less than 30 Watts or so, an insertion loss of 0.5 dB is quite acceptable. But a manufacturer of high power systems will usually try to reduce the insertion loss below 0.5 dB if economically feasible — for the power absorbed by a 0.5 dB filter at 100 Watts input will exceed 10 Watts.

Insertion losses are minimized by using coils of low dc resistance, and capacitors of low power factor.

(Apart from the insertion loss, a further loss of approx 3 dB will occur at the crossover frequency. This is because the amplifier power is divided more or less equally between the two speakers at this frequency. This loss is inevitable whenever a crossover network is used — but in practise it is hardly ever apparent to the ear.)

Speaker drive units used in multiple speaker systems should all have similar nominal impedances. Thus, if the bass driver is an eight ohm unit so also should be the mid-range (if used) and high frequency units. If twin units are used for any part of the spectrum — such as twin tweeters — then each speaker should be twice the nominal impedance of each of the remaining speakers, the twin speakers should then be connected in parallel. (Two 16

SYD 0134

HONEYWELL G265 TIME-SHARING SYD.

ON TTY 41 AT 18:52

USER NUMBER---S85004  
MODERN MAGAZINE H.

PROJECT ID---BARRY  
SYSTEM---BAS  
NEW OR OLD---NEW  
NEW FILE NAME--FILTERS  
READY.

TAPE  
READY

```

15 PRINT USING 18
18:FREQ.      C1      C2      C3      C4      C5      L1      L2      L3      L4      L5
20 PRINT
35 READ R
37 READ M
40 DIM F(25)
50 FOR I=1 TO 22
60 READ F(I)
70 LET W=2*3.14159*F(I)
80 LET Z=R*W
90 LET X=2E6/Z
110 GOSUB 500
120 LET C1=X
130 LET X=1E6/(Z+Z*M)
150 GOSUB 500
160 LET C2=X
170 LET X=1E6/Z
190 GOSUB 500
195 LET C3=X
200 LET X=1E6/(2*Z)
220 GOSUB 500
230 LET C4=X
240 LET X=1E6*(1+M)/Z
255 GOSUB 500
260 LET C5=X
270 LET X=1000*(1+M)*R/W
280 GOSUB 500
285 LET L1=X
290 LET X=1000*R/W
295 GOSUB 500
300 LET L2=X
310 LET X=1000*R/(2*W)
320 GOSUB 500
330 LET L3=X
340 LET X=2*1000*R/W
350 GOSUB 500
360 LET L4=X
370 LET X=1000*R/(W+W*M)
380 GOSUB 500
390 LET L5=X
450 PRINT F(I);TAB(10);C1;TAB(16);C2;TAB(22);C3;TAB(28);C4;TAB(34);C5;
460 PRINT TAB(40);L1;TAB(46);L2;TAB(52);L3;TAB(58);L4;TAB(64);L5
480 NEXT I
500 IF X<100 THEN 520
505 LET X=INT(X+.5)
510 GOTO 600
520 IF X<10 THEN 540
525 LET X=INT(X*10+.1)/10
530 GOTO 600
540 IF X<1 THEN 560
545 LET X=INT(X*100+.5)/100
550 GOTO 600
560 LET X=INT(X*1000+.5)/1000
600 RETURN
700 DATA 8
710 DATA 0.6
720 DATA 100,150,200,250,300,350,400,500,600,750,1000,1250,1500
730 DATA 2000,2500,3000,3500,4000,5000,6000,7500,10000
999 END

RUN
USED      33 UNITS
BYE

TOTAL TTY MINUTES =    35
TOTAL CRU'S USED. =    33

***OFF AT 19:27

```

*This is the computer programme used to calculate the circuit component values in the crossover networks. Copyright — Electronics Today International.*

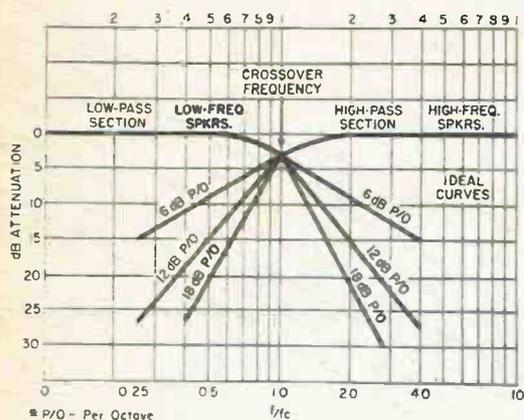


Fig. 1. Typical frequency response curves of a two-way crossover network. The graph shows three different pairs of filters, each having a different rate of attenuation.

ohm speakers connected in parallel will reflect an 8 ohm load impedance to the crossover network.)

## CALCULATING COMPONENT VALUES

Calculation of the component values for crossover networks is a long tedious business. Unless you have a computer that is!

The programme — specially written by our Engineering Manager, Barry Wilkinson — is reproduced elsewhere in this article. Component values have been calculated for both 'constant k' and 'm derived' filters and for speakers of both 8 ohm and 15 ohm nominal impedance; complete design data is given for all crossover frequencies from 100 Hz through 10 kHz.

## CONSTRUCTION

The actual construction of filter networks is quite simple. Air-cored coils are normally used (iron-cored coils can introduce distortion) and these are very simple to wind. (Design data for winding these coils is included in this article.)

Standard (non-polarized) electrolytic capacitors are not suitable for crossover networks, as even low leakage types have an unacceptable power factor for this application. Special non-polarized electrolytics are made by some firms specifically for crossover networks — but it is significant that many of the top speaker manufacturers will only use paper capacitors for this purpose.

Fig. 2

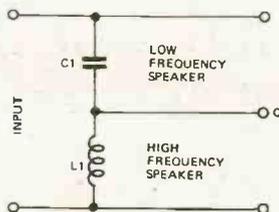


Fig. 4

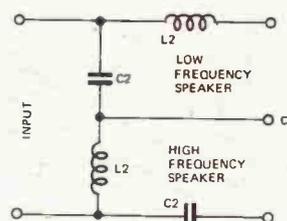


Fig. 3

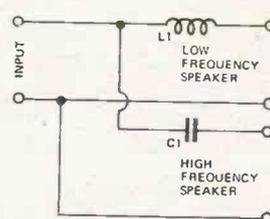
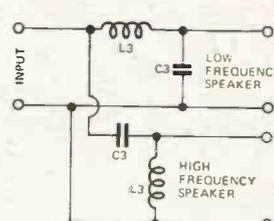


Fig. 5



Constant k crossover networks. Fig. 2 — series type, 6dB/octave. Fig. 3 — parallel type, 6dB/octave. Fig. 4 — series type, 12dB/octave. Fig. 5 — parallel type, 12dB/octave.

When choosing capacitors for crossover networks ensure that the capacitors' rated dc working voltage is never exceeded by the peak voltage of the signal.

## DESIGN PROCEDURE

1. Determine crossover frequencies and attenuation required (i.e. 6dB, 12dB or 18dB per octave).
2. Decide whether filter is to be series or parallel — as explained earlier the parallel type has some advantages over its series counterpart.
3. Select the appropriate circuit from Fig. 2 — 9.

4. Establish component values for required speaker impedance (and either 'constant k' or 'm derived' design) from computer print-out.
5. Design coils using data provided.

## COIL DESIGN

Any coil used in a crossover network has a certain amount of dc resistance — and this resistance will dissipate a proportion of amplifier power. Thus, the dc resistance of the coils should be as low as economically possible. A reasonable compromise — where the amplifier power does not exceed 30 Watts continuous power output per

FREQ.	C1	C2	C3	L1	L2	L3
100	199	281	141	12.7	9	18
150	133	188	93.8	8.49	6	12
200	99.5	141	70.3	6.37	4.5	9
250	79.6	113	56.3	5.09	3.6	7.2
300	66.3	93.8	46.9	4.24	3	6
350	56.8	80.4	40.2	3.64	2.57	5.14
400	49.7	70.3	35.2	3.18	2.25	4.5
500	39.8	56.3	28.1	2.55	1.8	3.6
600	33.2	46.9	23.4	2.12	1.5	3
750	26.5	37.5	18.8	1.7	1.2	2.4
1000	19.9	28.1	14.1	1.27	.9	1.8
1250	15.9	22.5	11.3	1.02	.72	1.44
1500	13.3	18.8	9.38	.849	.6	1.2
2000	9.95	14.1	7.03	.637	.45	.9
2500	7.96	11.3	5.63	.509	.36	.72
3000	6.63	9.38	4.69	.424	.3	.6
3500	5.68	8.04	4.02	.364	.257	.514
4000	4.97	7.03	3.52	.318	.225	.45
5000	3.98	5.63	2.81	.255	.18	.36
6000	3.32	4.69	2.34	.212	.15	.3
7500	2.65	3.75	1.88	.17	.12	.24
10000	1.99	2.81	1.41	.127	.09	.18

COMPONENT VALUES FOR CONSTANT-K LOUDSPEAKER CROSSOVER NETWORKS  
 INDUCTANCE IN MILLIHENRIES  
 CAPACITANCE IN MICROFARADS  
 SPEAKER IMPEDANCE = 8 OHMS

# SPEAKER CROSSOVERS

channel — is to keep the dc resistance below one ohm.

The coil design that will provide the highest inductance in proportion to the dc resistance is that shown in Fig. 10. In this drawing the radius of the circular winding bobbin is shown as 'x' and all other dimensions are related to this.

Construction of the bobbin is not critical and it can be made from cardboard, or a combination of a wooden core and cardboard cheeks. Metal must not be used.

The design procedure is as follows:—

1. Determine the bobbin size required. This is done by using Graph I. This graph indicates bobbin size ('x' measurement) required to accommodate coils of different sizes and dc resistances. For example a 5.5mH coil wound on a 0.75" former (remember that this refers to the measurement shown as 'x') will have a dc resistance of 1.4 ohms — if wound on a 1" former the resistance would be 0.7 ohms. As the dc resistance should be preferably less than 1.0 ohm, the 1" former should be used.
2. Graph II shows the number of turns required to provide the required inductance for various bobbin sizes. In our example 290 turns are required.
3. Graph III shows the wire gauge required. In our example 290 turns on a 1" bobbin would require 15.3 B&S. The nearest standard size is 16G so this is the wire size used.
4. The coil should be layer wound using enameled copper wire. As the operating voltage is quite low, no interlayer insulation is required.

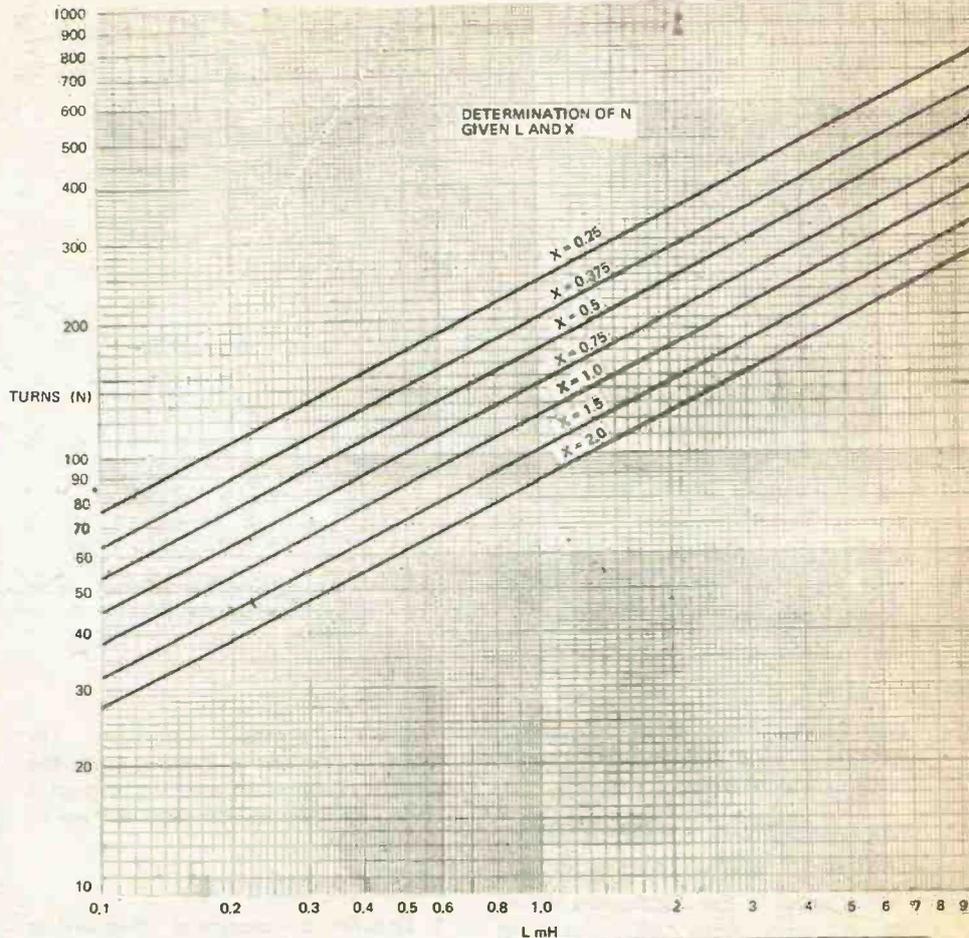
Graph IV shows the dc resistance of the coil given the wire gauge and former size, providing the former is filled completely. In our case the resistance shown is 1.0 ohm — but as we have only 290 turns whereas the filled bobbin accommodates about 350 turns the resistance would be about 0.8 ohm. This is sufficiently close to our design requirement and would be acceptable for a low power amplifier.

Figure 11 provides the approximate weight of wire used for fully wound coils of each size.

## THREE-WAY CROSSOVER NETWORKS

These differ from two-way networks only in that they include a mid-range filter.

A three-way 12 dB per octave parallel crossover is shown in Fig. 12.



The midrange section is a bandpass filter consisting of a low pass filter (L3B and C3B) and a high pass filter (L3A and C3A).

The design procedure is firstly to establish the values of the low pass section L3 and C3.

Circuit elements L3A and C3A, in the mid-range filter, must also cross over at the same frequency as the low pass filter, and thus have the same values as L3 and C3.

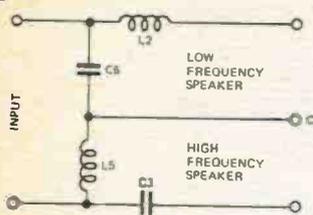
The values of L3B is the same as L3C as also are C3B and C3C. The values of

FREQ.	C1	C2	C3	L1	L2	L3
100	106	150	75	23.9	16.9	33.8
150	70.7	100	50	15.9	11.3	22.5
200	53.1	75	37.5	11.9	8.44	16.9
250	42.4	60	30	9.55	6.75	13.5
300	35.4	50	25	7.96	5.63	11.3
350	30.3	42.9	21.4	6.82	4.82	9.65
400	26.5	37.5	18.8	5.97	4.22	8.44
500	21.2	30	15	4.77	3.38	6.75
600	17.7	25	12.5	3.98	2.81	5.63
750	14.1	20	10	3.18	2.25	4.5
1000	10.6	15	7.5	2.39	1.69	3.38
1250	8.49	12	6	1.91	1.35	2.7
1500	7.07	10	5	1.59	1.13	2.25
2000	5.31	7.5	3.75	1.19	.844	1.69
2500	4.24	6	3	.955	.675	1.35
3000	3.54	5	2.5	.796	.563	1.13
3500	3.03	4.29	2.14	.682	.482	.965
4000	2.65	3.75	1.88	.597	.422	.844
5000	2.12	3	1.5	.477	.338	.675
6000	1.77	2.5	1.25	.398	.281	.563
7500	1.41	2				

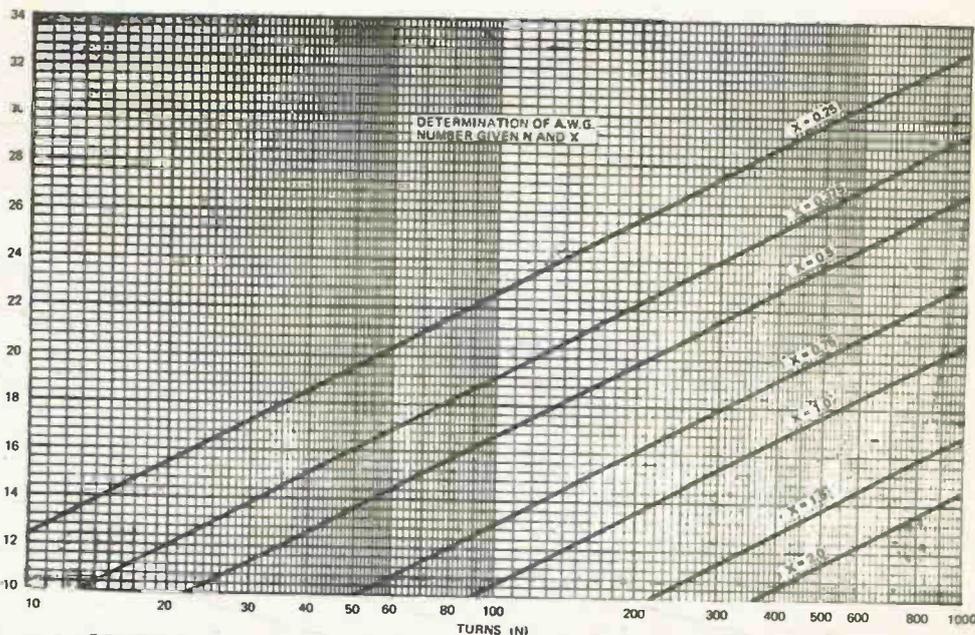
COMPONENT VALUES FOR CONSTANT-K LOUDSPEAKER CROSSOVER NETWORKS  
 INDUCTANCE IN MILLIHENRIES  
 CAPACITANCE IN MICROFARADS  
 SPEAKER IMPEDANCE = 15 OHMS

# SPEAKER CROSSOVERS

Fig. 6



A.W.G.  
(B&S)



GRAPH III

Fig. 7

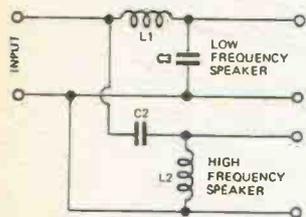


Fig. 8

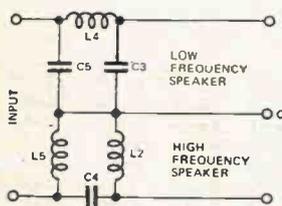


Fig. 9

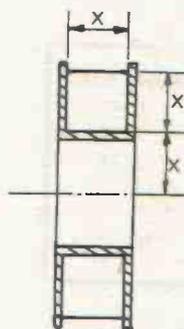
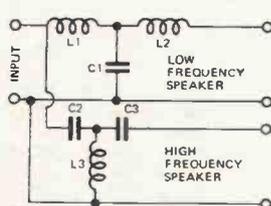


Fig. 10. Recommended dimensions of bobbin design. All dimensions are related to 'x' — thus if the measurement specified for 'x' is 1" then the winding width and depth are both 1", and the bobbin internal diameter is 2".

M-derived crossover networks. Fig. 6 — series type, 12 dB/octave. Fig. 7 — parallel type, 12 dB/octave. Fig. 8 — series type, 18 dB/octave. Fig. 9 — parallel type, 18 dB/octave.

these components are determined for the changeover frequency of the mid-range and tweeter speakers.

Four and five way networks are designed in a similar fashion.

The component values quoted for the three way 'constant k' network described above can be converted to an 'm derived network' by the following equations.

$$L3, L3B = (1 + m) \frac{R_0}{\omega C} \text{ Henry}$$

$$L3A, L3C = \left( \frac{R}{\omega C} \right) \text{ Henry}$$

$$C3A, C3C = \left( \frac{1}{1+m} \right) \left( \frac{1}{\omega C R_c} \right) \times 10^6 \mu F$$

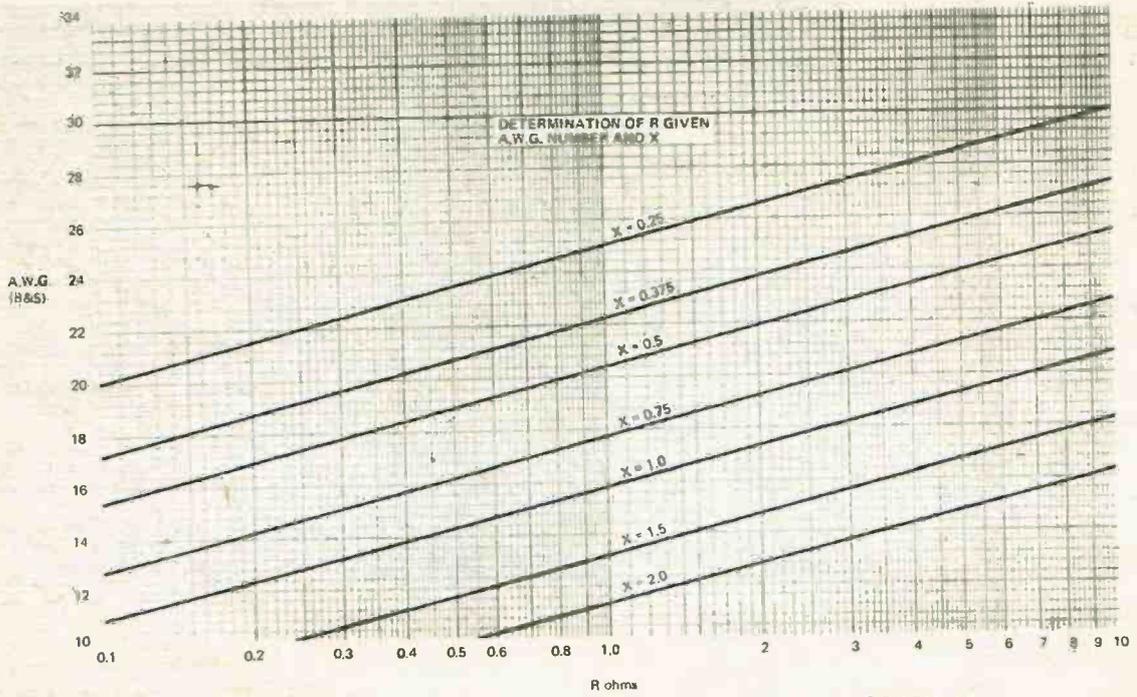
$$C3B = \left( \frac{10^6}{\omega C R_c} \right) \mu F$$

FREQ.	C1	C2	C3	C4	C5	L1	L2	L3	L4	L5
100	398	124	199	99.4	318	20.3	12.7	6.37	25.4	7.96
150	265	82.9	133	66.3	212	13.5	8.49	4.24	16.9	5.31
200	199	62.1	99.4	49.7	159	10.1	6.37	3.18	12.7	3.98
250	159	49.7	79.5	39.7	127	8.15	5.09	2.55	10.1	3.18
300	133	41.4	66.3	33.1	106	6.79	4.24	2.12	8.49	2.65
350	114	35.5	56.8	28.4	90.9	5.82	3.64	1.82	7.28	2.27
400	99.4	31	49.7	24.8	79.5	5.09	3.18	1.59	6.37	1.99
500	79.5	24.8	39.7	19.9	63.6	4.07	2.55	1.27	5.09	1.59
600	66.3	20.7	33.1	16.5	53	3.4	2.12	1.06	4.24	1.33
750	53	16.5	26.5	13.2	42.4	2.72	1.7	.849	3.4	1.06
1000	39.7	12.4	19.9	9.95	31.8	2.04	1.27	.637	2.55	.796
1250	31.8	9.95	15.9	7.96	25.4	1.63	1.02	.509	2.04	.637
1500	26.5	8.29	13.2	6.63	21.2	1.36	.849	.424	1.7	.531
2000	19.9	6.22	9.95	4.97	15.9	1.02	.637	.318	1.27	.398
2500	15.9	4.97	7.96	3.98	12.7	.815	.509	.255	1.02	.318
3000	13.2	4.14	6.63	3.32	10.6	.679	.424	.212	.849	.265
3500	11.3	3.55	5.68	2.84	9.09	.582	.364	.182	.728	.227
4000	9.95	3.11	4.97	2.49	7.96	.509	.318	.159	.637	.199
5000	7.96	2.49	3.98	1.99	6.37	.407	.255	.127	.509	.159
6000	6.63	2.07	3.32	1.66	5.31	.34	.212	.106	.424	.133
7500	5.31	1.66	2.65	1.33	4.24	.272	.17	.085	.34	.106
10000	3.98	1.24	1.99	.995	3.18	.204	.127	.064	.255	.08

COMPONENT VALUES FOR M-DERIVED LOUDSPEAKER CROSSOVER NETWORKS  
INDUCTANCE IN MILLIHENRIES  
CAPACITANCE IN MICROFARADS  
M=0.6  
SPEAKER IMPEDANCE = 8 OHMS

Electronics Today International would like to thank Mr I.C. Hansen for his very valuable assistance in providing design data for the coils described in this article.

# SPEAKER CROSSOVERS



GRAPH IV

'x' inches	Weight (ozs.)
0.25	0.5
0.375	1.75
0.5	4.25
0.75	14
1.00	33
1.50	110

Fig. 11 - This table shows the weight of wire used for fully wound coils of each size of 'x'.

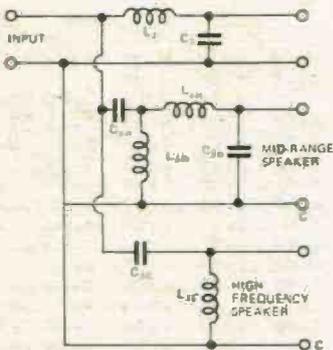


Fig. 12. Three-way 12 dB per octave parallel crossover.

FREQ.	C1	C2	C3	C4	C5	L1	L2	L3	L4	L5
100	212	66.3	106	53	170	38.2	23.8	11.9	47.7	14.9
150	141	44.2	70.7	35.3	113	25.4	15.9	7.96	31.8	9.95
200	106	33.1	53	26.5	84.8	19.1	11.9	5.97	23.8	7.46
250	84.8	26.5	42.4	21.2	67.9	15.2	9.55	4.77	19.1	5.97
300	70.7	22.1	35.3	17.6	56.5	12.7	7.96	3.98	15.9	4.97
350	60.6	18.9	30.3	15.1	48.5	10.9	6.82	3.41	13.6	4.26
400	53	16.5	26.5	13.2	42.4	9.55	5.97	2.98	11.9	3.73
500	42.4	13.2	21.2	10.6	33.9	7.64	4.77	2.39	9.55	2.98
600	35.3	11	17.6	8.84	28.3	6.37	3.98	1.99	7.96	2.49
750	28.3	8.84	14.1	7.07	22.6	5.09	3.18	1.59	6.37	1.99
1000	21.2	6.63	10.6	5.31	16.9	3.82	2.39	1.19	4.77	1.49
1250	16.9	5.31	8.49	4.24	13.5	3.06	1.91	.955	3.82	1.19
1500	14.1	4.42	7.07	3.54	11.3	2.55	1.59	.796	3.18	.995
2000	10.6	3.32	5.31	2.65	8.49	1.91	1.19	.597	2.39	.746
2500	8.49	2.65	4.24	2.12	6.79	1.53	.955	.477	1.91	.597
3000	7.07	2.21	3.54	1.77	5.66	1.27	.796	.398	1.59	.497
3500	6.06	1.89	3.03	1.52	4.85	1.09	.682	.341	1.36	.426
4000	5.31	1.66	2.65	1.33	4.24	.955	.597	.298	1.19	.373
5000	4.24	1.33	2.12	1.06	3.4	.764	.477	.239	.955	.298
6000	3.54	1.11	1.77	.884	2.83	.637	.398	.199	.796	.249
7500	2.83	.884	1.41	.707	2.26	.509	.318	.159	.637	.199
10000	2.12	.663	1.06	.531	1.7	.382	.239	.119	.477	.149

COMPONENT VALUES FOR M-DERIVED LOUDSPEAKER CROSSOVER NETWORKS  
 INDUCTANCE IN MILLIHENRIES  
 CAPACITANCE IN MICROFARADS  
 M=0.6  
 SPEAKER IMPEDANCE = 15 OHMS

# BATTERIES

System	Anode	Cathode	Electrolyte	Open current voltage,	Typical operating voltage,	Capacity Ah/lb <sup>a</sup> Wh/lb <sup>b</sup>		Remarks
<i>Primary systems</i> Leclanche	Zn(Hg)	MnO <sub>2</sub> (C)	NH <sub>4</sub> Cl-ZnCl <sub>2</sub>	1.6	0.9-1.4	23	30	most common form of the dry cell; extensive applications; sloping discharge curve suitable for greater drain rates than Leclanche cells; sloping discharge curve constant voltage during discharge; heavier drains and higher capacity than Leclanche cells higher capacity and voltage than Leclanche cell; 35-40% of magnesium consumed in hydrogen liberation during discharge utilizes oxygen from air; wet type used for railway signals, home radios; dry type available but moisture loss a problem one-shot, high-drain-rate reserve cell; also available in another form as a secondary battery one-shot, high-drain-rate reserve cell low operating voltage; used principally for railway signals one-shot, high-drain-rate cell one-shot, high-drain reserve types used for military applications [must be heated to melt electrolyte, one-shot, high-drain, military type cell one-shot, high-drain reserve cell; may be activated with sea water one-shot, high-drain reserve cell with very high capacity; may be activated with sea water available commercially as a bias cell providing stable voltage at zero current over long periods used as a voltage standard power source for radiation warning devices; charging source for low-leakage capacitors used extensively in foreign countries for telegraphy and telephone service; performs efficiently at low drains small sealed primary cell for electric wrist watch or hearing aid use; hermetically sealed electric wrist watch battery; hermetically sealed one of the more promising of a large number of organic compounds being considered; multistep discharge typical of a variety of similar systems that could be made commercially if applications warranted: thin flat cells providing 100 V/in. of battery length higher capacity and voltage than Leclanche cell; wasteful corrosion of aluminium remains a major problem operates 0.2-0.3 V lower potential than HgO-Zn system but otherwise has similar voltage-time discharge characteristics conventional lead storage cell much longer useful life than lead storage cell but lower capacity and more expensive available as a completely sealed cell very high capacity and suitable for high discharge rates but smaller number of cycles than Ni-Cd or lead storage cells greater number of cycles than Zn-AgO secondary cell available only as a completely sealed cell
Alkaline Zn-MnO <sub>2</sub>	Zn(Hg)	MnO <sub>2</sub> (C)	KOH	1.52	0.9-1.2	30	33	
mercury cell (Rubin)	Zn(Hg)	HgO(C)	KOH-ZnO	1.35	1.30	40	53	
Mg-MnO <sub>2</sub>	Mg	MnO <sub>2</sub> (C)	MgBr <sub>3</sub> , Li <sub>2</sub> CrO <sub>4</sub>	2.0	1.6-1.8	30	46	
air-depolarised Zn	Zn(Hg)	O <sub>2</sub> (C)	KOH	1.36	1.1-1.2	60 <sup>b</sup>	70 <sup>b</sup>	
Zn-AgO	Zn	AgO	KOH	1.8	1.4-1.5	35	53	
Zn-PbO <sub>2</sub>	Zn	PbO <sub>2</sub>	H <sub>2</sub> SO <sub>4</sub>	2.5	2.0-2.3	18	26	
Zn-CuO	Zn(Hg)	CuO	NaOH	1.06	0.9-1.0			
Zn-Cl <sub>2</sub> Pb-PbO <sub>2</sub> -HClO <sub>4</sub>	Zn Pb	ZnCl <sub>2</sub> PbO <sub>2</sub>	Cl <sub>2</sub> HClO <sub>4</sub>	2.1 2.0	1.5-1.9 1.6-1.8	11	19	
thermal cell	Ca	PbCrO <sub>4</sub> (Ni)	LiCl, KC1 fused	2.8	2.2			
Mg-Cu <sub>2</sub> Cl	Mg	Cu <sub>2</sub> Cl	MgCl <sub>2</sub>	1.4	1.1-1.3	25	30	
Mg-AgCl	Mg	AgCl	MgCl <sub>2</sub>	1.6	1.3-1.5	54	75	
Zn-V <sub>2</sub> O <sub>3</sub>	Zn	V <sub>2</sub> O <sub>3</sub>	NH <sub>4</sub> OH, ethylene glycol, boric acid	1.2	1.2			
Weston standard solid electrolyte	Cd(Hg) Ag	Hg CuBr <sub>2</sub>	HgSO <sub>4</sub> , CdSO <sub>4</sub> AgBr	1.019 0.75	1.2			
Fery cell	Zn	O <sub>2</sub> (C)	NH <sub>4</sub> Cl	1.2	0.7			
Zn-Ag <sub>2</sub> O	Zn	Ag <sub>2</sub> O	KOH or NuOH	1.6	1.5	33	49	
In-HgO	In	HgO	KOH	1.15	1.05			
<i>Experimental cells</i> (primary) organic cathode	Mg	<i>m</i> -dinitro- benzene	MgBr <sub>2</sub>	1.65	1.15 <sup>d</sup>	56	65	
solid electrolyte	Ag	V <sub>2</sub> O <sub>3</sub>	AgI	0.46	0.38-0.46 <sup>d</sup>			
Al-MnO <sub>2</sub>	Al	MnO <sub>2</sub>	AlCl <sub>3</sub> , (NH <sub>4</sub> ) <sub>2</sub> Cr <sub>2</sub> O <sub>7</sub>	1.7	1.3			
Mg-Bi <sub>2</sub> O <sub>3</sub>	Mg	Bi <sub>2</sub> O <sub>3</sub>	MgBr <sub>2</sub>	1.6	1.0	6		
<i>Secondary cells</i> lead acid Edison	Pb Fe	PbO <sub>2</sub> Ni oxides	H <sub>2</sub> SO <sub>4</sub> KOH	2.2 1.6	1.95-2.05 1.2-1.4	10 10	20 13	
Cd-NiO Zn-AgO	Cd Zn	Ni oxides AgO	KOH KOH, ZnO	1.35 1.8	1.1-1.3 1.4-1.5	10 35	12 53	
Cd-AgO	Cd	AgO	KOH	1.4	1.0-1.1	30	33	
MnO-Zn	Zn	MnO <sub>2</sub>	KOH	1.5		8	9	

<sup>a</sup> Based on total weight of commonly available size of commercial cells.

<sup>b</sup> Exclusive of O<sub>2</sub> consumed from air.

<sup>c</sup> Average voltage for light-drain application.

<sup>d</sup> 2 X 10<sup>-10</sup> A/in.<sup>2</sup> drain for the first 7 yr of cell life.

# CONVERSION TABLES

8		7		6		5		4		3		2		1	
Hex	Decimal	Hex	Decimal	Hex	Decimal	Hex	Decimal	Hex	Decimal	Hex	Decimal	Hex	Decimal	Hex	Decimal
0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
1	268,435,456	1	16,777,216	1	1,048,576	1	65,536	1	4,096	1	256	1	16	1	1
2	536,870,912	2	33,554,432	2	2,097,152	2	131,072	2	8,192	2	512	2	32	2	2
3	805,306,368	3	50,331,648	3	3,145,728	3	196,608	3	12,288	3	768	3	48	3	3
4	1,073,741,824	4	67,108,864	4	4,194,304	4	262,144	4	16,384	4	1,024	4	64	4	4
5	1,342,177,280	5	83,886,080	5	5,242,880	5	327,680	5	20,480	5	1,280	5	80	5	5
6	1,610,612,736	6	100,663,296	6	6,291,456	6	393,216	6	24,576	6	1,536	6	96	6	6
7	1,879,048,192	7	117,440,512	7	7,340,032	7	458,752	7	28,672	7	1,792	7	112	7	7
8	2,147,483,648	8	134,217,728	8	8,388,608	8	524,288	8	32,768	8	2,048	8	128	8	8
9	2,415,919,104	9	150,994,944	9	9,437,184	9	589,824	9	36,864	9	2,304	9	144	9	9
A	2,684,354,560	A	167,772,160	A	10,485,760	A	655,360	A	40,960	A	2,560	A	160	A	10
B	2,952,790,016	B	184,549,376	B	11,534,336	B	720,896	B	45,056	B	2,816	B	176	B	11
C	3,221,225,472	C	201,326,592	C	12,582,912	C	786,432	C	49,152	C	3,072	C	192	C	12
D	3,489,660,928	D	218,103,808	D	13,631,488	D	851,968	D	53,248	D	3,328	D	208	D	13
E	3,758,096,384	E	234,881,024	E	14,680,064	E	917,504	E	57,344	E	3,584	E	224	E	14
F	4,026,531,840	F	251,658,240	F	15,728,640	F	983,040	F	61,440	F	3,840	F	240	F	15
8		7		6		5		4		3		2		1	

## TO CONVERT HEXADECIMAL TO DECIMAL

1. Locate column of decimal numbers corresponding to left-most digit or letter of hexadecimal select from this column and record number that corresponds to position of hexadecimal digit or letter.
2. Repeat step 1 for next (second from left) position.
3. Repeat step 1 for units (third from left) position.
4. Add numbers selected from table to form decimal number.

## TO CONVERT DECIMAL TO HEXADECIMAL

1. (A) select from table highest decimal number that is equal to or less than number to be converted.  
(B) Record hexadecimal of column containing selected number.  
(C) Subtract selected decimal from number to be converted.
2. Using remainder from step 1 (C) repeat all of step 1 to develop second position of hexadecimal (and remainder).
3. Using remainder from step 2 repeat all of step 1 to develop units position of hexadecimal.
4. Combine terms to form hexadecimal number.

## Decimal-Hex-Octal-Binary-ASCII

Dec	Octal	Hex	Binary Bit Pattern 7 6 5 4 3 2 1	ASCII Character	Dec	Octal	Hex	Binary Bit Pattern 7 6 5 4 3 2 1	ASCII Character	Dec	Octal	Hex	Binary Bit Pattern 7 6 5 4 3 2 1	ASCII Character
0	0	0	0000000		43	53	2B	0101011	+	86	126	56	10101110	V
1	1	1	0000001		44	54	2C	0101100	,	87	127	57	10101111	W
2	2	2	0000010		45	55	2D	0101101	-	88	130	58	10110000	X
3	3	3	0000011		46	56	2E	0101110	.	89	131	59	10110001	Y
4	4	4	0000100		47	57	2F	0101111	/	90	132	5A	10110010	Z
5	5	5	0000101		48	60	30	01100000	0	91	133	5B	10110011	[
6	6	6	0000110		49	61	31	01100001	1	92	134	5C	10110100	/
7	7	7	0000111		50	62	32	01100010	2	93	135	5D	10110101	]
8	10	8	0001000		51	63	33	01100011	3	94	136	5E	10111000	↑
9	11	9	0001001		52	64	34	01100100	4	95	137	5F	10111001	←
10	12	A	0001010		53	65	35	01100101	5	96	140	60	11000000	_
11	13	B	0001011		54	66	36	01100110	6	97	141	61	11000001	a
12	14	C	0001100		55	67	37	01100111	7	98	142	62	11000010	b
13	15	D	0001101		56	70	38	01110000	8	99	143	63	11000011	c
14	16	E	0001110		57	71	39	01110001	9	100	144	64	11000100	d
15	17	F	0001111		58	72	3A	01110010	:	101	145	65	11000101	e
16	20	10	0010000		59	73	3B	01110011	;	102	146	66	11000110	f
17	21	11	0010001		60	74	3C	01110100	<	103	147	67	11001000	g
18	22	12	00100010		61	75	3D	01110101	=	104	150	68	11001001	h
19	23	13	00100011		62	76	3E	01110110	>	105	151	69	11001010	i
20	24	14	00100100		63	77	3F	01110111	?	106	152	6A	11001011	j
21	25	15	00100101		64	100	40	10000000	@	107	153	6B	11001011	k
22	26	16	00100110		65	101	41	10000001	A	108	154	6C	11001100	l
23	27	17	00100111		66	102	42	10000010	B	109	155	6D	11001101	m
24	30	18	00110000		67	103	43	10000011	C	110	156	6E	11001110	n
25	31	19	00110001		68	104	44	10000100	D	111	157	6F	11001111	o
26	32	1A	00110010		69	105	45	10000101	E	112	160	70	11100000	p
27	33	1B	00110011		70	106	46	10000110	F	113	161	71	11100001	q
28	34	1C	00110100		71	107	47	10000111	G	114	162	72	11100010	r
29	35	1D	00110101		72	110	48	10001000	H	115	163	73	11100011	s
30	36	1E	00110110		73	111	49	10001001	I	116	164	74	11100100	t
31	37	1F	00110111		74	112	4A	10001010	J	117	165	75	11100101	u
32	40	20	00100000		75	113	4B	10001011	K	118	166	76	11100110	v
33	41	21	00100001		76	114	4C	10001100	L	119	167	77	11100111	w
34	42	22	00100010		77	115	4D	10001101	M	120	170	78	11101000	x
35	43	23	00100011		78	116	4E	10001110	N	121	171	79	11101001	y
36	44	24	00100100		79	117	4F	10001111	O	122	172	7A	11110000	z
37	45	25	00100101		80	120	50	10100000	P	123	173	7B	11110001	_
38	46	26	00100110		81	121	51	10100001	Q	124	174	7C	11110010	-
39	47	27	00100111		82	122	52	10100010	R	125	175	7D	11110011	-
40	50	28	00101000		83	123	53	10100011	S	126	176	7E	11110100	-
41	51	29	00101001		84	124	54	10100100	T	127	177	7F	11110101	-
42	52	2A	00101010		85	125	55	10100101	U					

— means special characters or codes not used.

## CMOS/TTL COMPARISON

Logic family	Noise Immunity Volts	Prop. delay nS	Fan Out	Max. Toggle Speed MHZ	Supply Voltage			Power Diss. per package mW (typ)	Decoupling and other requirements
					Nominal V	Min. V	Max. V		
74 Series	0.4	9	10	15	5.0	4.75	5.25	40	0.1 uF Ceramic capacitor for every 8 packages to eliminate switching current spike
74H Series	0.4	6	10	40	5.0	4.75	5.25	60	
74S Series	0.3	3	10	125	5.0	4.75	5.25	40	
74LS Series	0.3	9	10	25	5.0	4.75	5.25	8	
CMOS	4.5	30	>50	10	—	3.0	18.0	0.01	

### TTL BIPOLAR LOGIC

The 74 Series of transistor-transistor logic is a medium speed family of saturating integrated circuit logic designed for general digital logic application requiring clock frequencies to 30MHz and switching speeds in the 7-11 nS range under moderate capacitive loading.

The circuits are identified by a multiple emitter input transistor and an active "pull up" in the upper output network. Clamp diodes are provided at each input to limit the undershoot that occurs in typical system applications such as driving long interconnect wiring. The active pull-up output configuration provides low output impedance in the high output state. The resulting low impedances in both output states ensures excellent a.c. noise immunity and allows a high-speed operation with capacitive loads.

### COMPLEMENTARY MOS (CMOS)

Complementary MOS is the newest of the general-purpose logic families.

The following are primary design features of the whole of the COS/MOS and McMOS ranges.

- Double diode protection on all inputs.
- Noise immunity typically 45% of VDD, 30% of VDD minimum.
- Buffered output compatible with MHTL and Low Power TTL.
- Low quiescent power dissipation: 25nW typ. per package.
- Wide power supply voltage: 3-18 Volt dependent on type.
- Single supply operation.
- High fanout: greater than 50
- High input impedance: 10<sup>11</sup> ohms typ.
- Low input capacitance: 5pf typ.

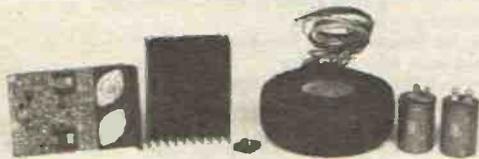
## TTL FUNCTIONS

Device	Description	Device	Description
7400	Quad 2-input Positive NAND Gate	7481	16-bit Active Element Memory
7401	Quad 2-input Positive NAND Gate (open collector o/p)	7482	2-bit Binary Full Addder
7401A	Quad 2-input Positive NAND Gate (open collector o/p)	7484	16-bit Active Element Memory
7402	Quad 2-input Positive NOR Gate (open collector o/p)	7485	4-bit Comparator
7403	Quad 2-input Positive NAND Gate (open collector o/p)	7486	Quad 2-input Exclusive Or Gate
7404	Hex Inverter	7489	64-bit RAM (16 x 4W)
7405A	Hex Inverter (open collector o/p)	7490	Decade Counter
7406	Hex Inverter/Buffer 30V o/p	7491	8-bit Shift Registers
7407	Hex Buffer 30V o/p	7492	Divide-by-twelve Counter
7408	Quad 2-input Positive AND Gate	7493	4-bit Binary Counter
7409	Quad 2-input Positive AND Gate	7494	4-bit Shift Registers (Parallel-In, Serial-Out)
7410	Triple 3-input Positive NAND Gate	7495	4-bit Right Shift, Left Shift Register
7412	Triple 3-input NAND Gate (open collector o/p)	7496	5-bit Shift Registers (Dual Para-in, Para-Out)
7413	Dual 4-input Schmitt Trigger	74100	8-bit Bistable Latch
7414	Schmitt Hex Inverter Buffer	74107	Dual J-K Master Slave Flip Flop
7416	Hex Inverter/Buffer 15V o/p	74121	Monostable Multivibrator
7417	Hex Buffer 15V o/p	74122	Monostable Multivibrator with reset
7420	Dual 4-input Positive NAND Gate	74123	Dual Monostable Multivibrator with reset
7421	Dual 4-input AND Gate	74124	Universal Pulse Generator
7426	Quad 2-input High Voltage Interface NAND Gate	74138	3 line to 8 line Decoder/Demultiplexer
7427	Triple 3-input NOR Gate	74141	BCD-to-Decimal Decoder/Driver
7428	Quad 2-input NOR Buffer (Fan Out 30)	74145	BCD-to-Seven Segment Decoder/Driver 15V output
7430	8-input Positive NAND Gate	74150	16-bit Data Selector
7432	Quad 2-input OR Gate	74151	8-bit Data Selector (with strobe)
7433A	Quad 2-input NOR Buffer 15V	74153	Dual 4 to 1 line Data Selector 1 MPX
7437	Quad 2-input NAND Buffer	74154	4 line to 16 line Decoder
7438A	Quad 2-input NAND Buffer 15V	74155	Dual 2-to-4 line Decoder/DeMPX (totem pole output)
7441A	BCD-to-Decimal Decoder/Nixie Driver	74156	Dual 2-to-4 line Decoder/DeMPX (open collector output)
7442	BCD-to-Decimal Decoder	74157	Quad 2 line to 1 line Selector
7445	BCD-to-Decimal Decoder/Driver 30V output o/c	74160	Synchronous Decade Counter
7446A	BCD-to-Seven Segment Decoder/Driver 30V/40mA	74162	Synchronous Decade Counter
7447	BCD-to-Seven Segment Decoder/Driver 15V/20mA	74163	Synchronous Binary Counter
7447A	BCD-to-Seven Segment Decoder/Driver 15V/40mA	74164	8-bit Shift Register, Serial In-Parallel Out
7448	BCD-to-Seven Segment Decoder/Driver	74165	8-bit Shift Register, Parallel In-Serial Out
7450	Expandable Dual 2 wide, 2 i/p AND-OR-INVERT Gate	74174	Hex Type "D" Flip Flop
7451	Dual 2 wide, 2 i/p AND-OR-INVERT Gate	74175	Quad "D" Flip Flop with common reset
7453	Expandable 4 wide, 2 i/p AND-OR-INVERT Gate	74180	8-bit Odd/Even Parity Generator/Checkers
7454	4 wide 2-input AND-OR-INVERT Gate	74181	4-bit Arithmetic Logic Out
7460	Dual 4-input Expander	74182	Carry-Look-Ahead Unit
7470	Positive Edge-triggered J-K Flip Flops	74190	Synchronous Up/Down Decade Counter (Single Clock Unit)
7472	J-K Master-Slave Flip Flops (AND inputs)	74191	Synchronous Up/Down 4-bit Binary Counter
7473	Dual J-K Master Slave Flip Flops	74192	Synchronous 4-bit Up/Down Counter
7474	Dual D-Type Edge Triggered Flip Flops	74193	Synchronous 4-bit Up/Down Counter
7475	4-bit bistable latch = Quad bistable latch	74195	Synchronous 4-bit Parallel Shift Register with J-K inputs
7476	Dual J-K Master Slave Flip Flops + preset and clear	74196	50Mhz Presettable Decade Counter/Latch (Bi-Quinary)
		74200	256-bit Random Access Memory (RAM)

## CMOS FUNCTIONS

Device	Description
CD4000	Dual 3-Input NOR gate plus Inverter
CD4001	Quad 2-Input NOR Gate
CD4002	Dual 4-Input NOR Gate
CD4006	18-Stage Static Shift Register
CD4007	Dual Complementary Pair Plus Inverter
CD4008	4-Bit full Adder with Parallel Carry
CD4009	Hex Buffer/Converter (Inverting)
CD4010	Hex Buffer/Converter (Non-Inverting)
CD4011	Quad 2-Input NAND Gate
CD4012	Dual 4-Input NAND Gate
CD4013	Dual "D" Flip-Flop with Set/Reset
CD4014	8-Stage Static Shift Register
CD4015	Dual 4-Stage Static Shift Register
CD4016	Quad Bilateral Switch
CD4017	Decade Counter/Divider
CD4018	Presetable Divide-By-"N" Counter
CD4019	Quad AND-OR Select Gate
CD4020	14-Stage Binary Ripple Counter
CD4021	8-Stage Static Shift Register
CD4022	Divide-by-8 Counter/Divider
CD4023	Triple 3-Input NAND Gate
CD4024	7-Stage Binary Counter
CD4025	Triple 3-Input NOR Gate
CD4026	Decade Counter/Divider
CD4027	Dual J K Master Slave Flip-Flop
CD4028	BCD TO-Decimal Decoder
CD4029	Presetable Up/Down Counter
CD4030	Quad Exclusive-OR Gate
CD4035	4-Stage Parallel IN/OUT Shift Register
CD4040	12-Stage Binary Ripple Counter
CD4042	Quad Clocked "D" Latch
CD4046	Micropower Phase-Locked Loop
CD4049	Hex Buffer/Converter (Inverting)
CD4050	Hex Buffer/Converter (Non-Inverting)
CD4051	Single 8-Channel Multiplexer
CD4052	Differential 4-Channel Multiplexer
CD4054	4-Line Liquid Crystal Display Driver
CD4056	BCD-7-Segment Decoder/Driver
CD4059	Programmable Divide-by-N Counter
CD4060	14-Stage Counter and Oscillator
CD4061	256-Word X 1-Bit Static Ram
CD4066	Quad Bilateral Switch
CD4068	8-Input NAND Gate
CD4069	Hex Inverter
CD4070	Quad Exclusive OR Gate
CD4071	Quad 2-Input OR Gate
CD4077	Quad Exclusive NOR Gate
CD4081	Quad 2-Input AND Gate
CD4082	Dual 4-Input AND Gate
CD4085	Dual 2-Wide 2-Input AOI Gate
CD4086	Expendable 4-Wide 2-Input AOI Gate
CD4093	Quad 2-Input NAND Schmitt Trigger
CD4099	8-Bit Addressable Latch
CD4510	BCD UP/DOWN Counter
CD4511	BCD TO 7-Segment Decoder/Driver
CD4514	1 to 16 Decoder (Output High)
CD4515	1 to 16 Decoder (Output Low)
CD4516	Binary UP/DOWN Counter
CD4518	Dual BCD UP Counter
CD4528	Dual Retriggerable Monostable
MC14502	Strobed Hex Inverter/Buffer
MC14517	Dual 64-bit Static Shift Register
MC14521	24 State Frequency Divider
MC14522	Programmable divide by N-4 bit Counter (BCD)
MC14526	Programmable divide by N-4 bit Counter (binary)
MC14534	Real Time 5-Decade Counter
MC14536	Programmable Timer
MC14543	BCD-to-Seven Segment Latch/Decoder/Driver
MC14553	Three-Digit BCD Counter
MC14566	Industrial time base Generator

## TWO NEW SUPERMODULES 170W INTO 4 OR 8 OHMS



By popular demand we have designed higher powered versions of our well known modules. The CE 1704 which gives 170W into 4 ohms and the CE 1708 which gives 170W into 8 ohms are physically similar to the original types and have the same combination of compatible performance features which makes CRIMSON amplification audibly superior to the competition and the only choice if you have an ear for music. We have also produced suitable power supplies which again use our superb TOROIDAL TRANSFORMERS, only 50mm high, with a 120-240 primary and single bolt fixing. Write of phone for more information and biased opinions.

	Home	Europe
<b>POWER AMPLIFIER MODULES</b>		
CE 606 60W/8 ohms 35-0-35V	£16.30	£16.60
CE 1004 100W/4 ohms 36-0-35V	£19.22	£19.30
CE 1008 100W/8 ohms 45-0-45V	£23.22	£23.00
CE 1704 170W/4 ohms 45-0-45V	£29.12	£28.46
CE 1708 170W/8 ohms 60-0-60V	£31.90	£31.04
<b>TOROIDAL POWER SUPPLIES</b>		
CPS 1 for 2 x CE 606 or 1 x CE 1004	£14.47	£18.40
CPS 2 for 2 x CE 1004 2/4 x CE 606	£16.82	£20.57
CPS 3 for 2 x CE 1008 or 1 x CE 1704	£17.56	£21.35
CPS 4 for 1 x CE 1008	£15.31	£19.15
CPS 5 for 1 x CE 1708	£22.68	£26.50
CPS 6 for 2 x CE 1704 or 2 x CE 1708	£23.98	£27.70
<b>HEATSINKS</b>		
Light Duty 50mm 2 C/W	£0.90	£1.30
Medium Power 100mm 1.4 C/W	£1.60	£2.40
Disc/Drop 150mm 1.4 C/W	£2.30	£3.65
<b>THERMAL CUT-OUTS</b>		
Recommended for improved reliability.		
70°C for use with free air heatsink	£1.60	£1.90
400°F for use with fan cooled heatsink	£1.60	£1.90

<b>T.M.D.</b>	Typically <.02% any power. 1kHz. 8 ohms
<b>T.I.D.</b>	Insignificant
<b>Slow Rate Limit</b>	: 20V/μs
<b>S/N Ratio</b>	: 110dB
<b>Freq. Response</b>	: 10Hz-35kHz. -3dB
<b>Stability</b>	: Unconditional
<b>Protection</b>	: Drives any load safely
<b>Sensitivity</b>	: 775mV (250mV or 100mV on Request)
<b>Size</b>	: 120 x 80 x 25mm

## CRIMSON ELEKTRIK

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Home prices include V.A.T. and postage. C.O.D. 90p extra. £100 limit. Export no problem. European prices include carriage, insurance and handling. Payment in Sterling by bank draft, P.O. International Giro or Money Order. Outside Europe, please write for specific quote by return. Send SAE or two International Reply Coupons for full literature. Favourable trade quantity price list on request. High quality pre-amp circuit 20p.

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NSB5LL30 1/2" Red stick of 4 **£4.32**, 5LT02 stick of 4 Green **£5.40**

**CLOCK CHIPS:** MK50253N Alarm. 12/24 hr. 4/6 dig. **£5.40**, MM5385 **£4.32**, MK50362N Calendar **£7.56**, 6 Digit Counter 50395/6/7 **£9.18**.

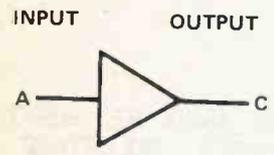
**MICRO PROCESSORS:** Z80 CPU **£21.60**, CTC & PIO **£15.12**, MK4096N 300ns 4K x 1 Dyn RAM **£6.75**, 1702A UV PROM 450ns **£10.80**.

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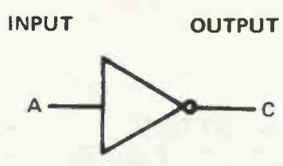
## TRUTH TABLES

### BUFFER

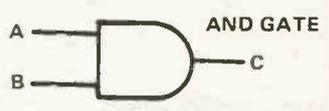


A	C
0	0
1	1

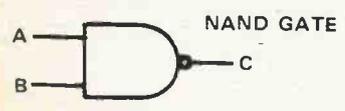
### INVERTER



A	C
0	1
1	0



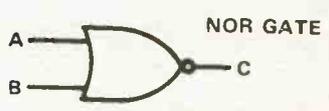
A	B	C
0	0	0
1	0	0
0	1	0
1	1	1



A	B	C
0	0	1
1	0	0
0	1	0
1	1	0



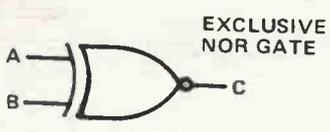
A	B	C
0	0	0
1	0	1
0	1	1
1	1	1



A	B	C
0	0	1
1	0	0
0	1	0
1	1	0



A	B	C
0	0	0
1	0	1
0	1	1
1	1	0

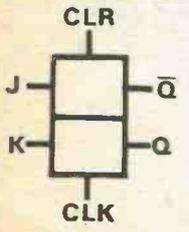


A	B	C
0	0	1
1	0	0
0	1	0
1	1	1

### LAWS OF BOOLEAN ALGEBRA

- $A + 0 = A$
- $A + 1 = 1$
- $A \cdot 0 = 0$
- $A + A = A$
- $A \cdot B + A \cdot C = A(B + C)$
- $A + B \cdot C = (A + B)(A + C)$
- $\overline{A \cdot B \cdot C} = \overline{A} + \overline{B} + \overline{C}$
- $\overline{\overline{A \cdot B \cdot C}} = A + B + C$
- $A \cdot A = A$
- $A + \overline{A} = 1$
- $A \cdot \overline{A} = 0$
- $A \cdot 1 = A$

### J.K. Flip Flop



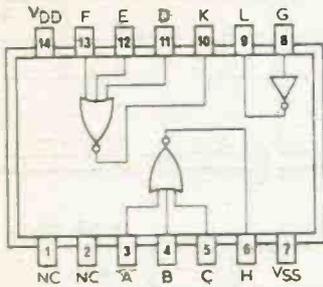
CLR	INPUTS			OUTPUTS	
	CLK	J	K	Q	$\overline{Q}$
0	X	X	X	0	1
1	High Pulse	0	0	$Q_0$	$\overline{Q_0}$
1	High Pulse	1	0	1	0
1	High Pulse	0	1	0	1
1	High Pulse	1	1	TOGGLE	

- High Pulse** - HIGH LEVEL PULSE, DATA IS TRANSFERRED ON FALLING EDGE OF PULSE.
- $Q_0$**  - THE LEVEL OF Q BEFORE INDICATED INPUT CONDITIONS WHERE ESTABLISHED.
- TOGGLE** - EACH OUTPUT CHANGES TO ITS COMPLEMENT ON EACH ACTIVE TRANSIENT (PULSE OF CLOCK).

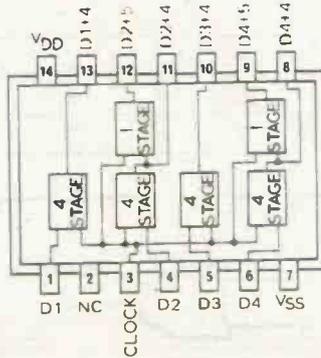
# LOGIC DATA

Every time we produce a circuit using CMOS chips we nag our readers to be careful how they treat them. Nonetheless we still receive countless telephone tales of woe to the common testimony of blown, melted, or otherwise deceased CMOS. So let us all kneel on our conductive-rubber mats facing magnetic north (wearing our earthed wrist-bands) and repeat the litany against static - Thou shalt not touch the pins, but ye shall leave the bloody thing in its black foam until the moment of the hot iron arrives. Got that? Good.

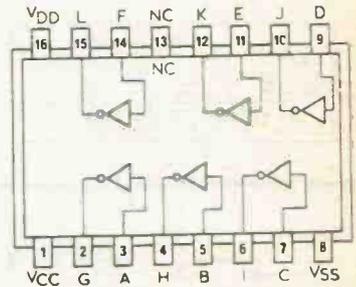
**CD 4000A DUAL 3-INPUT NOR GATE PLUS INVERTER**



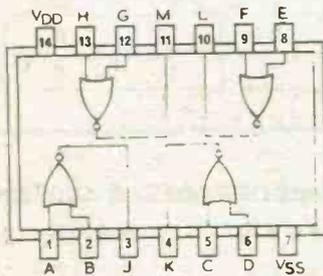
**CD4006A 18-STAGE STATIC SHIFT REGISTER**



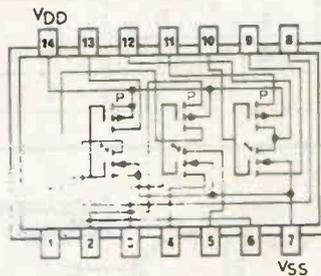
**CD4009A, 4049A HEX BUFFER CONVERTER - INVERTING**



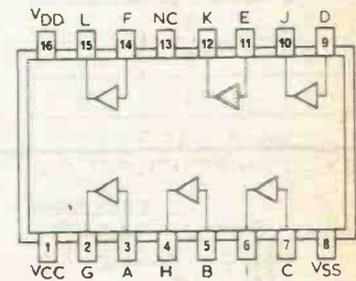
**CD4001A QUAD 2-INPUT NOR GATE**



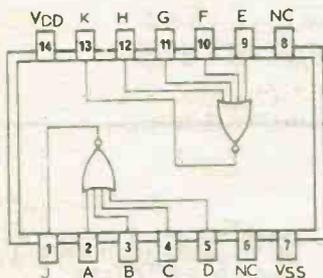
**CD4007A DUAL COMPLEMENTARY PAIR WITH INVERTER**



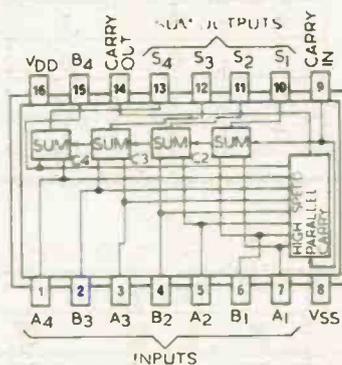
**CD4010A, 4050A HEX BUFFER CONVERTER - NON-INVERTING**



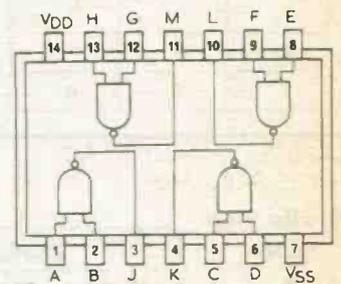
**CD4002A DUAL 4-INPUT NOR GATE**



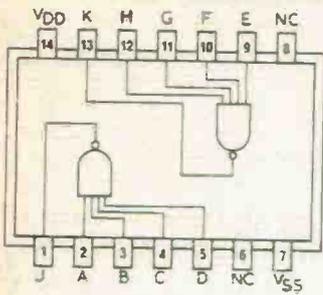
**CD4008A 4-BIT ADDER WITH PARALLEL CARRY**



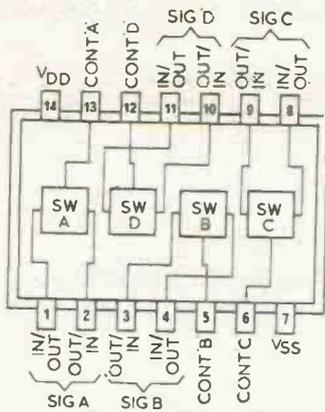
**CD4011A QUAD 2-INPUT NAND GATE**



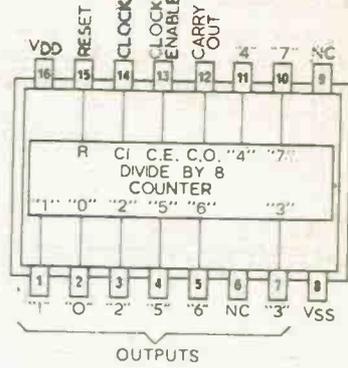
**CD4012A DUAL 4-INPUT NAND GATE**



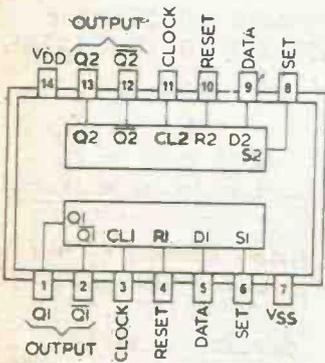
**CD4016A, 4066A QUAD SWITCH**



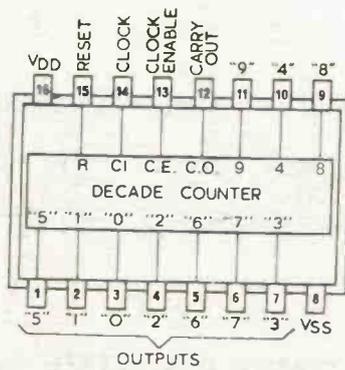
**CD4022A DIVIDE BY 8 COUNTER-DIVIDER**



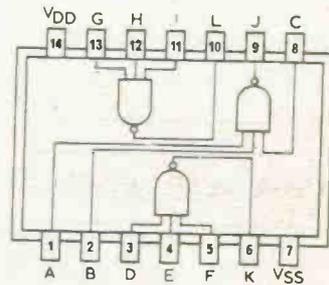
**CD4013A DUAL D-TYPE FLIP-FLOP WITH RESET**



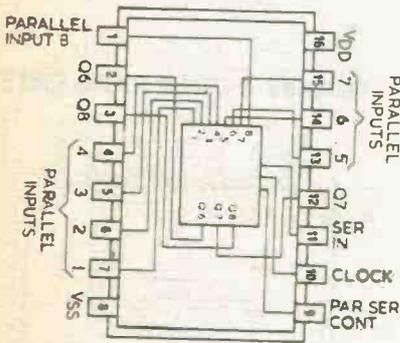
**CD4017A DECADE COUNTER**



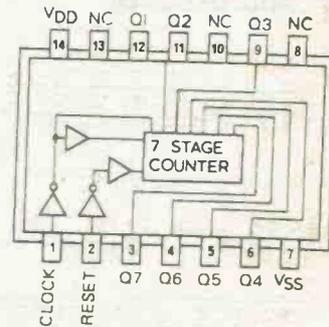
**CD4023A TRIPLE 3-INPUT NAND GATE**



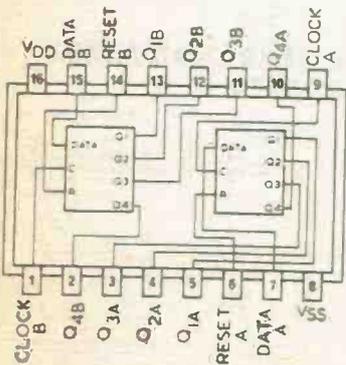
**CD4014A 8-STAGE STATIC SHIFT REGISTER**



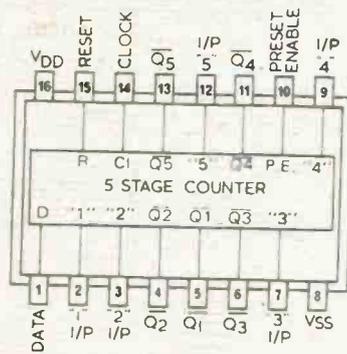
**CD4024A 7-STAGE BINARY COUNTER**



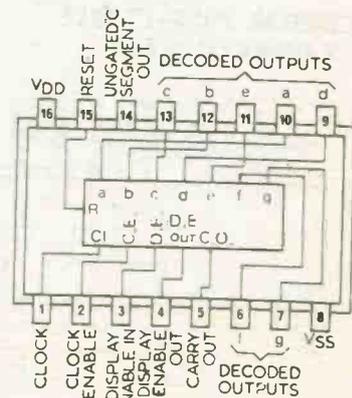
**CD4015A DUAL 4-STAGE SHIFT REGISTER**



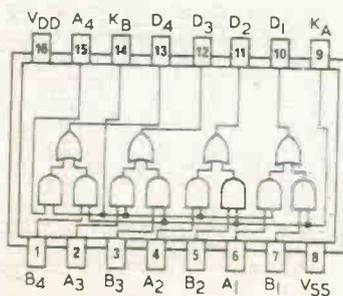
**CD4018A PRESETTABLE DIVIDE-BY-N COUNTER**



**CD4026A DECADE COUNTER-DIVIDER**

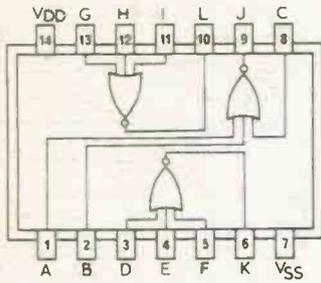


**CD4019A QUAD AND-OR SELECT GATE**

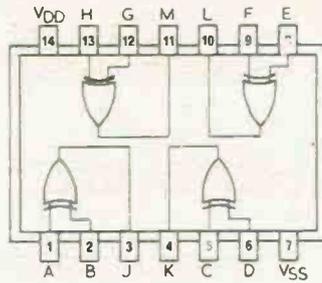


# LOGIC DATA

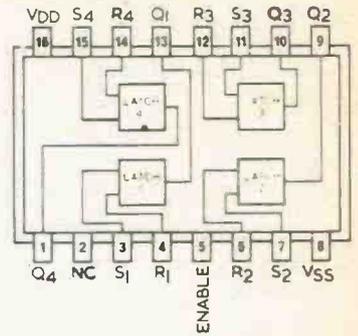
**CD4025A TRIPLE 3-INPUT NOR GATES**



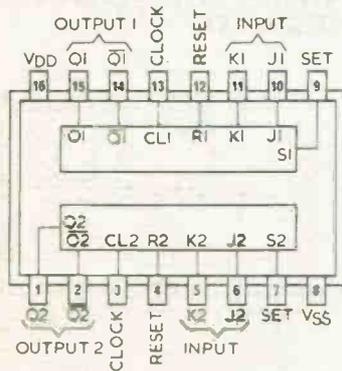
**CD4030A QUAD EX-OR GATES**



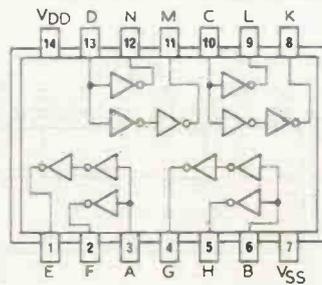
**CD4044A QUAD 3-STATE NAND R/S LATCH**



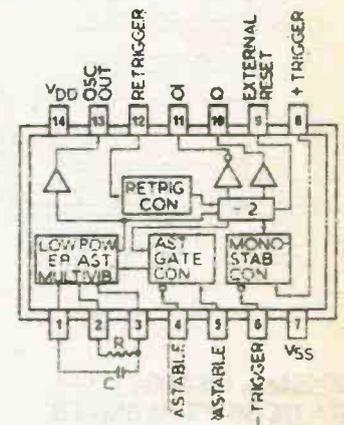
**CD4027A DUAL J-K FLIP-FLOP**



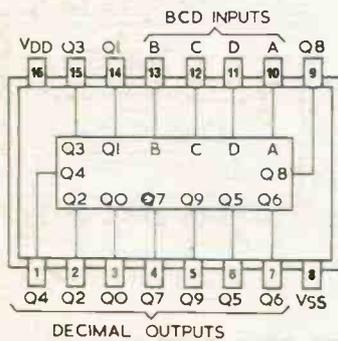
**CD4041A QUAD TRUE COMPLEMENT BUFFER**



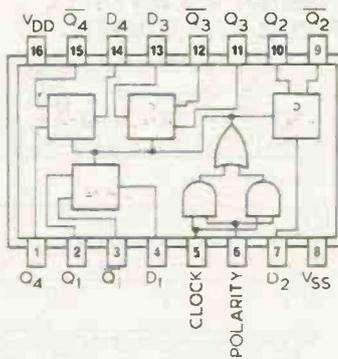
**CD4047A MONOSTABLE ASTABLE MULTIVIBRATOR**



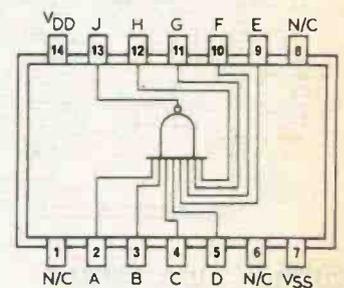
**CD4028A BCD TO DECIMAL DECODER**



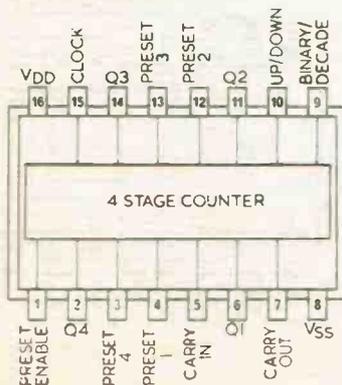
**CD4042A QUAD D-TYPE LATCH (CLOCKED)**



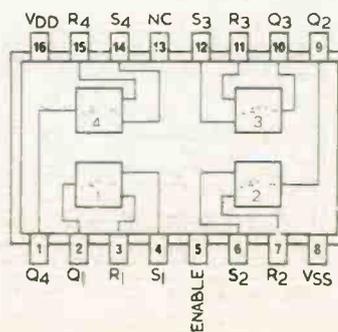
**CD4068B 8-INPUT NAND GATE**



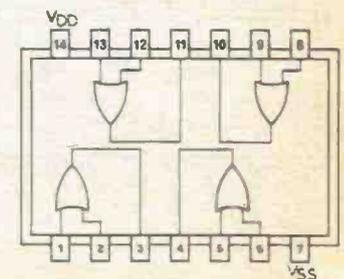
**CD4029A PRESETTABLE UP-DOWN COUNTER**



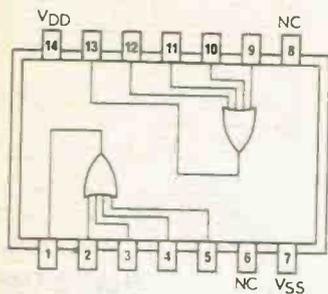
**CD4043A QUAD 3-STATE NOR R/S LATCH**



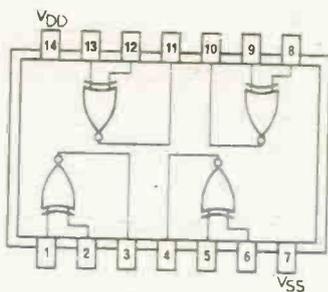
**CD4071B QUAD 2-INPUT OR GATE**



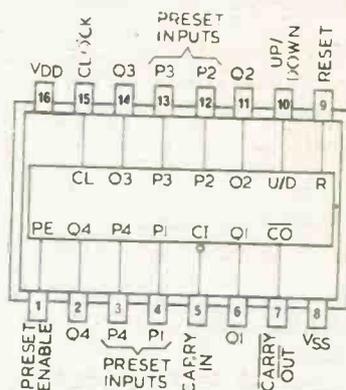
**CD4072B DUAL 4-INPUT OR GATE**



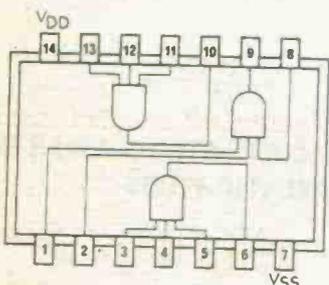
**CD4077B QUAD EX NOR GATES**



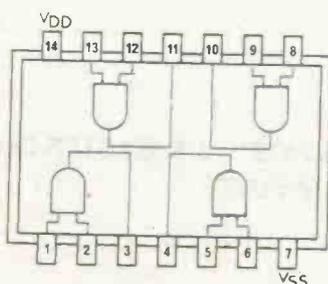
**CD4510B BCD UP-DOWN COUNTER**



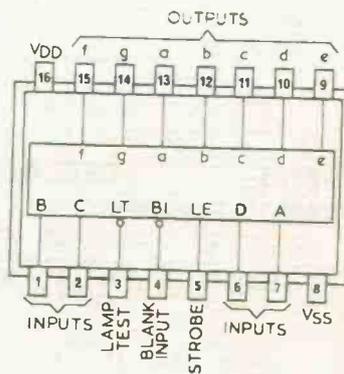
**CD4073B TRIPLE 3-INPUT AND GATE**



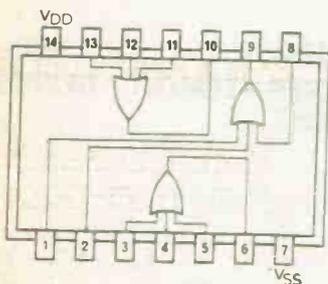
**CD4081B QUAD 2-INPUT AND GATE**



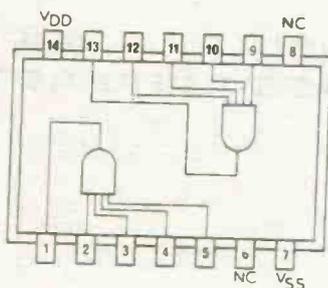
**CD4511B BCD-TO-7-SEGMENT DECODER/DRIVER**



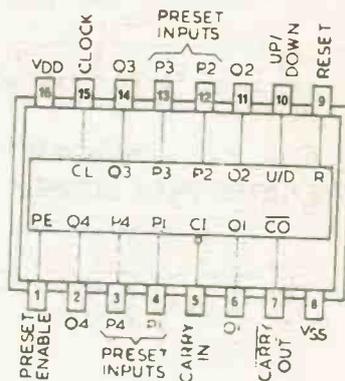
**CD4075B TRIPLE 3-INPUT OR GATE**



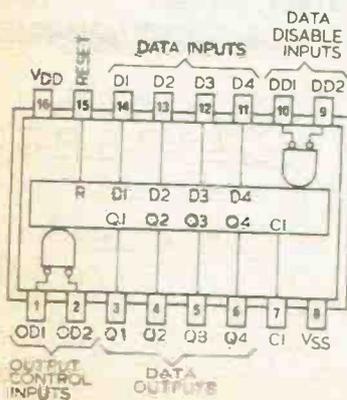
**CD4082B DUAL 4-INPUT AND GATE**



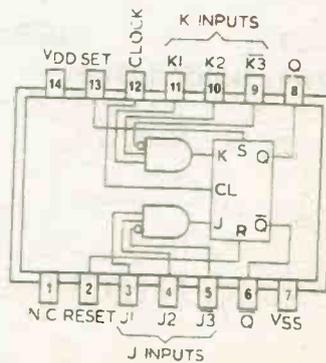
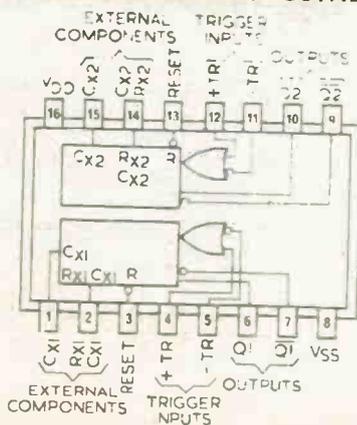
**CD4516B BINARY UP-DOWN COUNTER**



**CD4076B QUAD D TYPE FLIP-FLOP**

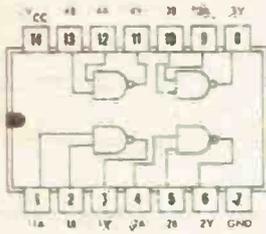


**CD4528 DUAL RETRIGGERABLE MONOSTABLE CD4096B GATED J-K FLIP-FLOP**

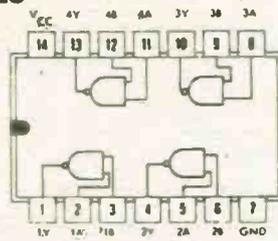


# LOGIC DATA

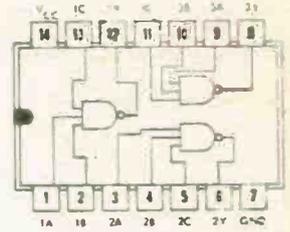
**SN7400 QUADRUPLE 2-INPUT POSITIVE NAND GATES**



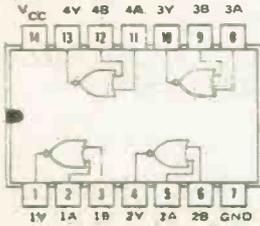
**SN7401 QUADRUPLE 2-INPUT OPEN-COLLECTOR NAND GATES**



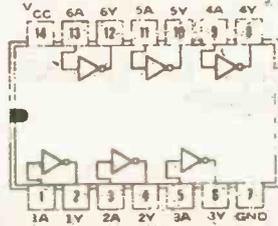
**SN7410 TRIPLE 3-INPUT POSITIVE NAND GATES**



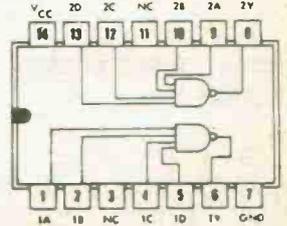
**SN7402 QUADRUPLE 2-INPUT POSITIVE NOR GATES**



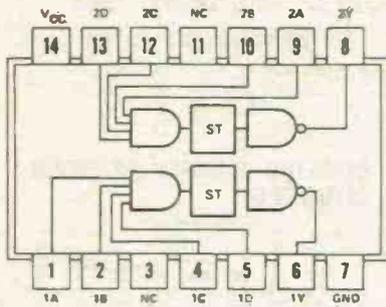
**SN7404 HEX INVERTERS**



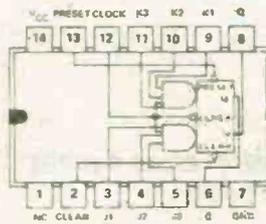
**SN7420 DUAL 4-INPUT POSITIVE NAND GATES**



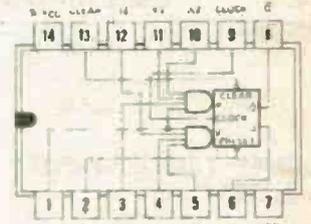
**SN7413 DUAL NAND SCHMITT TRIGGERS**



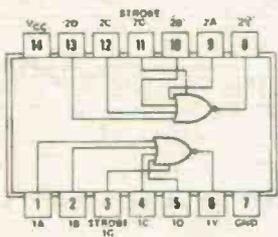
**SN7472 J-K MASTER SLAVE FLIP-FLOPS**



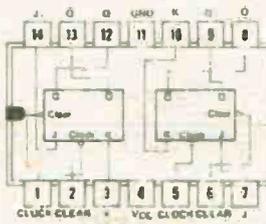
**SN74104 GATED J-K MASTER-SLAVE FLIP-FLOPS**



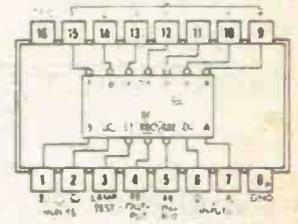
**SN7425 DUAL 4-INPUT NOR GATES WITH STROBE**



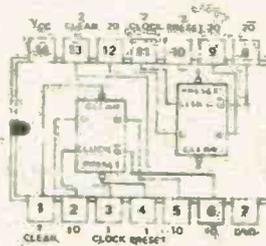
**SN7473 DUAL J-K MASTER-SLAVE FLIP-FLOPS**



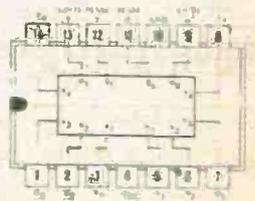
**SN7447A BCD-TO-SEVEN-SEGMENT DECODER/DRIVE**



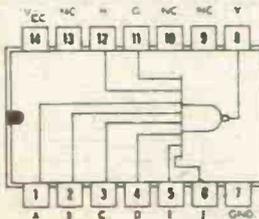
**SN7474 DUAL D-TYPE EDGE-TRIGGERED FLIP-FLOPS**



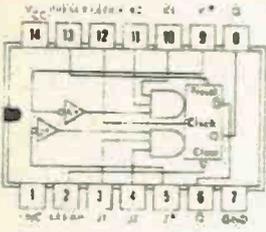
**SN7481, SN7484 16-BIT ACTIVE-ELEMENT MEMORIES**



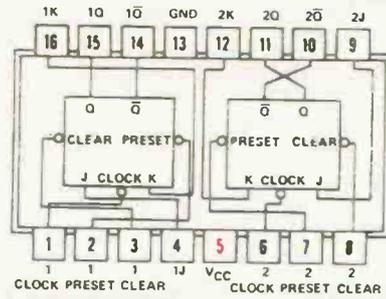
**SN7430 3-INPUT POSITIVE NAND GATES**



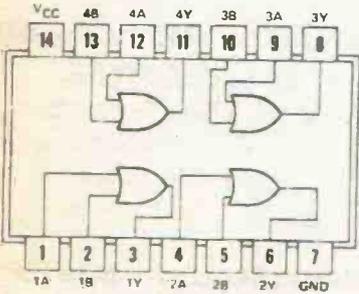
## SN7470 EDGE-TRIGGERED J-K FLIP-FLOPS



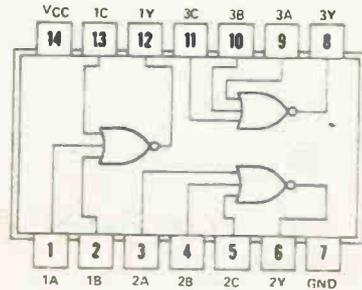
## SN7476 DUAL J-K MASTER-SLAVE FLIP-FLOPS WITH PRE-SET AND CLEAR



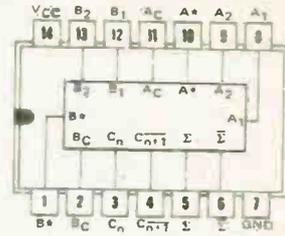
## SN7432 QUADRUPLE 2-INPUT POSITIVE-OR GATES



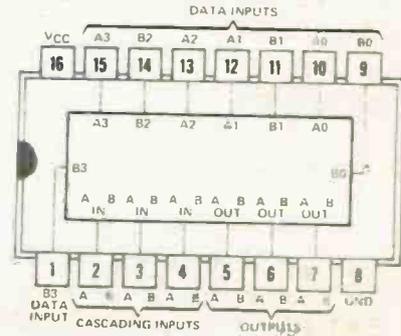
## SN7427 TRIPLE 3-INPUT POSITIVE-NOR GATES



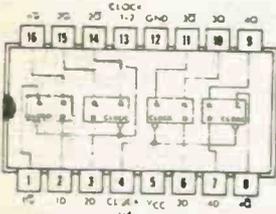
## SN7480 GATED FULL ADDERS



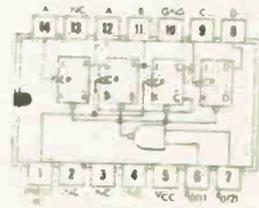
## SN7485 4-BIT MAGNITUDE COMPARATORS



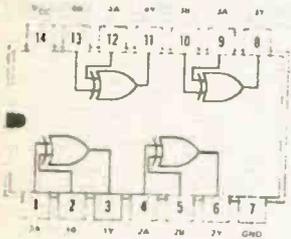
## SN7475 4-BIT BISTABLE LATCHES



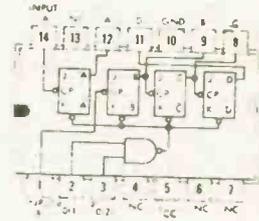
## SN7492 DIVIDE-BY-TWO AND DIVIDE-BY-SIX COUNTERS



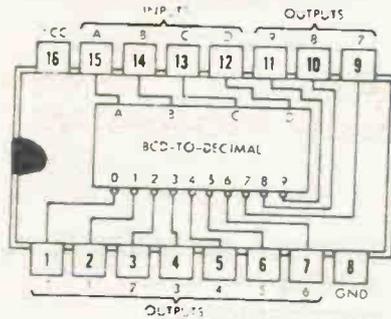
## SN7486 QUADRUPLE 2-INPUT EXCLUSIVE-OR GATES



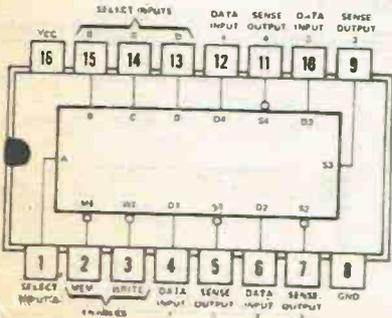
## SN7492 4-BIT BINARY COUNTERS



## SN7445, SN74145 BCD-TO-DECIMAL DECODER/DRIVERS

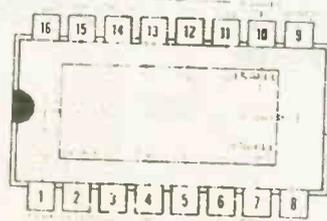


## SN7489 64-BIT READ/WRITE MEMORY

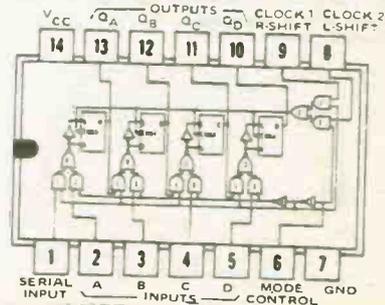


## SN74160 THRU SN74163 SYNCHRONOUS 4-BIT COUNTERS

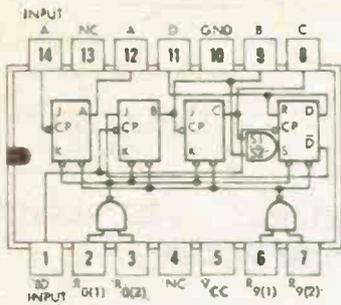
SN74160, SN74161 SYNCHRONOUS COUNTERS WITH DIRECT CLEAR  
SN74162, SN74163 FULLY SYNCHRONOUS COUNTERS



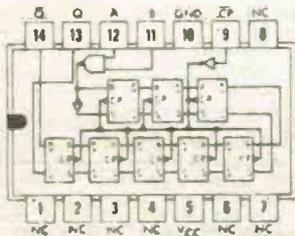
## SN7495A 4-BIT RIGHT-SHIFT LEFT-SHIFT REGISTERS



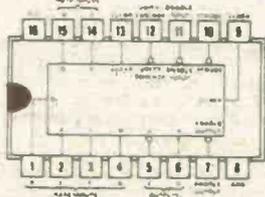
## SN7490 DECADE COUNTERS



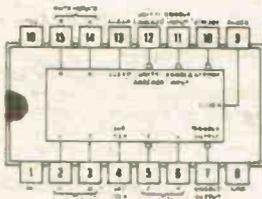
## SN7491A 8-BIT SHIFT REGISTERS



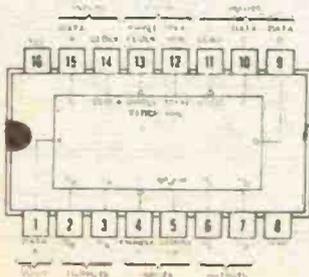
## SN7497 SYNCHRONOUS RATE MULTIPLIERS



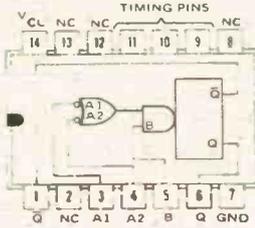
## SN74167



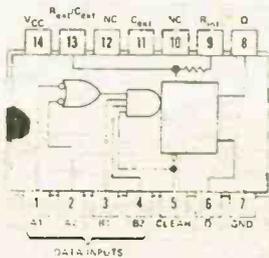
## SN74190, SN74191 SYNCHRONOUS UP/DOWN COUNTERS WITH MODE CONTROL



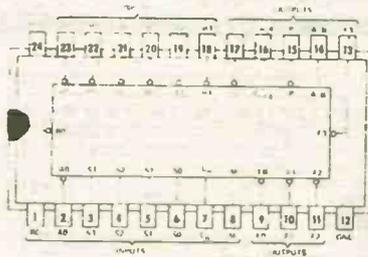
## SN74121 MONOSTABLE MULTIVIBRATORS



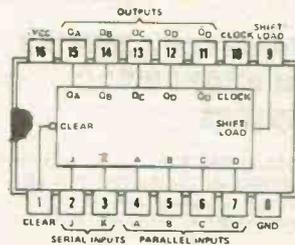
## SN74122 RETRIGGERABLE MULTIVIBRATORS WITH CLEAR



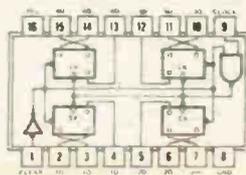
## SK74181 ARITHMETIC LOGIC UNITS



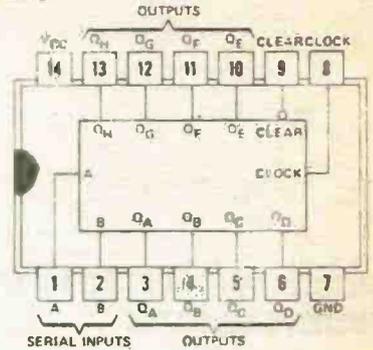
## SN74195A 4-BIT PARALLEL ACCESS SHIFT REGISTERS



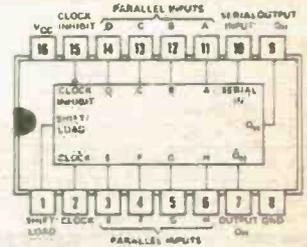
## SN74175 QUAD D-TYPE FLIP-FLOP



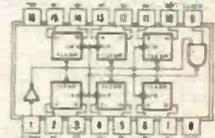
## SN74164 8-BIT PARALLEL-OUT SERIAL SHIFT REGISTERS



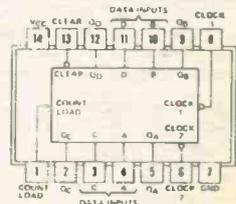
## SN74165 PARALLEL-LOAD 8-BIT SHIFT REGISTERS



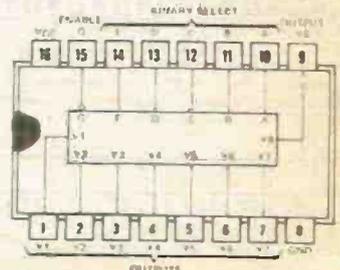
## SN74174 HEX D-TYPE FLIP-FLOP



## SN74176 & SN74177 PRESET DECADE & BINARY COUNTER



## SN74184, SN74185A CONVERTERS



## MPU Glossary

- ACCUMULATOR** The register where arithmetic or logic results are held. Most MPU instructions manipulate or test the accumulator contents.
- ACCESS TIME** Time taken for specific byte of storage to become available to processor.
- ACIA** Asynchronous Communication Inter-face Adapter. Inter-face between asynchronous peripheral and an MPU.
- ALU** Arithmetic and Logic Unit. The part of the MPU where arithmetic and logic functions are performed.
- ASCII** American Standard Code for Information Interchange. Binary code to represent alphanumeric, special and control characters.
- ASSEMBLER** Software which converts assembly language statements into machine code and checks for non valid statements or incomplete definitions.
- ASSEMBLY LANG** Means of representing programme statements in mnemonics and conveniently handling memory addressing by use of symbolic terms.
- ASYNCHRONOUS** Operations that initiate a new operation immediately upon completion of current one — not timed by system clock.
- BASIC** Beginner's All Purpose Symbolic Instruction Code. An easy to learn, widely used high level language.
- BAUD** Measure of speed of transmission line. Number of times a line changes state per second. Equal to bits per second if each line state represents logic 0 or 1.
- BAUDOT CODE** 5 bit code used to encode alphanumeric data.
- BCD** Binary Coded Decimal. Means of representing decimal numbers where each figure is replaced by a binary equivalent.
- BENCHMARK** A common task for the implementation of which programmes can be written for different MPUs in order to determine the efficiency of the different MPUs in the particular application.
- BINARY** The two base number system. The digits are 0 or 1. They are used inside a computer to represent the two states of an electric circuit.
- BIT** A single binary digit.
- BREAKPOINT** Program address at which execution will be halted to allow debugging or data entry.
- BUFFER** Circuit to provide isolation between sensitive parts of a system and the rest of that system.
- BUG** A program error that causes the program to malfunction.
- BUS** The interconnections in a system that carry parallel binary data. Several bus users are connected to the bus, but generally only one "sender" and one "receiver" are active at any one instant.
- BYTE** A group of bits — the most common byte size is eight bits.
- CLOCK** The basic timing for a MPU chip.
- COMPILER** Software which converts high level language statements into either assembly language statements, or into machine code.
- CPU** Central processor unit. The part of a system which performs calculation and data manipulation functions.
- CROM** Control Read Only Memory.
- CRT** Cathode Ray Tube. Often taken to mean complete output device.
- CUTS** Computer Users Tape System. Definition of system for storing data on cassette tape as series of tones to represent binary 1's and 0's.
- DEBUG** The process of checking and correcting any program errors either in writing or in actual function.
- DIRECT ADDRESSING** An addressing mode where the address of the operand is contained in the instruction.
- DMA** Direct Memory Access.
- DUPLEX** Transfer of data in two directions simultaneously.
- ENVIRONMENT** The conditions of all registers, flags, etc. at any instant in program.
- EPROM** Electrically Programmable Read Only Memory. Memory that may be erased (usually by ultra violet light) and reprogrammed electrically.
- EXECUTE** To perform a sequence of program steps.
- EXECUTION TIME** The time taken to perform an instruction in terms of clock cycles.
- FIRMWARE** Instructions or data permanently stored in ROM.
- FLAG** A flip flop that may be set or reset under software control.
- FLIP-FLOP** Two state device that changes state when clocked.
- FLOPPY (DISK)** Mass storage which makes use of flexible disks made of a material similar to magnetic tape.
- FLOW CHART** A diagram representing the logic of a computer program.
- GLITCH** Noise pulse.
- HALF DUPLEX** Data transfer in two directions but only one way at a time.
- HAND SHAKE** System of data transfer between CPU and peripheral whereby CPU "asks" peripheral if it will accept data and only transfers data if "answer" is yes.
- HARD COPY** System output that is printed on paper.
- HARDWARE** All the electronic and mechanical components making up a system.
- HARD WIRE** Circuits that are comprised of logic gates wired together, the wiring pattern determining the overall logic operation.
- HASH** Noisy signal.
- HEXADECIMAL** The base 16 number system. Character set is decimal 0 to 9 and letters A to F.
- HIGH LEVEL LANGUAGE** Computer language that is easy to use but which requires compiling into machine code before it can be used by an MPU.
- HIGHWAY** As BUS.
- IMMEDIATE ADDRESSING** Addressing mode which uses part of the instruction itself as the operand data.
- INDIRECT ADDRESSING** A form of indirect addressing which uses an Index Register to hold the address of the operand.
- INDIRECT ADDRESSING** Addressing mode where the address of the location where the address of the operand may be found is contained in the instruction.
- INITIALISE** Set up all registers, flag, etc. to defined conditions.
- INSTRUCTION** Bit pattern which must be supplied to an MPU to cause it to perform a particular function.
- INSTRUCTION REGISTER** MPU register which is used to hold instructions fetched from memory.
- INSTRUCTION SET** The repertoire of instructions that a given MPU can perform.
- INTERFACE** Circuit which connects different parts of system together and performs any processing of signals in order to make transfer possible (ie. serial — parallel conversion).
- INTERPRETER** An interpreter is a software routine which accepts and executes a high level language program, but unlike a compiler does not produce intermediate machine code listing but converts each instruction as received.
- INTERRUPT** A signal to the MPU which will cause it to change from its present task to another.
- I/O** Input/Output.
- K** Abbreviation for  $2^{10} = 1028$ .
- KANSAS CITY (Format)** Definition of a CUTS based cassette interface system.
- LANGUAGE** A systematic means of communicating with an MPU.
- LATCH** Retains previous input state until overwritten.
- LIFO** Last In First Out. Used to describe data stack.
- LOOPING** Program technique where one section of program (the loop) is performed many times over.
- MACHINE LANG** The lowest level of program. The only language an MPU can understand without interpreter.
- MASK** Bit pattern used in conjunction with a logic operation to select a particular bit or bits from machine word.
- MEMORY** The part of a system which stores data (working data or instruction object code).
- MEMORY MAP** Chart showing the memory allocation of a system.
- MEMORY MAPPED I/O** A technique of implementing I/O facilities by addressing I/O ports as if they were memory locations.
- MICRO CYCLE** Single program step in an MPUs Micro program. The smallest level of machine program step.
- MICRO PROCESSOR** A CPU implemented by use of large scale integrated circuits. Frequently implemented on a single chip.
- MICRO PROGRAM** Program inside MPU which controls the MPU chip during its basic fetch/execute sequence.
- MNEMONIC** A word or phrase which stands for another (longer) phrase and is easier to remember.
- MODEM** Modulator/demodulator used to send and receive serial data over an audio link.
- NON VOLATIVE** Memory which will retain data content after power supply is removed, e.g. ROM.
- OBJECT CODE** To bit patterns that are presented to the MPU as instructions and data.
- O/C** Open Collector. Means of tying together O/P's from different devices on the same bus.
- OCTAL** Base 8 number system. Character set is decimal 0-8.
- OP CODE** Operation Code. A bit pattern which specifies a machine operation in the CPU.
- OPERAND** Data used by machine operations.
- PARALLEL** Transfer of two or more bits at the same time.
- PARITY** Check bit added to data, can be odd or even parity. In odd parity sum of data 1's + parity bit is odd.
- PERIPHERAL** Equipment for inputting to or outputting from the system (e.g., teletype, VDU, etc.).
- PIA** Peripheral Interface Adapter.
- POP** Operation of removing data word from LIFO stack.
- PORT** A terminal which the MPU uses to communicate with the outside world.
- PROGRAMS** Set of MPU instructions which instruct the MPU to carry out a particular task.
- PROGRAM COUNTER** Register which holds the address of next instruction (or data word) of the program being executed.
- PROM** Programmable read only memory. Proms are special form of ROM, which can be individually programmed by user.
- PUSH** Operation of putting data to LIFO stack.
- RAM** Random Access Memory. Read write memory. Data may be written to or read from any location in this type of memory.
- REGISTER** General purpose MPU storage location that will hold one MPU word.
- RELATIVE ADDRESSING** Mode of addressing whereby address of operand is formed by combining current program count with a displacement value which is part of the instruction.
- ROM** Read Only Memory. Memory device which has its data content established as part of manufacture and cannot be changed.
- SCRATCH PAD** Memory that has short access time and is used by system for short term data storage.
- SERIAL** Transfer of data one bit at a time.
- SIMPLEX** Data transmission in one direction only.
- SOFTWARE** Programs stored on any media.
- SOURCE CODE** The list of statements that make up a program.
- STACK** A last in first out store made up of registers or memory locations used for stack.
- STATUS REGISTER** Register that is used to store the condition of the accumulator after an instruction has been performed (e.g., Acc = 0).
- SUB ROUTINE** A sequence of instructions which perform an often required function, which can be called from any point in the main program.
- SYNTAX** The grammar of a programming language.
- TRAP (Vector)** Pre-defined location in memory which the processor will read as a result of particular condition or operation.
- TRI STATE** Description of logic devices whose outputs may be disabled by placing them in a high impedance state.
- TTY** Teletype.
- TWO'S COMPLEMENT ARITHMETIC** System of performing signed arithmetic with binary numbers.
- UART** Universal Asynchronous Receiver Transmitter.
- VDU** Video Display Unit.
- VECTOR** Memory address provided to the processor to direct it to a new area in memory.
- VOLATILE** Memory devices that will lose data content if power supply removed (ie. RAM).
- WORD** Parallel collection of binary digits much as byte.

# MISCELLANEOUS DATA

## BIPOLAR TRANSISTORS

TYPE	CASE	POL MAT	V <sub>ce</sub>	V <sub>cb</sub>	IC mA	V <sub>ces</sub> @ IC mA	H <sub>fe</sub> @ IC mA	F <sub>T</sub> MHz @ IC mA	P <sub>tot</sub> mW	Use	Comparable Types			
AC107	GT3	NG	15	15	10		30-160	3	2	3	80	Low Noise Audio	AC125-2N406	
AC125	TO-1	PG	12	32	100		100	2	1.3	10	216	Audio Driver	2N406	
AC126	TO-1	PG	12	32	100		140	2	1.7	10	216	Audio Driver	2N406	
AC127	TO-1	NG	12	32	500		105	50	1.5	10	340	Audio O/P	AC187	
AC128	TO-1	PG	16	32	1000	.6	60-175	300	1	10	260	Audio O/P	AC188	
AC132	TO-1	PG	12	32	200	.35	115	50	1.3	10	216	Audio O/P	AC188	
AC187	TO-1	NG	15	25	2000	.8	100-500	300	1	10	800	Audio O/P	AC127	
AC188	TO-1	PG	15	25	2000	.6	100-500	300	1	10	220	Audio O/P	AC128	
AD149	TO-3	PG	30	50	3500	.7	30-100	1A	.3	500	32W	G.P. O/P	OC26, AU106	
AD161	PT1	NG	20	32	3000	.6	80-320	500	.02	300	4W	Audio Amp.	AD165, 2N1218, 2N1292	
AD162	PT1	PG	20	32	3000	.4	80-320	500	.015	300	6W	Audio Amp.	AD143, AD152, AD427	
AF114	TO-7	PG	15	32	10		150	1	75	1	75	H.F. Amp.	AF144, AF194, 2N3127	
AF115	TO-7	PG	15	32	10		150	1	75	1	75	H.F. Amp.	AF146, AF185, 2N2273	
AF116	TO-7	PG	15	32	10		150	1	75	1	75	H.F. Amp.	AF135, AF136, 2N3127	
AF117	TO-7	PG	15	32	10		150	1	75	1	75	H.F. Amp.	AF136, AF197, 2N5354	
AF118	TO-7	PG	20	70	30	5	35	10	175	10	375	V.H.F. Amp.	BFW20	
AS215	TO-3	PG	60	100	10A	.4	20-55	1A	.2	1A	30W	H.C. Sw.	OC28	
AS216	TO-3	PG	32	60	10A	.4	45-130	1A	.25	1A	30W	H.C. Sw.	OC29, AD138, AD723	
AS217	TO-3	PG	32	60	10A	.4	25-75	1A	.22	1A	30W	H.C. Sw.	OC35, AD424	
AS218	TO-3	PG	32	100	10A	.4	30-110	1A	.22	1A	30W	H.C. Sw.	OC36	
BC107	TO-18	NS	45	50	100	.2	110-450	2	300	10	300	S.S. Amp.	BC207, BC147, BC182	
BC108	TO-18	NS	20	30	100	.2	110-800	2	300	10	300	S.S. Amp.	BC208, BC148, BC183	
BC109	TO-18	NS	20	30	100	.2	200-800	2	300	10	300	Low Noise S.S. Amp.	BC209, BC149, BC184	
BC109C	TO-18	NS	20	30	100	.2	420-800	2	300	10	300	Low Noise High Gain	BC209C, BC184C, BC149C	
BC157	SOT-25	PS	45	50	100	.25	100	75-260	2	150	10	300	S.S. Amp.	BC177, BC307, BC212
BC158	SOT-25	PS	25	30	100	.25	100	75-500	2	150	10	300	S.S. Amp.	BC178, BC308, BC213
BC159	SOT-25	PS	20	25	100	.25	100	125-500	2	150	10	300	S.S. Amp.	BC179, BC309, BC214
BC177	TO-18	PS	45	50	100	.25	100	75-260	2	150	10	300	S.S. Amp.	BC157, BC307, BC212
BC178	TO-18	PS	25	30	100	.25	100	75-500	2	150	10	300	S.S. Amp.	BC158, BC308, BC213
BC179	TO-18	PS	20	25	100	.25	100	125-500	2	150	10	300	S.S. Amp.	BC159, BC309, BC214
BC182(L)	SOT-30	NS	50	10	200	.25	10	100-480	2	150	10	300	S.S. Amp.	BC107, BC207, BC147
BC183(L)	(TO-92/74)	NS	30	45	200	.25	10	100-850	2	150	10	300	S.S. Amp.	BC108, BC208, BC148
BC184(L)	(TO-92/74)	NS	30	45	200	.25	10	250-850	2	150	10	300	Low Noise, High Gain	BC109, BC209, BC149
BC186	TO-18	PS	25	40	200	.5	50	40-200	2	50	50	300	G.P. Amp.	BC213, BC177, BC158
BC207	TO-106	NS	45	50	200	.25	10	110-220	2	150	10	300	S.S. Amp.	BC107, BC182, BC147
BC208	TO-106	NS	20	25	200	.25	10	110-800	2	150	10	300	S.S. Amp.	BC108, BC183, BC148
BC209	TO-106	NS	20	25	200	.25	10	200-800	2	150	10	300	Low Noise, High Gain	BC109, BC184, BC149
BC212(L)	SOT-30	PS	50	60	200	.25	10	60-300	2	200	10	300	S.S. Amp.	BC307, BC157, BC177
BC213(L)	(TO-92/74)	PS	30	45	200	.25	10	80-400	2	200	10	300	S.S. Amp.	BC308, BC158, BC178
BC214(L)	(TO-92/74)	PS	30	45	200	.25	10	80-400	2	200	10	300	S.S. Amp.	
BC327	TO-92	PS	45	-	1000	0.7	500	100-600	100	100	10	800	O/P	2N3638
BC337	TO-92	NS	45	-	1000	0.7	500	100-600	100	200	10	800	O/P	2N3642
BC547	SOT-30	NS	45	50	100	.6	100	110-800	2	300	10	500	S.S. Amp.	BC107, BC207, BC147
BC548	SOT-30	NS	30	30	100	.6	100	110-800	2	300	10	500	S.S. Amp.	BC108, BC208, BC148
BC549	SOT-30	NS	30	30	100	.6	100	200-800	2	300	10	500	Low Noise S. Sig.	BC109, BC209, BC149
BC549C	SOT-30	NS	30	30	100	.6	100	420-800	2	300	10	500	Low Noise, High Gain	BC109C, BC149C
BC635	TO-92(74)	NS	45	45	1A	.5	500	40-250	150	130	500	1W	Audio O/P	BC639
BC636	TO-92(74)	PS	45	45	1A	.5	500	40-250	150	130	500	1W	Audio O/P	BC640
BC639	TO-92(74)	NS	80	100	1A	.5	500	40-160	150	130	1W	1W	Audio O/P	MU9610, TT801
BC640	TO-92(74)	PS	80	100	1A	.5	500	40-160	150	130	1W	1W	Audio O/P	MU9660, TT800
BCV70	TO-18	PS	40	50	200	5	50	50	10	250	50	350	G.P.	BC212
BCV71	TO-18	PS	45	45	200	5	50	100-600	10	200	50	350	G.P.	BC212
BCV72	TO-18	PS	25	25	200	5	50	50	10	200	50	350	G.P.	BC213
BD137	TO-126	NS	60	60	1A	5	500	40-160	150	250	500	8W	G.P. O/P	BD139
BD138	TO-126	PS	60	60	1A	5	500	40-160	150	75	500	8W	G.P. O/P	BD140
BD139	TO-126	NS	60	100	1A	5	500	40-160	150	250	500	8W	G.P. O/P	40409
BD140	TO-126	PS	80	100	1A	5	500	40-160	150	75	500	8W	G.P. O/P	40410
BD262	TO-126	PS	60	60	4A	2.5	1.5A	750	1.5A	7	1.5A	36W	High Gain Darl. O/P	BD266
BD263	TO-126	PS	60	60	4A	2.5	1.5A	750	1.5A	7	1.5A	36W	High Gain Darl. O/P	BD267
BD266A	TO-120	PS	80	80	8A	2	3A	750	3A	7		60W	High Gain Darl. O/P	
BD267A	TO-220	NS	80	100	8A	2	3A	750	3A	7		60W	High Gain Darl. O/P	
BDX64A	TO-3	PS	80	80	12A	2.5	5A	1000	5A	7	5A	117W	Darl. O/P	
BDX65A	TO-3	NS	80	80	12A	2.5	5A	1000	5A	7	5A	117W	Darl. O/P	
BDY20	TO-3	NS	60	100	15A	1.1	4A	20-70	4A	1	4A	115	Power O/P	2N3055
BF115	TO-72(28)	NS	30	50	30			45-165	1	230	1	145	V.H.F. Amp.	
BF167	TO-72(28)	NS	30	40	25			26	4	350	4	130	T.V. I.F. Amp.	
BF173	TO-72(28)	NS	25	40	25			37	7	550	5	230	T.V. I.F. Amp.	
BF177	TO-39	NS	60	100	50			20	15	120	10	795	T.V. Video Amp.	BF336
BF178	TO-39	NS	115	185	50			20	30	120	10	1.7W	T.V. Video Amp.	BF336
BF179	TO-39	NS	115	250	50			20	20	120	10	1.7W	T.V. Video Amp.	BF338
BF180	TO-72(25)	NS	20	30	20			13	2	675	2	150	U.H.F. Amp.	BF200
BF184	TO-72(28)	NS	20	30	30			75-750	1	300	1	145	H.F. Amp.	
BF185	TO-72(28)	NS	20	30	30			34-140	1	220	1	145	H.F. Amp.	BF195
BF194	SOT-25(1)	NS	20	30	30			65-220	1	260	1	250	H.F. Amp.	
BF195	SOT-25(1)	NS	20	30	30			35-125	1	200	1	750	H.F. Amp.	BF185
BF200	TO-72(25)	NS	20	30	20			15	3	650	3	150	V.H.F. Amp.	BF180
BF336	TO-39	NS	180	185	100			20-60	30	130	3W	Video Amp.		
BF337	TO-39	NS	200	300	100			20-60	30	130	3W	Video Amp.		
BF338	TO-39	NS	225	250	100			20-60	30	130	3W	Video Amp.		
BFY50	TO-39	NS	35	80	1A	.2	150	30	150	60	50	2.86W	G.P.	
BFY51	TO-39	NS	30	60	1A	.35	150	40	150	50	50	2.86W	G.P.	
BFY52	TO-39	NS	20	40	1A	.35	150	60	150	50	50	2.86W	G.P.	
MJ2501	TQ 3	PS	80	80	10A	2	5A	1000	5A			150W	Darl. O/P	
MJ2955	TO-3	PS	60	70	15A	1.1	4A	20-70	4A	4	500	115V	High Power O/P	2N4908, 2N4909, 2N5811
MJ3001	TO-3	NS	80	80	10A	2	5A	1000	5A			150W	Darl. O/P	

# MISCELLANEOUS DATA

TYPE	CASE	POL MAT	V <sub>ce</sub>	V <sub>cb</sub>	IC mA	V <sub>ces</sub> @ IC mA	Hfe @ IC mA	Ft MHz @ IC mA	Ptot mW	Use	Comparable Types	
MJE2955	90-05	PS	50	70	10A	1.1 4A	20-70 4A	2	500	90W	High Power O/P	TIP2955
MJE3055	90-05	NS	60	70	10A	1.1 4A	20-70 4A	2	500	90W	High Power O/P	TIP3055
MU9610	152	NS	30	40	2A	0.4 1.5A	80-400 350	70	250	1W	O/P	TT801
MU9611	152-01	NS	30	40	2A	0.4 1.5A	80-400 350	70	250	1W	O/P	TT801
MU9660	152	PS	30	40	2A	0.4 1.5A	80-400 350	70	250	1W	O/P	TT800
MU9661	152-01	PS	30	40	2A	0.4 1.5A	80-400 350	70	250	1W	O/P	TT800
NSD106	TO-202(35)	NS	100	140		2-9 100	50-150 100	80	50		Driver - O/P	
NSD206	TO-202(35)	PS	100	100		2-1 100	50-150 100	150	50		Driver - D/P	
OC26	TO-3	PG	30	50	3.5A	.7 3A	30-100 1A	.3	500	32W	G.P. O/P	AD149
OC28	TO-3	PG	60	100	10A	.4 10A	20-55 1A	.2	1A	30W	H. C. Switch	AS215
OC44N	TO-1	PG	5	15	10		45-225 1	7.5	1	85	R. F. Amp	AF 125, AF 135, AF 172
OC46	GT-3	PG	5	15	10		25-125 1	3	3	85	R. F. Amp	AF 132, AF 185, AF 196
OC70	GT-3	PG	10	30	50		30 5	5		125	G. P. Amp.	AC121, AC126, 2N1190
OC71	GT-3	PG	10	30	50		30-75 3	6		125	G. P. Amp.	AC126, 2N2429
OC72	GT-6	PG	16	32	250		45-120 10	35		165	Audio O/P	AC122, AC125, AC162
OC74N	TO-1	PG	10	20	300	.6 300	60-150 50	.1		550	Audio O/P	AC125, AC180, AC192
OC75	GT-3	PG	10	30	50		60-130 3	.1		125	G. P. Amp	AC173, AC192
TIP318	TOP-66	NS	80	80	3A	1.2 3A	20 500	3	500	40W	Power Amp - Sw.	
TIP328	TOP-66	PS	80	80	3A	1.2 3A	20 500	3	500	40W	Power Amp - Sw.	
TIP2955	TOP-3	PS	70	100	15A	1.1 4A	20 4A	.8		90W	Power Amp - Sw.	MJE2955
TIP3055	TOP-3	NS	70	100	15A	1.1 4A	20 4A	.8		90W	Power Amp - Sw.	MJE3055
TT797	TO-39	PS	60	60	1A		40-120 100	100		800	Audio Driver - Sw.	TT800, NSD206
TT798	TO-39	NS	60	60	1A		40-120 150	60		800	Audio Driver - Sw.	TT801, NSD106
TT800	TO-39	PS	60	60	1A		40-120 100	100		800	Audio Driver - Sw.	TT797, NSD206
TT801	TO-39	NS	60	60	1A		40-120 150	60		800	Audio Driver - Sw.	TT798, NSD106
2N301	TO-3	PG	32	40	3A		50 1A	2	1A	11W	Audio O/P	AT1138, OC26
2N306A	TO 18	NS	15	25	200		20 10	200		300	High Speed Sw	
2N2926	TO 92(74)	NS	25	25	100		150 2	100		200	G. P.	BC108 etc
2N3053	TO-39	NS	40	60	700	1.4 150	50-250 150	100	50	2.85W	G. P. Switch	BD137
2N3054	TO-45	NS	55	90	4A	1 200	25 500	8	20C	25W	Audio O/P	TIP318
2N3055	TO-3	NS	60	90	15A	1.1 4A	20 4A	.8	1A	115W	O/P - Sw	BDV20
2N3583	TO-106	NS	12	30	50		20-200 8	600	8	20C	R. F. - I. F. Amp.	BF173
2N3584	TO-106	NS	15	30	100	.3 20	20-600 15	400	15	20C	R. F. - I. F. Amp	BF167
2N3585	TO-106	NS	25	30	50	.35 1	150-600 1	40	1	20C	Low Level Amp	BC106, BC208
2N3586	TO-105	NS	30	40	200	1 100	150-600 10	40	30	100	G. P. Amp. & Sw.	BC183
2N3587	TO-105	NS	40	80	500	.25 150	40-120 1	60	50	300	G. P. Amp. & Sw	BC337
2N3588	TO-105	NS	60	80	500	.25 150	40-120 1	60	50	300	G. P. Amp. & Sw.	
2N3589	TO-105	NS	40	80	500	.25 150	100-300 1	60	50	300	G. P. Amp. & Sw.	
2N3638	TO-105	PS	25	25	500	.25 50	30 50	100	50	300	G. P. Amp. & Sw.	
2N3638A	TO-105	PS	25	25	500	.25 50	100 50	150	50	300	G. P. Amp. & Sw.	BC327
2N3640	TO-106	PS	12	12	80	.2 10	30-120 10	300	10	200	Saturated Switch	BC558
2N3641	TO-105	NS	30	60	500	.22 150	40-120	250	50	350	G. P. Amp. & Sw	BC337
2N3642	TO-105	NS	45	60	500	.22 150	40-120 150	250	50	350	G. P. Amp. & Sw.	BC337
2N3643	TO-105	NS	30	60	500	.22 150	100-300 150	250	50	350	G. P. Amp. & Sw.	BC337
2N3644	TO-105	PS	45	45	500	1 300	115-300 50	200	20	300	G. P. Amp. & Sw.	BC327
2N3945	TO-105	PS	60	60	500	1 300	115-300 500	200	20	300	G. P. Amp. & Sw	
2N3702	TO 92(74)	PS	25	40	200	.25 50	60-300 50	100	50	360	G. P. Amp. & Sw	BC213
2N3904	TO 92	NS	40	60	200		100-300 1mA			310	Low Level Amp.	BC167A, BF 194
2N4250	TO-106	PS	40	40	100	.25 10	250-400 1	50		200	Low Level Amp.	BC559
2N4258	TO-106	PS	12	12	50	.5 50	30-120 10	700	10	200	Saturated Sw	
2N4282	TO-92	NS	15	30	50	.6 10	20 3	600	4	200	Saturated Sw.	
2N4403	TO-92	PS	40	40	600		100-300 10			310	G.P.	
2N5580	MT-71C	NS	18	36	600		5 100	175→	3W	15W	H.F. Mobile R. F	BC307A 2N2904
2N5590	MT-72C	NS	18	36	2A		5 250	175→	10W	30W	H.F. Mobile R. F	
2N5591	MT-73E	NS	18	36	4A		5 500	175→	25W	70W	H.F. Mobile R. F	
2N5871	TO-3	PS	60	60	7A	1 4A	20-100 2.5A	4	250	100W	Power Transistor	2N5872, 2N4908, MJ2955
40260	TO-66	NS	50	50	4A	1.5 1.5A	25 100	1		29W	Power Transistor	2N9054
40408	TO-66	NS	80		700	1.4 150	40-200 200	100		1W	Power Transistor	BC639
40409	TO-39(H)	NS	80		700	1.4 150	50-250 150	100		3W	Power Transistor	BD139
40410	TO-39(H)	PS	80		700	1.4 150	50-250 150	100		3W	Power Transistor	BD140

FETS	CASE	BV <sub>ces</sub> V @ IC (mA)	V <sub>GS</sub> (OFF) Min Max @ V <sub>DS</sub>		I <sub>C</sub> (mA)	I <sub>DSS</sub> (mA) Min Max @ V <sub>GS</sub>		V <sub>GS</sub>	Y <sub>f</sub> (μmhos) Min I Max @ V <sub>GS</sub>		Ptot mW	Use/Comments					
			Min	Max		Min	Max		I	Max							
MPT102	TO-92(72)	25	10	5	8	15	2	1	20	15	0	2000	7500	15	310	N/CH Junction - VHF	
MPT103	TO-92(72)	25	10	5	6	15	1	2	5	15	0	1000	5000	15	310	N/CH Junction - Audio Sw.	
MPT104	TO-92(72)	25	10	5	7	15	1	2	9	15	0	1500	5500	15	310	N/CH Junction - Audio Sw.	
MPT105	TO-92(72)	25	10	5	8	15	10	4	16	15	0	2000	6000	15	310	N/CH Junction - Audio Sw.	
MPT106	TO-92(72)	25	10	5	4	15	10	4	10	15	0	2500	7000	15	310	N/CH Junction - RF	
2N5457	TO-92(72)	25	10	5	6	15	10	1	5	15	0	1000	5000	15	310	N/CH Junction - Audio Sw.	
2N5458	TO-92(72)	25	10	1	7	15	10	2	6	15	0	1500	5500	15	310	N/CH Junction - Audio Sw.	
2N5459	TO-92(72)	25	10	2	8	15	10	4	9	15	0	2000	6000	15	310	N/CH Junction - Audio Sw.	
2N5464	TO-92(72)	25	10	3	3	15	10	1	5	15	0	3000	6000	15	310	N/CH Junction - VHF	
2N5465	TO-92(72)	25	10	5	4	15	10	4	10	15	0	3500	7000	15	310	N/CH Junction VHF	
BFW10	TO-72(75)	30			8	20	15	0				6500			300	N/CH Junction Audio To H.F.	
BFW11	TO-72(75)	30			4	10	15	0				6500			300	N/CH Junction Audio To H.F.	
BFW12	TO-72(35)	25			2	20	15	0				6500			300	N/CH Junction Audio To H.F.	
MPT121	20H	7.70			4	15			5	30	15	0	10000	20000	15	500	N/CH Dual Gate MOS. VHF Amp
2N442	TO-106	20			5	10			12	30	10	0	6000	10	180	P/CH Junction - Audio, Sw.	

TYPE	MAT	V <sub>R</sub>	I <sub>F</sub> (A)	V <sub>F</sub>	I <sub>R</sub> (A)	I <sub>R</sub> (μA)	V <sub>R</sub>	USE
A 14 P	S	1000	2.5	1.25	2.5	0.5	1000	Transient Protected (Controlled Avalanche)
A 15 A	S	100	5	1.1	5	5	100	G.P. Rectifier
BYX11/201R	S	75	1	1.2	25	1.1	75	Automobile H. Duty
EM 4005	S	50	1	1.1	1	5	50	G.P. Rectifier
EM 401	S	100	1	1.3	1	5	100	"
EM 404	S	400	1	1.1	1	5	400	"
EM 410	S	1000	1	1.1	1	5	1000	"
IN 4001	S	50	1	1.1	1	5	50	"
IN 4002	S	100	1	1.1	1	5	100	"
IN 4004	S	400	1	1.1	1	5	400	"
IN 4007	S	1000	1	1.1	1	5	1000	"
IN 540M	S	1000	1	1	3	5	1000	"



# MISCELLANEOUS DATA

## RESISTOR AND CAPACITOR LETTER AND DIGIT CODE (BS 1852)

Resistor values are indicated as follows:

0.47 Ω marked	R47	100 Ω marked	100R
1 Ω	1R0	1 kΩ	1K0
4.7 Ω	4R7	10 kΩ	10K
47 Ω	47R	10 MΩ	10M

A letter following the value shows the tolerance.

F = ±1%; G = ±2%; J = ±5%; K = ±10%;

M = ±20%;

R33M = 0.33Ω ±20%;

6K8F = 6.8 kΩ ±1%.

Capacitor values are indicated as:

0.68 pF marked	p68	6.8 nF marked	6n8
6.8 pF	6p8	1000 nF	1u0
1000 pF	1n0	6.8 uF	6u8

Tolerance is indicated by letters as for resistors. Values up to 999 pF are marked in pF, from 1000 pF to 999 000 pF (= 999 nF) as nF (1000 pF = 1 nF) and from 1000 nF (= 1 uF) upwards as uF.

Some capacitors are marked with a code denoting the value in pF (first two figures) followed by a multiplier as a power of ten ( $3 = 10^3$ ). Letters denote tolerance as for resistors but C = ±0.25 pF. E.g. 123 J = 12 pF × 10<sup>3</sup> ±5% = 12 000 pF (or 12nF).

### Tantalum Capacitors

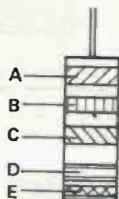
	1	2	3	4
Black	—	0	×1	10 V
Brown	1	1	×10	
Red	2	2	×100	
Orange	3	3	—	
Yellow	4	4	—	6.3 V
Green	5	5	—	16 V
Blue	6	6	—	20 V
Violet	7	7	—	
Grey	8	8	×0.01	25 V
White	9	9	×0.1	3 V
			(Pink 35 V)	



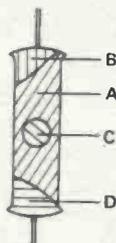
## RESISTOR AND CAPACITOR COLOUR CODING

Colour	Band A	Band B	Band C (Multiplier)		Band D (Tolerance)			Band e	Polyester Capacitors
			Resistors	Capacitors	Resistors	Capacitors	Resistors		
Black	—	0	1	1	—	2 p0	±20%	—	—
Brown	1	1	10	10	±1%	0 p1	±1%	—	—
Red	2	2	100	100	±2%	—	±2%	—	250 v w
Orange	3	3	1 000	1 000	—	—	±2.5%	—	—
Yellow	4	4	10 000	10 000	—	—	—	—	—
Green	5	5	100 000	—	—	0 p5	±5%	—	—
Blue	6	6	1 000 000	—	—	—	—	—	—
Violet	7	7	10 000 000	—	—	—	—	—	—
Grey	8	8	10 <sup>8</sup>	10 n	—	p25	—	—	—
White	9	9	10 <sup>9</sup>	100 n	—	1 p0	±10%	—	—
Silver	—	—	0.01	—	±10%	—	—	—	—
Gold	—	—	0.1	—	±5%	—	—	—	—
Pink	—	—	—	—	—	—	—	Hi-Stab.	—

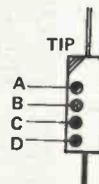
Note that adjacent bands may be of the same colour unseparated.



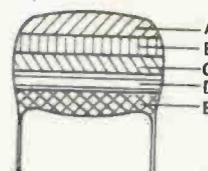
RESISTORS



RESISTORS (OLD TYPE)



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3.3	3.9	4.7	5.6	6.8	8.2	2.4	2.7	3.0	3.3	3.6	3.9	4.3	4.7	5.1
and their decades						5.6 6.2 6.8 7.5 8.2 9.1 and their decades								

## Reference Notes

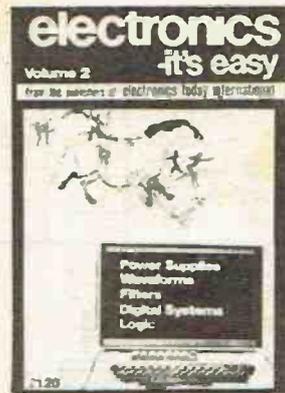
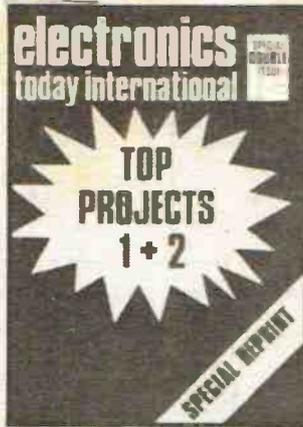
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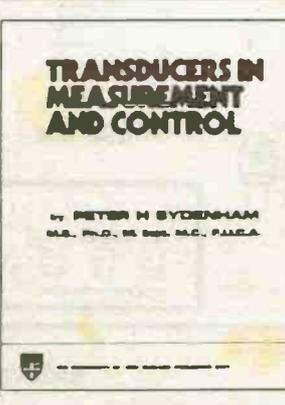
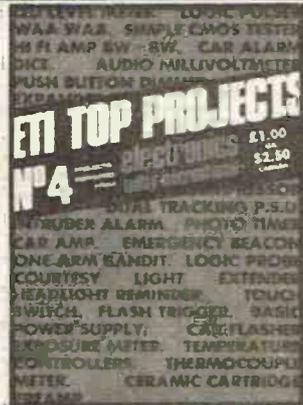
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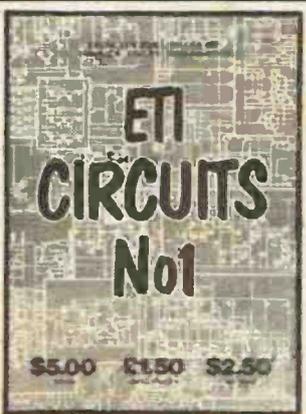
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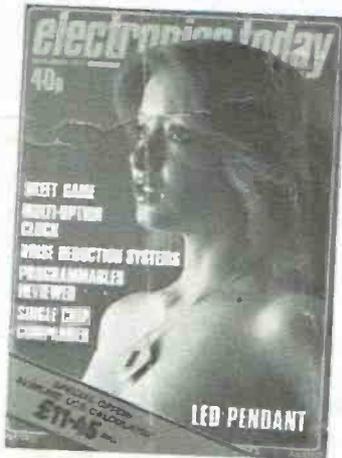
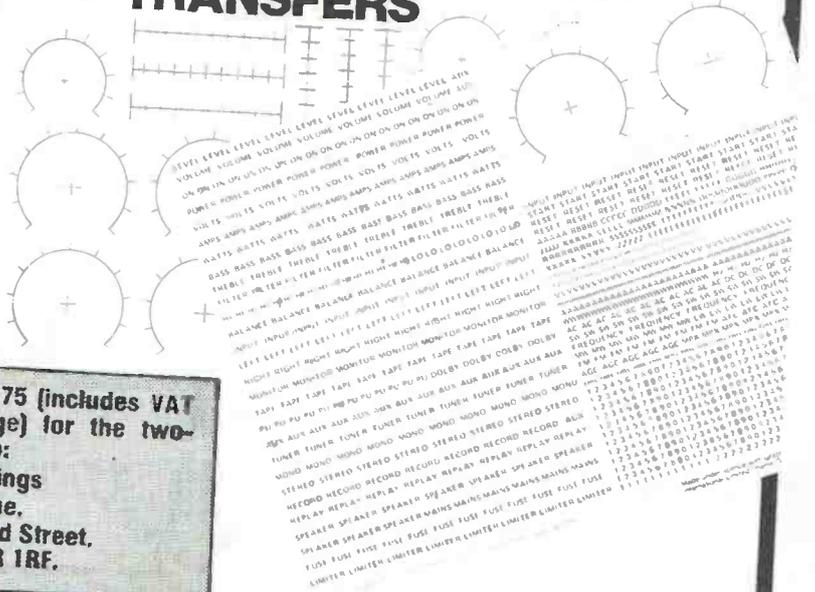
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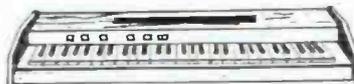
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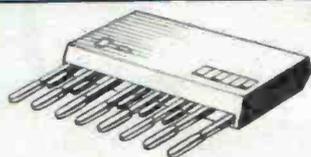
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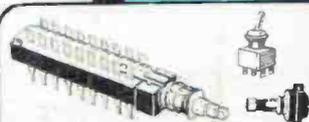
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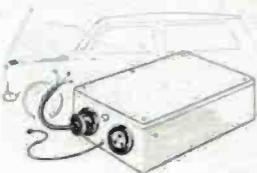
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