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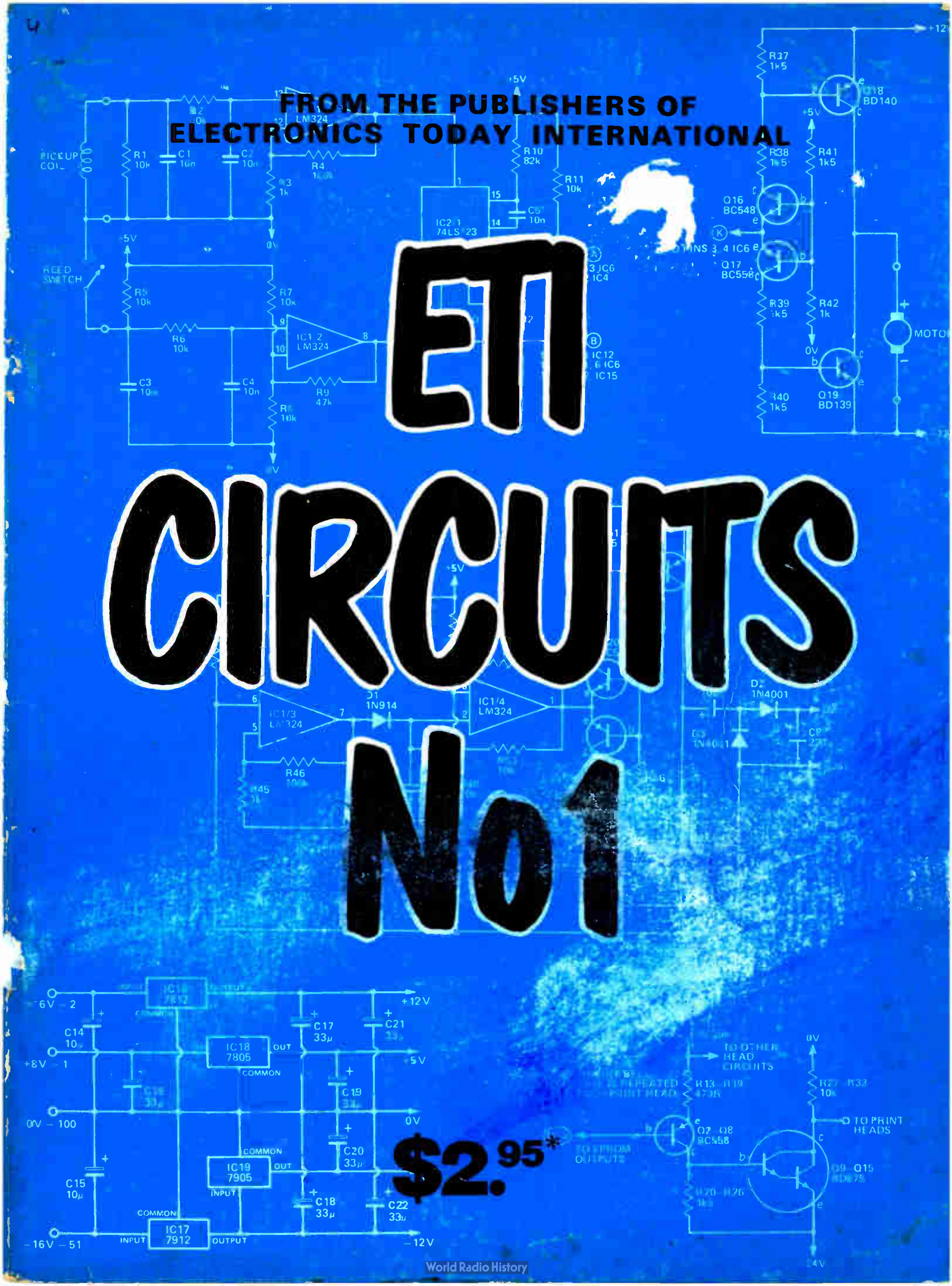


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ETI CIRCUITS BOOK: No.1

Ideas and data for experimenters

THE 'Ideas for Experimenters' section published each month in Electronics Today International has for many years been one of the most popular parts of the magazine. But by the very nature of being a monthly feature, it becomes impossible (without an elaborate filing system) to remember particular circuits — or compare them with similar ones.

Our answer to this problem is this Circuits Book. This book was first published in 1977 (and a second volume subsequently in 1978). Both sold out virtually within days and have now been reprinted by popular demand.

Highlighting the international aspects of ETI, this publication has been prepared from material originally published in the Australian edition, assembled in the UK and reprinted in Australia.

CIRCUITS NO 1:— is an ideas directory and is not meant for the beginner. We regret we cannot answer queries on any of these circuits.

ETI is always prepared to consider original circuits or ideas submitted by readers for inclusion in its publications. All items used are paid for. Drawings must be as clear as possible and text preferably typed. Circuits must not be subject to copyright.

Special Projects Editor: Jan Vernon
Managing Editor: Collyn Rivers

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INTRUDER

BASIC ALARM

The basic alarm circuit uses the minimum of components, has a very low standing current (less than 50μA) and thus may be operated from small dry batteries. The circuit has a lock-out system which prevents the alarm being stopped, except by disconnecting the battery. Any break in the detector loop allows the current through the 100kΩ resistor to switch on the transistors, pulling in the lock-out relay and sounding the alarm.

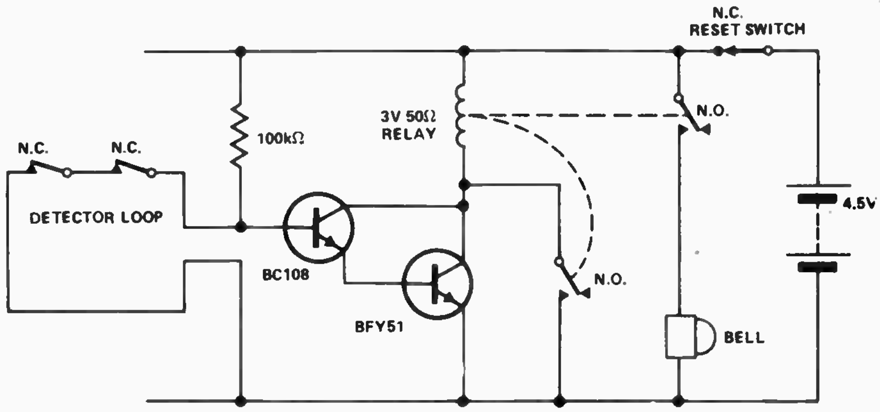


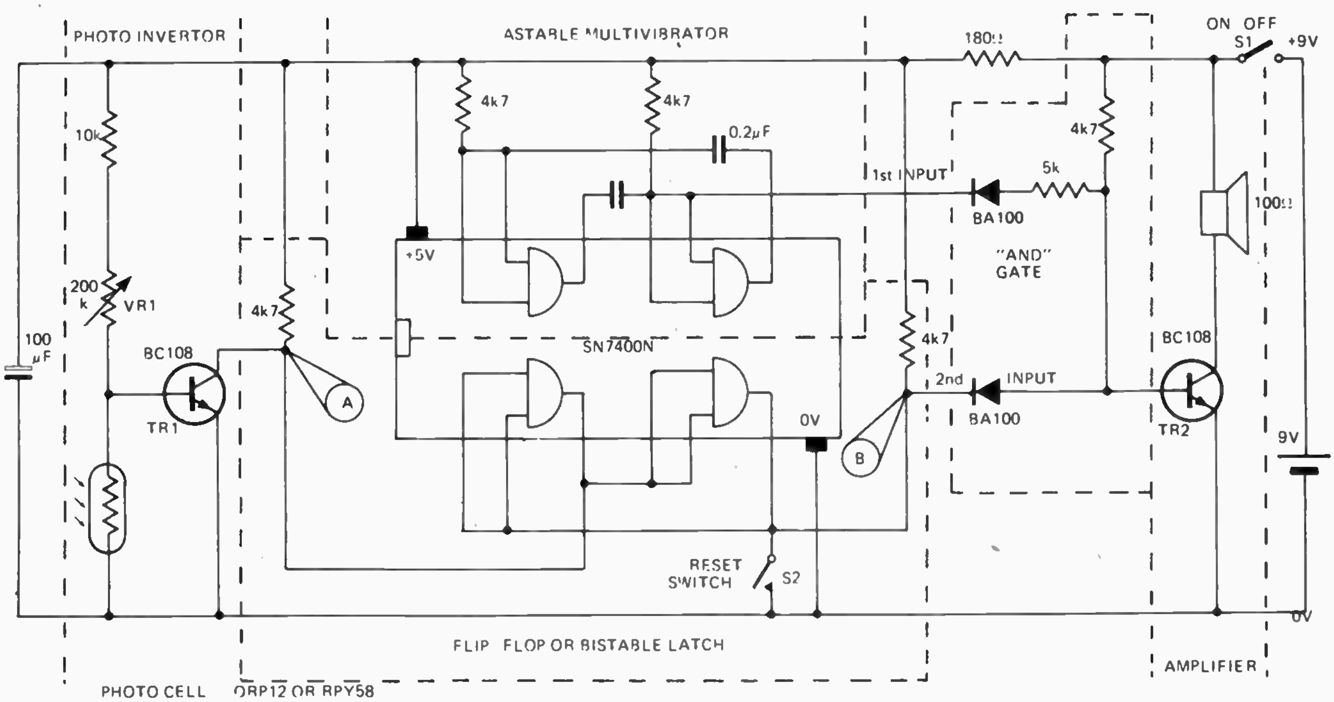
PHOTO INTRUDER ALARM

Designed so that when an object obscures the light following on a photocell an alarm is triggered.

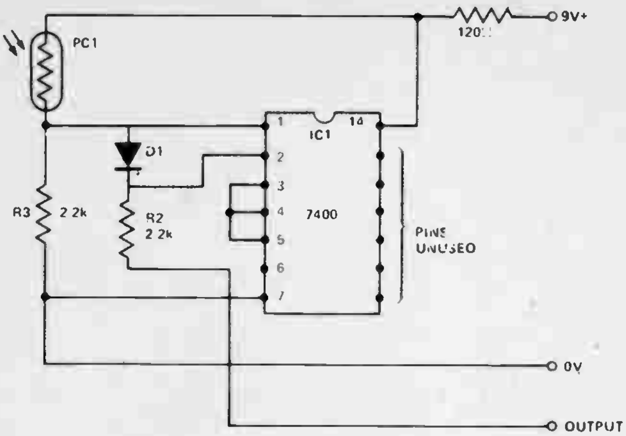
The circuit is reset on depression of the reset switch. IC's (SN7400) has been used for the astable multivibration and bistable latch instead of using transistors, making the finished circuit smaller and neater to build.

The circuit operates as follows. When the light on the photo cell fractionally decreases, the voltage on the collector of TR1 falls (point A). This triggers the bistable latch so that the voltage at B rises from 0V to +5V. Point B is the second input of a simple diode "AND" gate. An astable multivibrator (composed of the other two NAND gates of the IC) generates a square wave which is fed to the first input of the "AND" gate. This waveform is generated continually but does not appear at the output of the gate as long as the second input is held at 0V. When the second input rises to +5V the square wave appears across the emitter-base function of TR2 which drives the amplified tone into the speaker. To reset, the object in front of the cell is removed and the reset switch pressed.

To gain the maximum sensitivity VR1 is turned to its maximum. The reset switch is then pressed if the alarm is sounding. The tone should disappear. Now VR1 is slowly decreased until the alarm operates. VR1 is then increased very slightly until the alarm stops by pressing S2. The circuit is now at its maximum sensitivity for the given light level.



INTRUDER ALARM



Here two gates of a 7400 are used to provide photoelectric control in conjunction with an ORP 12 photocell. When light falls on PC1 the potential is applied to the trigger circuit consisting of ½ the 7400. The feedback provided ensures a positive output change at pin 6. The output, whilst PC1 is under illumination, is equal to the supply voltage. R1 enables a small 9V battery to be used. If PC1 is shaded the output at pin 6 is 0V. This may now be used to trigger a relay for an intruder alarm. If this is the case it is wise to use a small mains supply and to incorporate a diode across the relay coil, to prevent high back EMF from destroying the IC.

LEVEL

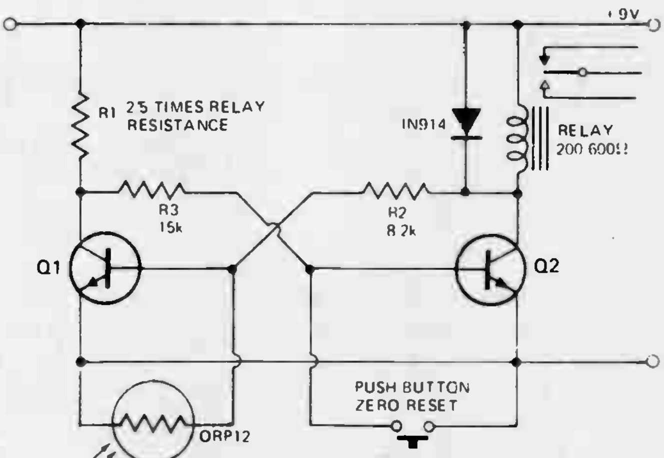
PHOTO ELECTRIC RELAY

There are many applications where photoelectric detection is used to switch a circuit on or off.

This simple circuit is a bistable multivibrator. The base resistor of Q1 is a photoresistor type ORP12. When not illuminated resistance is high, Q1 conducts and Q2 is off.

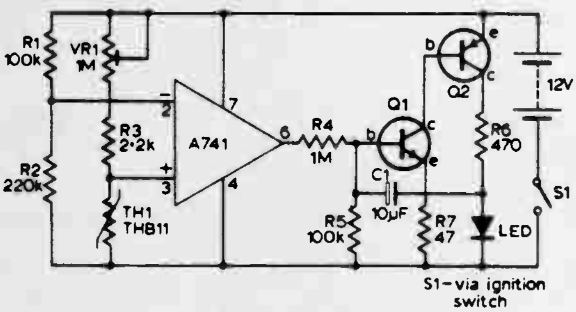
As the illumination on the OPR12 is increased the resistance drops till Q1 cuts off and Q2 turns hard on energising the relay coil.

The system is reset by the pushbutton. The diode across the relay coil can be any low power silicon type. It is for protecting Q2 from any spikes generated across the coil when de-energised.



Q1, Q2, BFY51 or similar

LOW–TEMPERATURE ALARM OR LIGHTS–OUT ALARM



The figure shows a general purpose 741 op amp acting as a sensitive level amplifier of the voltage across the points X and Y of the bridge circuit. Resistors R1 and R2 set the inverting terminal (2) of the op amp at about 6V with respect to ground. The voltage at the non-inverting terminal (3) is determined by the temperature of the thermistor Th1. As the temperature of the thermistor falls, its resistance rises, because it is a negative temperature coefficient type, and the voltage at pin 3 rises. If this voltage rises above that of pin 2, the voltage at the output pin 6 of the op amp goes positive. The temperature at which the output goes positive can be selected within limits by VR1.

When the voltage at pin 6 goes positive, the complementary pair of transistors Q1 and Q2 operate as an oscillator, the positive feedback being provided by C1. The LED flashes at a rate determined partly by it's own resistance but also dependent largely on the value of C1. The resistor R6 should be adjusted to maintain a current through the LED at a value less than it's rated maximum.

Switch S1 is the ignition switch on the car. Transistor Q1 is the npn type ZTX300 in the prototype and Q2 a pnp ZTX500, but other complementary pairs of medium-current audio frequency transistors can be used. The thermistor used is a glass bead type having a nominal resistance of 1MΩ at 25°C. The type indicated is from RS Components Limited.

LEVEL

TEMPERATURE SENSOR

Simple, inexpensive circuit has a resolution of 0.15°C.

THERE are many occasions when it is necessary to detect a difference in temperature between one point and another and to take some action should the temperature differential rise beyond some pre-set level. An application that immediately springs to mind is the control of cooling fans in equipment cabinets in response to the difference between the internal temperature of the cabinet and ambient.

Jim Barnes of Motorola's application laboratories at Phoenix, Arizona, has designed a simple, inexpensive circuit for just this purpose. This has a resolution of about 0.15°C and a differential range of 5.5°C.

In the circuit, temperature measurement is performed by general purpose silicon diodes (IN4001). These exhibit a temperature coefficient of about 2 mV/°C over a very wide temperature range. When two of the diodes are connected in the bridge configuration shown in Figure 1, a voltage appears between terminals A and B which is proportional to the temperature difference between the two diodes. The potentiometer provides a variable offset current giving a temperature offset range of about ± 5.5°C.

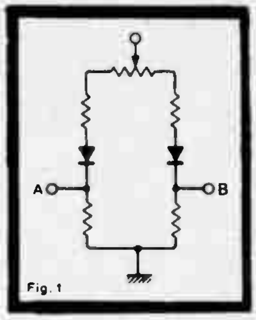


Fig. 1

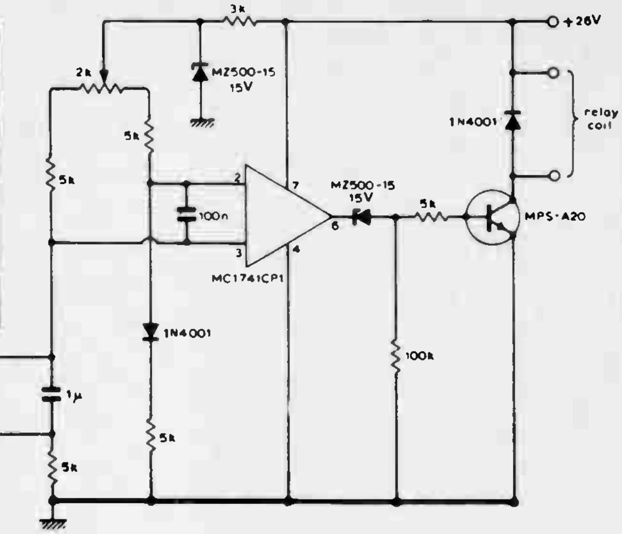


Fig. 2

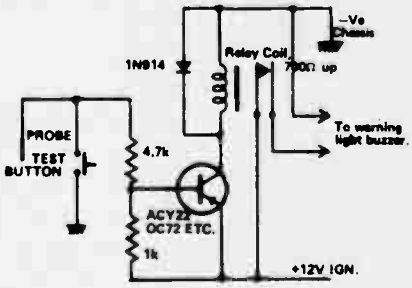
The bridge output of 2 mV/°C temperature differential must be amplified before it can perform a power switching function. A standard low-cost MC1741G operational amplifier was chosen for this purpose (see Figure 2) and used open-loop to provide a gain of some 100 000. A temperature differential of 0.15°C will therefore cause the output of the amplifier to swing almost the whole power supply voltage of 26 V.

The output current capability of the MC1741 (5 to 10 mA) is insufficient to drive most power relays, so a buffer transistor (MPS-A20) is included in the circuit for this purpose. A zener diode provides the necessary level shifting between the output of the op-amp and the input of the buffer.

COOLANT LEVEL WARNING DEVICE

A simple circuit is shown for indicating a drop in radiator coolant level. A variety of transistors and relays can be used and the probe can be made quite easily. The coolant and anti-freeze resistance to earth is about 100Ω and with the level below the probe, infinity.

On the mechanical side the hole in the top tank of the radiator was cut with a pair of sharp pointed dividers.

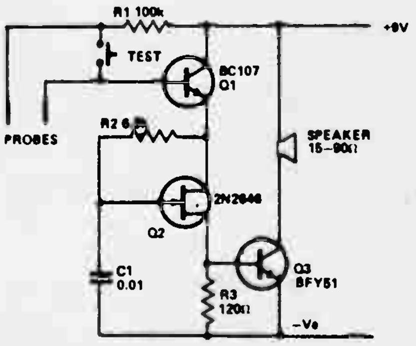


WATER LEVEL ALARM

The disadvantage with battery operated alarm circuits is the quiescent current that they draw. The circuit shown above draws so little current that the shelf-life of the battery is the limiting factor - the only current drawn is the leakage of the transistors.

The circuit is shown in the form of a water level alarm but by using different forms of probe can act as a rain alarm or shorting alarm; anything from zero to about 1MΩ between the probes will trigger it.

Q1 acts as a switch which applies current to the unijunction relaxation oscillator Q2. Alarm signal frequency is controlled by values and ratios of C1/R2. Pulses switch Q3 on and off, applying a signal to the speaker.

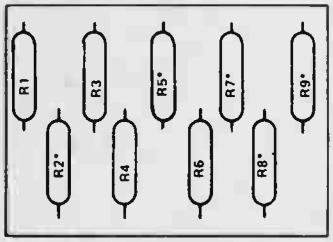


Almost any NPN silicon transistors can be used for Q1 and Q3 and almost any unijunction for Q2.

MISCELLANEA

ELECTRONIC LOCK

* REPRESENTS REED SWITCHES THAT HAVE TO BE TURNED ON I.E. MAGNETISED



REED SWITCHES ARE SHOWN IN ONE ROW IN CIRCUIT DIAGRAM BUT SHOULD BE WIRED AS SHOWN ABOVE TO AVOID CROSS TRIGGERING BY CLOSE REED MAGNETS

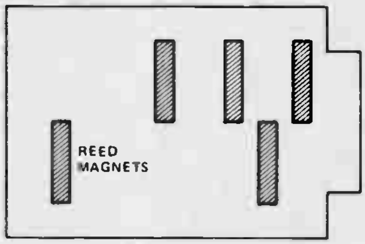


DIAGRAM OF KEY NEEDED TO OPEN THIS LOCK

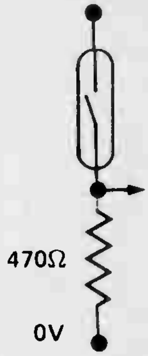


fig. 3

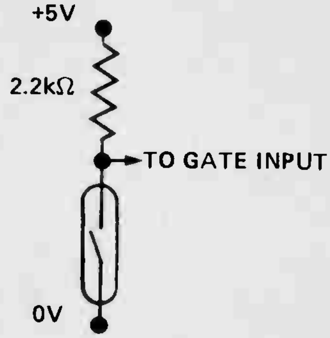
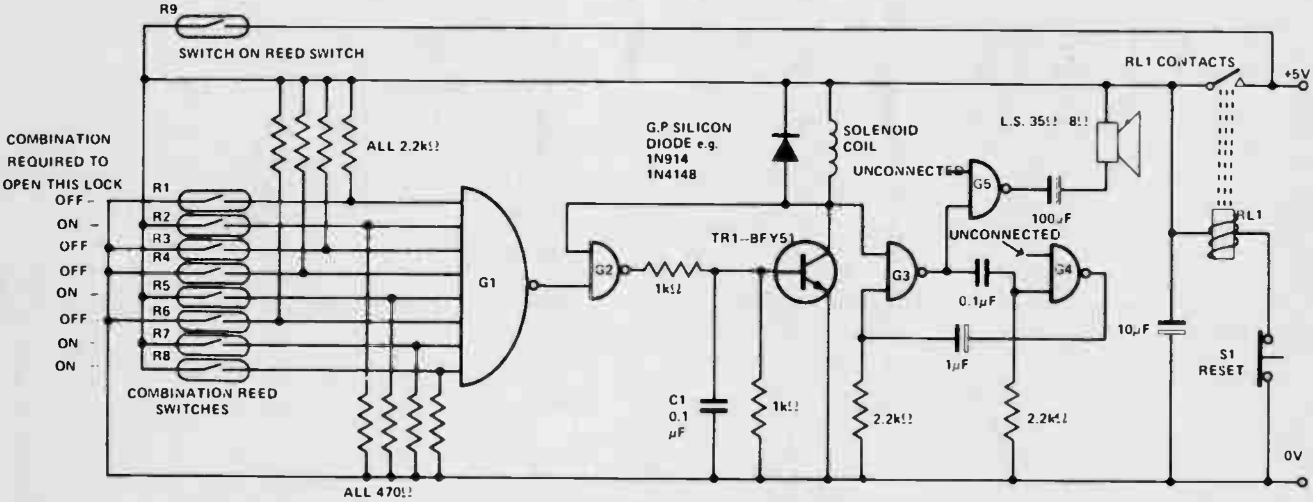


fig. 4



NOTE G1 = SN7430 8 INPUT NAND GATE I.C G2,G3,G4,G5 = SN7400 QUAD 2 INPUT NAND GATE I.C.

This device enables a solenoid to be switched on by means of an electronic key. If the correct key is used the circuit will latch, but if an incorrect key is inserted a warning tone rings until the correct key is used. The circuit has automatic switching to turn it on, but this can be replaced by a conventional ON-OFF switch if desired.

The main element of the circuit is G1 the eight-input NAND gate. If all inputs of the NAND gate are high (achieved by closing the right combination of reed switches) the output will be low. The low output is fed to G2 forcing its output high which turns on TR1 energising the solenoid coil. At the same time a low is fed, back from TR1's collector to G2's other input latching it. Thus once the solenoid is energised the key may be removed. C1 (0.1μF) ensures that TR1 is always 'off' on switch-on of the circuit.

TR1's collector is also connected to G3's input which along with G4 forms a multivibrator. When TR1's output is low the multivibrator is dis-enabled. However if an incorrect key is used TR1's output will be high and the multivibrator will oscillate. G5 acts as a buffer to drive a loudspeaker.

R9 with RL1 forms the automatic switch on circuit. When R9 is closed RL1 is energised pulling in it contacts to permanently connect the supply. If R9 is opened after this the circuit continues to operate. This means if an incorrect key is used and R9 is closed the alarm tone will continue to ring even if the key is removed.

Nine reed switches are used in the circuit. One to switch on the circuit and the other eight to provide the correct input combination. The lock opens only if all eight inputs to the NAND gate are high. To do this the circuit is wired so that some reed switches must be on and some must be off. The eight reed switches give 2^8 = 256 possible combinations.

Reed switches that have to be turned on are wired like this fig. 3. If the reed switch is not closed the 470Ω will pull gate input to low and the lock will not open.

Reed switches that have to be left open are wired like this fig 4. If the reed switch is closed it will put gate input to low and the lock will not open.

If preferred the reed switches could be replaced by simple ON/OFF switches then the circuit would act as a combination lock, like a tumbler lock.

The construction of the circuit is in no way critical. Veroboard provides a relatively cheap mounting. The connections of Vcc and gnd to the two ICs should not be forgotton!

MISCELLANEA

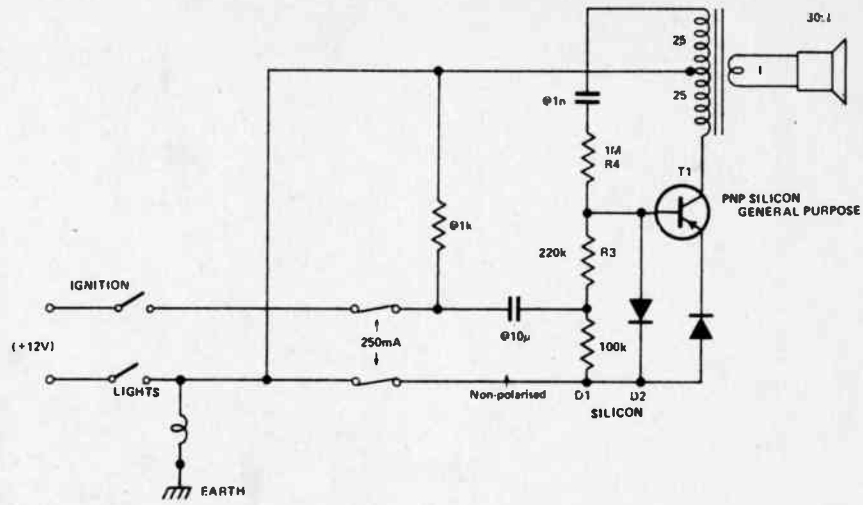
CAR BATTERY WATCHDOG

In winter, motorist are apt to emerge from work to face the inconvenience and perhaps expense of a totally flat battery due to having left their headlights switched on when parking.

This circuit provides an audible warning if the ignition is switched off with the lights left on, in the form of a few seonds of output of varying pitch. No switches are required and standby current is very small.

The audio oscillator is normally biased off, but when the ignition switch is opened it is temporarily biased on the charging action of R1, R2, C1.

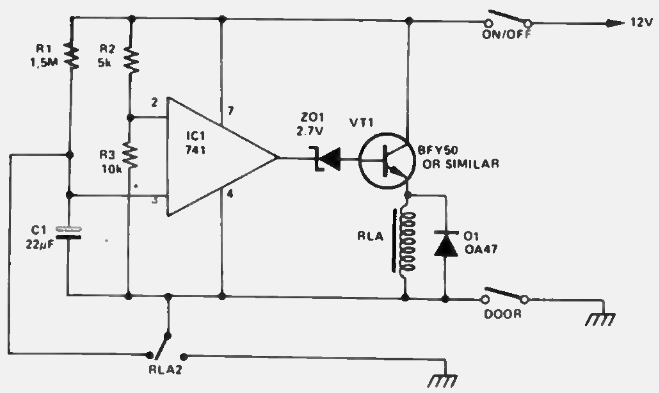
D1 in conjunction with R3, prevents damage to T1 due to spikes on the ignition line, etc. The fuses are an optional precaution against short circuits across ignition or lighting supplies.



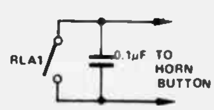
The oscillator circuit will no doubt depend, as will the transducer, on the contents of the experimenter's junk box. Basic requirements are that it should not be self-sustaining when the ignition switching transient in the base circuit has died away. The ratio R3/R4 was of course chosen to achieve this in the circuit shown, assisted by D2.

For negative earth operation T1 would of course be NPN and D1 and D2 would be reversed.

SIMPLE CAR ALARM



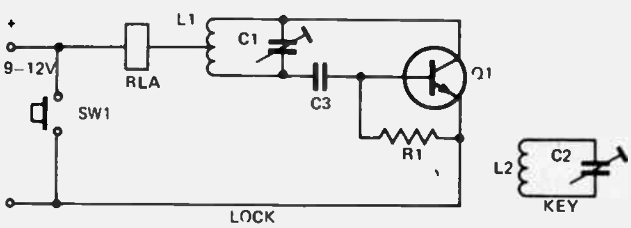
NOTE
RLA–12V ABOUT 200R
741 -VE SATURATION INSUFFICIENT TO TURN OFF VT1 IF ZO1 IS OMITTED SOME TYPES OF RELAY MAY NOT RELEASE.
ALL RELAY CONTACTS SHOWN RELEASED.



At the instant the door switch is operated the +Ve 741 input is at 2/3 the rail voltage. The –Ve 741 input is fully negative. The –Ve input is thus –Ve with respect to the positive input and an output near the rail voltage results turning on VT1 and operating RLA.

RLA1 sounds the horn and so must have heavy duty contacts. RLA2 shunts the door switch and also removes the short from C1. C1 now charges Via R1 until the voltage at IC1 –Ve is approaching that of the +Ve input. The output of IC1 now falls until –Ve saturation occurs and VT1 is turned off releasing RLA. This takes approximately C1 x R1 seconds. The components shown gave a 30 second delay in the prototype.

SIMPLE ELECTRONIC LOCK



Operation is very simple, when the key is brought near the lock, providing L1–C1 have the same resonant frequency as L2–C2, the reed relay will open as the key absorbs energy from the lock. After each operation of the lock, it should be reset by a short press on the reset button.

Setting up is best done with a resistor in place of the relay with a voltmeter across it, making the operating range of the circuit more apparent.

PARTS LIST

Q1	BC108
C1/C2	250pF trimmers
C3	1000pF ceramic
L1/L2	80 turns 34swg on 3/8" diameter ferrite (L1 is tapped at 40 turns)
R1	27k
RLA	Reed relay resistance around 5k
SW1	Push-to-make switch

P.S. Found that ETI's TIC-TAC radio could be used as key!

AUDIO : BUFFER

HIGH INPUT IMPEDANCE AMPLIFIER

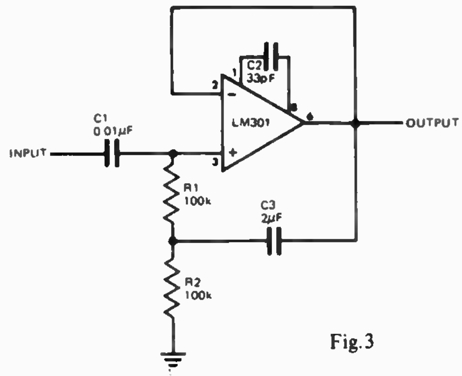
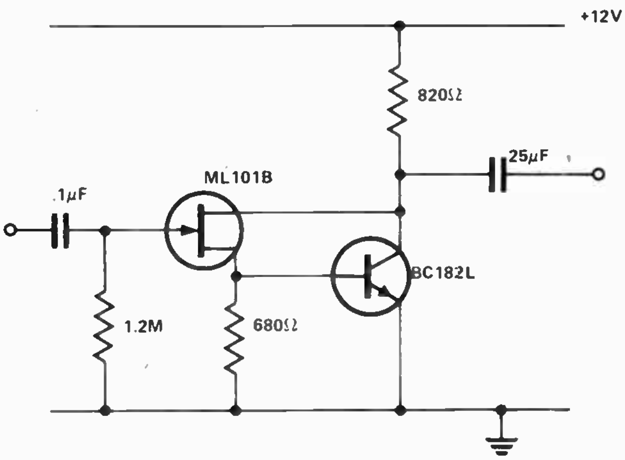


Fig.3

The LM301 may also be used to construct a simple high input-impedance ac amplifier as shown in Fig. 3. In this circuit even though the bias resistor is only 200 k, as required for good dc stability, the bootstrapping by C3 provides an input impedance of 12 M at 100 Hz increasing to 100 megohm at 1 kHz.

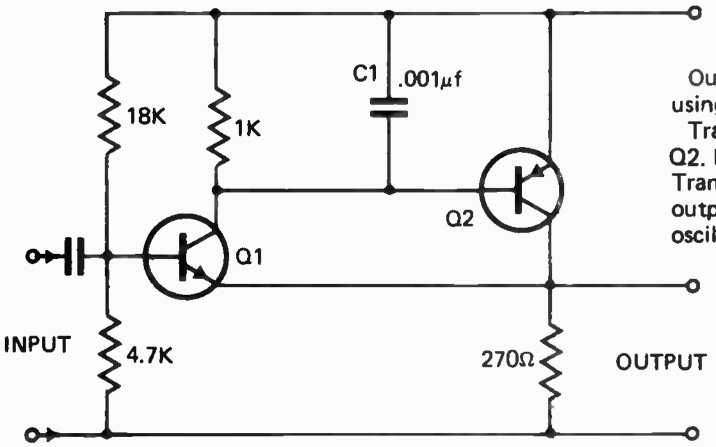
A HIGH IMPEDANCE BUFFER AMPLIFIER



This circuit has a voltage gain of just less than unity, but its power gain is very large indeed. It makes an ideal preamplifier for a high impedance source signal. The input impedance is about 800k with the FET specified, but if a FET without a built in gate protection diode is used, the input impedance will be largely controlled by the gate resistor. The circuit has a small signal output impedance of about 10 ohms and is capable of delivering about 7mA p-p into a capacitively-coupled 25 ohm load. The low-frequency breakpoint is about 240Hz, the upper breakpoint is in excess of 1MHz.

The principle of operation is simple. The circuit employs a FET front end to obtain the high input impedance, but the transconductance of the FET is too low to be useful on its own, and so it is boosted by the output transistor, the BC182L.

VERY LOW OUTPUT—IMPEDANCE

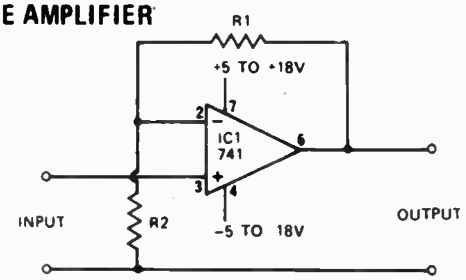


Output impedances as low as 0.05 ohm can be obtained by using this configuration.

Transistor Q1 is an ordinary emitter follower, assisted by Q2. Main load current is supplied by the collector of Q2. Transistor Q1 senses the difference between input and output voltage and regulates Q2 accordingly. C1 prevents oscillation.

AUDIO : BUFFER

HIGH INPUT IMPEDANCE AMPLIFIER



The circuit shown, using one op- amp and two resistors has a high input impedance (500 nanoamps input current) and a gain which may be programmed by R1 and R2.

$$G = \frac{R1 + R2}{R2}$$

Thus for G = 1 R1 = 0, R2 is not used
for G = 100 R1 = 100k R2 = 1k.

The frequency response decreases with increasing gain, eg, for G = 1 the amplifier is flat to 800 kHz, for G = 100 the response drops to 6 kHz.

AUDIO : EQUALISED

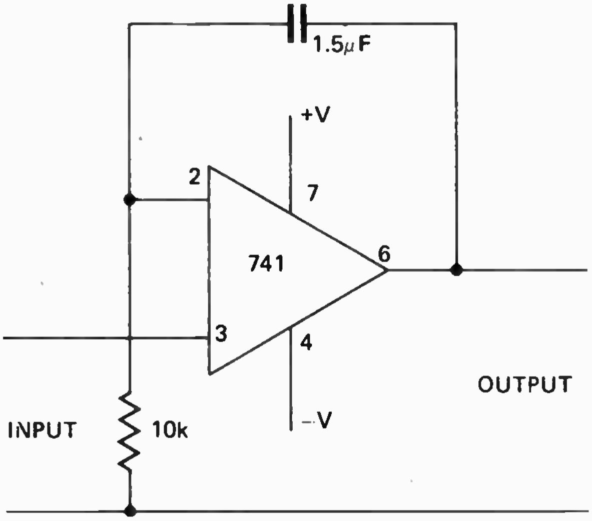
LOW FREQUENCY EXTENDER

In circuits which have a variable frequency input, e.g. optical tachometers, vibration measuring equipment etc., the low frequency response can leave a lot to be desired. The circuit shown brought the lower 3dB point of a measuring instrument down to 0.5Hz when placed in circuit between the transducer and the instrument.

Being of small size, the circuit may be fixed inside the case of the instrument it is to serve.

The gain of the circuit may be altered by means of the feedback capacitor to give a level response compatible with the instrument to which it is connected, i.e. a higher value will give a lower gain and vice-versa.

The 741 IC will operate at voltages between ±5 and ±15V.

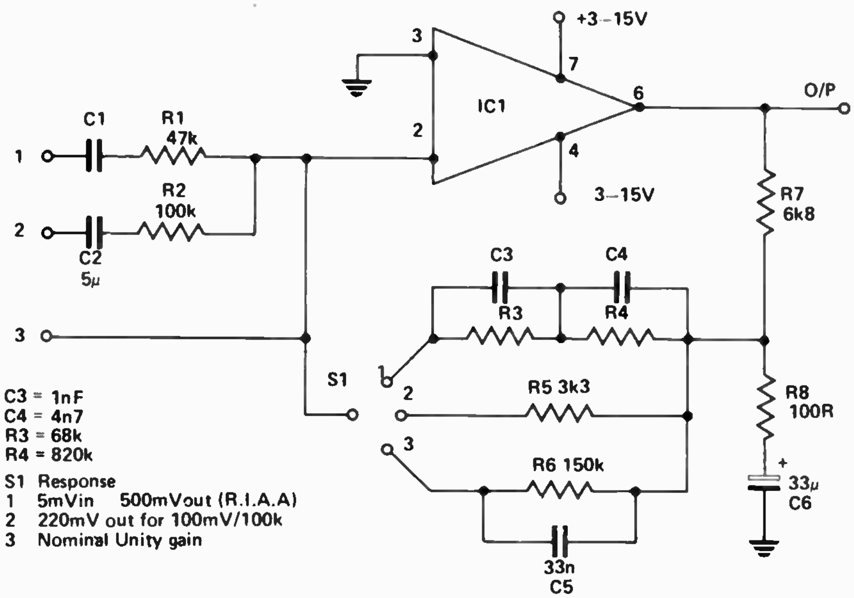


VIRTUAL EARTH PREAMP

This circuit caters properly for both magnetic and ceramic pickup cartridges and features an auxillary input with a flat response.

A 741 op amp is used in the inverting mode as a virtual earth amplifier. In order to maintain a reasonable closed loop gain the feedback is tapped from the junction of R7 and R8.

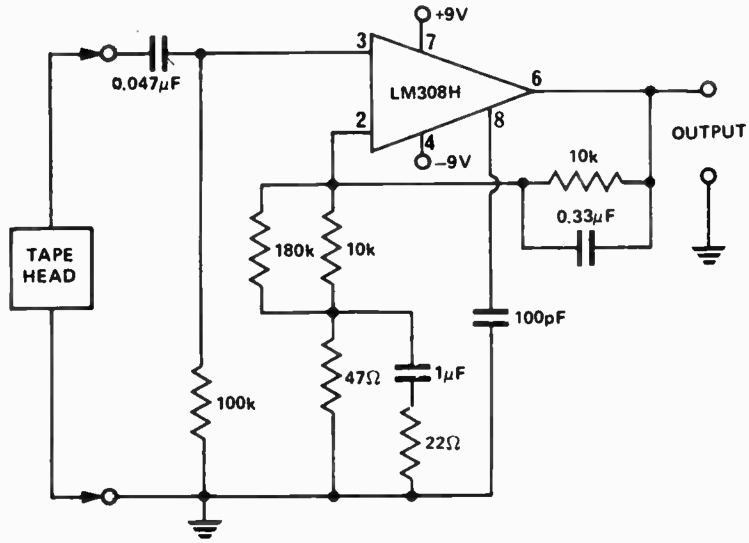
Advantage is taken of the low impedance summing point to directly connect the various inputs reducing the complexity of the switching arrangements. Of particular interest is the equalisation for ceramic cartridges employed. It is based on the charge amplifier principle. R6 in parallel with C5 introduces a –3dB cut at 40Hz and C1 performs the same cut at 40Hz and C1 performs the same function at the magnetic P.U. input.



C6 provides a further attenuation at this frequency providing, a steep cut rumble filter with a –6dB point at 40Hz.

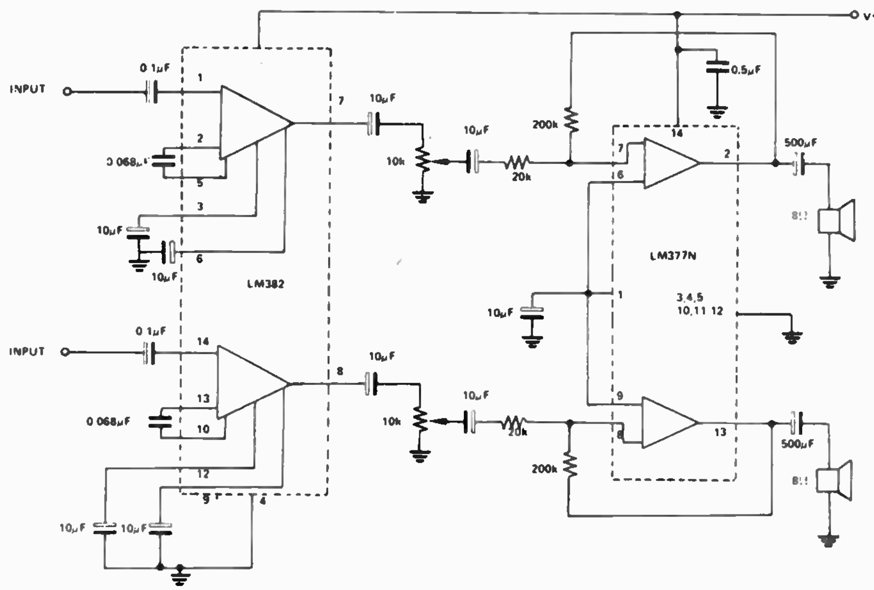
IC TAPE-HEAD PRE-AMP

This circuit is suitable for a tape speed of 3.75 inches/sec. and provides a rising gain at low frequencies (about 40 dB below 100 Hz) a minimum gain of about 15 dB around 2-3 kHz and a 6 dB boost (to about 21 dB) above 10 kHz for reasonable compensation. A low noise op-amp is used.



SIMPLE STEREO TAPE PLAYER

The circuit shown employs the National Semiconductor LM382 low noise dual preamplifier and the LM377 dual 2W power amplifier as a complete stereo tape player circuit providing the normal NAB equalization characteristic. The number of components employed is much smaller than in conventional circuits. The power supply voltage should be in the range 10V to 26V, but no special precautions are required to remove hum from the supply line, since the LM382 provides 120dB hum rejection and the LM377 80dB rejection.

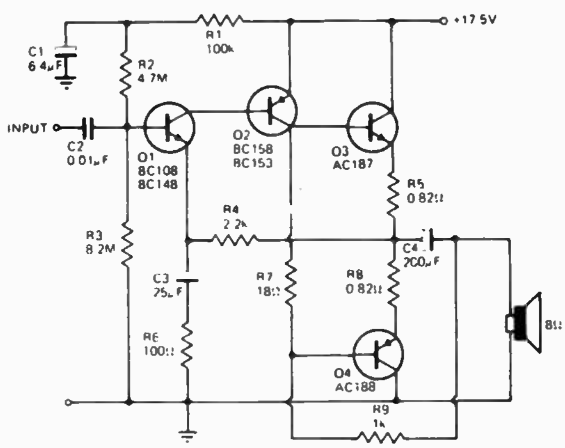


AUDIO : POWER

AUDIO AMPLIFIER

A high input impedance of 1.1 megohm is made possible in this amplifier by keeping the collector current of Q1 low, and by using a high level of ac and dc feedback. The input sensitivity is adjusted by altering the value of R3.

The quiescent current of the output stage is 2.5 mA and is stabilised by resistors R5 and R8. With a 17.5 volt supply the amplifier will deliver 2.5 watts across 8 ohms with a distortion of less than 1% at 1 kHz.



AUDIO : POWER

20W SLAVE AMP

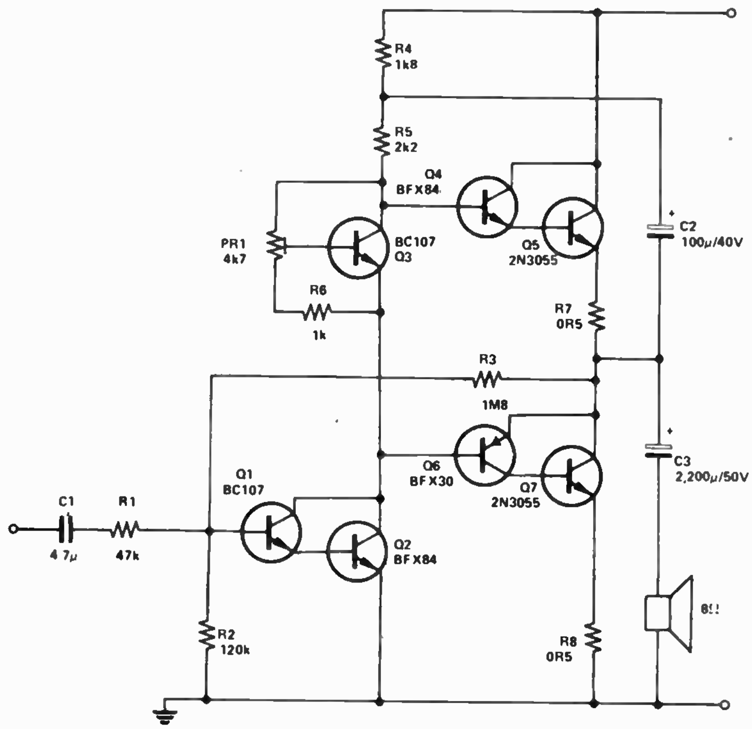
This amplifier is very simple to build and most of the parts will probably be available from the constructor's 'junk' box. The circuit consists of a Darlington pre-driver, Q1 and Q2, a V_{BE} multiplier Q3 and a quasi-complementary output stage Q4–7.

Overall shunt feedback is applied from the collector of Q7 to Q1's base via R3 which, in conjunction with R2, also provides DC feedback and input bias. The voltage gain, and hence the sensitivity of the amplifier, is set at 33 and 370mV by the ratio of R3 to R1.

Quiescent current through Q5 and Q7 should be set at 30mA by PR1.

The collector load of the Darlington, R4 and R5, is bootstrapped by C2 to provide a current drive for the output stage.

Although simple the amplifier is capable of good quality reproduction and will operate quite happily into a 4, 8 or 16R load.



AUDIO AMPLIFIER

In this circuit a 741 is used to drive a complementary output stage from a split supply of 9–0–9V.

The input signal is coupled to the non-inverting input of IC1 via C1.

The amplified output signal from the IC used to drive Q1, which is connected in the emitter follower mode.

Q2, in Q1's emitter circuit, provides sufficient bias for Q3 and Q4 to eliminate crossover distortion and prevents thermal runaway in these transistors.

R4 sets Q1's current at 44mA.

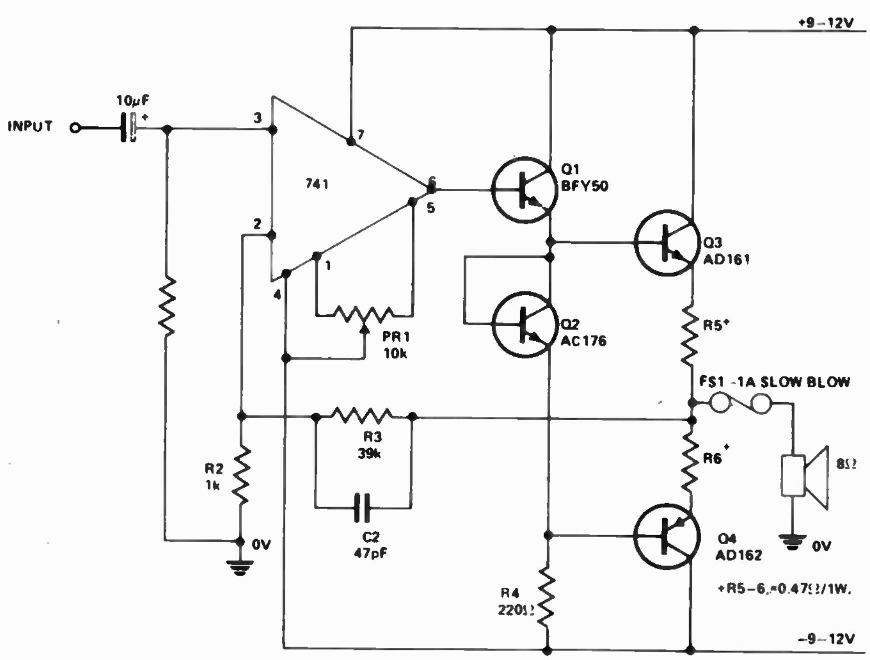
The low value resistors, R5 and R6 in the emitter circuit of Q3 and Q4 should be wire-wound types with a minimum rating of 1W.

Overall feedback is provided by R3 and R2, C2 is incorporated to roll off the response at R.F.

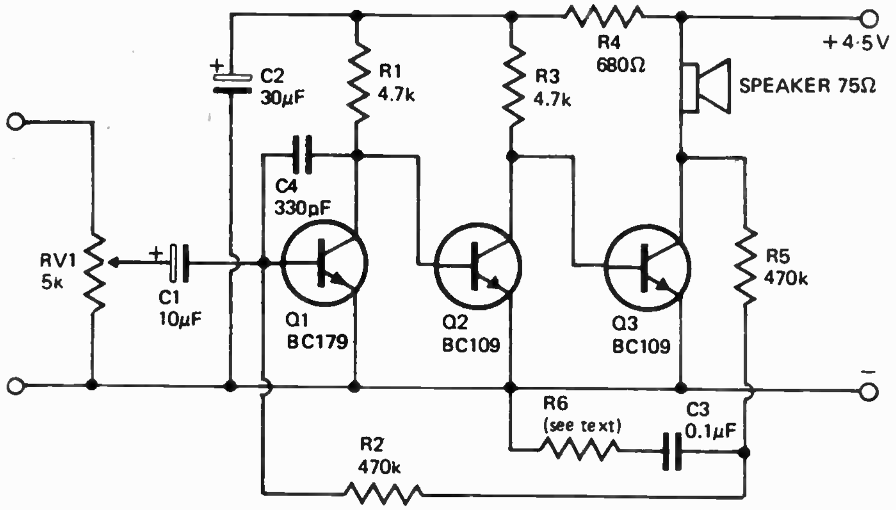
Q3 and Q4 should be a matched pair and require a heatsink of at least 12 sq.in. of 18swg aluminium sheet from which they should be insulated with mica washers and nylon bushes in the usual manner.

Before use the circuit must be correctly set up. This is accomplished by adjusting PR1 so that the voltage at the junction of R5 and R6 is 0V.

This must be done before connecting the speaker.



ECONOMY AMPLIFIER



When power output, harmonic distortion, frequency response are not the absolute parameters for an amplifier, such as in the case of small personal portable radios, operation of an amplifier in class 'A' does have a number of advantages.

The circuit shown uses only three transistors, does not require an output transformer, and gives an output of between 100 – 200 mW for a battery supply of only 4.5V.

RV1 provides volume control and couples into the amplifier through C1. The following three stages are directly coupled.

Q1 base bias is established by resistors R2 and R5. R1 – Q1 act as a bias potential divider for Q2 base and similarly R3 – Q2 bias base of Q3.

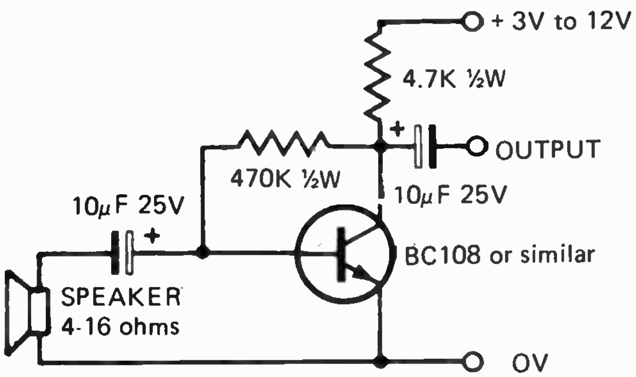
R2 and R5 also form part of an overall negative feedback loop improving frequency response and reducing distortion.

A compromise between gain and quality results in a choice of values for R6 and C3. C3 is a decoupling capacitor and R6 is adjusted by trial and error. (Minimum value should be 22 k.)

AUDIO : SPECIAL

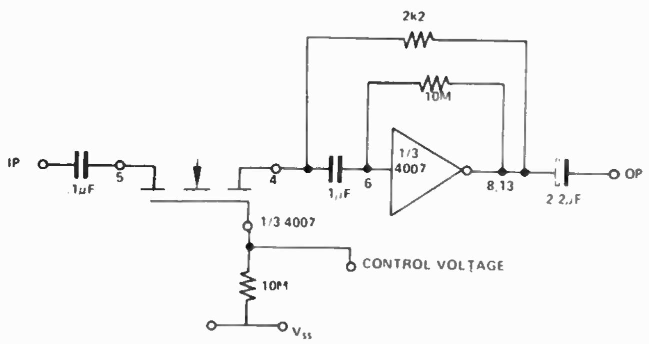
LOUOSPEAKER MICROPHONE

A small loudspeaker may be used as a very effective microphone for intercoms etc. if it is coupled via the circuit shown above.
Output is sufficient to drive practically any transistor or valve audio amplifier.



VOLTAGE CONTROLLED AMPLIFIER

When the voltage at the gate of a n-channel MOSFET is varied from 0V – supply volts its resistance varies from about 1kΩ to several tens of megohms. This fact is utilised in the following VCA. The inverter is biased into linear operation by the 10mΩ resistor. When feedback is applied the gain is set by $\frac{RF}{RIN}$. By allowing a MOSFET to be RIN and RF fixed, with the values shown as the control voltage varies from V_{DD} – V_{SS} the gain of the amplifier varies from cut-off to just over unity.



R.F. & VIDEO

WIDEBAND AMPLIFIERS

IT IS not commonly known that some digital ICs can be used in the linear mode to obtain performance equal, or superior, to some more conventional components.

A typical example is the use of a MECL logic gate as a wideband amplifier. Such an amplifier based on the Motorola MC 1023 of the MECL 2 family provides a gain of 5.2 over a frequency range from zero to 125 MHz (at the 3 dB points). A still wider bandwidth of zero to 350 MHz may be obtained by using the MC 1660 from the MECL 3 family.

The method used to bias MECL gates for linear operation is shown in the inset of Fig. 1. The NOR output is connected back to the input. This can be done over one, or over several, gates. The external 'self-biasing' network feeds back only the dc component of the output signal. Therefore the dc input current is furnished by the output of the same gate. Assuming that the voltage drop across the biasing resistors is small, the input and output voltages are identical. This is only possible in the centre of the gates' transition region. The main advantage of this very simple biasing method is that the circuit automatically compensates for all offset and bias voltage variations. In addition, the method is very economical, especially when a cascade arrangement of gates is needed.

The response depends on how many inputs are connected in parallel, there is a disadvantage however in connecting several inputs together to increase gains. The offset voltage between input and output increases with the number of inputs that are paralleled. It therefore depends on the individual application, if a slightly higher offset voltage can be tolerated then a higher gain can be achieved.

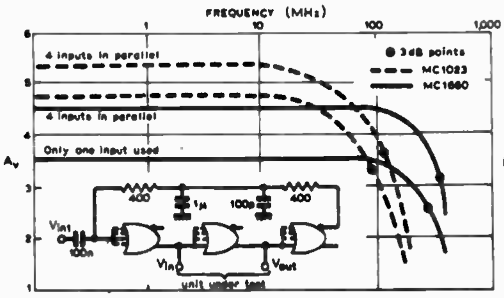


Figure 1

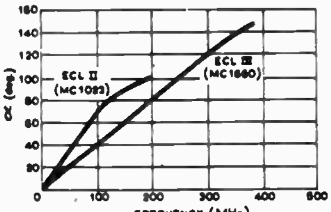


Figure 2

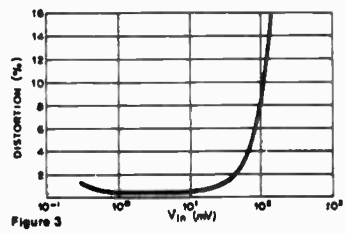
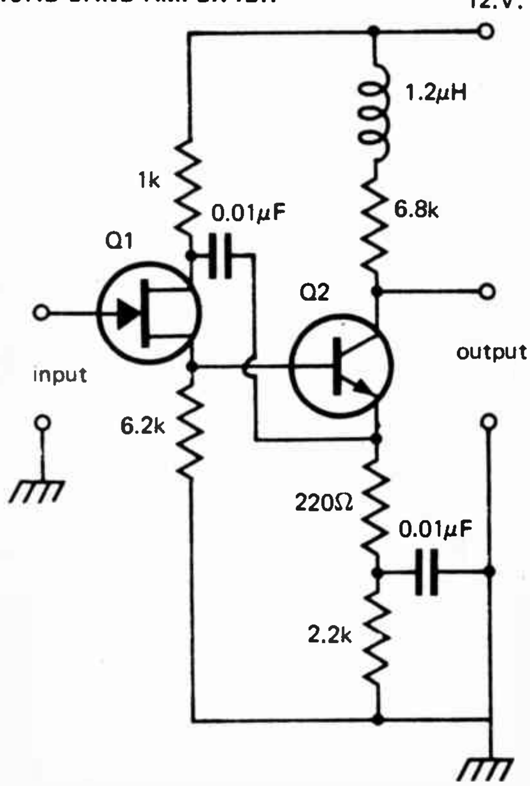


Figure 3

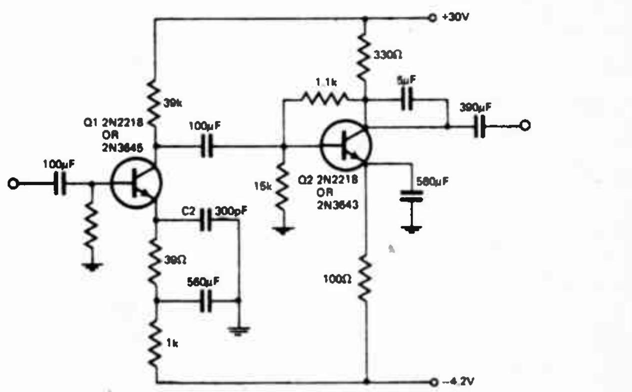
Fig. 2 shows the phase shift curves for the two gates and Fig. 3. is a plot of distortion against input voltage.

BROAD BAND AMPLIFIER



This circuit has a typical gain of 10dB and bandwidth of 90 MHz.

VIDEO POWER AMPLIFIER



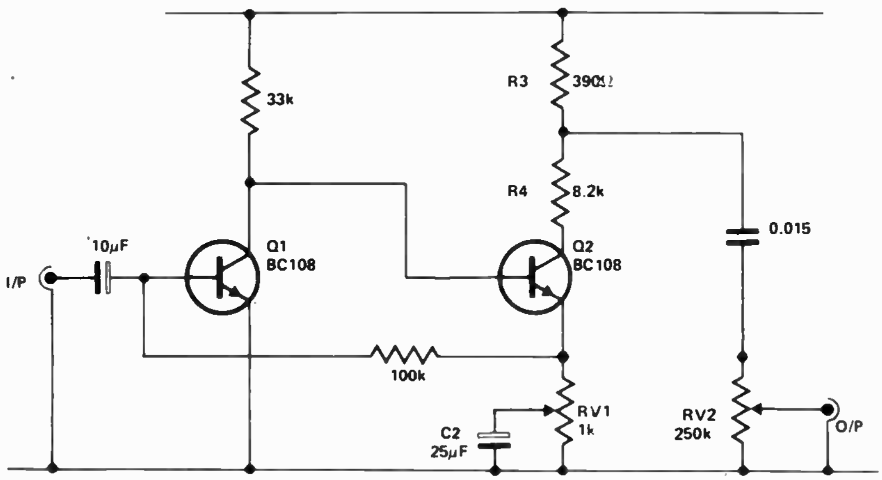
The amplifier shown has a frequency response of from 5Hz to 30MHz and is capable of giving a 10 volt output into a 100 ohm load. The circuit provides 26 dB gain and has excellent stability and linearity.

Input impedance is around 10 megohms in parallel with 1.0pf. Output impedance is reasonably high and depends mainly on Q2; output capacitance will be around 2 to 3pf with careful construction.

FET, Q1, should be an n-channel type with low gate source capacitance and a high cutoff frequency.

The Transistor, Q2, should have a high gain-bandwidth product and low collector-emitter capacitance. Careful selection can extend the bandwidth beyond 100 MHz.

FUZZ BOX FOR ELECTRIC GUITARS



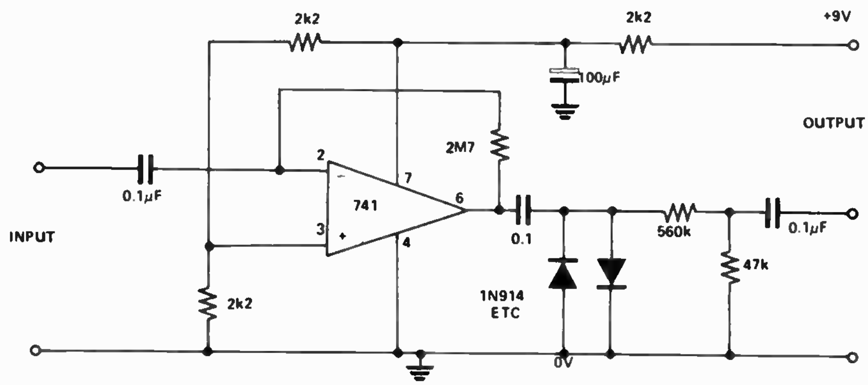
This circuit has been well tried by many musicians and has proved very successful.

Q1 and Q2 form a voltage amplifier which has sufficient gain to be 'overdriven' by a relatively low input, such as an electric guitar. The result is that the output from Q2 is a 'Squared-Off' version of the input, giving the required fuzz sound.

RV1 adjusts the amount of negative feedback inserted into the circuit by C2, and thus the amount of squaring of the signal. The purpose of R3 and R4 is to lower the output voltage to a suitable level, which is then adjusted as required with the volume control VR2.

GUITAR FUZZ

The 741 normally has a gain of 20,000, but the circuit is so designed that the IC's gain is 2,700,000 which then distorts the output. This distortion gives the fuzz effect. The two diodes clip the output to drop the level, also lowered by the potential divider. This circuit also sustains the notes, due to clipping, giving a totally new sound.

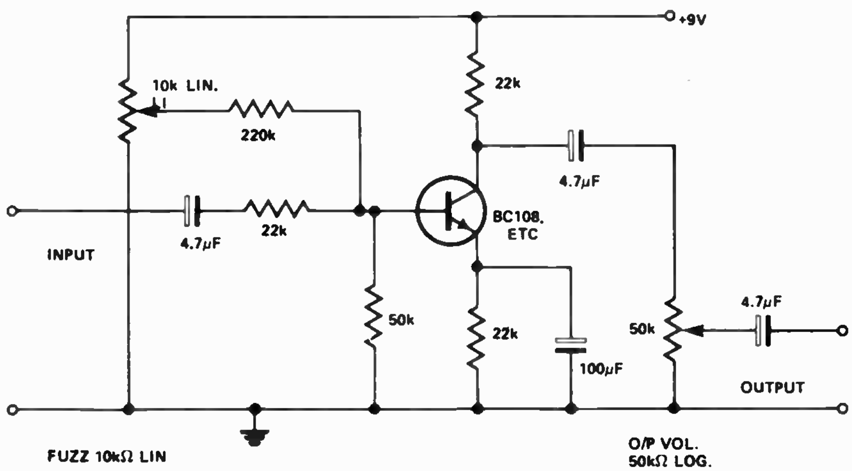


FUZZ BOX

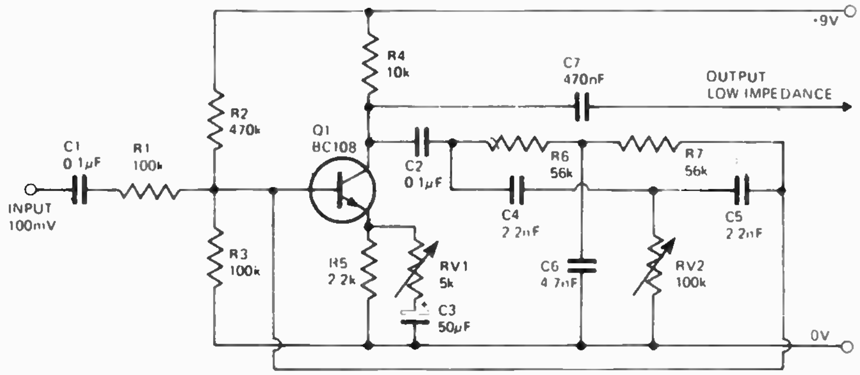
A quick look at a commercial one will show that fuzz-boxes are grossly overpriced for what they are. This general principle is that the input is split, and one part of it is distorted, then the two signals are mixed, variably, providing variable "fuzz". But why not cut costs again by simply varying the distortion of a one-transistor stage.

None of the components are particularly critical in value or quality, as distortion is the sole object!!

The transistor could be BC107-8-9, 2N2926, etc. A PP3 battery completes the "fuzz-box" which fits easily into a small plastic box with two jack sockets for the input and output and an on-off switch. The unit could be made easier to operate by reducing the value of the "fuzz" control and adding two series resistors.



SIMPLE WAA-WAA CIRCUIT



This circuit can be incorporated in guitar amplifiers or electronic organs.

A phase shift RC oscillator makes up the basic circuit. C4, C5, C6 and R6, R7, RV2 make up the components of the bridge that determine the operating frequency. Negative feedback is obtained by feeding part of the signal back to the base via C2.

The waa-waa effect is achieved as certain frequencies are amplified more than others.

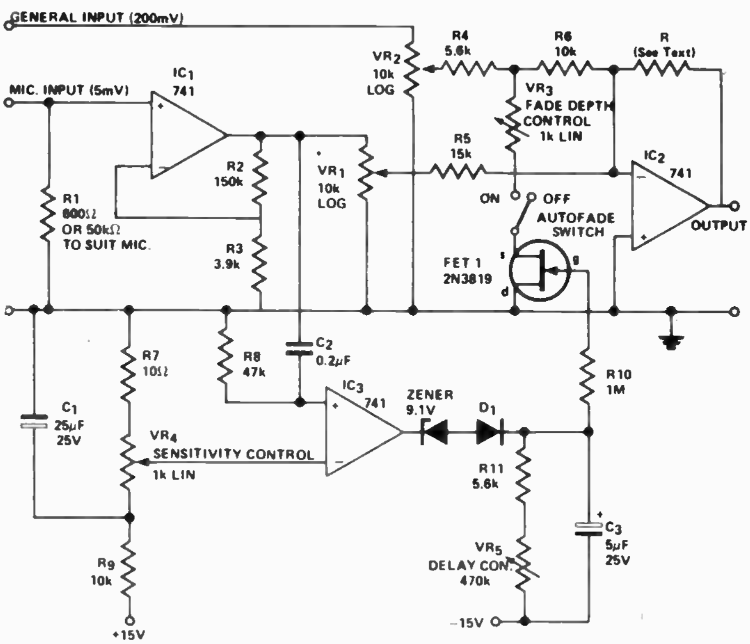
The transistor used is not critical however it should have a gain of more than 150. An NPN type such as the BC108 or BC109 is suitable.

The values of C4, C5, C6 are chosen so as to emphasise waa-waa effect on the higher audio frequencies. This gives the sound its brilliance. These values can be changed quite freely till the specific desired effect is achieved.

When adjusting the unit initially, RV1 is turned to its minimum value. RV2 is now adjusted to and fro till a point is found at which an audible whistle appears indicating oscillation. RV1 is then adjusted till the oscillation just disappears. RV2 is turned over its whole range and if at any point oscillation occurs again, RV1 is again advanced till it ceases.

It should be possible to set RV2 to any value over its range of adjustment without any oscillation being apparent; this should also be achieved with the minimum possible value of RV1. The unit is now ready for operation.

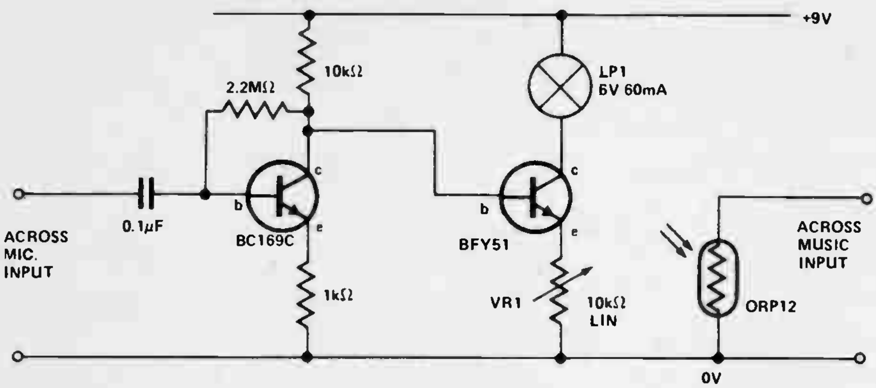
DISCO AUTOFADE



This is an autofade circuit for use in discotheques and the like. This autofade unit has advantages over VCG ICs which introduce distortion and noise (the 741s may be substituted by low noise op-amps) by using a FET to switch the signal gain characteristic. The principle may be easily adapted into existing mixers.

The microphone is amplified by IC1 and fed to the input mixer (IC2 the gain of which is set by R) and to the comparator IC3. If the input is large enough (larger than the voltage on the wiper of VR4) the output swings positive and charges C3 (in about 4mS). When the voltage across C3 is sufficient the FET is turned fully on and the fade depth control is grounded hence attenuating the signal. The FET turn off time is determined by R11, VRs and C3 and may hence be varied between 25mS and 2.5S.

SIMPLE DISCO AUTO FADE

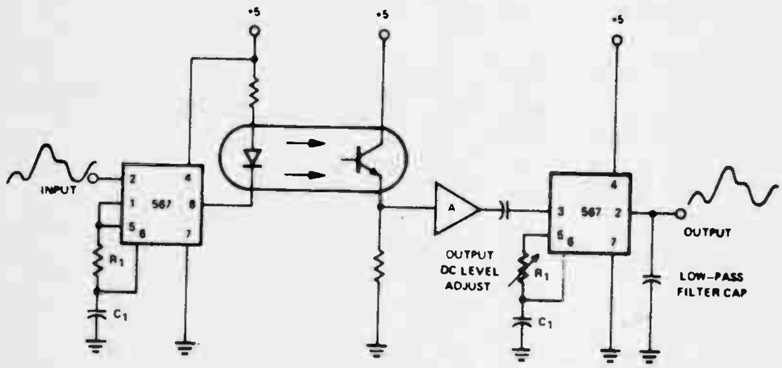


When a DJ has to make an announcement over a record, the normal procedure is to fade out the deck, fade in the microphone, and vice versa at the end. If this unit is used, however, the operator need only speak into the microphone, the deck being faded out automatically.

The lamp and LDR need to be taped together, preferably with black tape, to exclude light. VR1 is used to set the brightness of the lamp. With no signal on the input, VR1 is set for no attenuation of the music signal. When speaking at normal volume through the mic. the music should fade down until it can be heard quietly in the background. Some microphones may not produce enough signal to do this. If this happens, a simple pre-amp can be added to the input stage, as was done in the prototype.

OPTO-ISOLATED

INFORMATION TRANSFER



SOMETIMES it is necessary to transfer an analogue signal from one system to another without making any electrical connections.

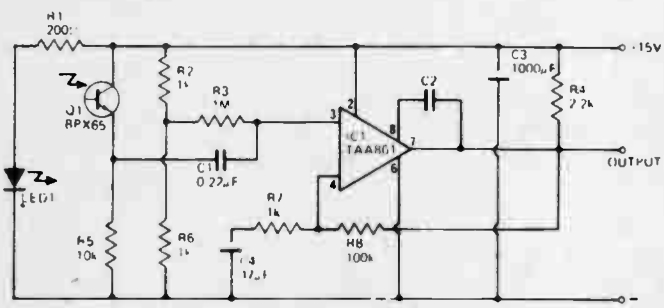
This can be done with two phase-locked-loops in an fm system using light as the transmission medium. Because of the high degree of electrical isolation obtained, low level signals can be transmitted without interference, even if there is a large potential difference between the sending and receiving circuits.

The circuit is shown above right.

Transmitter is an NE 567 phase-locked-loop IC operating as a voltage controlled oscillator which drives the LED section of an opto-coupler. The LED will flash at the operating frequency of the oscillator which is in turn dependent on the input signal level and the values of R, and C.

The output signal from the opto-coupler drives an amplifier which provides an output of sufficient amptitude (50 to 200 mV) to drive the receiving NE 567 phase-locked-loop. The receiver operates as an fm detector which demodulates the output of the opto-coupler to provide the original input signal. The inherent non-linearity of the transfer function in the two phase-locked-loops cancel one another out to give an extremely linear information transfer.

OPTICAL PULSE CONDITIONER



This circuit generates a fast rise time pulse each time the illumination of the BPX65 photo transistor by the LED is broken by a small object or rotating disc segment etc.

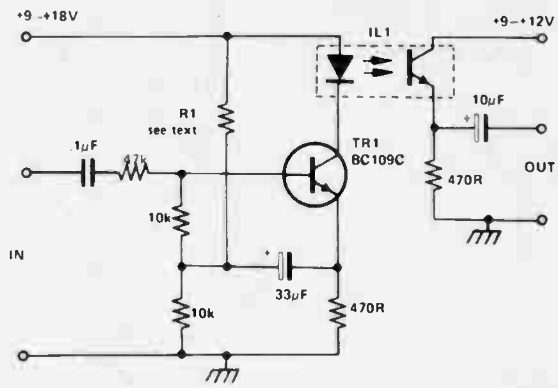
The operational amplifier, Siemens type TAA861, amplifies the signal from the photo-transistor and generates the fast rise time pulse the duration of which is determined by the value of C2: 5 microseconds when C2 = 47 pF, 1 microsecond for C2 equal to 16 pF and 0.4 microsecond when C2 equals 5 pF. The amplitude of the output pulse is 200 millivolt.

The circuit is thus ideal for generating pulses for an electronic counter. The maximum separation between the LED and the photo-transistor is around 20 mm.

SIGNAL PROCESSORS

OPTO-ISOLATED

LOW COST TELEVISION SOUND PICKOFF



The idea behind this circuit is the fact that in many modern TV set the sound IF and detector stages are up to Hi-Fi standard but the output stage and loudspeaker are not, so it makes sense to extract the signal before the output stage can distort it.

This circuit uses a cheap opto-isolator to couple the sound from the output of the sound detector of a television to a Hi-Fi system. The signal from the detector is buffered and amplified by TR1 and fed to the LED in the opto-isolator. This modulates the light emitted and this is detcted by the phototransistor in the opto-isolator.

There is no electrical connection between the two halves of the circuit and the isolator can withstand a voltage of 2.5kV between its input and output.

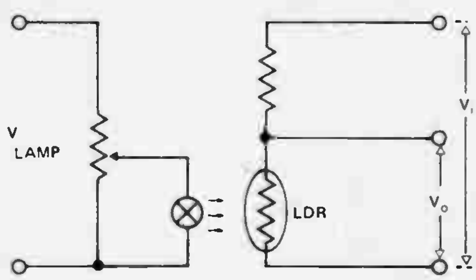
The power for the input side is taken from the television, a supply of about 6mA at any voltage between 9 and 18V will do, and to 36V if a BC107C is used for TR1. (R1 should be chosen to give a voltage of about 3V at the base of TR1).

If the TV uses a transistor IF strip, the power can normally be taken from it, otherwise a dropping resistor and zener can be used to take the power from the HT supply. The entire unit can be built for around £2.

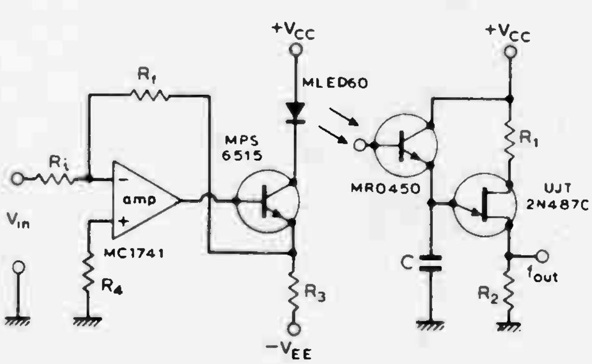
CRACKLE-FREE POTENTIOMETER OPTO-ISOLATOR

The resistance of a light dependent resistor (LDR) varies as the light falling upon it varies – hence its name! When wired into a divider network it forms an excellent crackle-free potentiometer.

It also provides excellent electrical isolation of the manual control – often a valuable feature where high voltage circuits must be isolated from low voltage circuits.



LIGHT-COUPLED VOLTAGE-TO- FREQUENCY CONVERTER

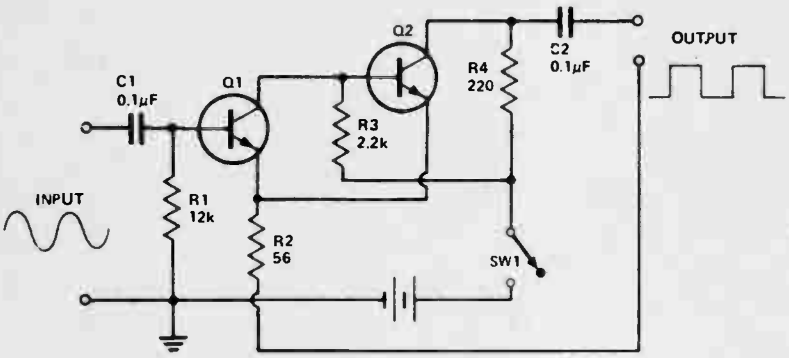


The output takes the form of a signal, the frequency of which is proportional to the applied voltage.

The functioning of the circuit is more or less self-evident. An operational amplifier drives a l.e.d. to provide a light output proportional to the applied input. The operational amplifier's scaling resistors are chosen to suit the application. At the receiving end the impedance of the photo-transistor alters the time constant in a conventional UJT relaxation oscillator circuit in sympathy with the level of incoming light beam to alter the output frequency.

CONVERTORS

SINE/SQUARE WAVE CONVERTER



Many audio generators only give a sinusoidal output. However a square-wave output is often useful too.

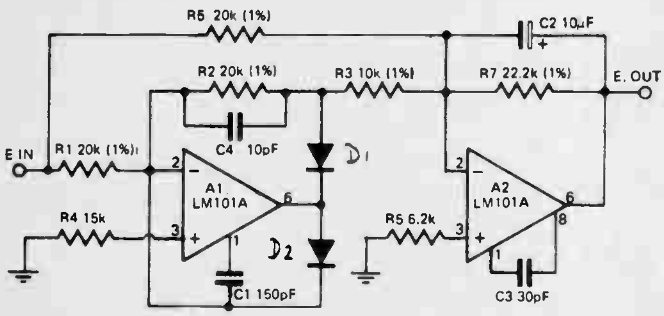
This circuit will square any sinusoidal input over the range of 20 Hz to 30 kHz with an output of about one volt, input signal should be about 400 mV.

The waveform obtained is of much better purity than obtained by a diode squaring circuit. The circuit is in fact suitable for use where square waves with a fast rise-time are required.

Transistors are germanium NPN types such as AC 127.

The power supply is 1.5 V and consumption is in the region of one to 2 mA.

PRECISION AC TO DC CONVERTER



The circuit shown provides better than 1% conversion accuracy of ac signals up to 100 kHz. The output is calibrated to read the rms value of the sine-wave input with less than 1% ripple at 20 Hz.

Amplifier A1 with diodes D1 and D2 forms a precision half-wave rectifier and the amplifier A2 sums the half-wave rectified signal and the input signal to provide a full-wave output. For negative input signals, the output of A1 is zero and no current flows through R3. Neglecting the effect of C2,

the output of A2 is $-\frac{R7}{R6}$ E in.

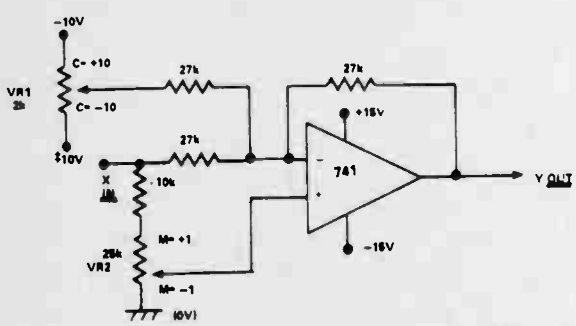
For positive input signals A2 sums the currents through R3 and R6

$$E\,out = R7\left(\frac{Ein}{R3} - \frac{Ein}{R6}\right)$$

If R3 is ½ R6, the output is $\frac{R7}{R6}$ Ein

Hence the output is always the absolute value of the input.

VOLTAGE PROCESSOR



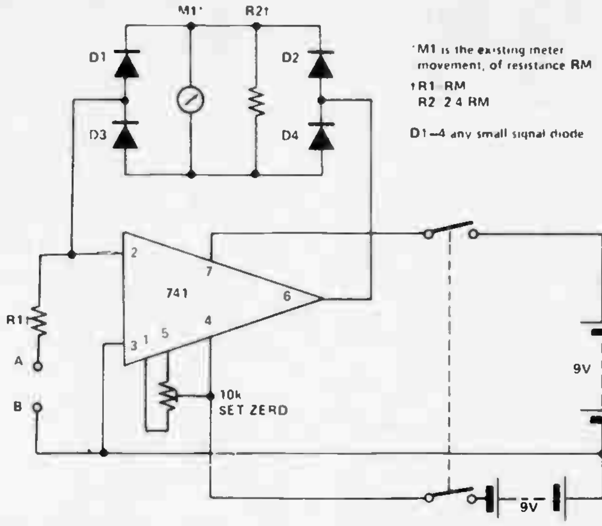
This circuit takes an input voltage x, and outputs a voltage y in accordance with the general linear equation y=mc+c. The offset is variable between +10V and −10V, by setting VR1; and gain m can be set anywhere between +1 and −1 (including zero) using VR2. There is no interaction between these two controls.

The circuit is very useful for processing electronic music synthesiser control voltages, and for coupling tuner to IF strips that have AGC systems operating in different senses. The input voltage should be derived from a low impedance (less than 1kΩ) source.

SIGNAL PROCESSORS

RECTIFIERS

UNIVERSAL METER RECTIFIER



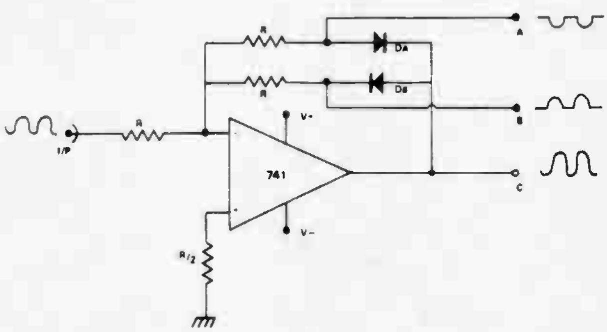
This circuit can be built for about £1 but could save pounds in multimeter repair costs.

The meter movement is removed from the meter circuit, it's place being filled by the input (terminals A and B) of the circuit shown. Pin 2 of the 741 remains at the same potential as pin 3, so the input signal "sees" R1 as it's load. However, the current which flows through R1 does not flow into pin 2, but through D1–D4, the original meter movement M1 and RMS correction resistors R2, to pin 6. Hence the circuit is current controlled, and so unaffected by the non-linearity of the rectifier, D1–D4.

R2 should only be in the circuit if it is desired to measure RMS AC values, all measurements are made on the DC ranges of the instrument.

R1 and R2 should be close tolerance types for accuracy; the circuit is accurate up to 100kHz.

A DOUBLE PRECISION-RECTIFIER CIRCUIT



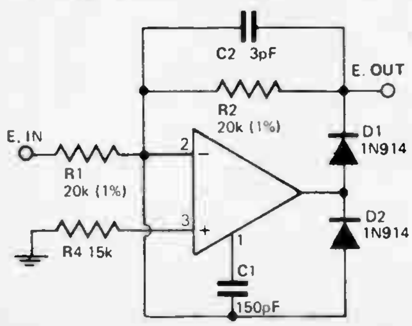
This circuit separates the positive and negative halves of an input waveform, and presents them (both phase-inverted) at separate output terminals. When the input swings positive, terminal A swings negative by an equal amount, terminal B remaining at zero voltage due to the reverse-biasing of D_B and the virtual-earth action of the op-amp. For negative inputs, terminal B swings positive by an equal amount, terminal A remaining at zero due to the reverse-biasing of D_A.

The insertion of a resistance of value R/2 in series with the non-inverting input gives partial cancellation of drift due to input offset-current changes. A suitable general-purpose value for R is 10K; note that the input-impedance is always R due to virtual-earth action.

A waveform appears at terminal C (at low impedance) that is the inverted input waveform with the addition of two diode-voltage-drops, in such a way that the central part of the waveform is "stretched". If the input is an audio source resulting distortion provides an interesting electronic music effect.

The circuit is particularily useful because positive and negative of a waveform can be operated on separately, and then combined in a simple virtual-earth mixer. For example, using two single-polarity log converters on the two outputs would provide bipolar logarithmic conversion upon recombination.

FAST HALF-WAVE RECTIFIER

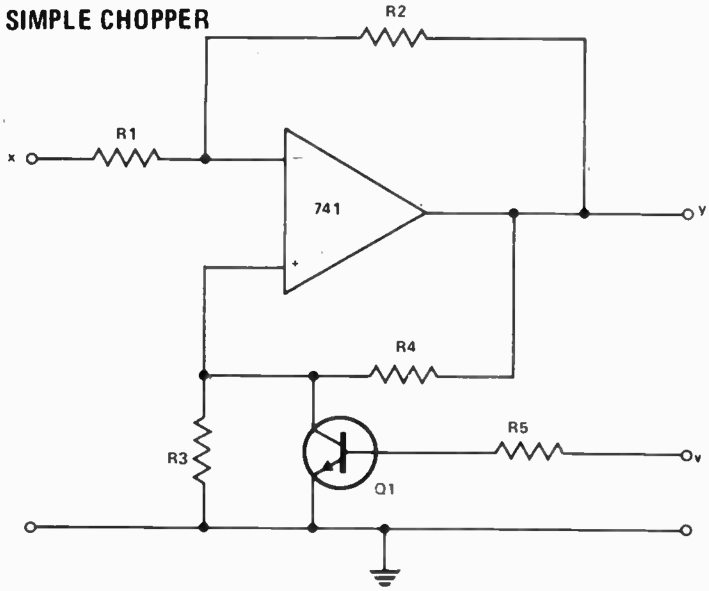


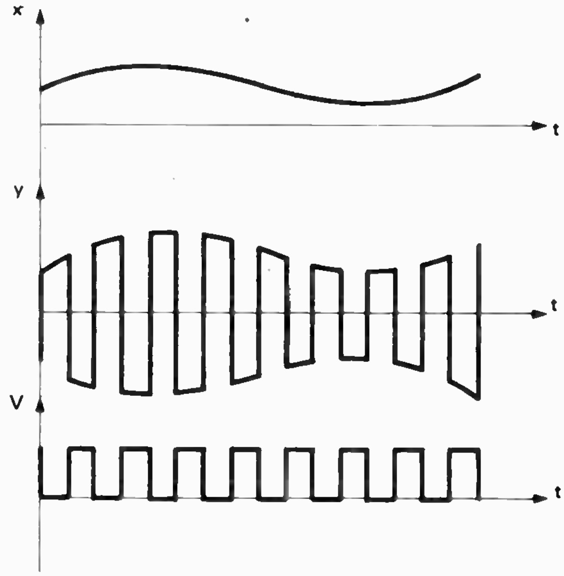
A precision half wave rectifier using an operational amplifier may be constructed as shown. This will have a rectification accuracy of 1% from dc to 100 kHz.

The input signal is applied through R1 to the summing node of an inverting operational amplifier. When the signal is negative, D1 is forward biased and develops an output signal across R2. As with any inverting amplifier the gain is R2/R1. When the signal goes positive, D1 is non-conducting and there is no output. The path through D2 reduces the negative output swing to –0.7V, and prevents the amplifier from saturating.

MISCELLANEA

SIMPLE CHOPPER



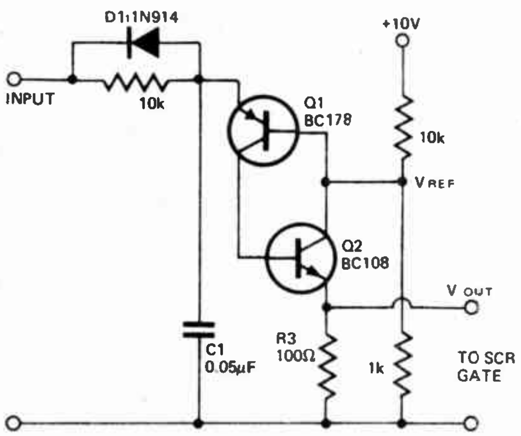


The circuit shown is a simple amplifier, the gain of which can be switched between two precisely controlled values by the application of a signal voltage: V. If V is such that Q1 is saturated, then the voltage gain of the circuit is simply $-\frac{R2}{R1}$. If the transistor is cut off, then the voltage gain becomes $(1+\frac{R3}{R4})/(\frac{R3}{R4} - \frac{R1}{R2})$.

One obvious application of the circuit is if the resistors are adjusted such that the two gains are equal in magnitude, but opposite in sign, (e.g. R1 = 20k, R2 = 10k, R3 = 50k, R4 = 10k gives voltage gains of +2 and −2) Then the circuit could be used as a chopper for the input of a DC amplifier.

The value of R5 is largely arbitrary, depending on the magnitude of the chopper signal, V. Its sole purpose is to prevent excess current being drawn by the base of Q1. Suitable components for the op amp and transistor are the 741 and BC182L.

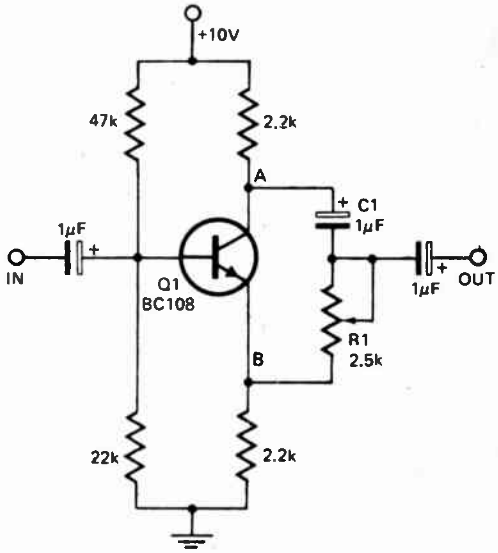
NOISE REJECTING SCR TRIGGER



When switching inductive loads, unreliable triggering is sometimes encountered due to feedback of switching transients.

The circuit shown overcomes this problem by using an integrator together with a voltage comparator to eliminate transients. Data pulses should be of 8 volt amplitude and 0.5 millisecond duration. Discrimination against noise pulses will depend on their energy content. For example a 70 volt 10 microsecond wide pulse will not cause triggering, but a 100 microsecond pulse must not exceed 20 volts amplitude.

SINGLE TRANSISTOR PHASE SHIFTER

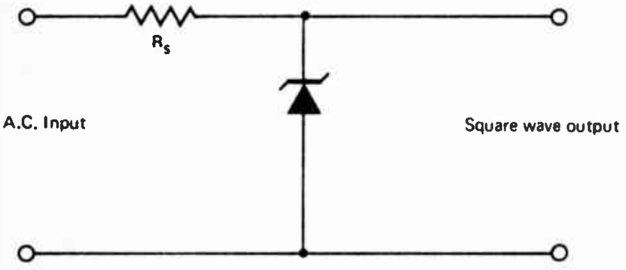


The circuit shown provides a simple means of obtaining phase shifts between zero and 170°. The transistor operates merely as a phase splitter, the output at point A being 180° out of phase with the input. Point B is in phase with the input phase. Adjusting R1 provides the sum of various proportions of these and hence a continuously variable phase shift is provided.

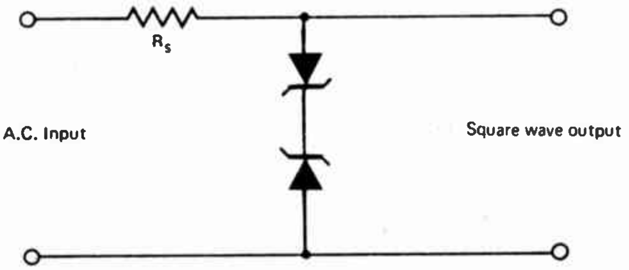
The circuit shown operates well in the range 600Hz to 4kHz.

SQUAREWAVE

SIMPLE SQUARE–WAVE GENERATOR

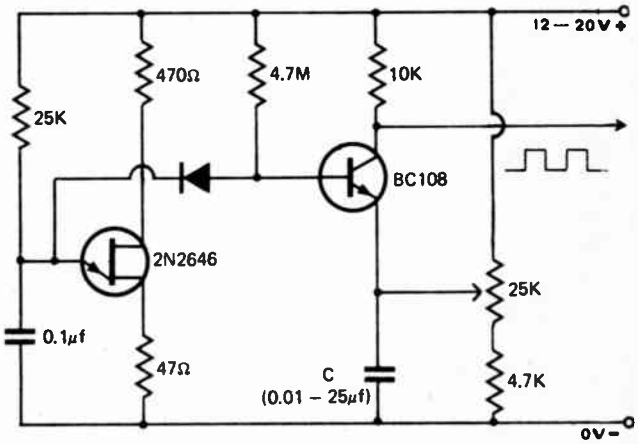


A zener diode clips one half-cycle of the input sine-wave in this simple square-wave generator. The zener voltage must be a small fraction of the input voltage for acceptable results.



If a larger output is required, opposed zener diodes may be used to clip both halves of the input sine-wave.

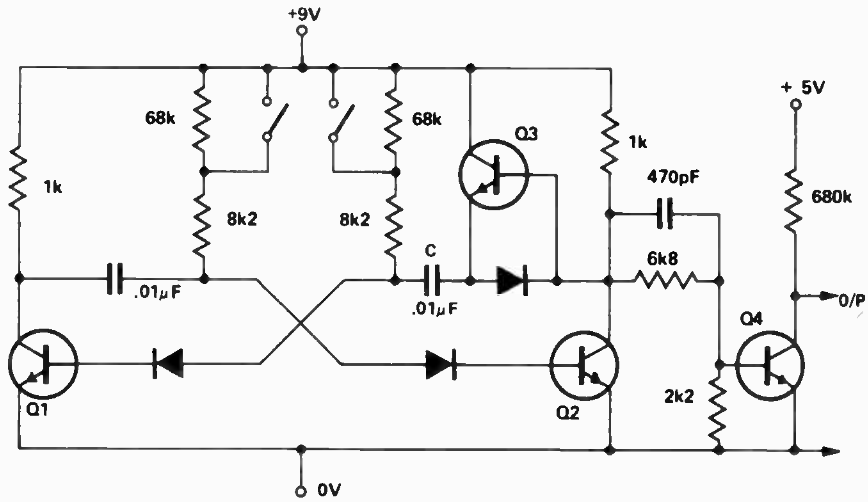
VARIABLE DUTY SQUAREWAVE GENERATOR



A variable duty cycle squarewave can be obtained from this unijunction circuit. The light loading imposed on the emitter timing circuit preserves frequency stability.

Faster rise and fall times can be attained by including bypass capacitor C. The value of this capacitor should be chosen to suit the pulse shape required.

FAST-EDGE SQUARE-WAVE GENERATOR

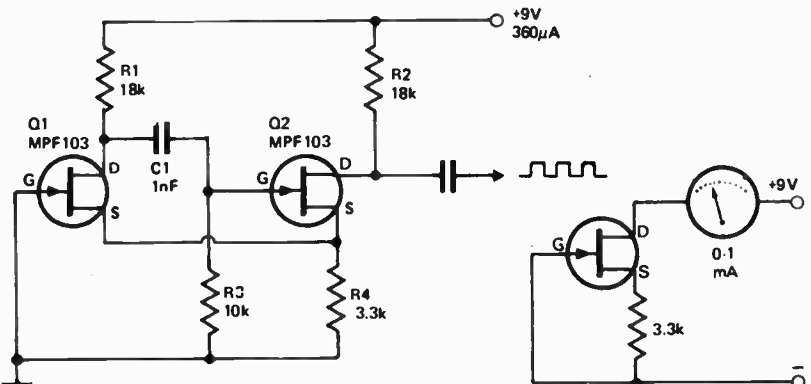


The circuit shown above generates a clean square-wave with very fast rise and fall edges; such a signal is essential for some applications such as the testing of amplifier transient responses, and the reliable driving of TTL.

The multivibrator circuit shown is unusual in that it produces a waveform with fast risetime as well as fast fall time. The standard astable multivibrator has a slow risetime as capacitor C is charged relatively slowly through the collector load of Q2; in the modified circuit C is charged very quickly through Q3. Diode D ensures that Q3 is only turned on when Q2 is off. The final stage (Q4) increases rise and fall speed still further; at the output the rise time is 100 nanosec. and the fall time 300 nanosec. When the output is used for driving TTL the collector load must be returned to a voltage no greater than +5V.

The two-pole switch gives an output at about 1kHz when open and about 10kHz when closed; these two frequencies being standard for checking the stability of audio amplifiers.

FET SQUARE WAVE GENERATOR



FET SQUARE WAVE GENERATOR

Field effect transistors lend themselves readily for use in astable multivibrator circuits.

The output square wave yields an amplitude close to the power supply voltage, and battery drain is low.

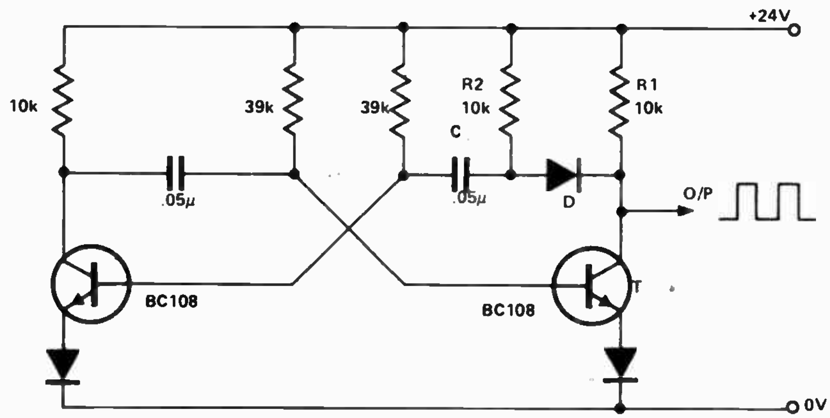
In this circuit the battery supply is 9 V. Drain is a minimal 360 μA. The waveform shows very good symmetry and this is achieved by matching the FETs by means of the circuit (b); transistors are matched up for equal drain currents.

Frequency of operation is set by R3 and C1. The values in the circuit give a frequency in the region of 15 kHz.

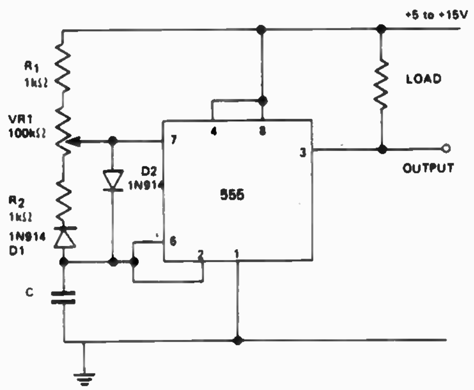
IMPROVED MULTIVIBRATOR

Conventional astable multivibrators suffer from the disadvantage that they do not produce a good square-wave output; the leading edge of the waveform has a very slow rise since the collector resistor R1 is tied to a slowly charging capacitor C when the transistor T turns off.

This circuit prevents this effect and thus generates a clean square-wave with 400nS rise-times and 100nS fall-times. This is because diode D turns off when the output begins to rise in voltage, and a fast rise is then possible. C is charged by a separate resistor R2, and apart from this multivibrator action is normal. The components shown give an operating frequency of about 700Hz.



VARIABLE DUTY CYCLE OSCILLATOR



The circuit shown enables a rectangular wave output to be obtained with a duty cycle which can be varied over a wide range by the setting of the potentiometer VR1.

The well known 555 integrated circuit is used as a monostable device. The capacitor C charges from the positive line through R1, part of VR1 and D2. When the voltage across this capacitor rises to two-thirds of the power supply voltage, the state of the 555 is switched so that the capacitor C discharges through D1, R2 and the other parts of VR1 into pin 7 of the 555 device. The diodes therefore enable the charging and discharging paths to be separated; the effective value of the charging and discharging resistors can therefore be set independently of one another.

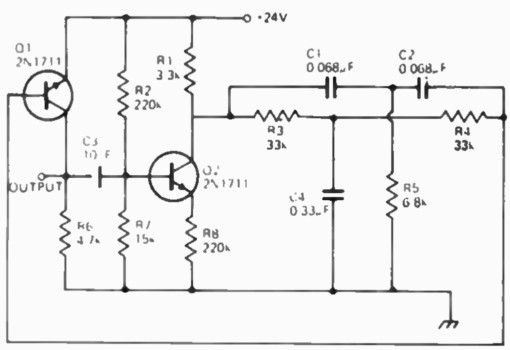
When the slider of VR1 is near to R2, the discharging time is very short and the output spends only a small fraction of its time in the low voltage state. In this case short negative pulses will be obtained at the output. Similarly, short positive pulses are obtained when the slider of VR1 is near to R1.

One great advantage of this type of circuit is that the frequency is almost independent of the setting of VR1 over most of its travel. If VR1 is in the centre of its track, the duty cycle will be approximately 1:1. The frequency is almost independent of the output current up to the recommended maximum of 200mA.

The value of C is chosen according to the frequency required. The latter can be as great as 100kHz or very low indeed - one cycle in a few minutes.

SINE WAVE

STABLE RC OSCILLATOR



The frequency of oscillation of this circuit is determined by a twin T network and is stable to within 0.05% for ± 10% supply variation.

A temperature stability of 0.2% from −20°C to 80°C will be obtained if polycarbonate capacitors are used throughout.

With the values show the circuit oscillates at 60 Hz. It will operate at very low frequencies for which the values required are given by the formula:—

$$F = \frac{0.159}{R3C1}$$

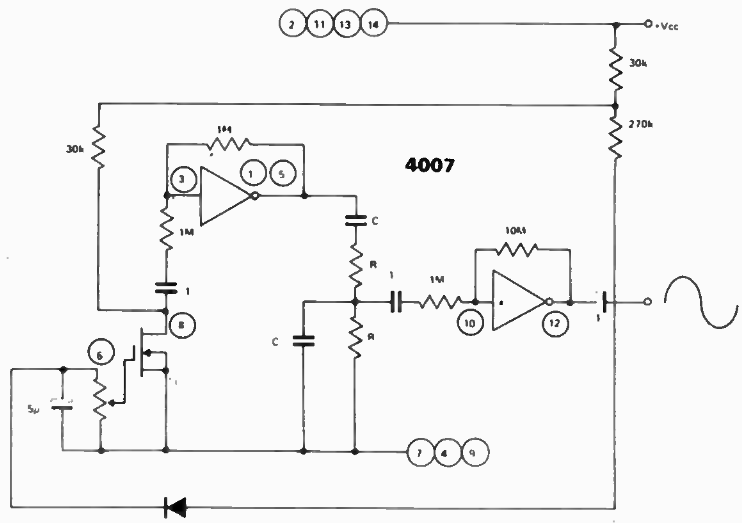
where F = frequency in hertz and R_3 is in ohms C1 in farads C1 = C2 = ½ C4 and R3 = R4 = 2R5.

CHEAPO OSCILLATOR

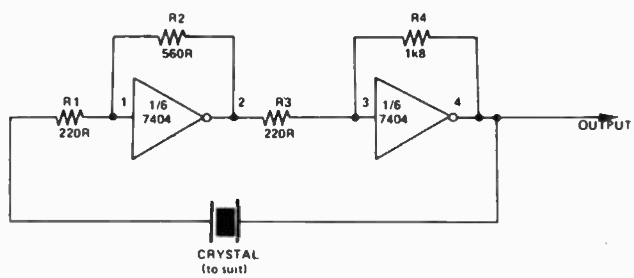
The two inverters give a gain of 100; the MOSFET reducing this to a necessary level to just sustain oscillation.

Supply voltage can be between 5V and 15V. (10V is recommended). Do not try to obtain more than 1mA from the circuit. At 5V supply a buffer is required.

The supply should be ripple free, as any ripple will be passed to the output.



A SIMPLE TTL CRYSTAL OSCILLATOR



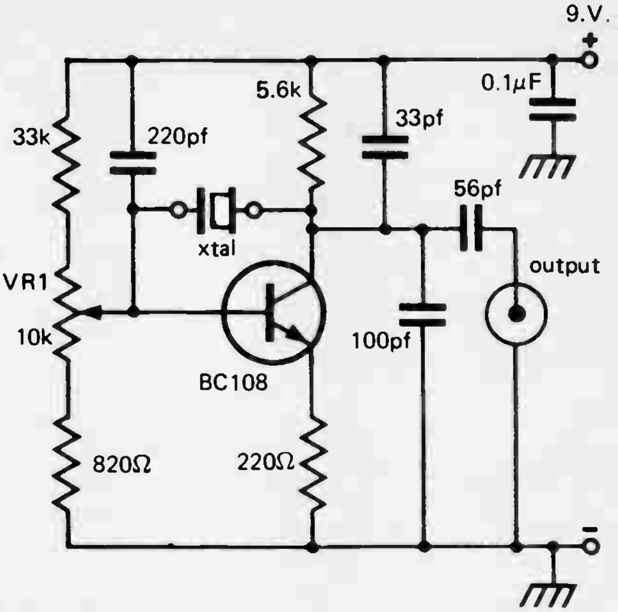
This is possibly the simplest and cheapest crystal oscillator it is possible to make, comprising one third of a 7404, four resistors and a crystal. It was originally designed for a battery operated timer.

The two inverters are biased into their linear regions by R1 to R4, and the crystal provides the feedback. Oscillation can only occur at the crystals fundamental frequency.

Note that the oscillation occurs at fundamental frequency, whereas many crystals (particularly higher frequency ones) are stamped with the overtone frequencies. These will operate at the fundamental and not the marked value.

Trimming capacitors for fine adjustment can be added if required.

UNCRITICAL CRYSTALS

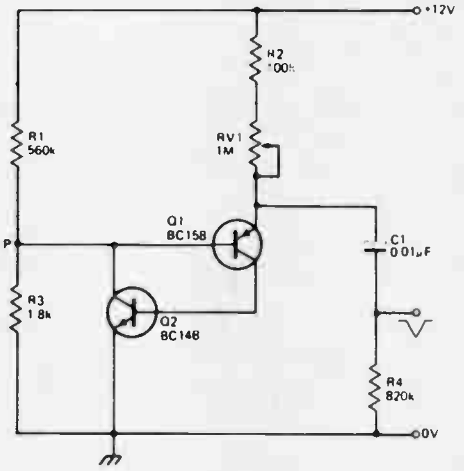


Surplus crystals are very useful in a variety of applications. But it is sometimes difficult to obtain reliable oscillation in many standard circuits.

This circuit overcomes this problem. It will provide reliable oscillation and provides an output close to one volt peak-to-peak. Power consumption is around 1mA from a nine volt supply.

PULSE & SAWTOOTH

PULSE GENERATOR

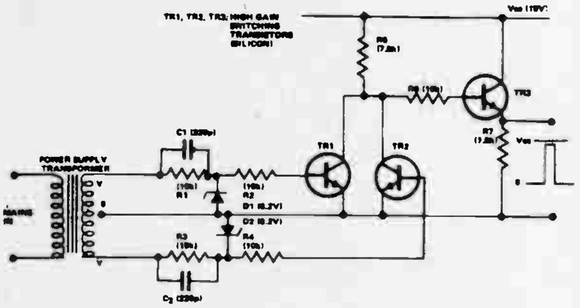


This simple pulse generator produces 100 nanosecond negative pulses of 8 volts amplitude.

At switch on, Q1 and Q2 are off, and C1 charges through R2, RV1 and R4. When the potential across C1 becomes 0.7 volts above point P, the transistors saturate, discharging C1 through R4. A negative pulse is thus generated across R4.

When the capacitor is fully discharged the transistors turn off and the cycle repeats. Pulse spacing may be adjusted between 1.5 and 15 milliseconds by RV1 and the pulse duration may be altered by using a different value for C1.

ZERO–CROSSING PULSE GENERATOR



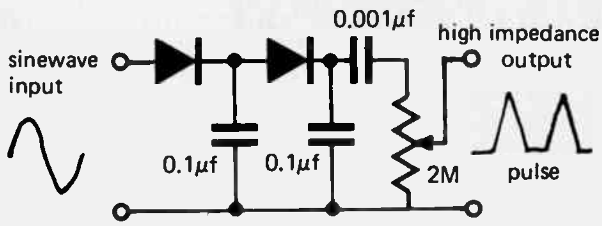
The circuit was originally used to provide a narrow zero-crossing pulse for switching on triacs in a mains sound to light converter. A narrow zero-crossing pulse is required to reduce RFI (caused by switching large current) by switching when there is no voltage across the load. As a bonus the life of the bulb is substantially improved.

The circuit operation is as follows: TR1 and TR2 form a NOR gate, the output of which is high only when both inputs to transistor bases are low (i.e. at the zero-crossing points of the mains cycle). TR3 as an output buffer. R1-4 and D1-2 are included to provide voltage and current protection for the bases of TR1 and TR2. Note for large values of transformer voltage 'V' the pulse width tends to zero.

SIGNAL GENERATORS

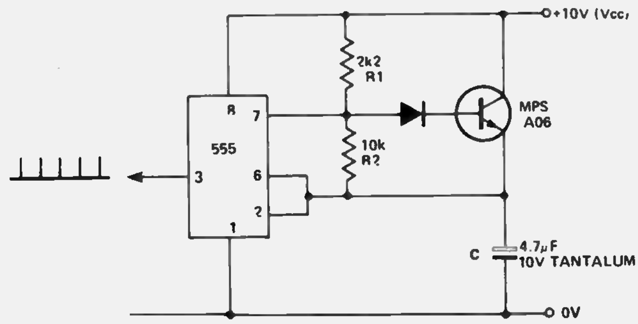
PULSE & SAWTOOTH

SIMPLE PULSE GENERATOR



This circuit is useful in adjusting noise blankers for producing time marker pulses for a CRO or chart recorder.
The high impedance output is adjustable by the 2 Megohm potentiometer. Peak inverse rating of the diode should be high enough for input voltage used.

NEEOLE PULSE GENERATOR

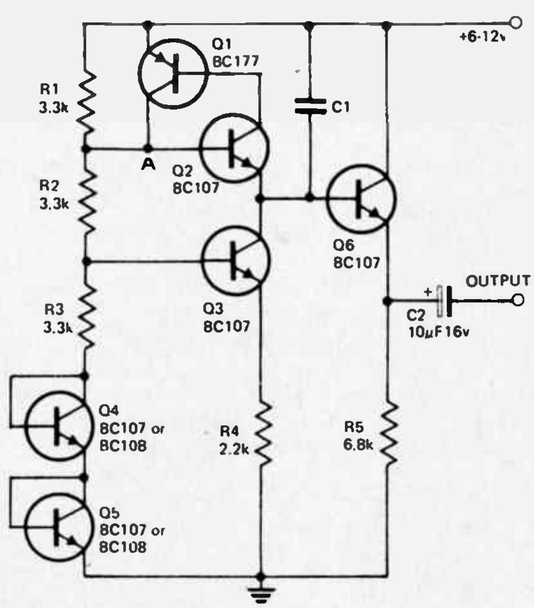


This circuit generates very short positive pulses at long time intervals – useful for strobing sample-and-hold circuits etc.

In the discharge part of the cycle, capacitor C discharges slowly through R2, as reset pin falls below 1/3 Vcc, the bistable (internal) switches, and the short between pin 7 and earth is removed. The transistor is then turned hard on by current flowing through R1, and C charges very rapidly – when the voltage across it exceeds 2/3 Vcc the 555 switches again, and the discharge cycle begins again.

The "charge" portion of the cycle is very short, about 120µS, while the discharge time depends entirely on the value of R2. For example, with R2= 2M2, a 120µS pulse is produced about every 10 seconds; a mark/space ratio of 100,000 to one!

STABLE HIGH-LINEARITY SAW-TOOTH GENERATOR



In this circuit two transistors Q1 and Q2 are connected so that they operate as a unijunction transistor.

Capacitor C1 is charged by a constant current source made up of transistors Q3, Q4 and Q5. This ensures a linear voltage rise. As soon as this voltage rises to the value as found at a point A of the circuit (less the base emitter voltage of Q2), the transistors Q1 and Q2 become conducting and C1 is discharged very rapidly.

The voltage rise across C1 is very linear and is applied to an emitter follower stage. This ensures that there is no shunting effect by the load circuit which could be detrimental to the linearity of the output waveform.

The potential divider chain (R1-R2-R3) contains two transistors strapped as diodes and ensures good stability of the oscillator for variations in both supply voltage and temperature.

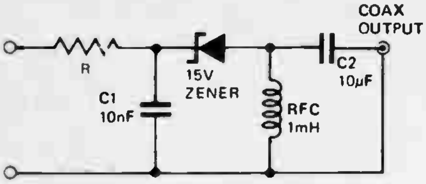
Frequency of operation can be calculated from the formula:

$$f = 1/R4.\ C1$$

where C and R are expressed in ohms and farads and f in Hz.

NOISE

ZENER DIODE NOISE GENERATOR



Zener diodes generate quite intense internal noise. This noise level is in the region of 30 dB above the inherent thermal noise and extends out to 150 MHz or so.

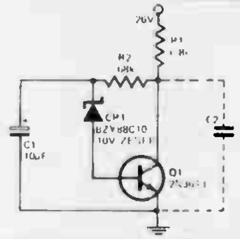
The circuit shown here may be used for adjustment of VHF converters and other receiving equipment.

It is energised by a dc source. The resistance R is adjusted so that between 6 and 8 mA flows through the circuit.

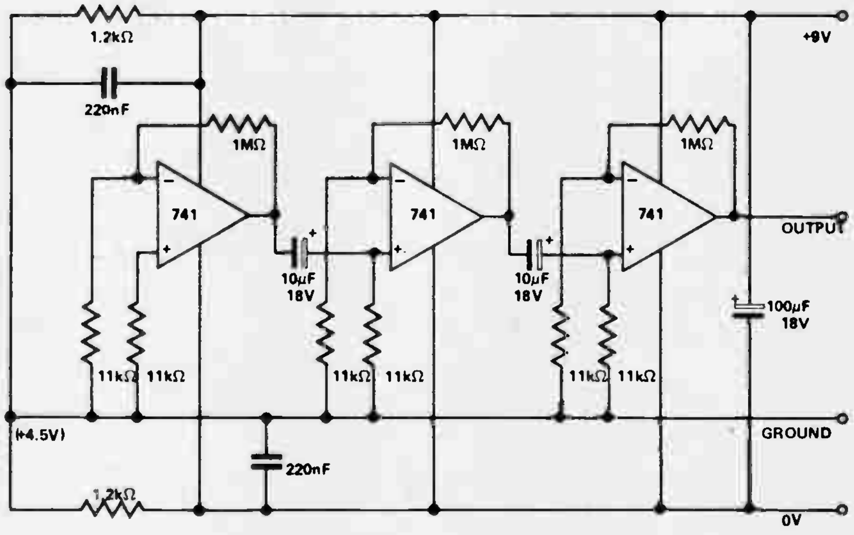
Capacitor C2 should preferably be ceramic. The signal output should be fed via a coaxial cable. ●

NOISE GENERATOR

In this circuit the Zener diode, as well as providing a source of noise, stabilizes the amplifier transistor collector operating point. The gain of the transistor is about 75 and the noise output of the circuit is about 15 volts. Capacitor C2 may be added to filter out high frequency noise – in which case the output drops. For example with C2=0.1μF, the output falls to 0.5 volt.



PINK NOISE GENERATOR



A useful method of frequency response testing of audio equipment is to excite the system with a random noise electrical signal and then analyse the output into its various frequency components using narrow band filters. The ideal noise signal is one having unit power per unit bandwidth (this is termed "white" noise). The system will be effectively driven by all frequencies at once. The frequency spectrum of the output will then be the required frequency response of the system.

However, the most common form of frequency analyzer uses filters with a constant percentage bandwidth (often one-third-octave). Thus an analysis of true white noise would give a frequency spectrum rising at 3dB per octave, because the power in a noise signal is directly proportional to the measuring bandwidth.

Pink noise was developed to give a flat frequency spectrum into such filters. The output of a pink noise generator falls at 3dB per octave. Normally they are made by installing a –3dB per octave filter after a white noise source. By a fortunate coincidence, the electrical noise from a 741 operational amplifier, when connected as shown, does have a pink noise frequency spectrum.

The circuit is simply three high gain operational amplifier stages cascaded. The first stage generates internal electrical noise which is amplified to approximately one volt r.m.s. by the two following stages. The circuit must be laid out as closely as possible to the schematic diagram, and carefully screened, because the input stage is very sensitive to extraneous signals and could pick up hum or oscillate due to capacitive coupling with the output. The prototype is run from a PP9 battery, mounted inside the case, to further reduce any possibility of hum pick-up.

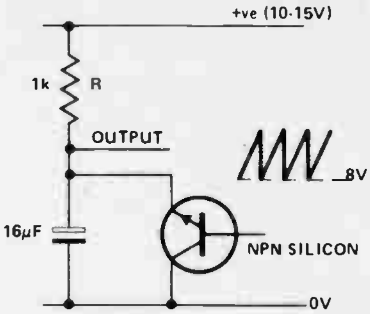
The output does have a slight roll-off from a pink noise characteristic, starting at about 100Hz. This is caused by the a.c. coupling between stages in the circuit which is necessary to prevent d.c. fluctuations from saturating the output. Also there is a roll-off at the high frequency end caused by the internal compensation in the operational amplifiers. There is, nevertheless, useable output up to 25kHz.

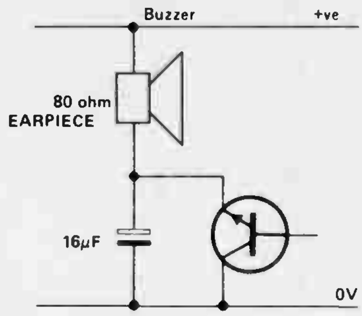
My apologises are due to the manufacturers concerned for using their devices in this unorthodox fashion. It may be useful to point out that the cheaper brands of 741 op-amp are likely to have higher noise levels and thus be more useful for this particular purpose.

SIGNAL GENERATORS

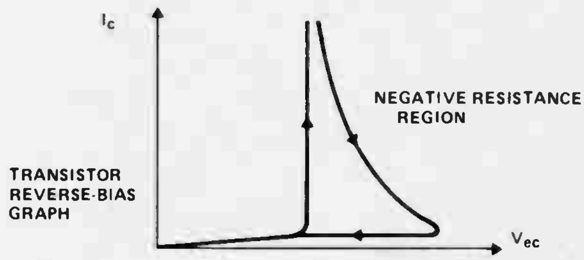
SPECIAL

SIMPLE OSCILLATOR

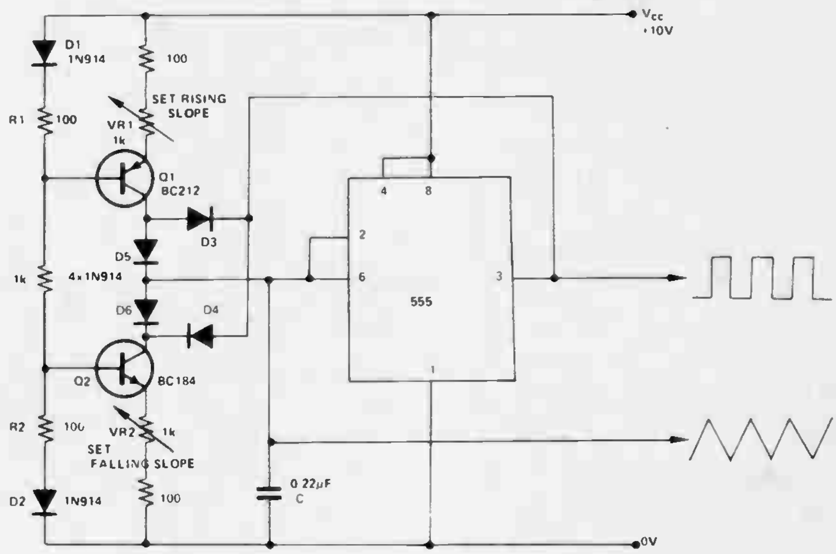




The negative resistance region of a reverse-biased silicon transistor can be used in a relaxation oscillator circuit. Its advantage is that a surplus transistor is used instead of a UJT (which is more expensive) and it does provide a minimum of components. The frequency is governed by the time constant RC, the power supply voltage and the size of the negative-resistance region. The latter also governs the signal amplitude, so various transistors (from a surplus batch) should be tried for best results. The output is a sawtooth waveform with a mean dc level around 8V. Replacing the resistor with an 80 ohm earpiece makes an effective buzzer.



TRIANGLE GENERATOR WITH INDEPENDENT SLOPE SETTING

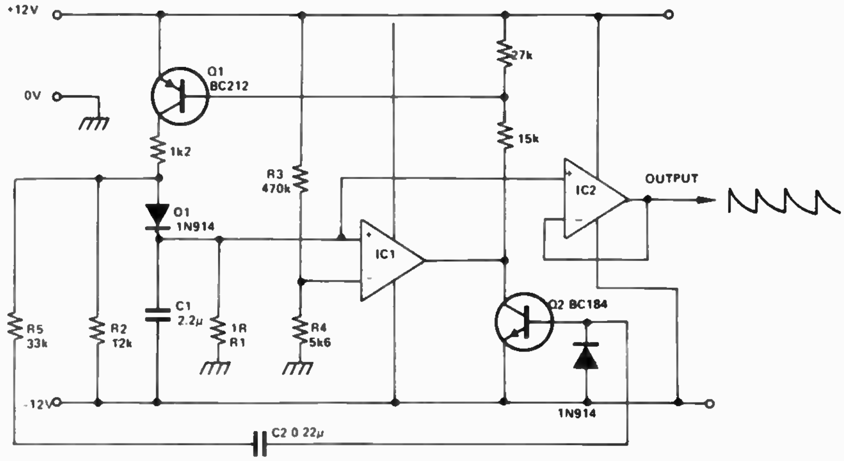


This free-running oscillator circuit generates a triangle waveform, the rising and falling slopes of which may be set by completely independent controls. Simultaneously the 555 output (pin 3) provides a rectangular waveform at low impedance that is synchronised with the triangle waveform.

Assuming the 555 output is low, the output of constant-current source Q1 is shorted to earth via diode D3, and diode D5 is reverse biased. During this time current source Q2 linearly discharges the timing capacitor C through D6, D4 being reverse biased. Eventually the voltage on C falls to $1/3V_{CC}$ (set by the internal potential divider that biases the two comparators in the 555) and the 555 output goes high. Now the output of Q2 is shorted away from the timing capacitor, and Q1 is allowd to linearly charge it up; when the capacitor voltage reaches $2/3V_{CC}$ the 555 output goes low again, and the cycle repeats.

The biasing networks R1D1 and R2D2 compensate for the changes in V_{CC}.

EXPONENTIAL WAVEFORM GENERATOR



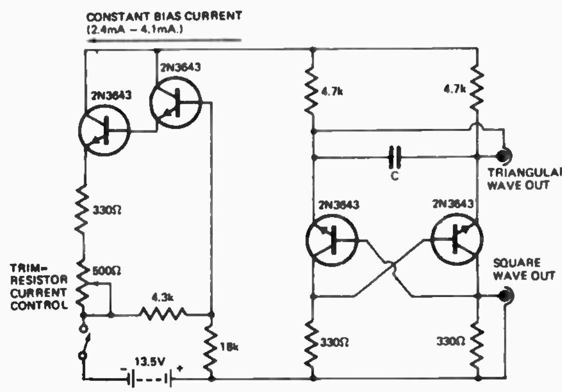
IC1 . An open-collector O/P comparator, eg LM339. IC2 .. 741 or similar.

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

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

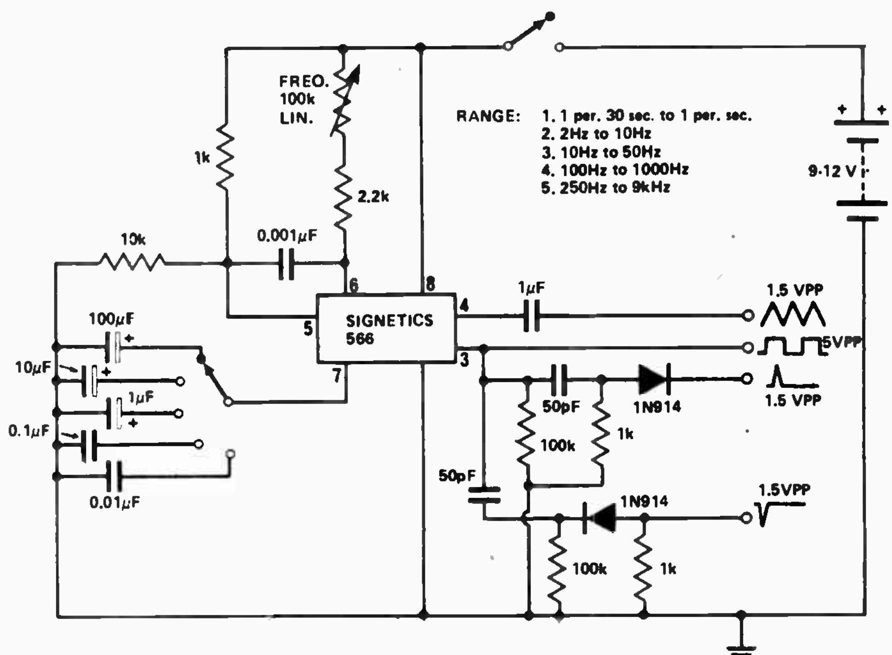
WIDE RANGE MULTIVIBRATOR



In the circuit shown the multi-vibrator section is driven by a constant current generator. This causes the square wave across C to be flat and the triangular wave across resistor R to be linear. When the constant current is varied, the repetition rate of the multi is varied by 70% with the current control, and from 5Hz with C equal to 100 microfarad, to over 2.5mHz with C equal to 330pF.

Voltage to frequency conversion may be performed by injecting an analogue voltage into the base of the first current regulator transistor.

VARIABLE FREQUENCY MULTIPLE WAVEFORM GENERATOR



Signetics 566 IC chip lends itself ideally as a test generator by utilising its internal voltage controlled oscillator (VCO).

The circuit will deliver separate outputs giving triangular and square waves and both positive and negative going spikes.

The square wave amplitude is 5 V pk-pk, all other waveforms are 1.5 V pk-pk.

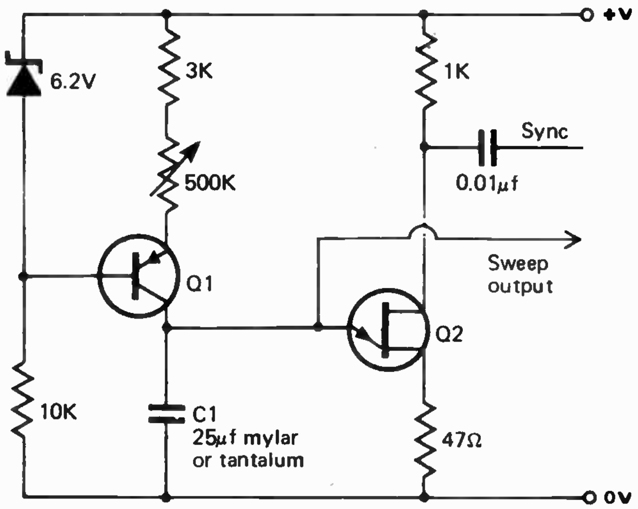
Frequency is determined by the value of the capacitor connected to pin 7.

It is preferable to use tantalum capacitors rather than electrolytics.

The outputs are designed to operate into high impedance loads. A transistor buffer stage is needed to match to low input impedance devices.

SPECIAL

LINEAR SWEEP GENERATOR

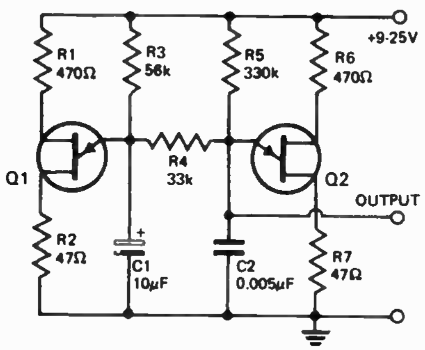


This circuit provides a linear time-base with a sweep time adjustable from a few milliseconds to over one minute.

The constant-current effect of emitter follower Q1 causes C1 to be charged at a constant rate.

The increasing voltage across C1 will be essentially linear (displacement error is less than 1%). The sweep is terminated when the increasing capacitor voltage reaches the peak valley point of unijunction Q2, when capacitor C1 will discharge through the current limiting resistor R1.

STEP FREQUENCY OSCILLATOR



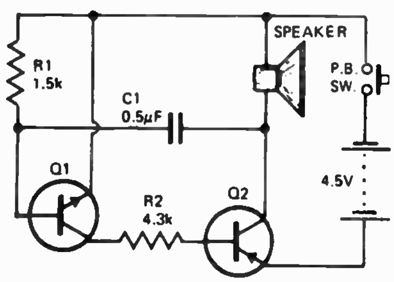
This circuit will produce a continuous sequence of increasing frequencies (in steps) until the highest is reached. The system then resets itself and starts again.

Two unijunction relaxation oscillators are cross-coupled together. On switching on capacitors C1 and C2 start to charge up through R3 and R5. The time constant C2-R5 is shorter; Q2 fires first and discharges C2. As C2 charges up again it will draw current through R5 and R3-R4. This will shorten the Q2 time constant, and in progressive cycles, as C1 charges up slowly, the Q2 time constant will keep shortening till Q1 fires, at which stage C1 will discharge and the whole cycle begins again.

Various sound effects can be obtained by varying R3, R4, C1 and C2.

SPECIAL EFFECTS

TRANSISTORISED BEEPER



This circuit consists of an assymetric multivibrator activated by a pushbutton. The loudspeaker is a transistor radio type with a voice coil impedance of about 25 to 40 ohms. Earpieces up to 500 ohms can be used for lower power output. R1 varies frequency over the audio range.

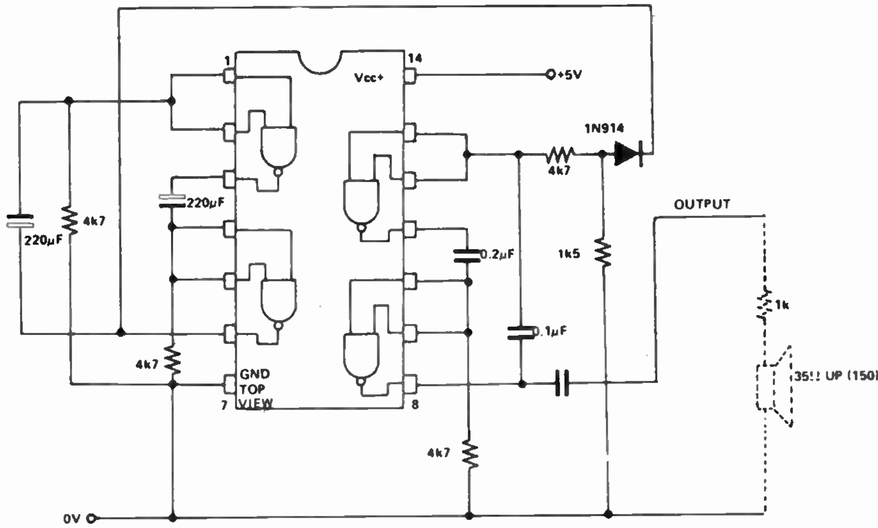
Transistor Q1 can be any LF small signal type (NPN), either germanium or silicon. (AC127, BC107, BC108 etc). Q2 is a small signal germanium type of up to 1A collector current. (AC128, AC132, AC188 etc). The battery size should be determined by the drain current of Q2.

7400 SIREN

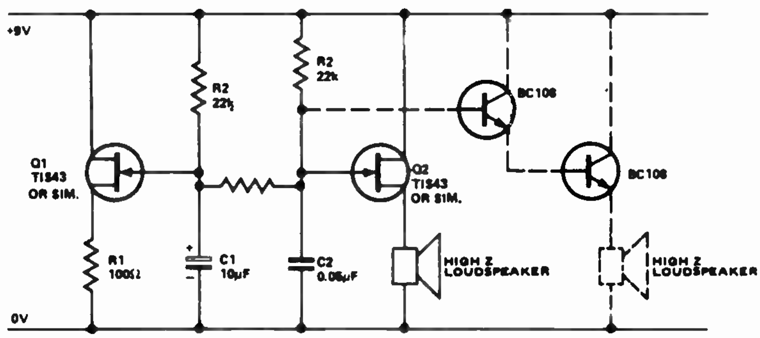
The circuit uses the NAND gates as Hex invertors. Two of these are used for the oscillator, and two as the control. If the two tone speed needs to be altered, the 220μF capacitors can be changed (larger for slower operation).

When the control oscillator output is at logic "0" it effectively shorts out the 1k5 resistor, giving a low note. When the control oscillator output goes to logic "1" the diode blocks the output. So the "1" condition gives high note, (as if the control oscillator was disconnected).

If the frequency of the oscillator is to be changed, the capacitors can be varied and the value of R1 can be increased. To change frequency range between the two notes alter the 1k5 resistor. (Note, this changes the oscillator frequency altogether). DTL-946s could be used instead of the 7400s but they give a higher pitched note.



SIMPLE SIREN



The circuit consists of two unijunction relaxation oscillators, Q1 for low frequency and Q2 for audio frequency. R3 couples the slow rising voltage across C1, determined by the time constant C1 and R2, to the audio frequency across C2, determined by the time constant of C2 and R4. The effect is that the audio frequency generated by Q2 rises in pitch as the slow rising voltage across C1 is applied, via R3 to the time constant C2/R4.

This type of sound carries much further than a continuous note from a single oscillator. Extra amplification can be achieved, by adding two transistors in a super-alpha arrangement as shown dotted. R5 should be replaced by a 100 ohm ¼W resistor.

Connected to a pressure mat (from across C2), this unit would make an excellent baby snatch alarm for prams.

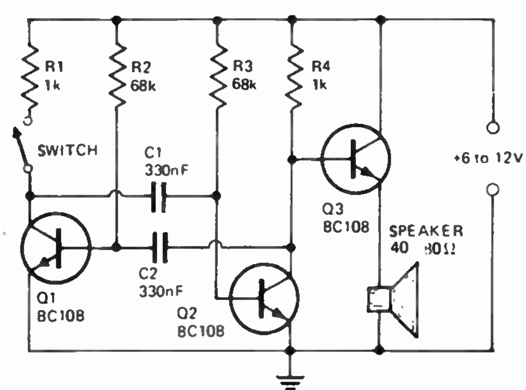
ELECTRONIC SHIP SIREN

This circuit will give a sound like a ship's siren. It can be used with the low power output source for model ships if fed into a more powerful amplifier/speaker, as an alarm tone.

The circuit consists of a multivibrator (Q1 & Q2), and a low power output stage Q3. The speaker should have an impedance in the region of 40 to 80 ohms.

C1 and C2 determine the pitch of the siren and the values specified will provide a tone of about 300 Hz. Quiescent current is negligible.

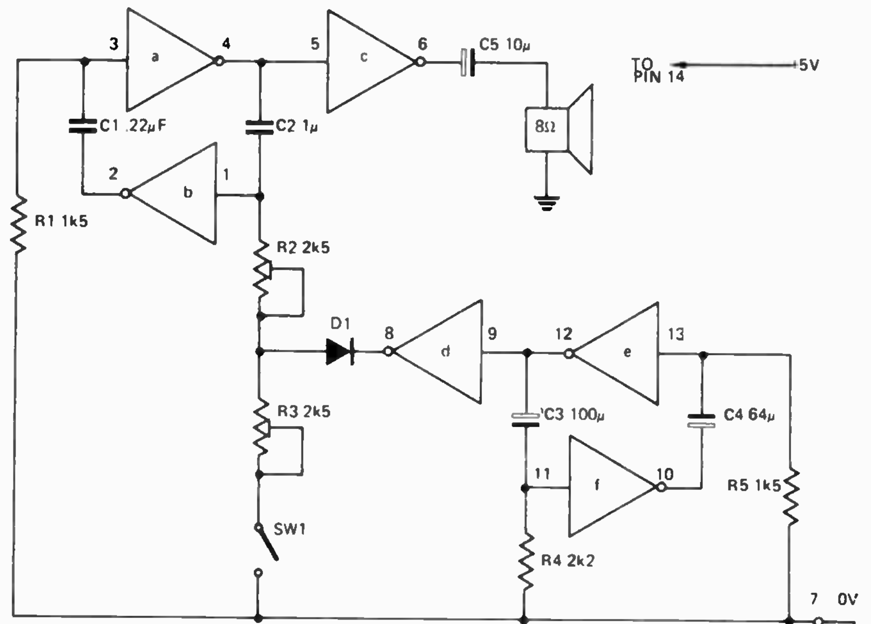
Should a more powerful output be desired then the output at the collector of Q2 can be fed into an amplifier input via a 1 μF electrolytic, in series with a 12 k resistor.



SIGNAL GENERATORS

SPECIAL EFFECTS

TWO TONE OSCILLATOR

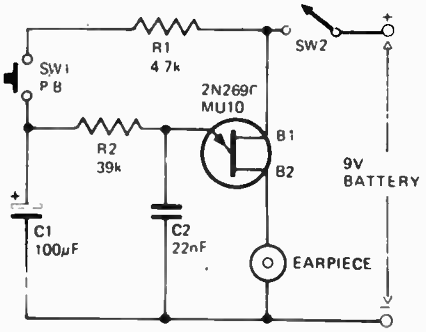


ALL GATES = SN7404 SW1 OPEN = ONE TONE SW1 CLOSED = TWO TONE
D1 = ANY GENERAL PURPOSE DIODE eg. 1N4001

This simple circuit uses only 1 SN7404 IC and can either generate an altering two tone signal or a pulsed single tone, either of the two tones can be varied and the circuit will drive a loudspeaker or a crystal earpiece.

In the circuit gates 'A' and 'B' generate an AF signal, the frequency of which depends on the resistance of the two inputs to earth (Pins 1 and 3). Gates 'F' and 'E' switch at a lower frequency (typically .5Hz) and their o/p is fed to gate 'D'. Whenever pin 8 is at logic '0', D1 is forward biased and the effective resistance to earth from pin 1 is R2 plus the resistance of D1 and the invertor 'D'. Whenever pin 8 goes to logic '1', D1 is reverse biased and the effective resistance to earth from pin 1 changes to that of R2 + R3 therefore the AF output frequency changes, when the circuit is switching a high and low tone alternately, R2 sets the frequency of the high tone and R3 that of the low tone, when the circuit isgiving a single frequency (i.e. SW1 open) then the o/p frequency is set by R2.

TOY SIREN



This circuit can be built small enough to be fitted inside a toy.

With a little manual skill on the part of the operator it can be made to sound like the sirens on such vehicles as fire-trucks, ambulances etc.

The transducer used is an earpiece which will give a scaled down sound in the proximity of the toy, without being annoyingly loud.

The circuit consists of a relaxation oscillator utilising one unijunction transistor. (2N2646, MU10, TIS43) R2 and C2 determine the frequency of the tone.

On pushing the button SW1 the capacitor C1 charges up and the potential at the junction of R2 and C2 rises thus causing an upswing in the frequency of oscillation; if one now releases the pushbutton the charge on C2 will drop slowly with a proportional reduction in the frequency of oscillation.

Manual operation of the button at intervals of approximately 2 sec will give a siren sound.

SIREN CIRCUITS FOR CHILDRENS TOYS

This circuit was originally designed to produce the sound of a police siren for my son's pedal car. It uses two 555 timers connected as oscillators (see Fig. a). The first oscillator IC1 is set for a period of 6 secs, 3 on and 3 off. Diode D1 is included to give equal mark-space ratio. This oscillator determines the rise and fall time of the siren.

The square wave output on pin 3 is turned into an exponential rise and fall by R3 and C3. This is reproduced at a low impedance by the emitter follower TR1 at pin 5 of IC2. The 555 timer has the facility for its timing period to be controlled externally by means of a control voltage applied to pin 5. IC2 is set for a nominal frequency of oscillators of about 1kHz, but this is pulled above and below the set frequency by the exponential waveform on pin 5. The output wave form starts at a low frequency, rises over 3 secs to a high frequency, falls over 3 secs to a low frequency and so on.

The loudspeaker used was a 75 ohm ex mobile radio handset speaker. This gave more than adequate volume off a 9V battery. Any loudspeaker can be used, provided a resistor is put in series with it to keep the total impedance above 45 ohms (for a 9V supply).

As originally designed the circuit gives an American-type police siren. It can easily be changed to give other types of siren: If R3, C3, TR1, R4 are omitted, and IC1 pin 3 is linked to IC2 pin 5 by R7 as shown in Fig. 1b, the "De-Dah" sound used by the British police is given.

If the values of R1, R2 are changed and D2 is added as shown in Fig. 1c we get the Star Trek "Red Alert". The values of R1 and R2 give a highly unsymetrical output from IC1. C3 now

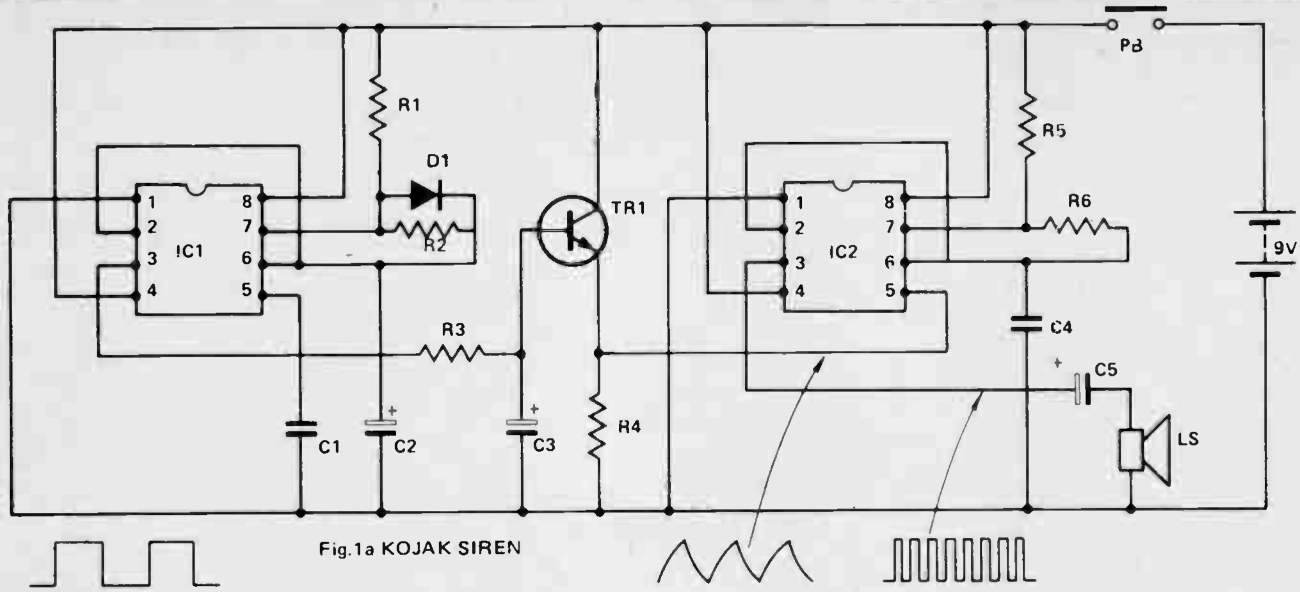


Fig.1a KOJAK SIREN

Component Values

R1	220k	R4	1k0
R2	220k	R5	4k7
R3	220k	R6	4k7

IC1,IC2 555 (or one 556)
TR1 Any G.P. NPN Silicon (eg BC107)
C1 0.01μF C2 250μF C3 250μF
C4 0.1μF C5 250μF 25V

D1 Any G.P. Silicon diode
R7 5k6 R2c 220k
R1c 1k
D2 Any G.P. Silicon diode

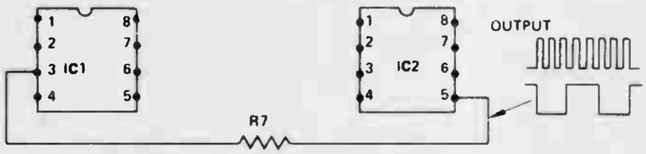


Fig.1b Modification to give Z-CARS SIREN

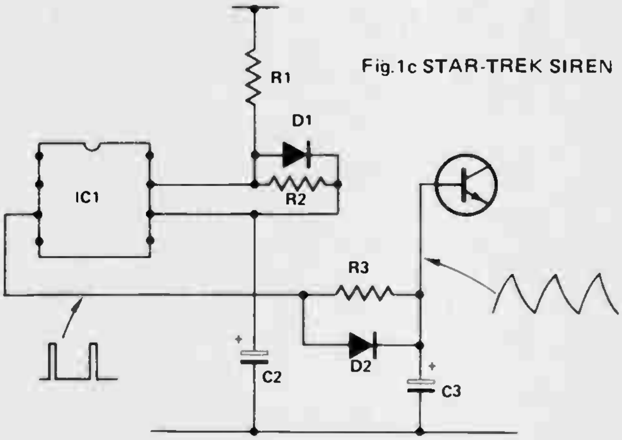


Fig.1c STAR-TREK SIREN

continued from previous page

gets a rapid charge via D2 during the short positive output from IC1, but discharges through R3 during the long low output time. The wave form at IC2 pin 5 thus approximates to a saw tooth, and the resulting output starts at a low frequency rises up to a high frequency over a period of 3 secs then falls abruptly to the low frequency again, and so on.

The circuits were originally built with 555 timers because I had a box full of them. A more elegant circuit can be made, however, by using the 556 dual timer. IC1 and IC2 can thus be obtained in one chip. The circuit works equally well with the 556, but the device has a slightly lower current rating than the 555. The loudspeaker impedance should be kept above 60 ohms by a series resistor as described above.

SPECIAL EFFECTS

SOUND EFFECT GENERATOR

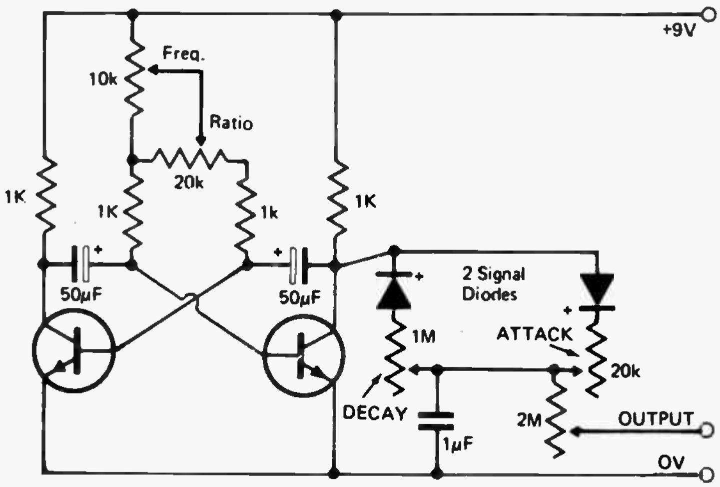


FIG 1

FIG. 2

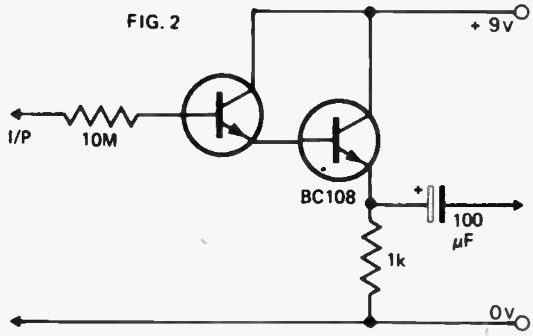
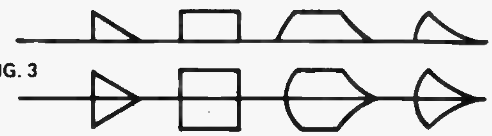


FIG. 3



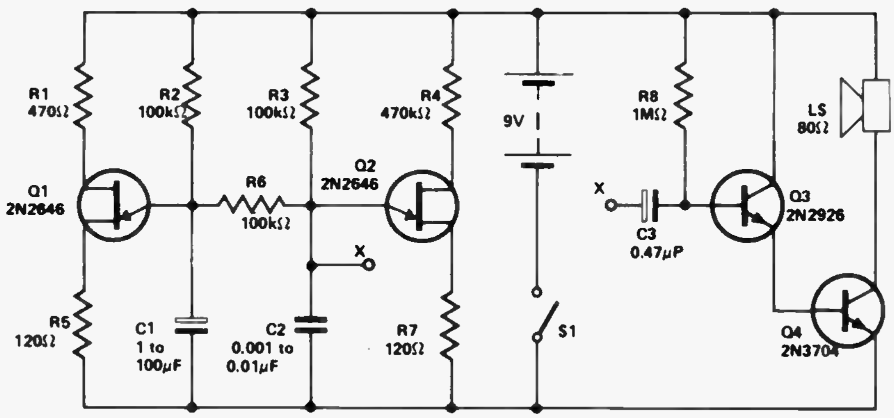
AFTER MODULATION

The waveshape generator shown in this circuit will interest those readers experimenting with sound effects.

Basically the circuit is a slow running oscillator with variable attack and decay. A variable amplitude (high impedance) output is available via the 2 meg potentiometer. Figure 2 shows an add-on circuit which should be used if a low impedance output is required.

Some of the output waveforms that can be produced are shown in Fig 3.

SOUND EFFECTS GENERATOR

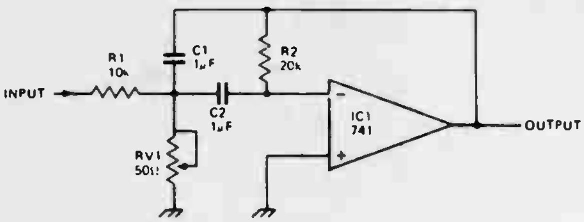


Interesting sound effects are produced by the circuit shown in the figure. The sounds generated by the speaker LS arise from the coupling of two unijunction transistor oscillators by means of the resistor R6. The circuit allows considerable scope for experimenting with the values of the timing capacitors C1 and C2.

As soon as switch S1 is closed, both C1 and C2 begin to charge via resistors R2 and R3, respectively. Owing to the smaller time constant of the R3/C2 combination, UJT Q2 discharges before Q1, the pulse being fed to the speaker via C3 and the Darlington-pair amplifier consisting of Q3 and Q4. Meanwhile C1 is much more slowly charging so that the next time that C2 begins to charge there is small voltage already on its upper plate and a shorter time elapses before Q2 fires once again. Thus Q2 fires faster and faster until the voltage on C1 is sufficient to fire Q1. The cycle then repeats. The sound heard consists of a tone of rising pitch which abruptly stops as Q1 fires and then repeats. For starters, make C1 10μF and C2 0.1μF. Note that R2 and R3 may be varied and these could be pots. R8 may require some adjustment to suit the gain of the Darlington Pair amplifier, Q3 and Q4.

AUDIO : ACTIVE

ACTIVE BAND PASS FILTER



This active filter has a gain of unity (0 dB) and is useful over the range 0.01 Hz to 3 kHz. The centre frequency of the passband is set by potentiometer RV1 and the bandwidth is determined by the values of R1, C1 and C2. The values shown in the circuit provide a bandwidth of about 15 Hz. With RV1 set to mid-position the centre frequency is approximately 220 Hz.

ACTIVE FILTERS

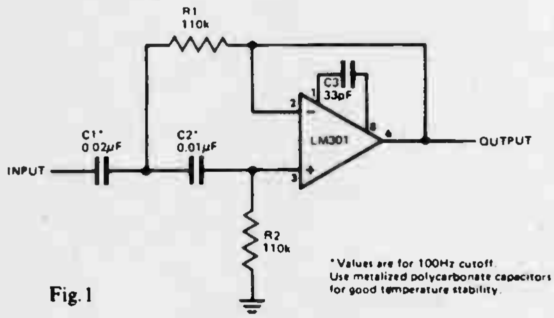


Fig.1

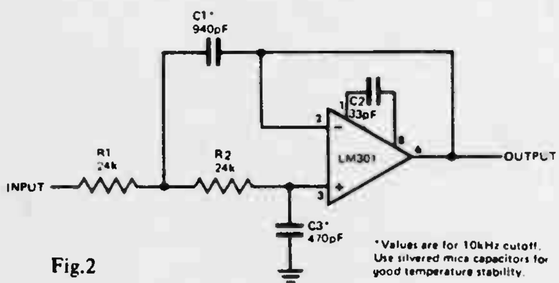


Fig.2

ACTIVE RC filters using operational amplifiers are increasingly being used to supplant LC filters because of the small size and ever-decreasing cost of integrated circuit operational amplifiers. Here are two useful general purpose circuits which may be readily incorporated into other circuitry where needed.

Figure 1 shows one of the simplest forms of filter, the low pass. The circuit has the same characteristic as two isolated RC filter sections with the additional advantage of a buffered low impedance output.

The attenuation is 12 dB per octave at twice the cut off frequency with an ultimate of 40 dB per decade.

There are two basic designs for this filter, the Butterworth (maximum flatness), and Linear Phase (minimum settling time for pulse input). The equations for the Butterworth design are:—

$$C_1 = \frac{R_1 + R_2}{\sqrt{2}\ R_1\ R_2\,\omega C}$$

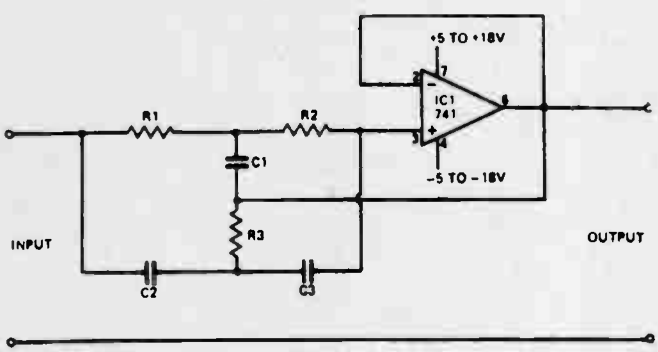
and

$$C_2 = \frac{\sqrt{2}}{(R_1 + R_2)\,\omega C}$$

For the Linear Phase design simply substitute $\sqrt{3}$ for $\sqrt{2}$ in the above equations.

To make a high pass filter we merely substitute resistors for capacitors and capacitors for resistors, as shown in Fig. 2, and apply the same formulae.

REJECTION FILTER



This narrowband filter using the 741 operational amplifier can provide up to 60 dB of rejection. Setting resistors equal to 100 k and capacitors equal to 320 pF the circuit will reject 50Hz.

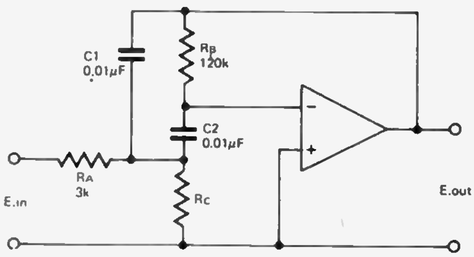
Frequencies within the range 1 Hz to 10 kHz may be rejected by selecting components in accordance with the formula'—

$$F = \frac{1}{2\pi RC}$$

To obtain rejections better than 50 dB resistors should be matched to 0.1% and capacitors to 1%.

AUDIO : ACTIVE

ACTIVE BANDPASS FILTER



A simple bandpass filter may be constructed using an op-amp and a few discrete components. The circuit shown has a constant gain and bandwidth, and the centre frequency may be adjusted from 1.6 kHz to 2.4 kHz by changing Rc from 1100 ohms to 400 ohms. Gain is 26dB at centre frequency and bandwidth is 775 Hz at 10dB down.

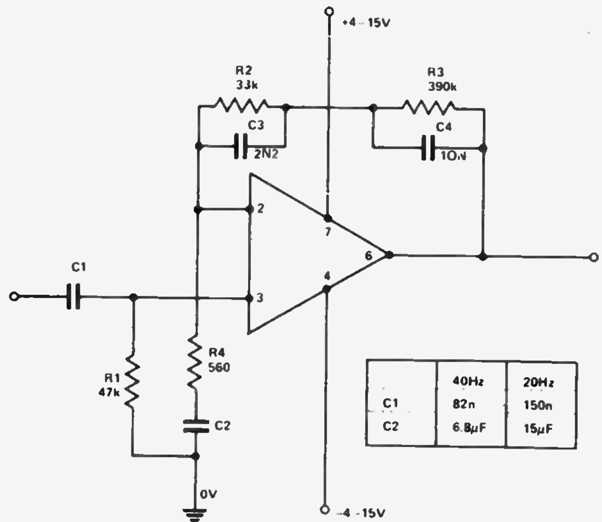
CARTRIDGE EQ AND RUMBLE FILTER

In this circuit a 741 op amp is used to provide standard RIAA equalisation for a magnetic pickup cartridge. The input signal is coupled via C1 into the non-inverting input of the IC. R1 damps the inherently high impedance at this point and provides the correct load for the cartridge. Feedback from the output, pin 6, is taken through the equalisation network R2, C3, R3 and C4 to the inverting input.

The ratio of R2 to R4 sets the midband gain at 65, 35dB. C1 and C2 together form a steep cut rumble filter whose cut off point can be set at 20 or 40Hz by selecting the appropriate component values in the table.

C2 also reduces the dc gain of the circuit to unity so that the output offset voltage will be ±5mV with reference to 0V.

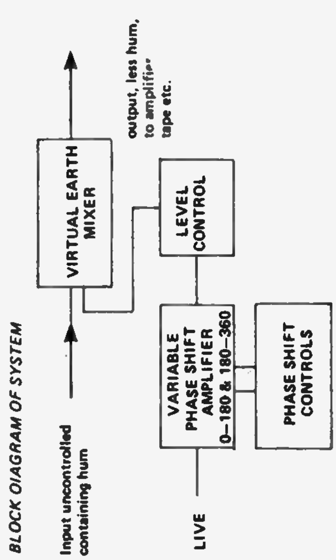
One of the major disadvantages of discrete equalisers is overload distortion. Although the output of a magnetic cartridge may be only about 5mV normally a musical peak may well force the cartridge output to 100mV. Clearly unless a large signal swing is possible the sound emmanating from the speaker is not going to be Hi-Fi.



	40Hz	20Hz
C1	82n	150n
C2	6.8μF	15μF

This circuit, operating from a ±15V supply, has an overload factor of +35dB refered to a nominal input of 5mV, equivalent to a maximum input of 325mV!

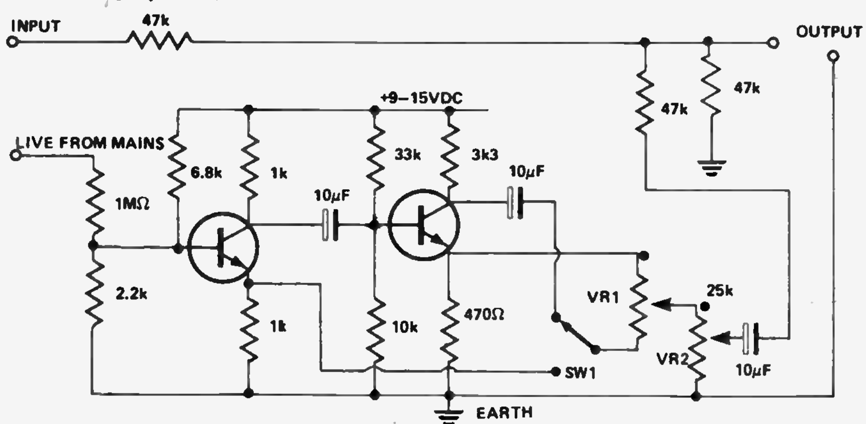
HUM STOPPER



Hum can be removed from an audio signal to great effect by mixing an antiphase hum of equal level.

In the circuit below all the transistors can be cheap or surplus npn (low or high gain) types.

VR1 is adjusted with VR2 low (not off) until the hum is at a minimum, SW1 may have to be changed over, then the level VR2, is altered until the hum is removed.



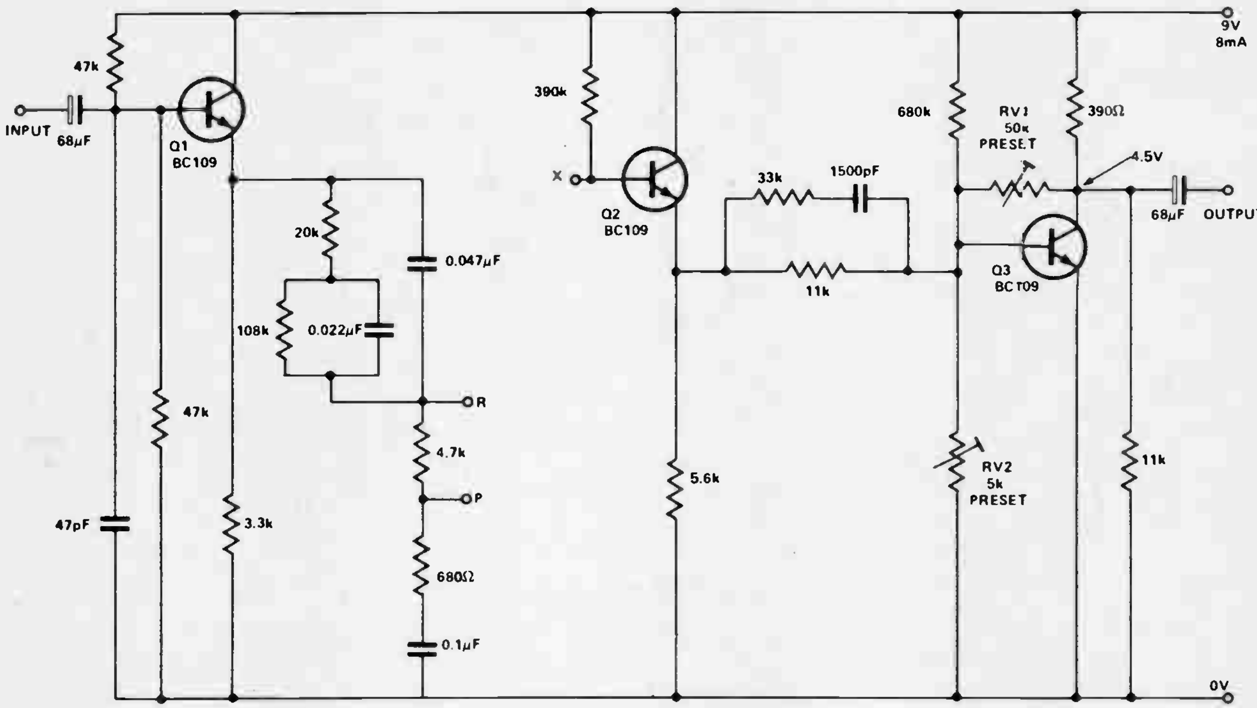
TAPE HISS REDUCTION CIRCUIT

The circuit in Fig. 1. is used to either boost or cut frequencies. When making a recording, point X is wired to point R so that treble signals are boosted by 10dB, and then during playback, point X is wired to point P so that the signal from the tape, including the hiss, has the treble cut by an equivalent amount. The circuit values are such that the overall frequency response, from record through playback, is flat over the range 20Hz–20kHz. Thus the output signal after playback is identical with the input signal before recording, but the hiss is cut by 10dB.

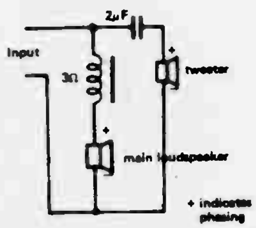
RV1 sets the gain of the circuit to be unity at low frequencies (<500Hz); RV2 is adjusted so that the collector voltage of Q3 is half the positive rail voltage. When this is set, the circuit will function without apparent distortion with an input voltage of up to 1.5V r.m.s.

If monitoring during record is not required, the same circuit may be used for record and playback, with X switched between P and R as necessary. If monitoring during record is required, two circuits are needed, one with X wired to R and the other with X wired to P.

For stereo, two circuits are required.



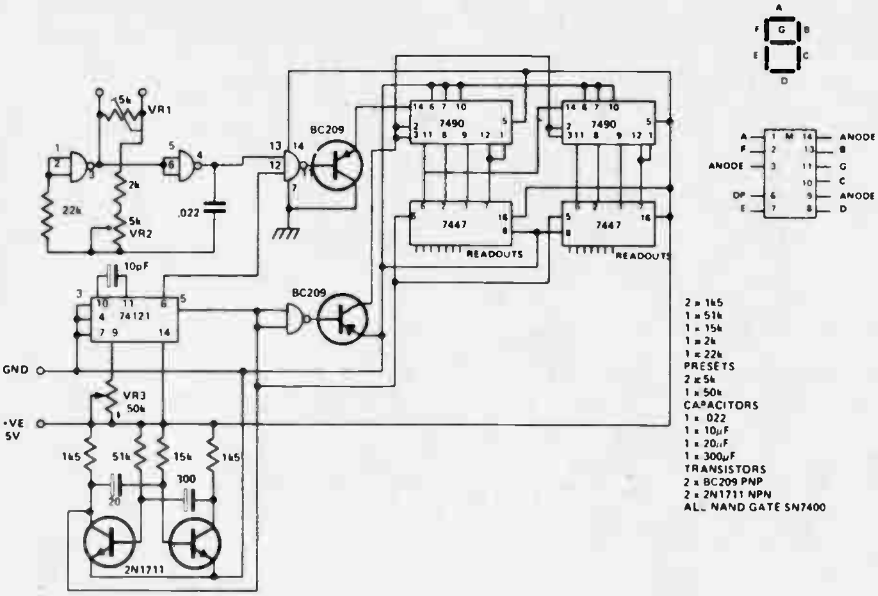
SIMPLE CROSSOVER



The construction of an inexpensive crossover for improving the performance of a cheap loudspeaker need not be complicated. The crossover enables a tweeter to be used; in the prototype an 8Ω miniature speaker from a pocket-type transistor radio was used, these make good tweeters.

The crossover can be built using the 3Ω secondary of a sound output transformer from a scrap T.V. set, the primary is left unconnected. The capacitor is a 2μF 63V polyester (non-electrolytic type).

DIGITAL THERMOMETER



DESCRIPTION

The frequency of the CMOS Multivibrator depends on the resistance of the thermistor, which is determined by the ambient temperature. Thus, if the temperature increases, the frequency of the multivibrator goes up and vice versa. Trimmer pot VR1 is used to adjust linearity.

The two transistor multivibrator automatically resets the 7490 decade counters and triggers the monostable 74121. When the 74121 operates, it closes the CMOS 'NAND' gate and allows the output of the temperature dependent multivibrator to pass to the counters. The length of time that the 74121 is on, is determined by the value of C2 and the setting of trimmer VR3.

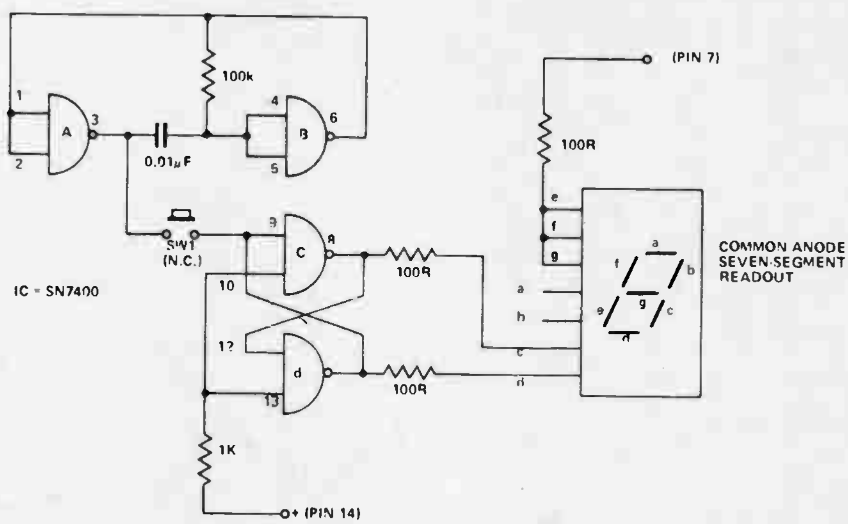
CALIBRATION

Fill a glass with ice cubes and top it up with cold water. Fill another glass with water that is as close to 90°F as possible. (Use an accurate thermometer). Place the thermistor in the ice water, adjust VR3 until display reads 32. Place the thermistor in the 90°F water, adjust VR2 until display reads 90°F. Repeat adjustment until accurate. Adjust VR1 for linearity. The digital thermometer is accurate to within 1°F between 32°F and 90°F.

HEADS OR TAILS CIRCUIT

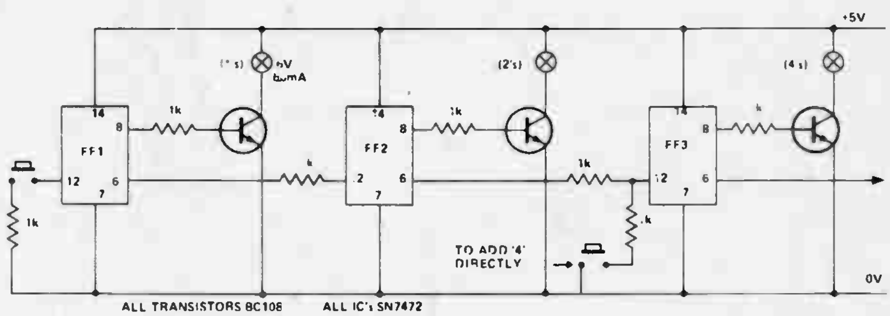
The two gates (A) and (B) of the SN7400 IC are connected as a astable multivibrator, whilst (C) and (D) are connected as a bistable.

Output from the multivibrator is taken via a 'spin' switch (SW-1) to pin 9. If the input to this pin 9 is high, and since pin 10 is already connected to a logic-1 through the 1kΩ resistor, the output at pin 8 will be at logic-0. This causes segment "C" of the readout to light forming the letter "h" for heads. Since the output of pin 8 is at logic-0 and connected to pin 12, the output of pin 11 will be at logic-1 which will light segment "d" to cause it to indicate the letter "t" for tails.



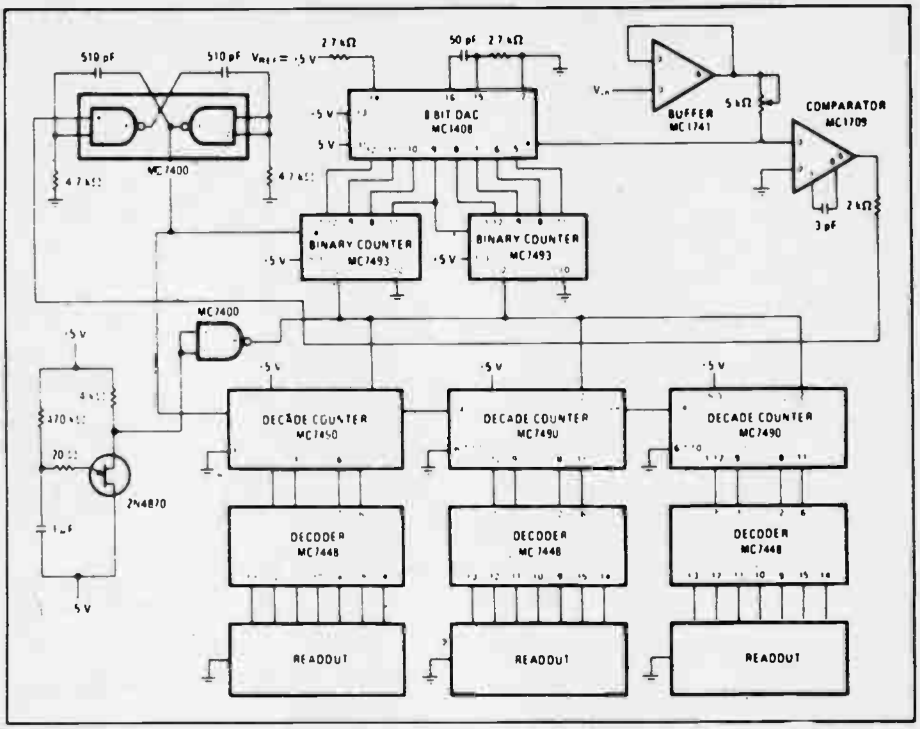
BINARY CALCULATOR

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.

SIMPLE DIGITAL VOLTMETER



This meter, is a closed-loop system that uses a clocked binary counter feeding a digital-to-analogue converter to produce a staircase ramp. The output of the converter is compared to the unknown input signal, and the clock pulses are terminated when the input signal level and the staircase function level are equal. The number of clock pulses occurring during the comparison process are therefore proportional to the voltage of the unknown input signal.

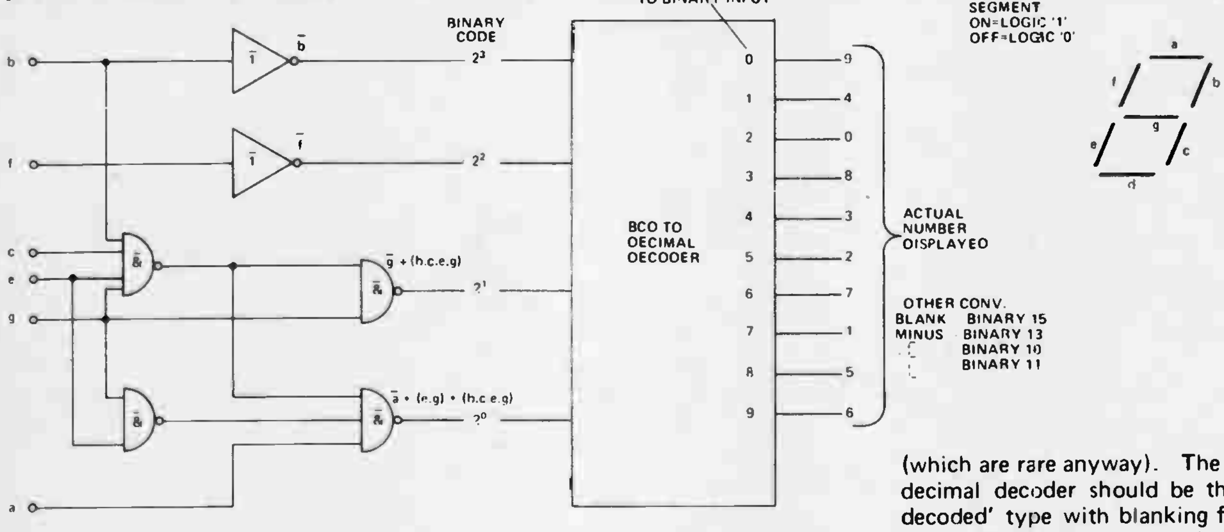
Clock pulses are generated by two cross-coupled TTL NAND gates at a frequency of 330 kHz so that 256 pulses can be counted in less than a millisecond. Such a high-speed clock has two main advantages: counting can be done without causing display flicker and the need to have latches to store the previous total count while the system is sampling is obviated. The clock pulses are applied to two sets of counters — a binary counter chain in the feedback loop that controls the converter, and a binary-coded-decimal counter chain that provides an easy interface with the seven-segment digital readouts.

The D/A (MC1408) converter generates an output sink current that is proportional to the value of the applied digital word. The maximum full-scale value of this current, which is typically 2 mA, is set by a reference voltage and a reference resistor. The convertor's output current is compared with the current from an input buffer amplifier which, in addition to giving the meter a high input impedance, supplies an output current of up to 2 mA for comparison with the output of the converter.

A second amplifier acts as a high-gain comparator to stop the clock when the current ramp from the converter exceeds the current from the input buffer amplifier. A unijunction-transistor oscillator is used to reset both sets of counters so that the unknown voltage is resampled about every 0.5 seconds, and BCD-to-seven-segment decoders convert the outputs of the BCD counters to the proper format for the seven-segment light-emitting-diode displays.

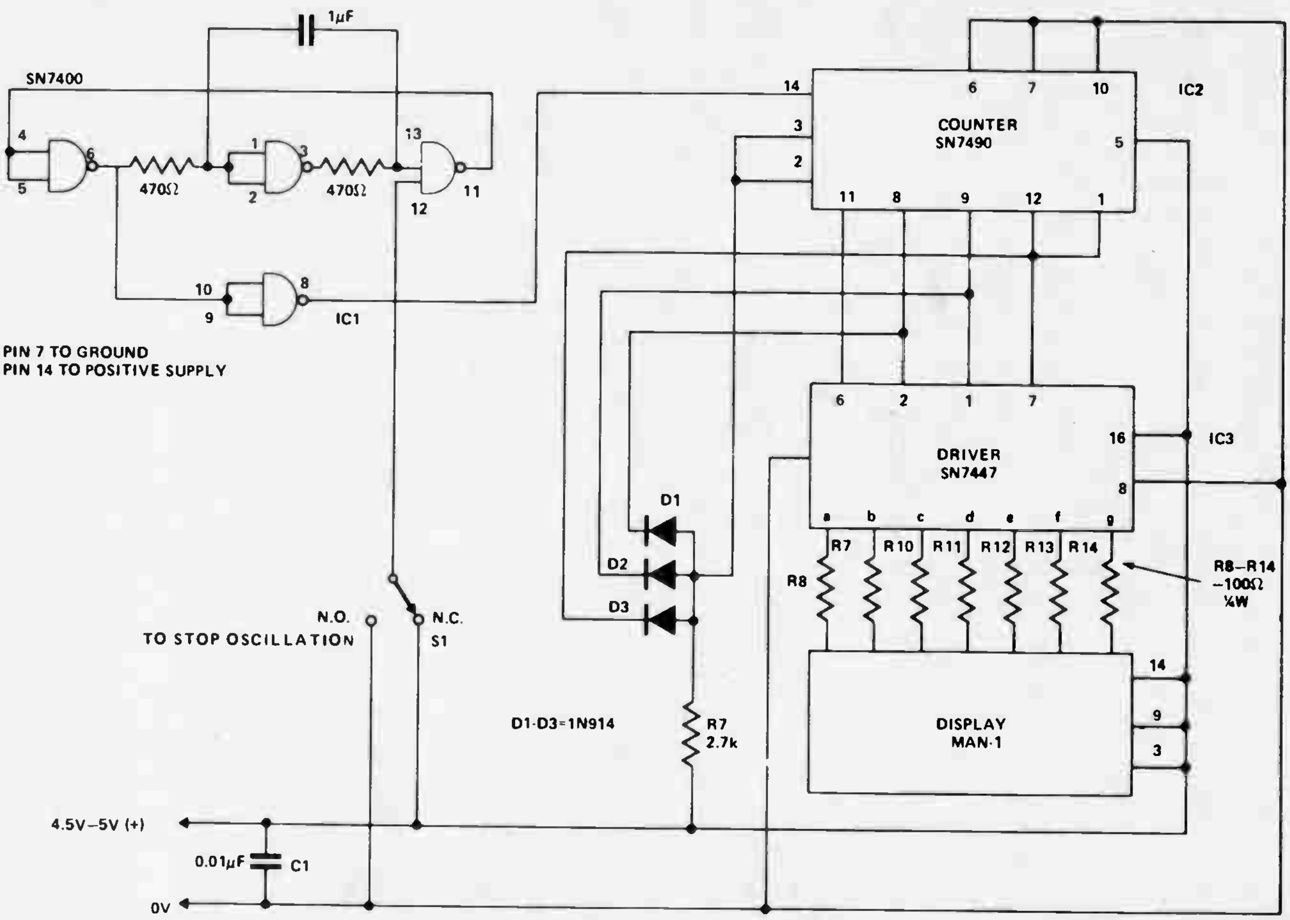
For the components used here, the meter can measure up to 2.55 V (to within ± millivolts) in 10 mV steps. Different full-scale values can be obtained by using suitable input voltage dividers or by providing the appropriate fixed-gain, rather than the unity-gain, input buffer shown.

SEVEN SEGMENT TO DECIMAL CONVERTER



Note that the output from the gates is not straight' BCD so the outputs from the BCD to decimal decoder are transposed. It will convert 6's and 9's with or without the top and bottom bars respectively but not 'hooked' 7's (which are rare anyway). The BCD to decimal decoder should be the 'fully decoded' type with blanking for BCD inputs over 9 since a blank is encoded as binary 15, hence a 74141 instead of 7441. Some other conversions which result from this circuit are shown.

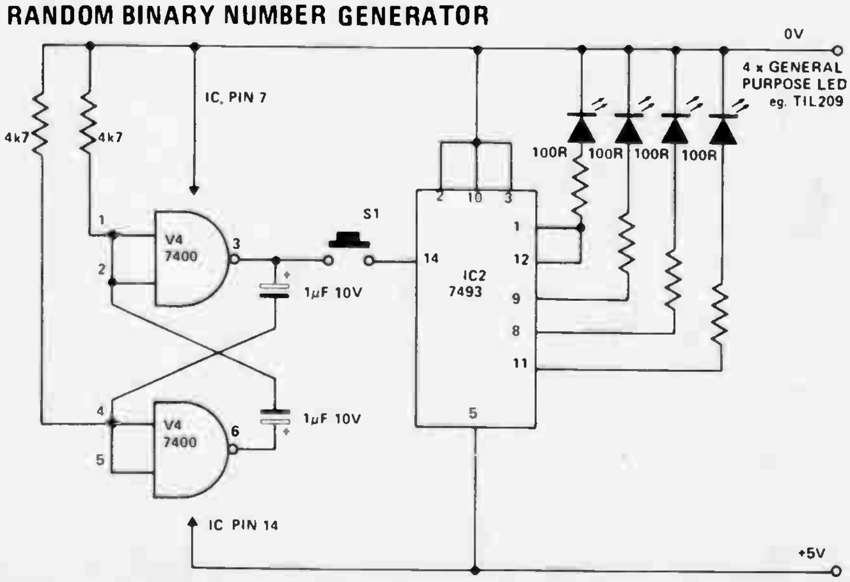
DIGITAL DIE



This device is based on a multivibrator (ICI) which has a frequency of about 1kHz. Oscillation continues as long as the input to pin 12 is high; as soon as the input is taken low or connected to earth it stops the cycle. This is used as a stop switch which causes the LED to indicate the random digit.

A diode AND gate, made up of D1 D2, D3 and R7 is used to reset the còunter (IC2) to zero so that only 0-6 are counted. To stop the 0, pin 5 of IC3 is connected to the negative supply. A 4.5v or 5v supply may be used; capacitor C1 reduces the noise on VCC line when TTL outputs switch logic states.

RANDOM BINARY NUMBER GENERATOR



The circuit shown is a random indicator providing an output from one of 16 states.

It consists of a BCD counter driven by a multivibrator. As the multivibrator's frequency is relatively high, one can say that the output from the counter, IC2, is random.

IC2 has a fan-out capability of 10 normal TTL loads and so can operate the LED displays directly. The four 100 ohm resistors are used to limit the current through the LEDs and so prevent them and IC2 from being damaged.

The unit is operated by depressing S1, which will cause the LEDs to flash, and when S1 is subsequently released the last number held in the counter will be displayed in BCD (Binary Coded Decimal) form.

CMOS ELECTRONIC DIE

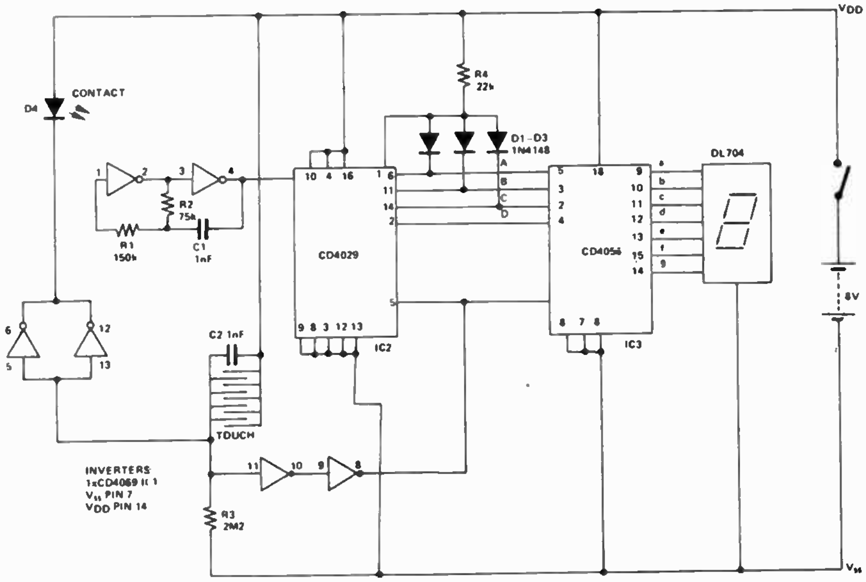
This circuit gives a readout of a random number on a 7-segment LED display. As shown, it will generate a number from 1 to 6 as does an ordinary cubic die. However the 4029 can be made to reset on another BCD number (one greater than the higiest number required) by changing teh arrangement of the diode AND gate. by changing the logic level on the preset pin (Q1:4, Q2:12, Q3:13 and Q4:3), the counter will reset to the pre-programmed number. If these features are not to be used (i.e. count is 0–9) pin 1 on IC2 should be taken to V_{SS}.

Alphabetical symbols (L,H,P and A) will be produced on the display if pin 9 of IC2 is taken to V_{DD} causing the counter to produce the binary numbers from 10–15 as well.

A finger on the touch plate causes pins 5 and 1 on ICs 2 and 3 to go high, stopping the counter and enabling the latch in IC3. When the finger is removed the number is stored in the latch and the counter restarted.

If a common anode display (DL707) is used instead of the DL704, taking pin 6 of IC3 to V_{DD} will invert all its outputs, suitable for a common anode device.

The touch switch may be replaced by a single pole push to make if required, otherwise it is a small veroboard cutoff with alternate strips wired together.

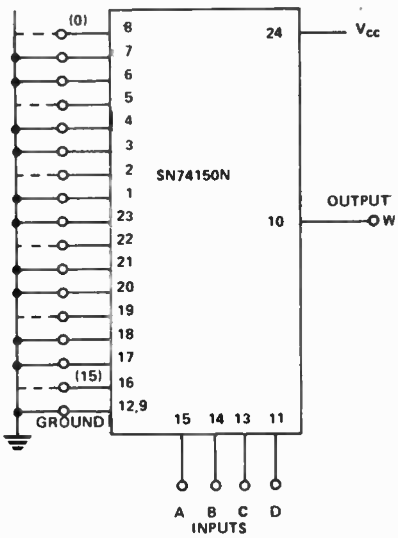


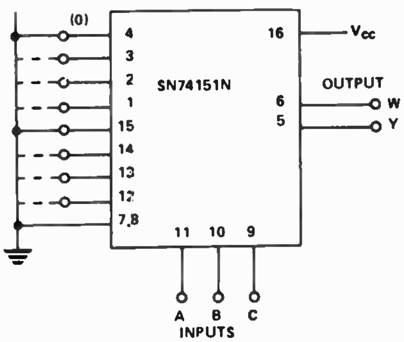
SELECTOR MULTIPLEXER HINTS

This is a method of implementing arbitrary logic functions with an absolute minimum of wiring up and maximum reliability.

The circuits are based on logic data selector/multiplexers, either TTL or CMOS (TTL shown). The first diagram shows the arrangement for producing a function of four variables. The four input variables are decoded by a 74150 16-line to 1-line data selector and used to govern which of sixteen data inputs is used to control the output state of the selector. The output is the complement of the selected input. This in effect forms a low cost, hard-wired PROM (Programmable Read-Only-Memory) which is programmed by wired links or switches as shown, (or by inputs from other logic gates).

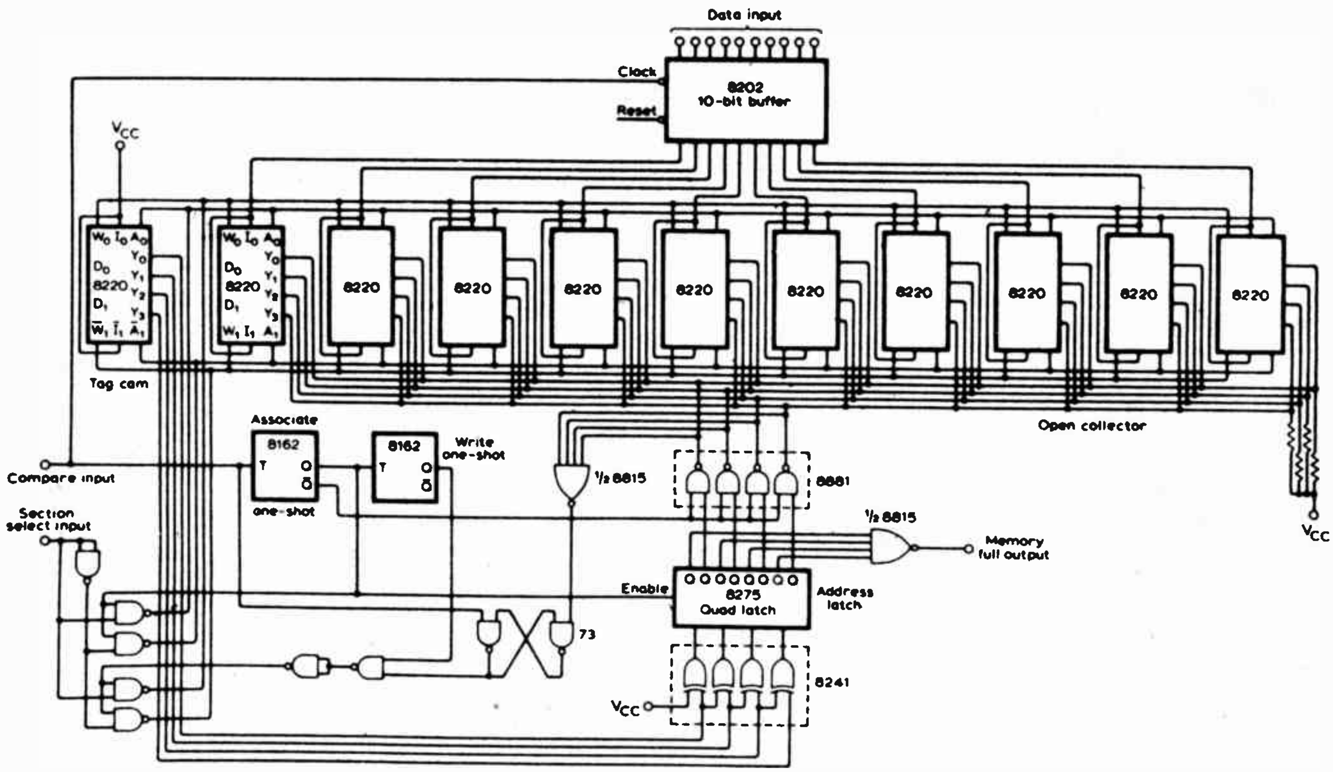
To programme the inputs, for a given set of variables, the input, number corresponding to the binary number formed by variables ABCD, (1-2-4-8) is connected high for a low output or low for a high output. In practice only connections to low need be made if it is more convenient, although it is good practice to tie the high inputs to Vcc (or VDD) via a 1k resistor. (The links as shown in the example implement a function to produce an output only if the binary number ABCD is exactly divisible by three.) If it is preferred, each input may be tied to a high level via its own resistor which may then be left in circuit even if required to short the input down to earth.





An ideal form of switch for programming infrequently changed functions which must nevertheless be easily changed is the modern PCB mounting dual-in-line switch which takes up little room. A typical application of this might be for adjusting clock rate timers. A similar arrangement is shown for a function of three variables using the SN74151 IC. In this case the output is available in both true (Y) and complement (W) forms. (In the example shown the majority function is produced. Output is high if two or more inputs are high, otherwise low.)

A MEMORY THAT LEARNS



There is currently a great deal of interest in Content Addressable Memories (CAMs) and they are finding use in more and more applications. However, it is probably true to say that the novel characteristics of the CAM have not yet been fully realised.

The CAM is simply a memory with the ability to make a comparison between data already stored and data which is presented to the input. When several CAMs are connected in an array it is possible to apply feedback in such a way as to make a word which has just been read from the memory the next address. Circuits such as these can be made to generate or recognise sequences of digital words.

In the recognition mode, for instance, a CAM array with associated external logic could recognise a dangerous sequence of events in a process control system .and could be made to take the appropriate action or altert staff to the impending danger.

Here is a memory constructed from Signetics 8220 CAMs, which has the ability to reject or accept new data depending on what is already in the memory. Once the memory has learned a data word, it will not accept another identical word. In addition, the memory automatically decides at what address new acceptable data is to be stored and ensures that new information is not written into locations which are already occupied.

Each 8220 is a CAM capable of storing four words of two-bits and the memory as a whole can store eight 10-bit words. Although the storage capacity of the memory is 80-bits (8 x 10), eleven CAMs are employed which together have a capacity of 88-bits (11 x 2 x 4). The eleventh CAM has been called the tag CAM because it keeps track of the locations within the memory which are occupied, and allocates a new address for acceptable information.

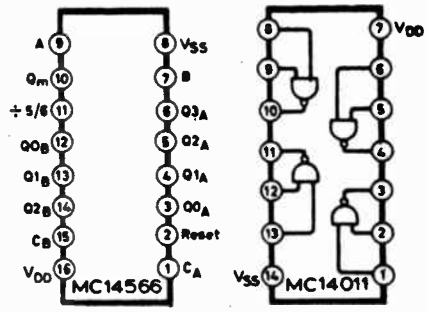
The memory is sub-divided into the two sections of equal capacity and either of the two sections can be selected using the "section select input". Input data is presented to the 10-bit buffer and the "compare input" is activated. This clocks the data into the buffer and initiates a comparison process in which each word already stored within the memory is compared with the data in the buffer. If a location within the memory is found to carry data identical to that within the buffer, one of the 8220's Y outputs will go 'high' and the write command will be inhibited. If no accurate match is found, the data in the buffer is written into the address specified by the 'tag' CAM. Exclusive-OR gates connected to the Y outputs of the tag CAM specify the next available address and ensure that memory locations are filled successively. The address at the outputs of the exclusive-OR gates is latched into the quadlatch before the 'write' command is available to the CAM array. Thus the Y lines of unavailable memory locations are forced to logic '0'.

CMOS CLOCK

A new, and unique, addition has been made to Motorola's rapidly expanding family of CMOS logic circuits. It is the MC14566 time base generator which consists of two pulse shapers, a divide-by-ten ripple counter, a divide-by-5 (or 6) ripple counter and a monostable multivibrator on a single chip. A single MC14566 can be connected to divide by 50 or 60 (÷ 5 and ÷ 10 or ÷ 6 and ÷ 10) to produce one output pulse per second when fed with a 50 or 60 Hz input. In addition, a binary coded decimal output indicating tenths-of-seconds is available.

A second MC14566 can be connected in cascade with the first (arranged to divide-by-ten and then by six) to provide one output pulse per minute and a BCD output of up to 59 seconds. A third cascaded MC14566 will then provide a minute's BCD output and one pulse per hour.

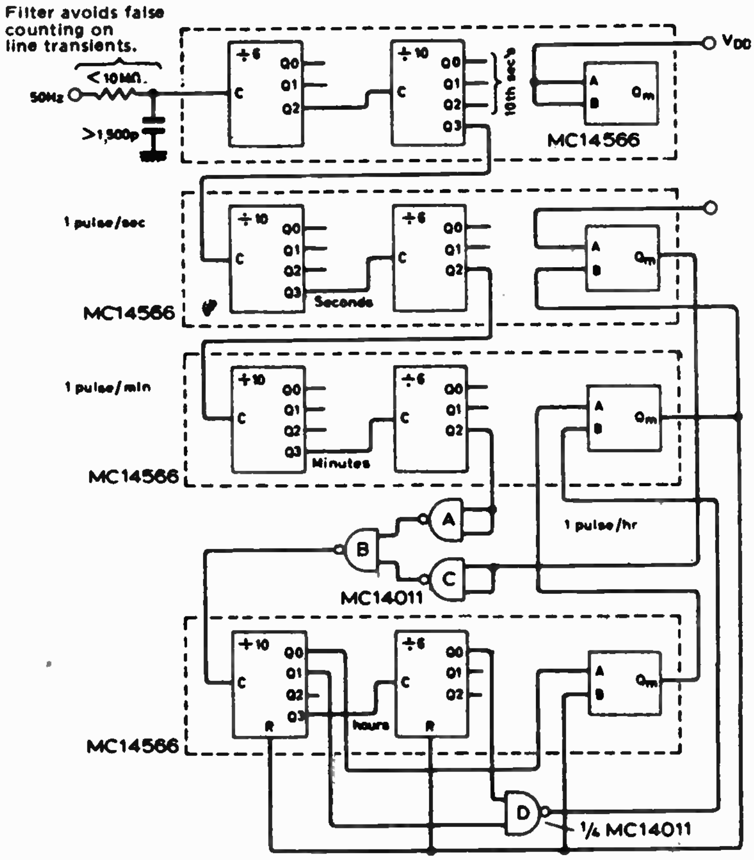
Although the devices can be used to construct electronic digital clocks – as shown in the circuit diagram – their main application will be to provide timing signals in industrial process control, data-logging and computing equipment from 50 or 60 Hz line supplies.



V_{DD} = +3V to +18V
V_{SS} = 0V
Connect pin 11 to V_{SS} to divide by six and to V_{DD} to divide by five
All unused reset pins must be connected to V_{SS}

Quad two-input NAND gate

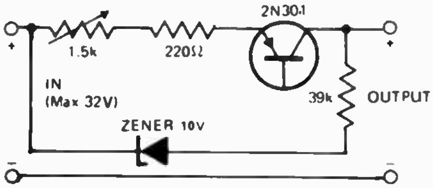
Available in a plastic package (suffix P) or a ceramic package (suffix L), the MC14566 has Zener diode protection fitted to all inputs and is available for operation over the extended industrial temperature range (–40 to 85°C) or the full military temperature range (–55 to 125°C). As with all members of the Motorola CMOS family the power supply voltage can be from 3 to 18 V, the noise immunity is typically 45% of V_{DD} and an input capacity of 5 pF is standard for all inputs. Quiescent power dissipation at 5 V supply voltage is 25 nW, rising to about 1.5 mW at a clock frequency of 1 MHz when working into a 15 pF load. Normally, when used as a timer, power consumption would be less than this since the clock frequency would be either 50 or 60 Hz. Maximum operating frequency is typically 4.2 MHz at V_{DD} = 15 V.



POWER SUPPLIES

CURRENT

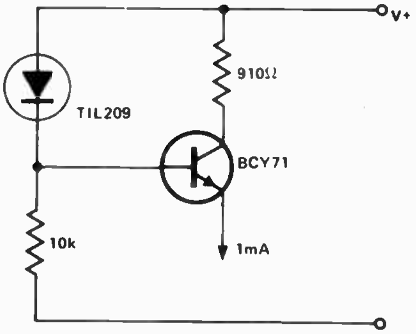
SIMPLE CONSTANT CURRENT CIRCUIT



A series transistor is used as a variable resistor for this constant current supply.

The output current is held within 10% over a range of loads from a short circuit to 500 ohms. The required current is set by the potentiometer. The transistors specified will handle voltages up to 32 V.

TEMPERATURE-STABLE CURRENT SOURCE



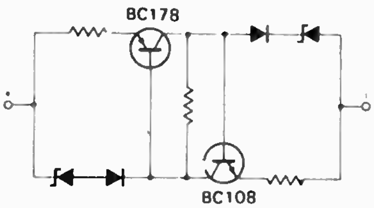
This current source is very temperature-stable; the output current varies by less then 1% over the temperature range −55°C to +100°C. This is possible because the transistor is biased by an LED, whose forward voltage drop has a temperature coefficient of −2mV/°C, the same as the base-emitter voltage of a silicon transistor. Hence near-perfect temperature compensation is possible, a great improvement over conventional methods of biasing with zener diodes.

The circuit values shown give an output current of about 1mA, though wide variation is possible by altering the value of emitter resistance. They are good for supply voltages in the range 25V to 5V.

CONSTANT CURRENT SOURCE

This unique two terminal circuit can be used to define a constant current in the same manner as a Zener diode may be used to define a constant voltage.

The values of R1 and R2 shown are for a current of 1mA. Maximum applied voltage with transistors shown should be limited to 50 volts. Minimum should be at least 8 volts.

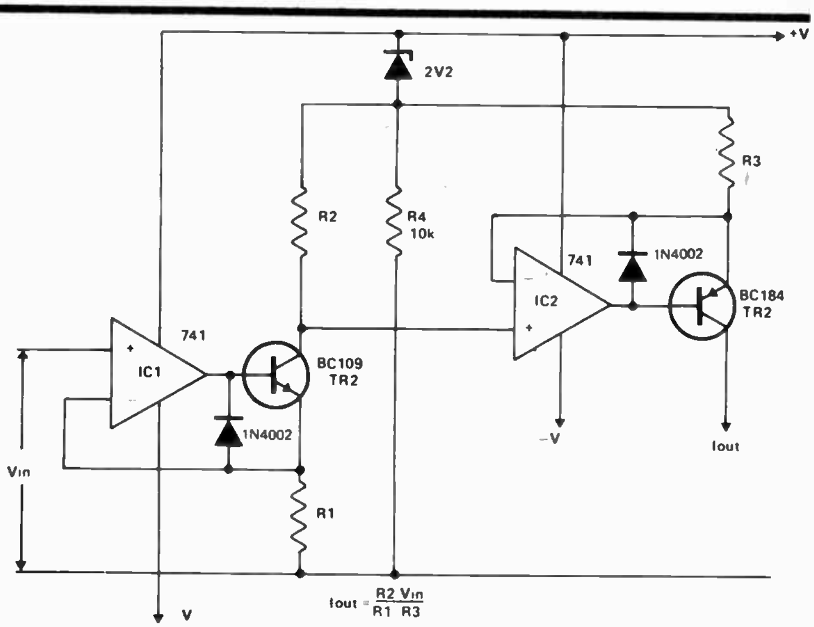


VOLTAGE CONTROLLED CURRENT SOURCE

The voltage follower IC1, buffered by TR1 provides a current at the collector proportional to the input voltage due to R1. This current is applied to R2 which means that the voltage across it will be Vin $\frac{R2}{R1}$.

IC2 forces the voltage across R3 to equal that across R2. The zener prevents IC2's inputs from operating at the supply rail.

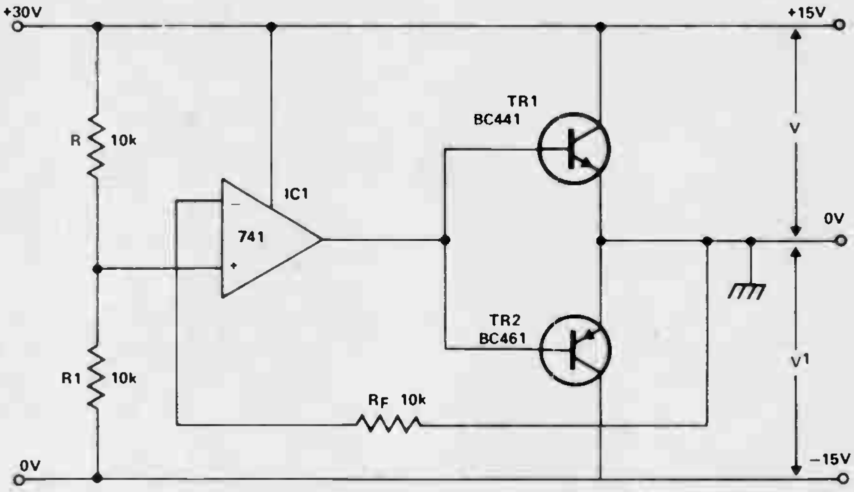
In this circuit the input voltage is generated relative to the 0V rail. For maximum output voltage capability the voltage across R2 and R3 at the maximum proposed output current should be kept small. However offsets in the IC's have more effect on the linearity as the max resistance of VR2 becomes smaller — these should be nulled out.



DUAL

PRECISION VOLTAGE DIVIDER

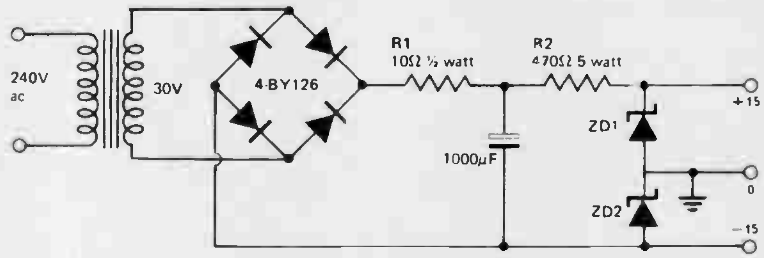
This circuit has the advantage over the simple 'two resistor' voltage divider in that the voltage ratio V:V' does not depend on the current taken from it. The ratio of resistances R:R' sets the voltage ratio. The OP AMP detects any change in this ratio via R_f and provides correction. The actual voltages used will be limited by the upper and lower operating voltages of the OP AMP. The circuit shown was designed to provide 15V for operational amplifiers from a single supply.



DUAL POLARITY SUPPLY POWERS OP-AMPS

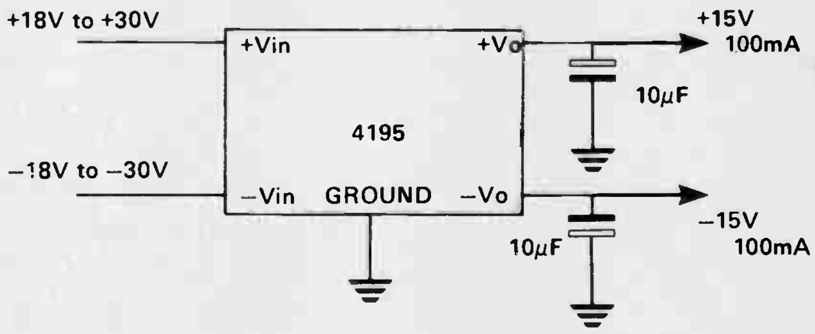
Ever been bothered by the lack of a dual supply for your op-amp circuits? This simple circuit gives positive and negative supply from a single transformer winding and one full-wave bridge.

Two Zener diodes in series provide the voltage division and their centerpoint is earthed. (N.B. the smoothing electrolytic must not be earthed via its case).



ZD 1 and ZD 2 = 15V ZENER DIODES – BZX70 C15

SIMPLE BALANCED REGULATOR



When experimenting with operational amplifiers and other circuits, one often requires balanced positive and negative power supplies of about ±15V. One can, of course, employ two separate stabiliser circuits employing an integrated circuit voltage regulator.

A simpler solution is shown in which the new Raytheon 4195 ±15V dual-tracking voltage regulator is employed. Only two capacitors are required in addition to the regulator device, so it is convenient to employ one of these regulators on each printed circuit board.

The circuit can supply up to 100mA from each output and is fully protected against short circuits. In addition, the device switches itself off if the temperature of the chip exceeds +175°C, so there is no danger of thermal damage. The short circuit is typically 220mA.

As the load current varies from 1 to 100mA, the output voltages stay constant to 5mV in a typical case. If the input voltage varies between the limits shown, the typical output voltage is 2mV. The temperature coefficient of the output voltage is about 0.005% per °C. The ripple rejection is about 75dB and the output noise voltage 60μV r.m.s.

LOW VOLTAGE

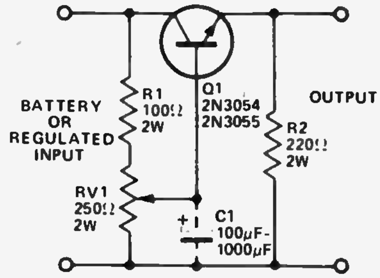
REGULATED VOLTAGE DIVIDER

IC's requiring 3.6 or 6 volts can be run from a battery or fixed-regulated supply of a higher voltage by using the circuit shown.

The transistor should be mounted on a heatsink as considerable power will be dissipated by its collector.

Additional filtering can be obtained by fitting a capacitor (C1) as shown. The capacitance is effectively multiplied by the gain of the transistor. A ripple of 200 mV (peak to peak) at the input can be reduced to 2 mV in this fashion.

Maximum output current depends on the supply rating and transistor type (with heatsink) used.



LOW REGULATED DC VOLTAGES

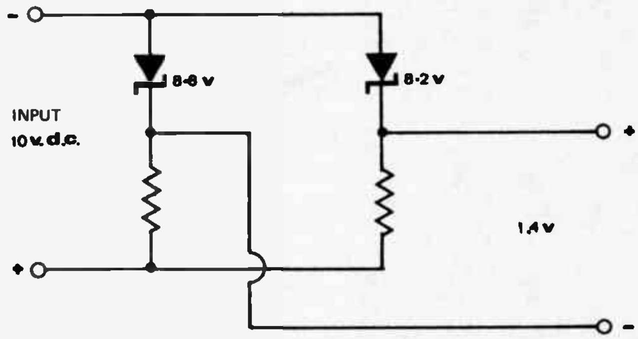
For some purposes it is necessary to use regulated dc voltages lower than can be obtained from commercially available zener diodes.

This can be achieved by using two zener diodes of different voltages and utilising the potential difference between them.

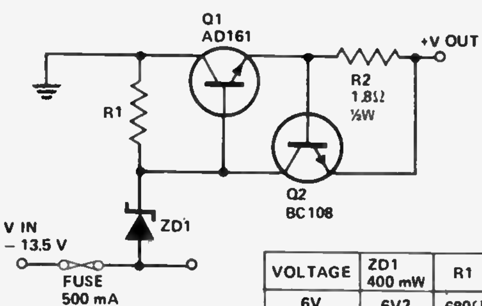
In the example shown, 6.8 volt and 8.2 volt zeners provide the required 1.4 volt difference.

One such application is the supply voltage for gallium arsenide emitters which, typically, require 1.4 volts ± 0.05 volt.

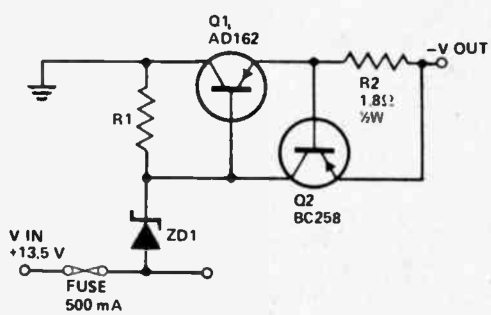
In general, temperature compensation is excellent as both zeners tend to drift in the same direction — either positively or negatively, depending upon voltage. It is inadvisable to use zeners which bracket 5.6 volts, as below this voltage zeners have a negative temperature coefficient and above 5.6 volts they have a positive temperature coefficient.



LOW VOLTAGE STABILIZERS HAVE SHORT CIRCUIT PROTECTION



VOLTAGE	ZD1 400 mW	R1
6V	6V2	680Ω
7.5V	7V5	390Ω
9V	9V1	220Ω



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.

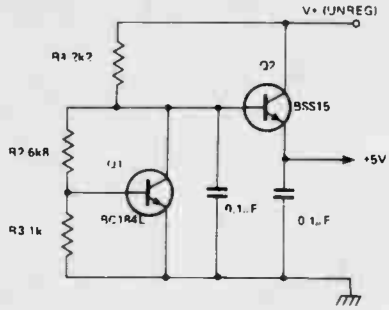
Both circuits are protected against overload or short-circuits.

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.

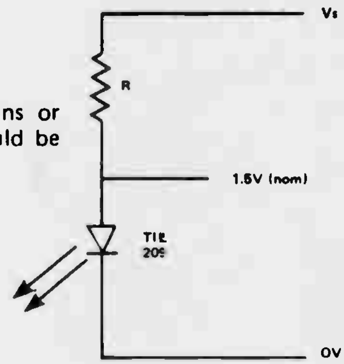
SIMPLE SUPPLY FOR TTL



When it is necessary to power a few TTL packages, a simple zener diode — emitter follower stabliser is often used. If a zener of the appropriate value is not available, then the circuit shown may be used instead. Q1 is a standard V_{BE}-multiplier; the current through Q1 increases until its V_{BE}-drop is established across R3. Hence about 8 times V_{BE} is established across Q1, and this voltage is used as a reference for emitter follower Q2.

The unregulated input voltage may be between 20 and 8V; many different types of transistor will work satisfactorily.

ZN 414 POWER SUPPLY



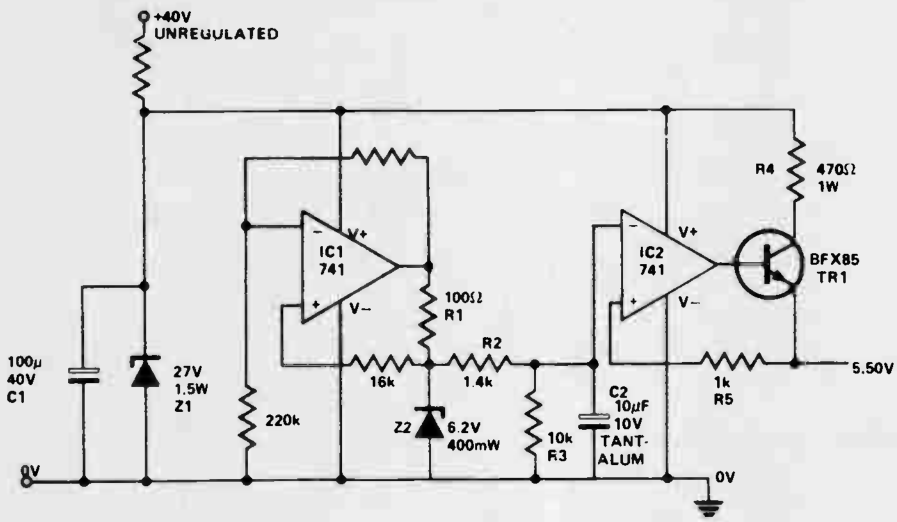
The Ferranti ZN 414 radio IC requires 1.1-1.8V at up to 1mA. A light-emitting diode can serve a dual purpose as a low-voltage zener diode and an "on" indicator, as shown. A suitable type is the Texas TIL 209 which gives a reasonable light output at 2-3mA consumption, so that the idea is suitable for mains or battery-powered radios. The resistor value should be

$$\frac{V_s - 1.5}{3}$$

kΩ approximately, where Vs is the supply voltage.

SPECIAL

STABLE REFERENCE-VOLTAGE SUPPLY



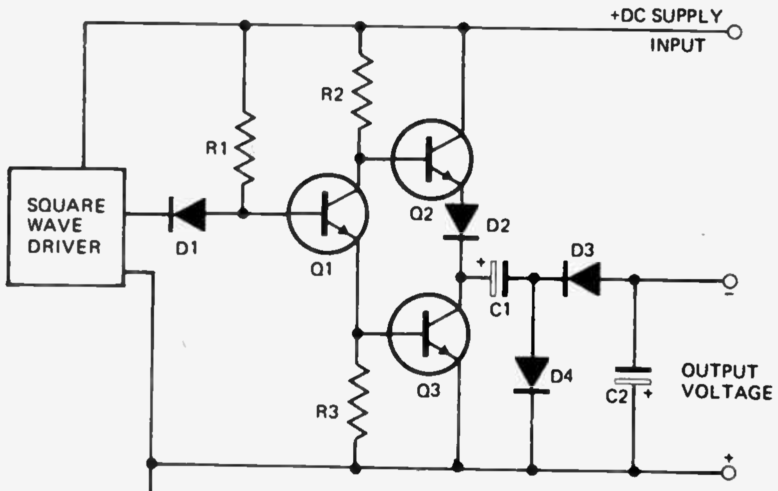
This circuit was evolved to provide a highly stable and ripple-free voltage to act as the reference for the stabilised power-supplies of an electronic music synthesiser; the stability of its voltage-controlled oscillators depended directly on the constancy of supply-rail voltages.

The 40V unregulated supply is derived from conventional full-wave rectification and smoothing of the ac from a 25Vrms transformer winding, and pre-regulated by Z1; C1 provides some further smoothing. IC1 and IC2 are powered from the 27V rail thus generated. IC1 drives a constant current of 5mA through Z2 by acting as a differential amplifier sensing the voltage-drop across R1. Z2 is a 6V zener since diodes of about this voltage have the lowest voltage/temperature coefficient.

The stable voltage across Z2 is then reduced to the desired value by the potential divider R2, R3. This network has a fairly high output impedance and so C2, although fairly small, has a large smoothing effect. C2 must be tantalum as a conventional electrolytic may inject more noise than it removes, in this application. The voltage across C2 is then buffered by IC2 and TR1; R4 forms a simple but foolproof protection against short-circuits. The prototype was designed for a maximum rated output of 30mA.

The reference voltage provided by the circuit as shown above is 5.5V. Different values may be obtained by altering the value of R2; if a voltage above 6.2 is required then R5 must be connected to a potential divider across the output.

TRANSFORMERLESS INVERTER



D1 – Small signal diode D2, 3, 4 – Rectifier diodes
Q1 – Small signal N.P.N. Q2, Q3 – Output transistors

This transformerless inverter chops the dc supply voltage then rectifies the resulting square wave using a conventional voltage doubling circuit.

The square wave source can be a simple IC multivibrator which is used to drive phase splitter Q1 through coupling diode D1. Base bias is established through R1 and the collector and emitter load resistors are R2 and R3 respectively.

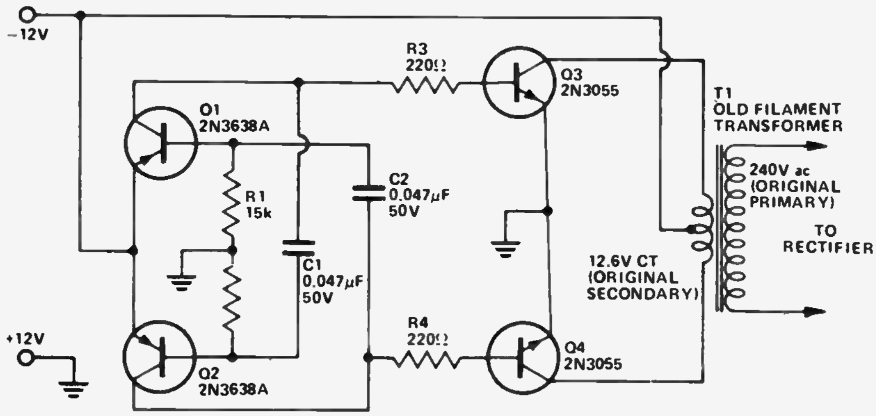
The output power transistors Q2 and Q3 in conjunction, with diode D2, serve as a simple high level switch developing a square wave whose peak to peak amplitude is near that of the dc supply voltage.

The output square-wave is coupled via C1 (which must be a suitably chosen high-value electrolytic) and fed into a voltage doubler circuit thus producing a dc output of reverse polarity.

Capacitor C2 is the output ripple filter.

Resistor values are dependent on the original supply voltage – the drive frequency is not critical but signals in the kilohertz region are preferred (2-6 kHz).

DC TO DC/AC INVERTER



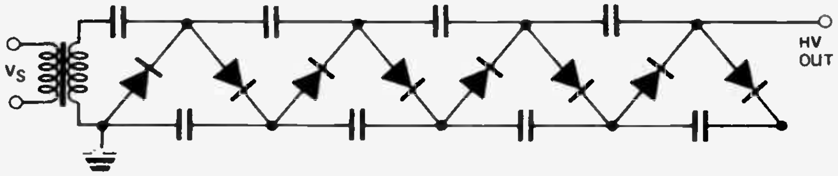
This inverter uses no special components such as the torodial transformer used in many inverters. Cost is kept low with the use of cheap, readily available components.

Essentially, it is a power amplifier driven by an astable multivibrator. The frequency is around 1200 Hz which most 50 Hz power transformers handle well without too much loss. Increasing the value of capacitors C1 and C2 will lower the frequency if any trouble is experienced. However, rectifier filtering capacitors required are considerably smaller at the higher operating frequency.

The two 2N3055 transistors should be mounted on an adequately sized heatsink.

The transformer should be rated according to the amount of output power required allowing for conversion efficiency of approx. 60%.

VOLTAGE MULTIPLIER

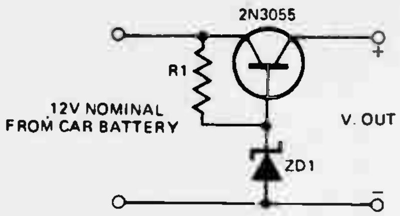


Sometimes a very high voltage is required for applications such as for ionisers or a CRT supply. secondary of an ordinary power transformer can have its output voltage multiplied any number of times determined by the number of stages "cascaded".

It is important to note that the rating of individual diodes and capacitors should be twice the transformer output voltage V_s.

The capacitor value and diode rating are determined by the required output current.

12V–9, 7.5, or 6V CONVERTER (AUTOMOBILE)



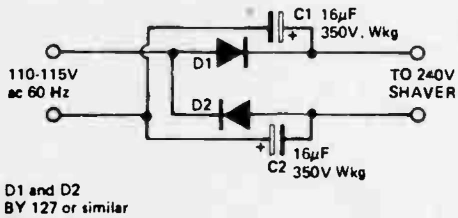
OUTPUT VOLTAGE	9	7.5	6
R1 (½ WATT)	180Ω	270Ω	330Ω
ZENER DIODE (250 mW)	10V	8V1	6V6

Many transistorised items such as radio, cassettes and other electrical items operate on batteries. Usually these are in the 6–12V range and sockets are provided for external power supply.

This circuit enables these devices to be operated from a car's electrical supply.

The table gives values for resistors and specified diode types for different voltage. Should more than one voltage be required a switching arrangement could be incorporated. For high currents the transistor should be mounted on a heatsink.

TRAVELLER'S SHAVER ADAPTOR

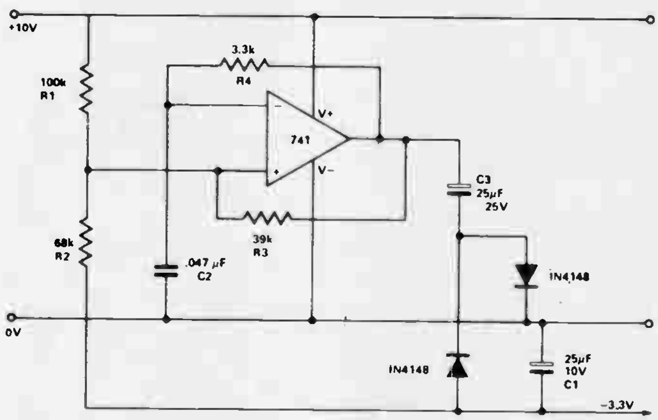


Many overseas countries have 115 volts mains supplies. This can be a problem if your electric shaver is designed for 220/240 volts only.

This simple rectifier voltage doubler enables motor driven 240 volt shavers to be operated at full speed from a 115 volt supply.

As the output voltage is dc the circuit can only be used to drive small ac/dc motors. It cannot be used, for example, to operate vibrator-type shavers, or radio sets unless the latter are ac/dc operated.

SIMPLE DC–DC CONVERTER



Often in circuit design it is handy to have a low-current negative rail available to bias FETs etc. This circuit generates a supply rail 2 to 5V below its 0V line.

If the lower end of R2 is connected to 0V, then the circuit is seen to be an op-amp relaxation oscillator driving a pair of diodes that charge C1 negatively. R3 provides positive feedback, changing the switching point of the op-amp, according to whether C2 is charging or discharging through R4. When the voltage on C2 reaches the switching point, the circuit changes state and the C2 voltage sets out for the other switching point.

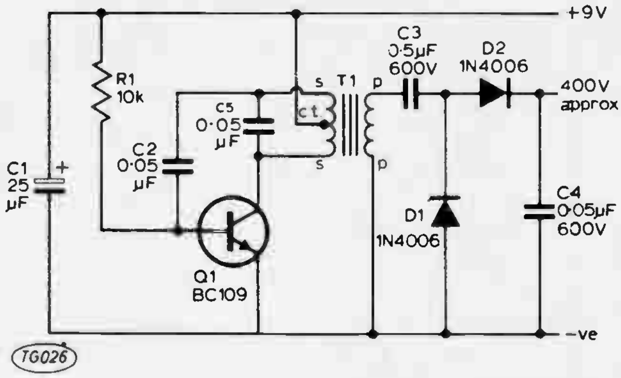
When the lower end of R2 is attached to the negative output, then as the negative charge on C1 increases, the operating range of the oscillator is pulled down until it is outside the operational range of the 741, and the charging ceases. This provides a form of switching regulation of the output voltage, roughly halving the output impedance. The output voltage can be set to the desired value by altering R2. For 3.3V output the prototype showed about 10mV ripple on full design load of 1mA.

The output is inherently short-circuit protected by the current-limiting action of the 741.

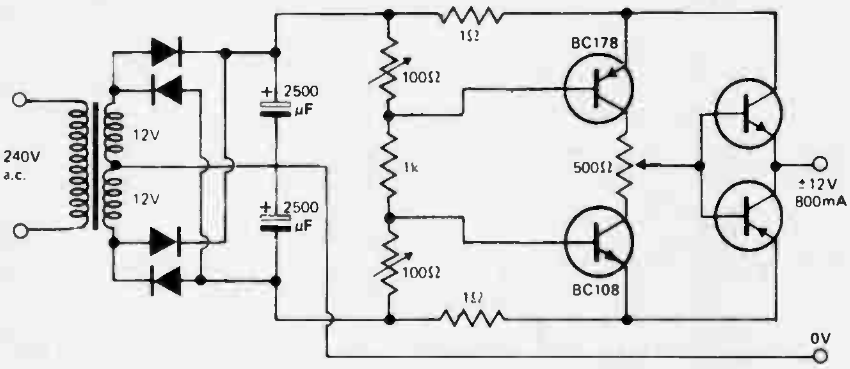
SPECIAL

HIGH VOLTAGE FROM A BATTERY

It is a simple matter to obtain up to 400V from the smallest 9V battery using the circuit shown. The transformer is widely available – it is a 250V to 9-0-9V (or similar) type. The 9-0-9V connections are connected to the transistors in a Hartley oscillator configuration. The 250V connection is taken to a voltage doubler which will give about 400V, albeit at very high impedance and is not all that dangerous. The secondary voltage can be varied by inserting a potentiometer (5k) in the supply line.



VARIABLE POWER SUPPLY GIVES POSITIVE OR NEGATIVE OUTPUT



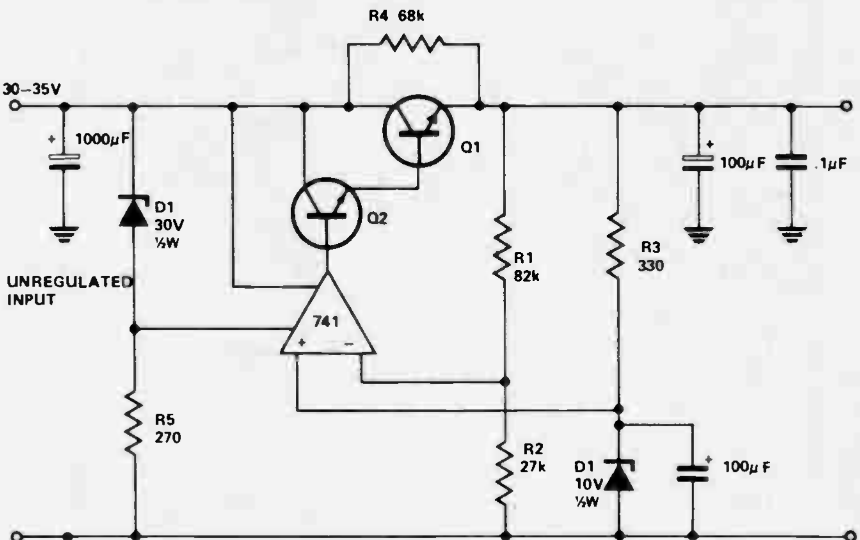
A variable power supply using complementary output transistors is capable of swinging the voltage at the output from +12 V through zero to –12 V.

The two output transistors can be types BD 135 – BD 136. These are both cut off when the 500 ohm potentiometer is centred.

Rotating the potentiometer in either direction will give positive or negative output voltages up to 12 V and 800 mA. The series resistors (1 ohm) monitor output current and when this exceeds a level preset by the 100 ohm trimmers will current limit the output.

STABILISED

STABILISED POWER SUPPLY



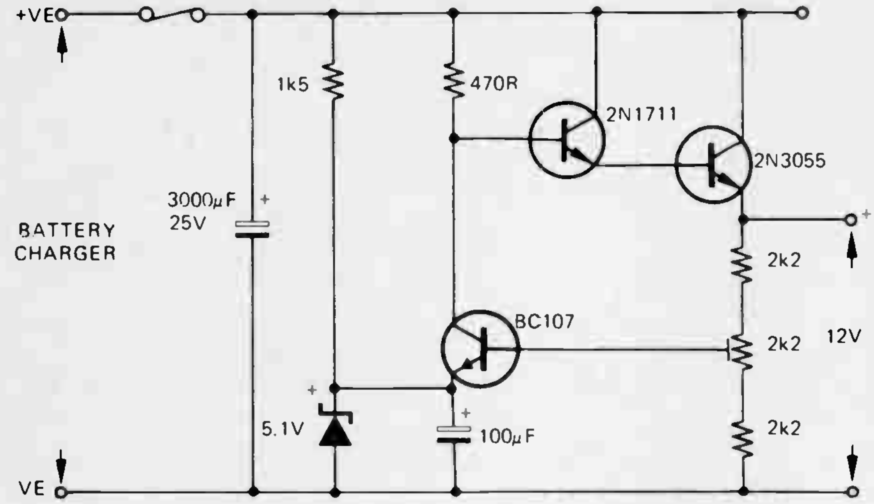
Q1: Any NPN power series stabiliser eg: TIP3055, 2N4923, 2N4921
Q2: Any NPN transistor capable of driving the above eg: BC182L

The operation of the circuit is quite simple and straightforward, as regulated power supplies can be considered merely as special kinds of feedback amplifier. Here, the output signal is sampled by R1 and R2, and compared with a reference voltage supplied by D2. The resultant correction signal is fed back via the 741 to the series pass element Q1. Note that the stability of the circuit is improved by supplying the reference source R3-D2 from the stabilised output as opposed to from the unstabilised input as is usual. In order that the circuit operates when turned on, a leakage resistance R4 is put in parallel with the series pass element. This ensures that the feedback loop starts to operate. No regulation is lost as a consequence of R4, because it is the overall output that is sampled by R1-R2, and so the effect of the ripple current flowing through R4 is corrected by the feedback.

The output may be made variable by replacing R1-R2 with a potentiometer, but in its present form, the circuit cannot be made to regulate below the zener voltage of D2. If continuous variation is required, the reference source R3-D2 must be supplied from the unregulated input, with consequent slight loss of stability.

The amount of power the circuit can deliver is limited chiefly by the current rating of Q1 and the rated output of the unregulated supply.

12V SUPPLY FROM A BATTERY CHARGER

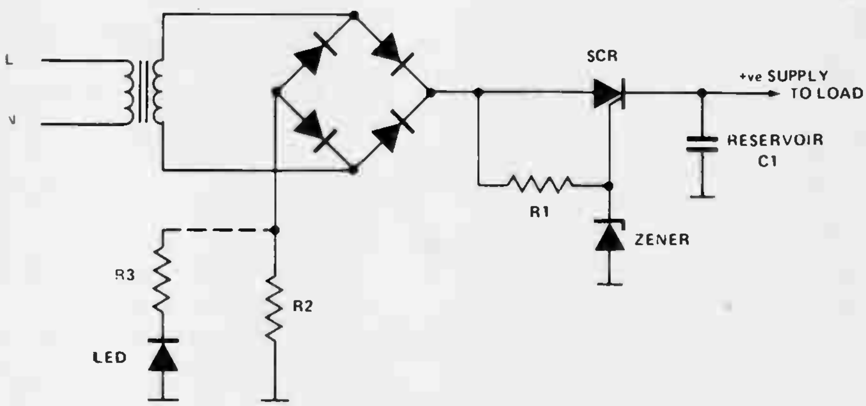


This 12V regulator unit was designed to enable bench testing of mobile equipment (radios, tapes, C–D units, etc,) using a battery charger, thus avoiding the expense of a complete bench supply and the inconvenience of a car battery.

The charger output is smoothed by the 3000mfd capacitor. The BC107 is a comparator, sampling part of the output voltage while the reference zener holds the emitter at approximately 5V (4.7V or 5.6V zeners could be used). The 2N1711 supplies the necessary current gain to drive the 2N3055 series regulator.

A heatsink of at least 16 sq.ins. should be provided for the 2N3055 but the 2N1711 will run cool without a heatsink.

'BUCKET' REGULATOR



There are a number of applications where a simple cheap form of regulated power supply, giving a supply regulation of the order of 5–10%, is useful. One such application is the class B audio amplifier. The cost of the additional components required to achieve regulation is more than offset by the saving in cost and size of the electrolitic capacitors alone.

Fig. 1. shows the circuit of the regulated supply. The only additional components required to affect regulation are the SCR, R1, and the zener.

At switch on the the reservoir C1 is discharged and the cathode of the SCR is at zero volts. The positive going output from the bridge rectifier will cause gate current to flow via R1 triggering the SCR. The reservoir C1 starts to charge. At the end of the half-cycle the SCR will turn off.

The following half-cycles will repeat the process charging C1 until the supply voltage approaches the zener voltage. However the maximum positive potential at the SCR gate is determined by the zener, so there comes a time when the reservoir will have charged to the point where the SCR gate cannot be driven positive with respect to its cathode. At this stage the SCR will stop firing and no further charging current will be delivered to the reservoir. The reservoir will discharge via the load, whatever power is being supplied, until the gate is once more positive, when the SCR will fire again. One or more half cycles are sufficient to raise the reservoir voltage sufficiently to prevent further firing.

Thus the SCR fires as necessary to keep the reservoir "topped up" and it is this topping up action which gave the regulator its name. The number of times it fires in any particular interval being dependant on the load current taken from the supply.

There are two particularly attractive features of this type of supply. First its efficiency is high, there are none of the power losses associated with either series or shunt regulators.

The second is that it is possible to obtain very simply an indication of the current being delivered, This may be obtained by connecting a LED (in series with a current limiting resistor R3) across the main current limiting resistor R2. The LED will flash each time the SCR conducts and hence the rate at which the flashes occur will depend on load current, the flash rate varying from once every few seconds when only leakage is being made good - to continuously under full load conditions.

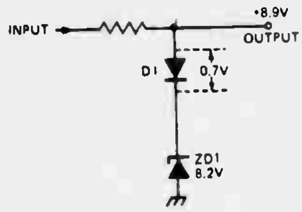
As an indication of circuit values the following where used for a 25V, 1.5A supply: R1=1.2k, R2=2Ω, R3=330Ω, C1=5000μF, 25V, Transformer 30V.

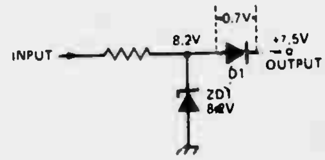
The SCR and bridge rectifier should be rated at full load current, but for many music and speech applications the transformer can be derated as much as 50%.

POWER SUPPLIES

ZENERS

DIODE ADJUSTS ZENER VOLTAGE

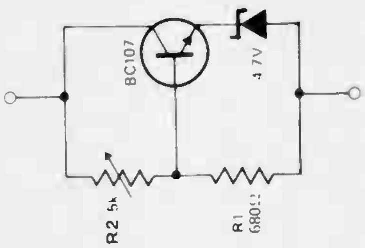




A silicon diode when forward biased has a constant voltage drop of 0.7 volts. A germanium diode has 0.2 volts.

This characteristic may be used to trim a Zener reference voltage as shown. A silicon diode in series with the Zener will raise the output voltage by 0.7 volt, and in series with the supply it will be reduced by 0.7 volt. Make sure that the diode used will carry the required current.

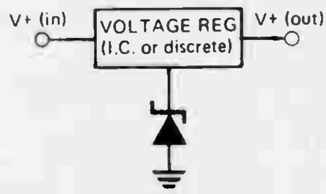
A VARIABLE ZENER DIODE

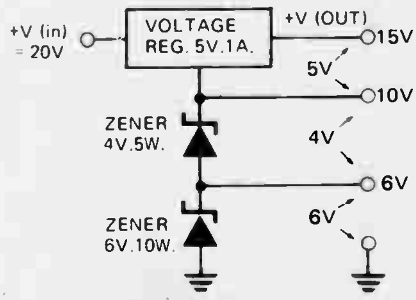


The circuit shown behaves like a Zener diode over a large range of voltages. The current passing through the voltage divider R1 – R2 is substantially larger than the transistor base current and is in the region of 8 mA. The stabilising voltage is adjustable over the range 5 – 45 V by changing the value of R2. The total current drawn by the circuit is variable over the range 15 mA to 50 mA. This value is determined by the maximum dissipation of the Zener diode. In the case of a 250 mW device this is of the order of 50 mA.

When stabilising higher voltages or operating at higher currents it is necessary to fit a small heatsink to the transistor.

ZENER BOOSTS OUTPUT VOLTAGE OF REGULATOR





In this circuit the zener diode raises all voltages – with respect to earth – by the zener voltage, i.e.

Vin (max) ≏ voltage regulator Vin (max) + zener voltage

Vin (working min) ≏ voltage regulator Vin (min) + zener voltage

Vout = voltage regulator Vout + zener voltage

As the voltage regulator dissipates all excess power while the zener merely clamps the output voltage above its own voltage, a low wattage zener (250 mW) should be adequate – unless lower voltage taps are used, as in the second example in which the total output is one amp.

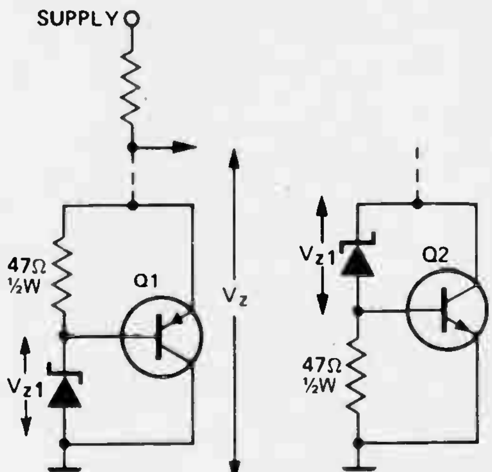
For other value zeners, wattages can be worked out by the formula W = zener voltage X current.

HIGH-POWER ZENERED VOLTAGE FROM LOW POWER SOURCES

A power transistor can be used to provide a high power zenered voltage from a low wattage zener. A 400 mW zener can be used where a 10 watt zener is required or a 1 W zener can be used where a 50 to 80 watt zener is required, by using appropriate transistors for Q1 and Q2 in the circuits shown.

Where low rating is required Q1 would be a ASZ 15 (germanium) or an AY9140 (silicon). Q2 could be a 2N3054 (silicon). For higher powers Q1 could be an ASZ18 (germanium) or a 2N2955 (silicon) and Q2 a 2N3055 (silicon) or an AY8149 (silicon).

A heatsink on the transistor is required. The circuit in (a) has the advantage that power transistors can be bolted directly on to a chassis which may serve as a heatsink.



(a)PNP (b) NPN

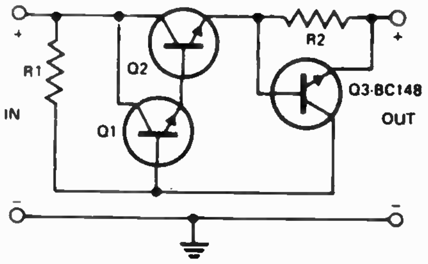
$V_{z1} = V_z - V_{be}$

Q1, Q2 – GERMANIUM OR SILICON POWER TRANSISTOR

V_{be} – GERMANIUM = 0.3V
V_{be} – SILICON = 0.7V

PROTECTION

ELECTRONIC FUSE



Imax	R1	R2	Q1	Q2
5.0A	100Ω	0.12Ω	BFY50	2N3055
0.5A	1k	1.0Ω	BC107	BFY50
0.1A	4.7k	3.7Ω	BC107	BFY50

Here is a circuit for protecting modern transistorised gear which requires a faster action than can be provided by an orthodox fuse.

Transistor Q2 is saturated by base current supplied by Q2, which is itself turned on by R1. The overall voltage drop between input/output is in the region of 2V. If a momentary surge in current or a short circuit in the load appears then the voltage drop across R2 will increase and when it reaches about 0.7V, Q3 will begin to conduct and its collector emitter voltage will drop to about 0.3 V. This in turn cuts off Q1 and Q2 thus breaking the supply current.

The tables gives circuit values for various currents. These are suitable for supply voltages up to 45 V.

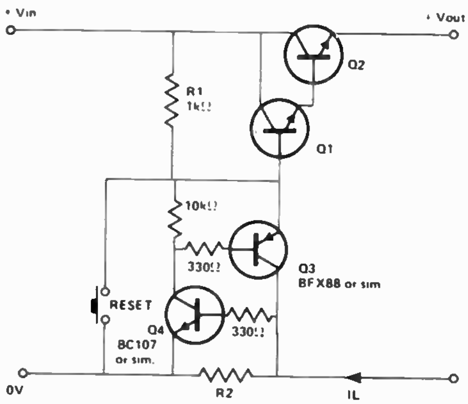
BETTER ELECTRONIC FUSE

The electronic fuse implies that load current is disconnected in the event of an overload. In fact it merely limits the load current to a value given by $\frac{0.7}{R2}$ amps. The following circuit will actually cause the load current to fall to zero.

If it increases so that $I_L R_2 > 0.7V$, Q4 will turn on, supplying base current to Q3. Q4 thus turns on, supplying further base current for Q4. Regenerative action continues until Q4 and Q3 are saturated. Q3 will then remove all base current from Q1, thus switching Q2 off making the load safe.

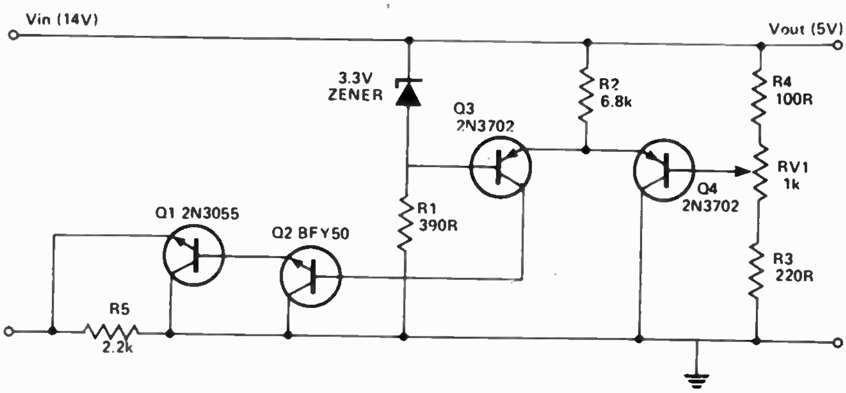
If the reset button is depressed, all current drive will be removed from Q3 and Q4, bringing them out of saturation. On releasing the reset button, the circuit will either revert to normal if the overload cause has been removed, or will snap off again if it is still present.

Care should be taken with earthing to avoid shorting R2.



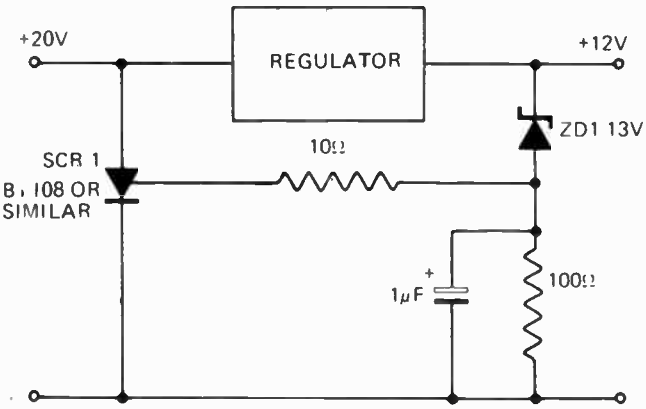
VOLTAGE REGULATOR AND ELECTRONIC FUSE

This circuit improves on those previously published in that current cut-off is achieved, it is self-resetting once that overload is removed and it is an efficient voltage regulator. Choose Z to be about $2/3V_{out}$ and R_1 to supply enough current for stabilization of the Zener voltage. Choose R_2, which determines the cut-off current, I_{max} such that $I_{max}R_2 = (V_Z - 0.5) \times (\beta Q1+Q2)$ and the values of R_3, RV_1 and R_4 so that the base of Q4 is at the same voltage as the base of Q3 and a large current (100 times) passes down the resistor chain compared to the base current of Q4 which is $(V_Z - 0.5)/R_2\beta Q4$. Altering RV1 gives fine control over V_{out}. R5 (200 ohms to 2.2k) allows switch-on under no load conditions. Component values are given for a 5V supply with a 2A cut-out. For low current applications, Q1 can be a BFY50 with Q2 omitted.



PROTECTION

FAST ACTING PSU PROTECTION

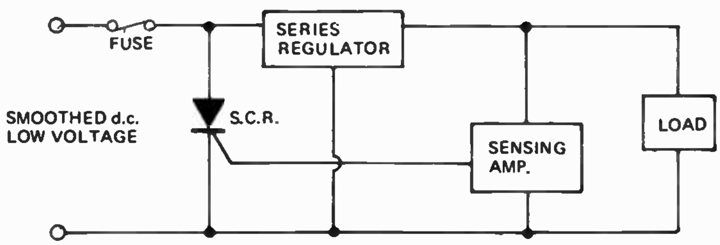


When using a regulated PSU to reduce a supply voltage there is always the danger that component failure in the PSU might lead to a severe overvoltage condition across the load. If the load is a circuit using ICs they can be permanently damaged under such conditions.

A fuse prevents excess current from flowing in a circuit but would generally be too slow to cope with overvoltage situations. The circuit shown is designed to protect the load under overvoltage conditions. Component values given are for a 20V supply with regulated output at 12V. The zener diode can be changed according to whatever voltage is to be the maximum.

If the voltage at the regulator output rises to 13V or above, the zener diode breaks down and triggers the thyristor which shorts out the supply line and blows the main fuse.

IMPROVED SCR CROWBAR PROTECTION



Conventional SCR crowbar power supply overvoltage protection circuits have some drawbacks.

In the conventional circuit the SCR is connected directly across the output of the power supply. For normal operation the negative gate bias on the SCR is such that it remains in the non-conducting condition.

In the event of an overvoltage at the load terminal the sense amplifier applies a positive voltage to the gate, causing the SCR to conduct and effectively short circuit the output so protecting components in the load circuit.

Provided that the series regulator has a current-limiting circuit and that it has not failed, the SCR should maintain its protection until the mains input has been disconnected. In fact unless the overvoltage is caused by a fault in the power supply unit itself, interruption of the mains input is all that is necessary to reset the system.

However, the protection should be fully effective even in the event of a failure in the power supply, as it is potentially the most hazardous in terms of damage to the load.

Internal power supply faults must be considered because they are usually of a sustained nature. In addition to component failure, faults can arise from external causes such as the ingress of swarf or moisture when the equipment is unattended, so that the SCR in a conventional arrangement may have to carry a significant overload for a fairly long period.

It must also be remembered that a fault in a power supply may prevent the current-limit circuit functioning, but the current drawn may not be sufficient to blow the fuse.

By transferring the SCR from the output to the input of the series regulator, full protection against power supply faults is obtained. In the event of an overvoltage the SCR will pass the full short-circuit unregulated current, so that the fuse will blow every time.

In addition the heavy current is only passed momentarily so that complex heat sinking is not required.

This arrangement also gives complete protection against damage due to mains voltage surges, not only to the load circuit but to the power supply as well. However, this arrangement provides only minimal protection against incorrect connection of a separate high voltage source.

Some protection is afforded when a momentary high voltage is applied via a fairly high source impedance because the series regulator emitter-follower would be subjected to a reverse voltage when the SCR went into conduction and would act as a moderately low impedance diode.

Virtually no protection is provided against the application of sustained spurious voltage but even with the conventional arrangement little protection would be provided against a fault of this kind.

For instance, the load and the sensing amplifier are likely to be damaged before the SCR operates, or if it does operate and the incorrect voltage is not removed quickly it would probably be destroyed.

In the light of experience the engineers at Weir Electronics claim that the modified configuration provides better protection than the conventional method. Some degree of compromise is inevitable, but the fact that a positive fuse replacement action is required to restore the supply every time with the second method is in itself a safety factor.

VOLTAGE/POLARITY PROTECTION CIRCUIT

Many circuit, i.e., car radios, can be destroyed if improper voltage or polarity is applied. A simple yet effective technique using only two transistors avoids this possibility.

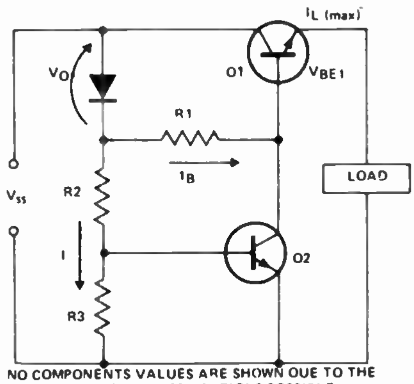
The circuit shown prevents accidental destruction of a load circuit caused by incorrect supply voltage or polarity. This is accomplished without shorting the supply as in SCR and zener protectors. Under normal supply voltage, Q1 is on and Q2 is off provided that:

$$R_1 \leqslant \frac{\beta_1 [V_D + V_{BE1}]}{I_{L(max)}}$$

$$I \geqslant \frac{I_{L\,(max)}}{\beta_1 \beta_2}$$

$$R_3 \leqslant \frac{V_{BE2}}{I}$$

In case the supply voltage exceeds V, Q2 turns on, diverting the base current I_B to ground thus turning off Q1. In the case of wrong polarity, Q1 does not turn on due to the absence of base current I_B which is blocked by diodes D.



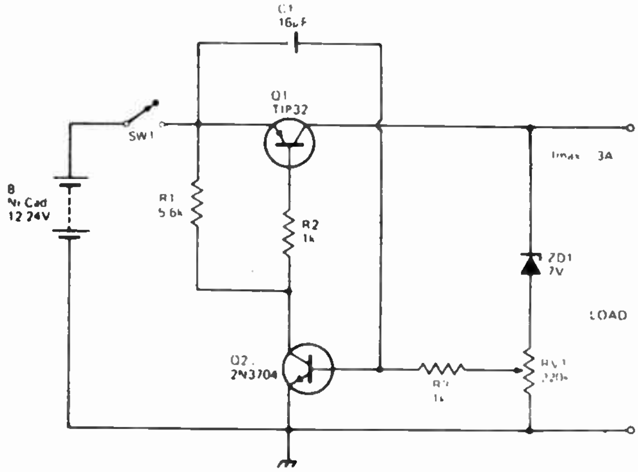
NO COMPONENTS VALUES ARE SHOWN DUE TO THE NUMEROUS CIRCUIT APPLICATIONS POSSIBLE

$$R_2 = (V - V_D)I - R_3$$

NI-CAD DISCHARGE LIMITER

Nickel-cadmium batteries should never be completely discharged as this leads to shortened life. The circuit shown may be used to disconnect the battery from the load when ever output voltage falls below a preset level.

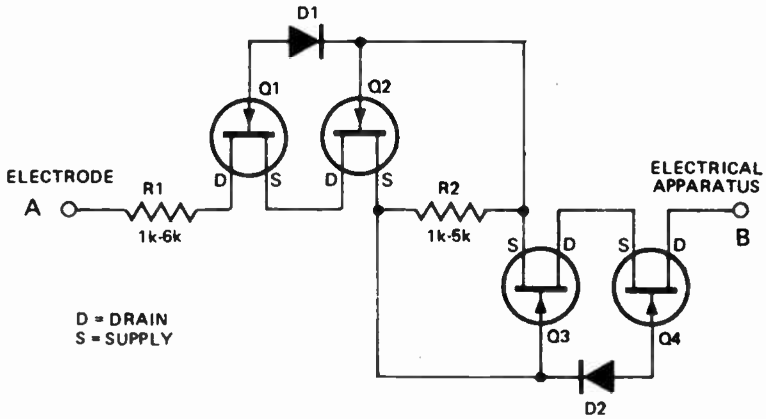
In operation C1 charges through R1 and turns on Q2, the collector current of which flows through R2, turning Q1 on. Thus the battery is connected to the load. When the output voltage falls below a point set by RV1, Q2 turns off, Q1 turns off and further discharge of the battery is prevented.



CURRENT LIMITING CIRCUIT

Danger of accidental shock exists during the use of electrocardiographs and other electrical apparatus that are connected directly to the patient. As part of the Skylab program, a circuit was developed to prevent accidental shock through electrodes to the test subjects.

The circuit allows undistorted signal voltage transfer, as long as the current remains low. If a high current begins to flow from the electrode terminal A toward apparatus terminal B, it will produce a potential difference across resistor R2 (left side of R2 will be at a higher potential than the right side). This potential biases the gate electrodes of the field-effect transistors, Q3 and Q4, to produce an extremely high impedance. Similarly, a current flow in the opposite direction is cut off by a bias on the gates of Q1 and Q2.

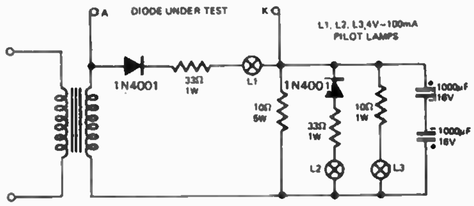


This circuit effectively protects the patient from dangerous shock that could be caused by a failure in the electrical apparatus. When a 1000 Hz signal at 141 Vac (rms) is applied to the terminals of the network, the current is limited to approximately 87µA.

This circuit can also be used to protect sensitive electrical measuring instruments.

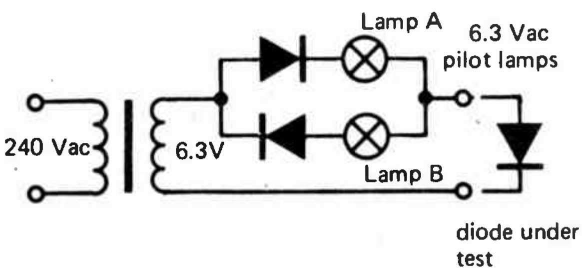
DIODE, TRANSISTOR, SCR

DIODE CHECKER



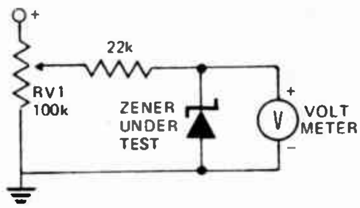
The diode to be checked is connected across the points shown as A and K (observing the polarity indicated). If the diode is functioning correctly, both lamps will light; if the diode is shorted, lamp L2 will light; if the diode is open circuit, Lamp L1 will light.

GO/NO–GO DIODE TESTER



If lamp A or lamp B is illuminated the diode is serviceable. If both light the diode is short circuit. If neither light, diode is open circuit.

ZENER DIODE CHECK

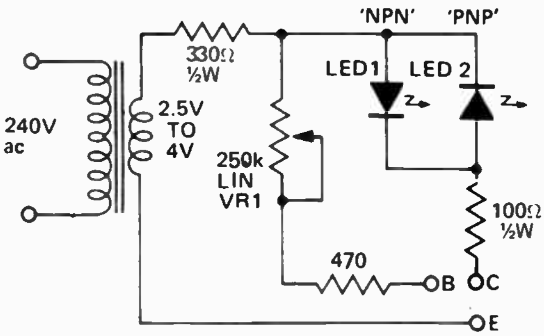


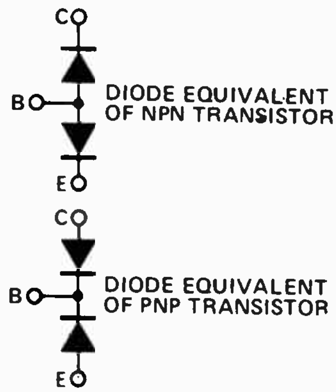
Unmarked Zener diodes may be tested using this simple circuit.

An external power supply giving a voltage higher than the highest expected rating of the Zener diodes to be tested is required.

Potentiometer RV1 is adjusted until the meter reading stabilizes. This reading is the Zener diode's breakdown voltage.

GO/NO-GO DIODE/TRANSISTOR CHECKER





A diode can be checked by connecting it between C and E. If LED 1 lights the diode is OK and its anode is connected to C. If LED 2 lights its cathode is connected to C. If both light it is a short circuit suitable only as a link!

To check transistors with known pin connections, set VR1 at maximum resistance and connect the transistor. Advance VR1 until one LED lights. If LED 1 lights it is NPN, PNP if LED 2 lights. If both light you have a three-legged link. If neither light you have a three-legged fuse!

To check transistor connections, if unknown, short two of its leads together and check as for a diode making note of which lead/leads respond as anodes. Short two other leads together and do it again. Refer to diagrams above.

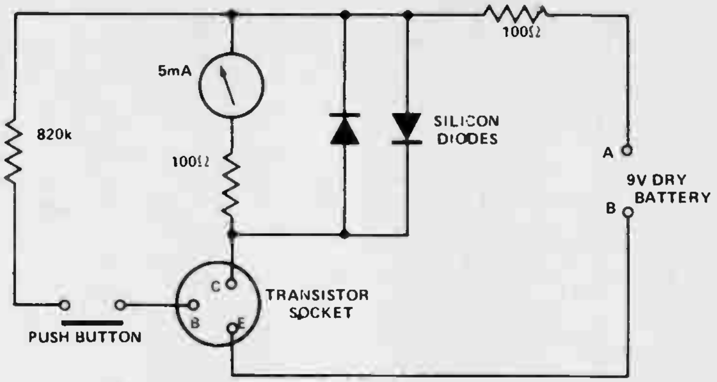
QUICK JFET TEST

A quick test of an N or P-channel JFET is possible using only a standard mulitmeter ohmmeter.

With the ohmmeter connected between source and drain (polarity unimportant) the channel resistance (about 200Ω) will be read. If the gate is now touched with a finger once or twice, the channel resistance should rise to about 10MΩ indicating pinch off. If this does not happen the FET may be assumed not working. Electrostatic pickup from the "mains" charges the gate capacitance and pinches off the FET. The time it takes for the channel resistance to return to normal gives an indication of the gate leakage resistance of the FET.

The relatively low gate leakage resistance, and the high resistance between the finger and the mains helps to prevent destruction of the FET whilst it is being tested in this way.

TEST TRANSISTOR CURRENT GAIN

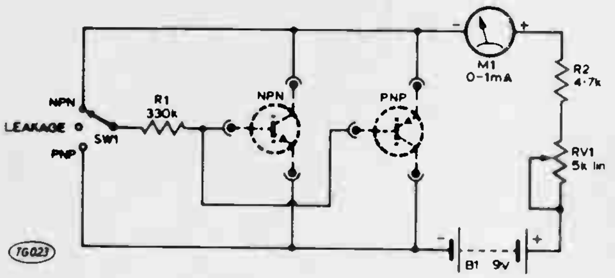


A reasonable estimate of current gain can be obtained from the above circuit. Before the button is pressed, the meter should give negligible deflection. Closing the contacts gives approximately 10μA to the base of the transistor, so every mA indicated by the meter has to be multiplied by 100 to obtain the current gain. The resistors and diodes are to protect the meter in the event of a short circuit transistor being tested.

For NPN transistors, A & B should be + & – whilst for PNP, A & B should be – & +. The meter also needs to be reversed with the change of polarity.

The changeover for both meter and battery could be carried out with a two-way, four pole switch.

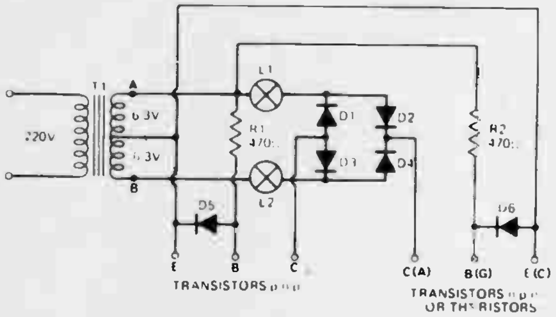
BASIC TRANSISTOR TESTER



There are a number of parameters which determine a transistor's performance but the main ones for low voltage use are leakage and gain. The circuit shown above will indicate both. Two transistor sockets are required, one for NPN devices, the other for PNP; this simplifies switching. With the switch SW1 set on leakage the current passing through the device is indicated on the meter: modern transistors, even germanium types should show only the tiniest reading if any at all. For gain R1 is applied between base and collector. RV1 should be adjusted so that short-circuiting the emitter-collector contacts just registers full scale deflection.

Gain can be directly calibrated onto the scale but this is best done by noting deflections given by inserting transistors of known quality.

SIMPLE TRANSISTOR/SCR TESTER

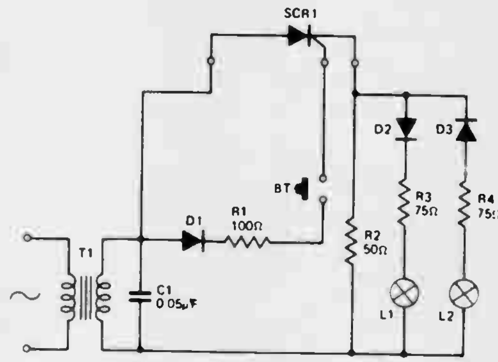


The 6.3-0-6.3V winding of T1 is bridge rectified by D1-D4, the two ac arms of the bridge being connected through L1 and L2 (75 mA maximum). The rectified waveform is applied to the collector of the transistor (or anode of SCR) under test.

The diodes D5 and D6 provide the correct drive polarity for the transistor base or SCR gate.

When testing a pnp transistor, for example, the collector and base are both driven negative when point A of the transformer swings negative. With a good transistor both functions will conduct, the transistor will saturate and L1 will be lit. If the base-collector junction is open circuit L1 will be off and if there is a collector-emitter short both lamps will be on.

SCR TESTER



With a good SCR, lamp L1 (6.3 volt 0.25A) will come on, and stay on, only whilst push button BT is depressed. If Lamp L1 comes on before the push button is pressed the thyristor defective – most probably due to an internal short.

If both lamps turn on simultaneously then the SCR is completely short circuited.

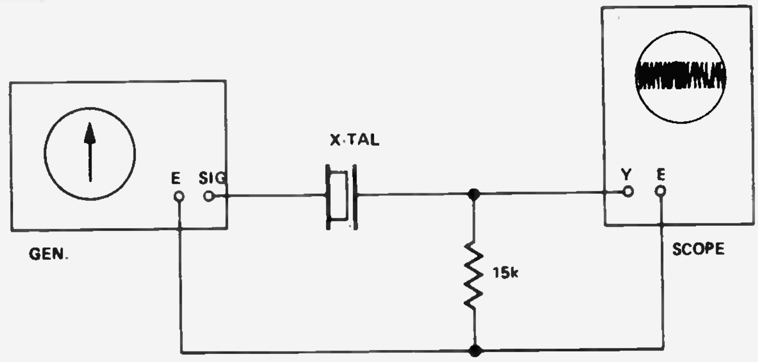
The same circuit may be used to test power diodes rated at 500 milliamp or more. In this case of course a good diode will light lamp L1 and a shorted one will light both lamps. If neither lamp lights the diode is open circuit. The polarity of a good diode will be indicated by which lamp turns on.

Diodes D1-D3 should be capable of carrying 300 mA and transformer T1 should have a 25 volt 300 mA secondary.

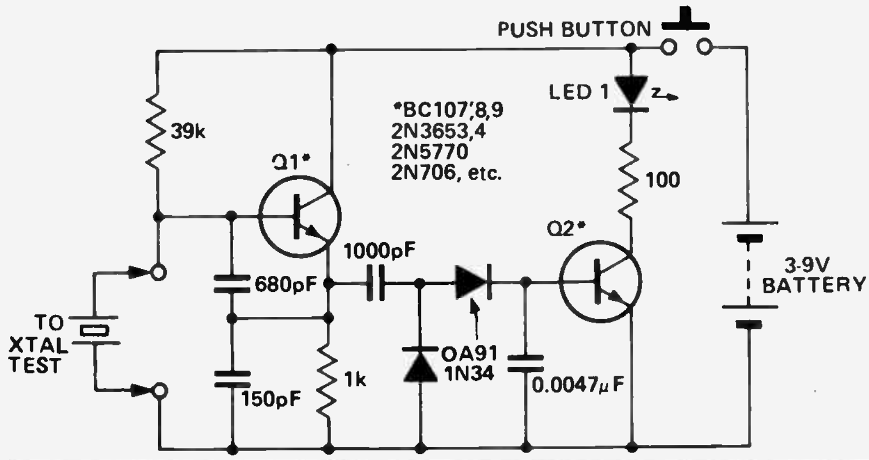
CRYSTALS, BATTERIES

CRYSTAL CHECK-UP

If one has access to a signal generator and oscilloscope, the hook-up shown will check both the generator and crystal. As the frequency is increased, the low impedance series vibration of the Xtal can be observed by a sharp increase in Y amplitude. This is followed by a dip as the Xtal goes into the high impedance parallel mode. The harmonic activity can be checked by comparison with the fundamental.

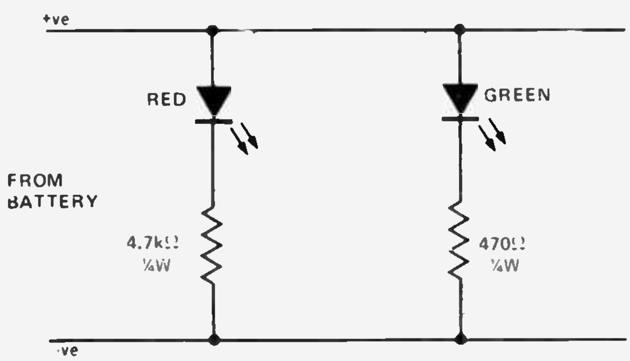


CRYSTAL CHECKER



For checking fundamental HF crystals on a 'Go-No-Go' basis, the above circuit works quite well. An untuned Colpitts oscillator drives a voltage multiplier rectifier and a current amplifier. If the crystal oscillates, Q2 conducts and the LED lights. A3 or 6V, 40mA bulb could be substituted for the LED.

GOOD/NO GOOD BATTERY TESTER

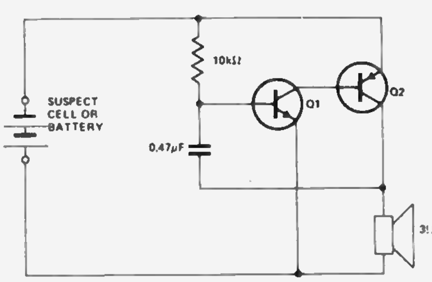


This is a simple tester for use with a PP3 or similar battery.

It is wired to a PP3 battery clip remembering that red is connected to −ve of battery and black to the +ve. It uses 3 small LEDs of the same size: one red, one green. Due to the fact that the green LED needs a far greater current, the green will glow only if the battery is in reasonable condition. The red will glow even if battery is down. If the red glow is very faint the battery is no good.

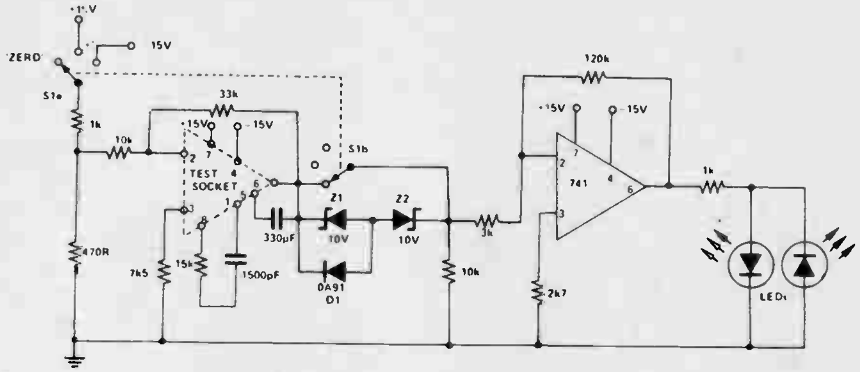
BATTERY TESTER

This device tests the condition of dry cells. The circuit consists of a simple oscillator whose output frequency is relatively independent of supply voltage, but varies greatly with changes in supply impedance. Thus, with the component values shown, a fresh battery or cell will give a note of about 500Hz, whereas an exhausted cell will give a note above 1kHz. The device has been tested with battery voltages between 1.5V and 14V, using a 2N2923 as Q1, and an OC81D as Q2. The unit is undamaged by reversed supply potentials.



OP-AMPS

OP AMP TESTER



The design illustrated is intended for 709, 741 and similar amplifiers which can use ±15V rails, and ±15V *must* be used with the component values shown. For checking amplifiers such as the CA 3130 as well, the rails would have to be dropped to ±8V and Z1 and Z2 reduced accordingly.

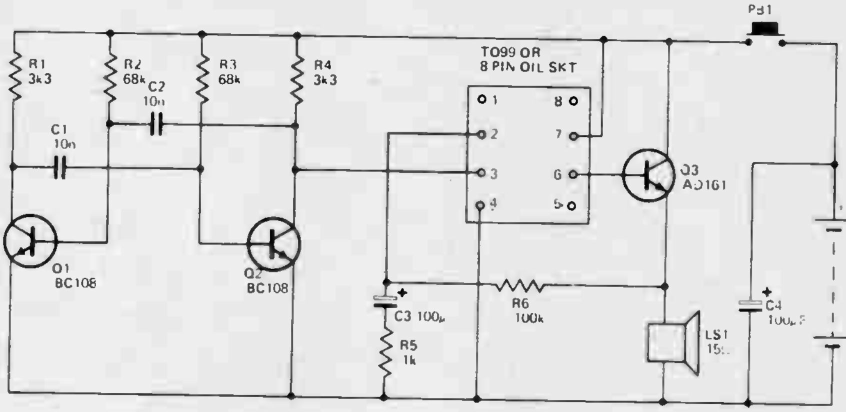
Circuit operation is as follows: with S1 in the 'zero' position neither LED should be on, and one LED on indicates excessive offset, both LEDs on oscillation.

In the other two positions S1 the LEDs can only light if the output of the amplifier on test exceeds the zener voltage. A good amplifier should light the LED corresponding to the position of S1.

D1 is a Ge diode (low voltage drop) which helps to allow for the unequal output drive of 741s by reducing the forward voltage drop across Z1.

S1 can be replaced by two toggle or push-button switches if more convenient.

OP AMP CHECKER



This circuit allows a quick and accurate GO/NO GO test to be made on 741 op-amps.

A 1kHz square wave is generated by the astable multivibrator Q1 and Q2 and the associated components R1–R4 and C1–2. This is fed into pin pin 3 of a standard 8 pin DIL or TO99 IC socket.

Assuming a working IC has been inserted this signal will enter it's non-inverting input and appear amplified at the output, pin 6.

Q3, an inexpensive germanium power transistor is connected as a class A output stage whose load is a 15Ω speaker.

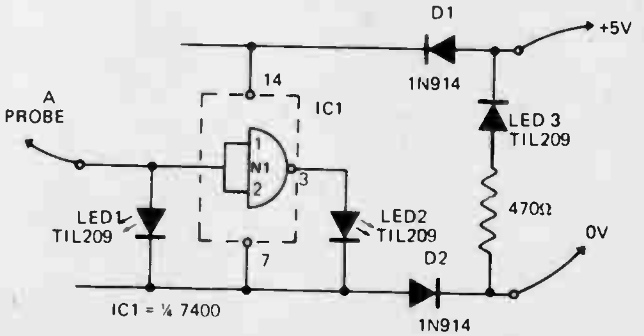
R5 and R6 form the feedback loop to the inverting input whilst C3 isolates this pin from ground.

The battery employed should be a large type and periodic checking of its voltage must be made since a 741 will not work below 6V.

On pushing the test button a good IC will produce a loud note from the speaker. A faulty one will produce little or no output and should be discarded.

LOGIC

CHEAP LOGIC PROBE



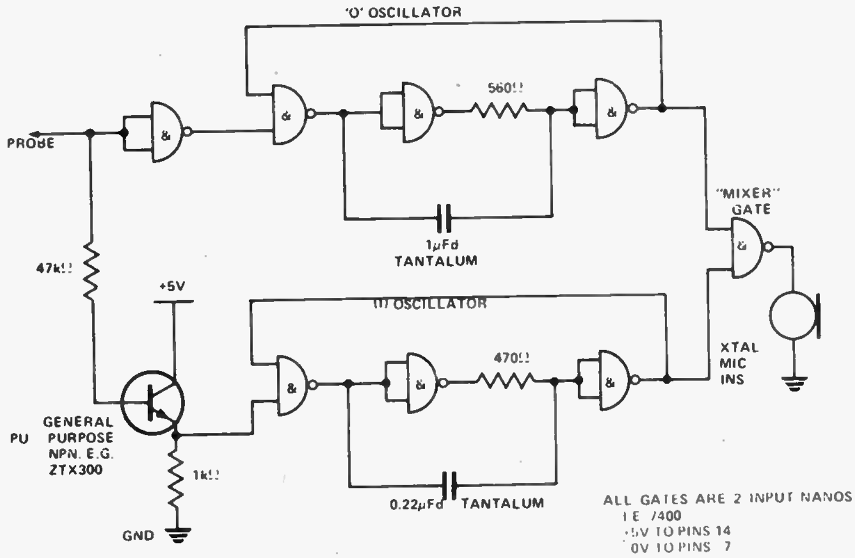
When point A is at '1', LED1 is on indicating the 'Hi' state. When point A is at '0', output of N1 is '1', so LED2 is on indicating the 'Lo' state. Diodes D1 and D2 protect the circuit should it be wrongly connected to the supply, LED3 coming on to indicate wrong polarity.

AUDIBLE TTL PROBE

With many LED TTL logic probes it is difficult to watch the LEDs and the circuit one is testing. The following circuit is an attempt to produce a new sort of probe.

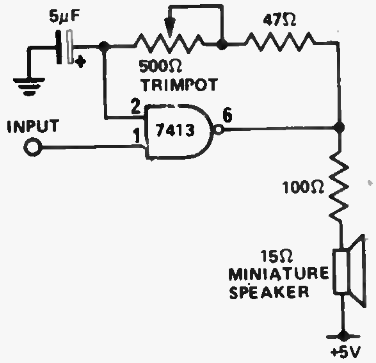
When the probe is in contact with a TTL low (0) the probe emits a low note. With a TTL high (I) a high note is emitted.

The whole circuit uses only two TTL ICs and several auxiliary components. This low component count makes it easy to miniaturise the unit. The power is supplied by the circuit under testing so no battery is required.



MAKING SLOW LOGIC PULSES AUDIBLE

For monitoring slow logic pulses a Schmitt trigger is connected as an oscillator. The trimpot controls the pitch of the output. Very useful as a keying monitor or digital clock alarm. When the input goes high, the 7413 will oscillate.



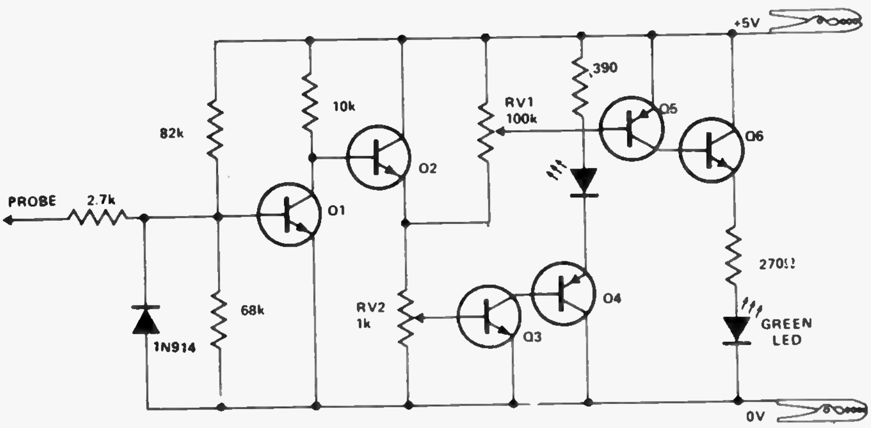
LOGIC PROBE

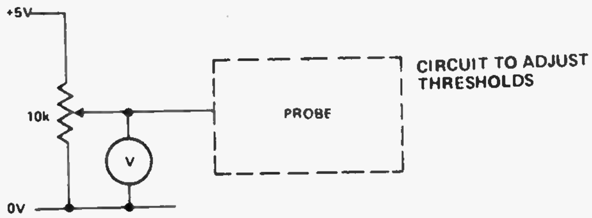
Transistors Q1 and Q2 form a simple voltage buffer providing the probe with a reasonable input impedance.

Q3 and Q4 form a level detecting circuit as the voltage across the base-emitter junction of the Q3 rises above 0.6V the transistor turns on thus turning on Q4 and lighting the red (high) LED.

Q5 and Q6 perform the same function but for the green (low) LED.

Q1, Q4, Q5 are all pnp general purpose silicon transistors (BC178 etc). Q2, Q3, Q6 all pnp general purpose silicon transistors (BC108 etc). The threshold Low ≤0.8V, the threshold High ≥2.4V.



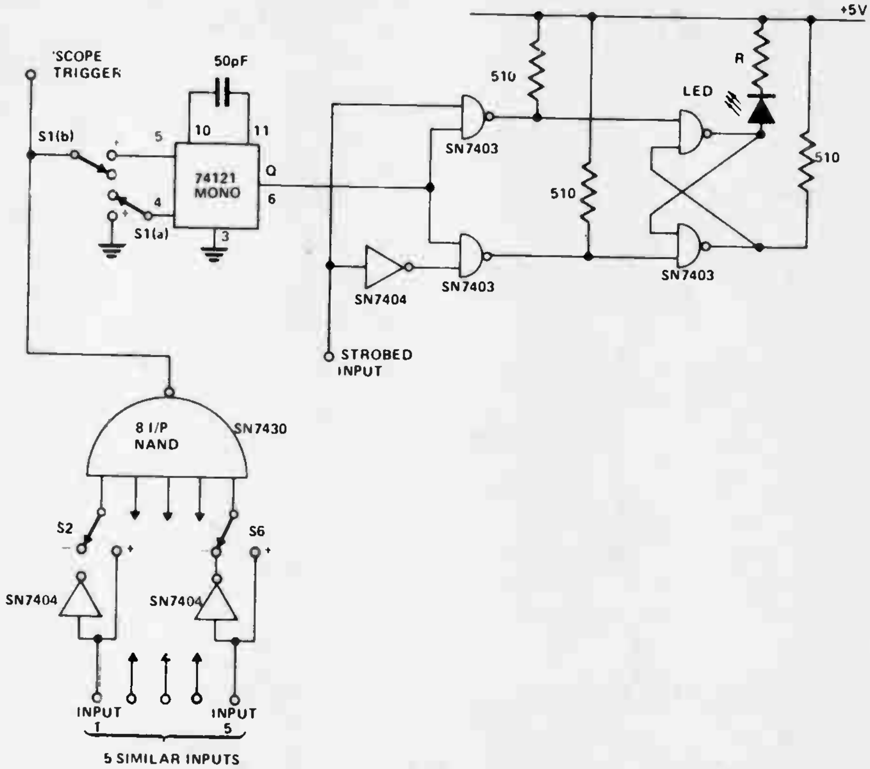


LOGIC ANALYSER

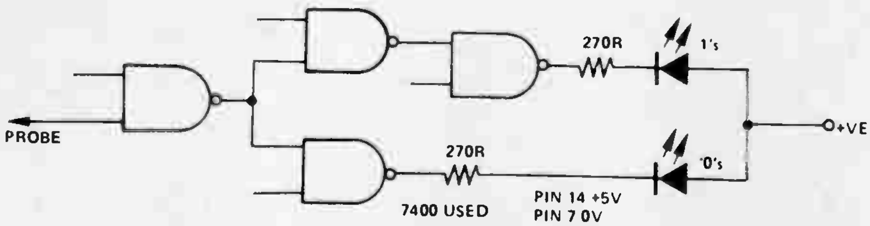
This circuit has been found useful for in-situ testing of TTL logic elements and general circuit development. An LED is used to indicate the state of the strobed input a short time (about 100nS) after the output of the 8 I/P NAND gate changes. Switch S1 is used to control whether the strobe occurs on the leading or trailing edge of the NAND gate.

The five control inputs and associated switches S2 to S6 determine when the strobe occurs relative to logic states in the circuit under test. In use, one or more of the inputs would be connected to appropriate points in the circuit, with switches S2 to S6 as described. For example, the state of a particular circuit point can be strobed each time a selected set of five other circuit points have the logic level 1,0,1,1,0 by setting the switches to +,–,+,+,–, respectively.

An output to trigger an oscilloscope is provided to overcome a common problem in logic circuit testing of finding suitable trigger points in the circuit.



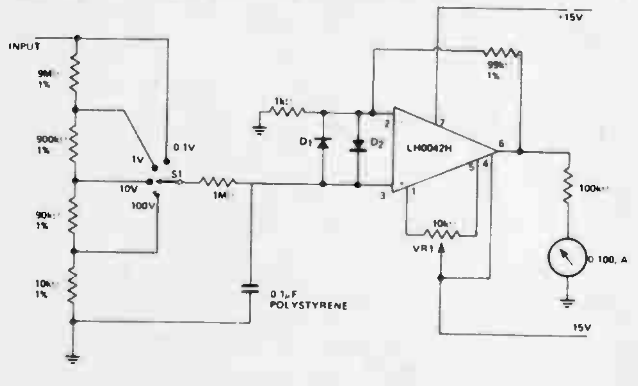
LOGIC PROBE WHICH DISPLAYS 1's & 0's



The logic probe automatically goes to logic 1 when not in use and uses the power supply of the circuit under test.

MISCELLANEA

SIMPLE HIGH IMPEDANCE VOLTMETER



The circuit shown can be used as a high impedance voltmeter with ranges having full scale deflections of 0.1V, 1V, 10V and 100V. The range is selected by the input potential divider connected to S1.

The meter is set to zero by means of the potentiometer VR1 when no voltage is applied at the input. D1 and D2 are low leakage silicon protective diodes, whilst the LH0042H is a relatively low cost FET operational amplifier manufactured by the National Semiconductor Company.

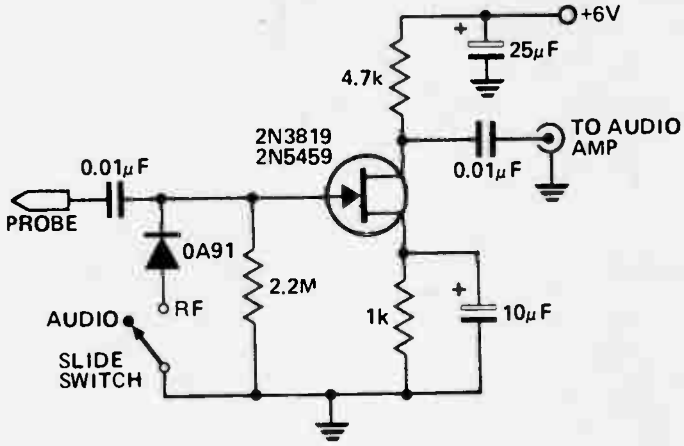
The circuit consumes only about 20mW of power and can be supplied from batteries. Further ranges can be added by including appropriate additional resistors in the input circuit and additional positions in the switch S1.

MISCELLANEA

AUDIO-RF SIGNAL TRACER PRE-AMP

This economical signal tracer is very useful for servicing and alignment work in receivers and low power transmitters. It is easily constructed on a small piece of matrix board which can be mounted inside a commercially-available probe case or homemade probe. The slide switch can be mounted on the probe housing. A miniature toggle switch could be used as a substitute.

When switched to RF, the modulation on any signal is detected by the diode and amplified by the FET. A twin-core shielded lead can be used to connect it to an amplifier and to feed 6 volts to it.

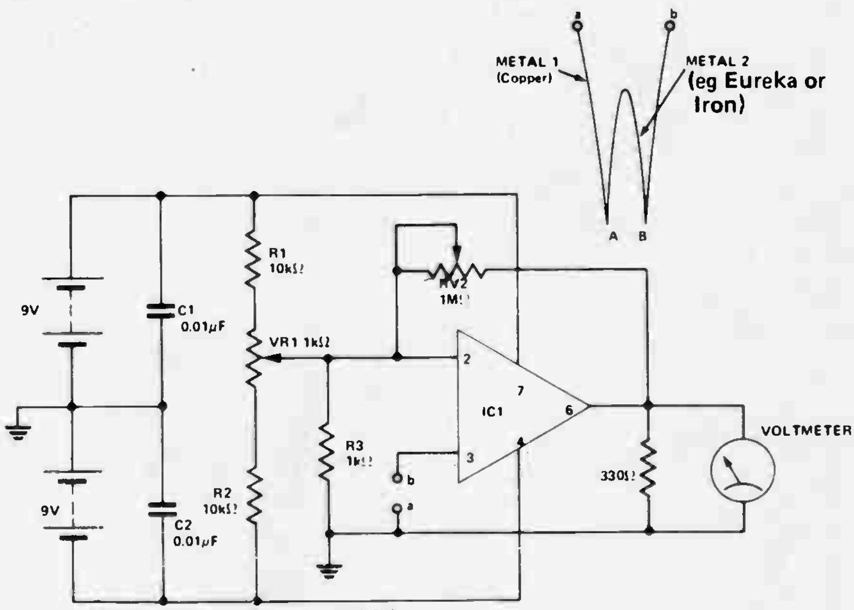


THERMOCOUPLE THERMOMETER

The circuit illustrated was devised to provide a low-cost, sensitive thermometer for measuring temperature differences. The transducer used is a thermocouple consisting of two wires of the same metal, often copper, joined at the two points A and B by a wire of different metal. This thermocouple pair generates a small voltage difference across the points A and B when a temperature difference exists between the junctions a and b. This voltage varies almost linearly with temperature for differences up to about 100°C, although this assumption should not be made in calibrating the thermometer for accurate measurement.

A 741 is used (IC1) for amplifying the small voltage difference between the points a and b enabling a rugged voltmeter to be used to display the temperature difference. The potentiometer is used to set the meter to zero; values of 1kΩ makes setting easy when measuring small temperature differences. However, it may prove necessary to adjust the value of R1 or R2 if zero setting cannot be obtained. If fairly large temperature differences are being measured, VR1 could be increased to 1kΩ.

The sensitivity of the circuit is controlled by the full scale deflection of the voltmeter chosen, on the setting of VR2 (the voltage gain is the ratio VR2/R3), and on the choice of metals in the thermocouple. If the gain of the circuit is set high (at 1,000), electrical noise pick-up and drift become serious problems and it is advisable to assemble the circuit in a metal, earthed box and to ensure the unit is kept at constant temperature.

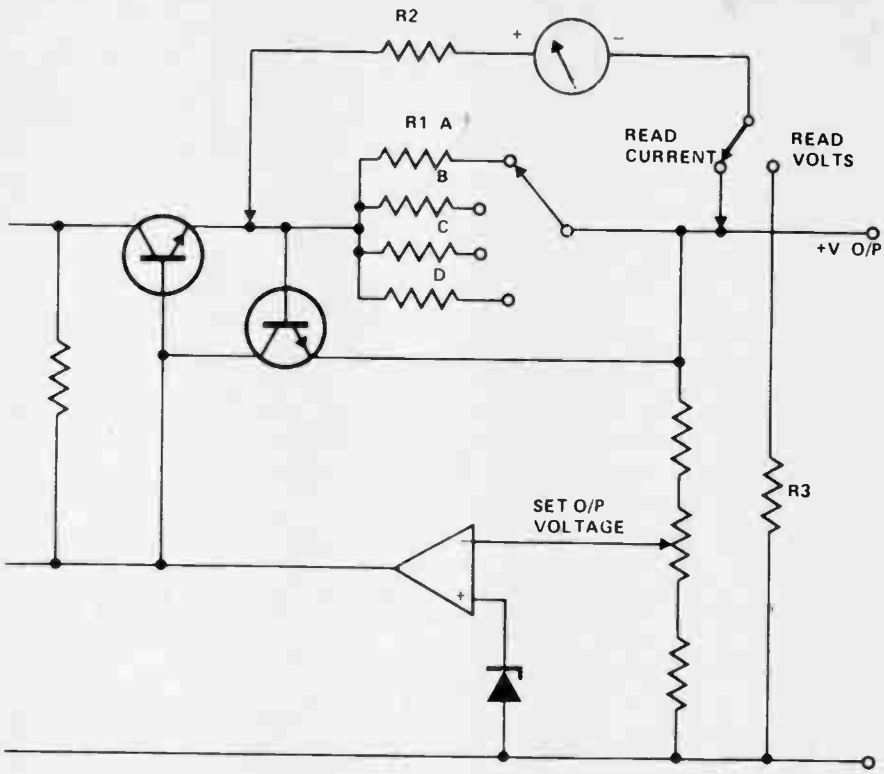


Connections to IC1 are for 8 pin DIL version of '741'

For best results, the power supplies should be stabilised and balanced Capacitors C1 and C2 filter out any electrical noise on the power supply leads; if the thermocouple leads are long, a similar value capacitor across a and b should be used for the same reason.

Calibration and use of the thermometer is carried out by immersing one junction in a liquid at a reference temperature, say melting ice, and using the other junction to monitor the changing temperature.

METERING A STABILISED POWER SUPPLY



It is not easy for the home constructor to make shunts of the correct value for a meter when wishing to alter its current range.

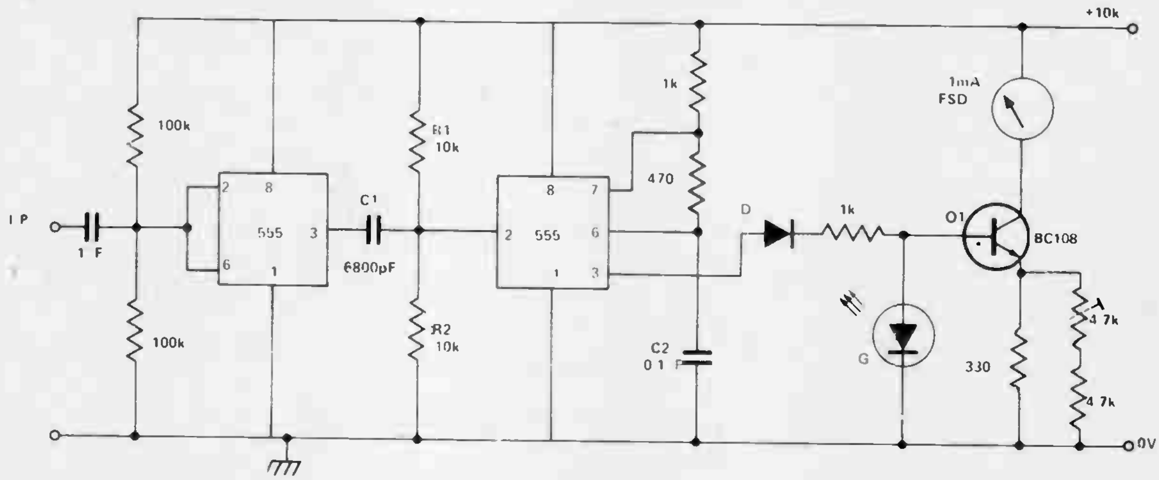
One way to monitor the current supplied by a current limited power supply is simply to measure the voltage drop across the current limiting resistor. This is usually of the order of .65V, and if the series meter (R2) is calculated to give fsd when .65V is applied, will indicate the limiting current at fsd no matter what the value of R1 may be. In effect, the limiting resistor becomes the meter shunt.

For a basic 1mA meter, the series resistor R2 will need to be about 560Ω, for a 50μA meter it will be about 12kΩ, for a 5mA meter about 120Ω and so on.

Unless individual adjustment of resistors and calibration of range is undertaken, this method cannot be absolutely accurate, but it will show whether a circuit is drawing something like its expected current. The method does have the advantage however, of the meter being within the feedback loop, and will therefore not add to the power supply output impedance, which can be important in some applications.

The addition of a single pole C/O switch as indicated will enable the meter to be used to set up the desired output voltage also, though it must be remembered that this will include the voltage drop across R1 if any current is being drawn, which could lead to a difference of .5V or so between indicated and actual output voltage. R3 should be chosen to give fsd at maximum output voltage and the meter scaled accordingly.

SIMPLE ACCURATE FREQUENCY METER



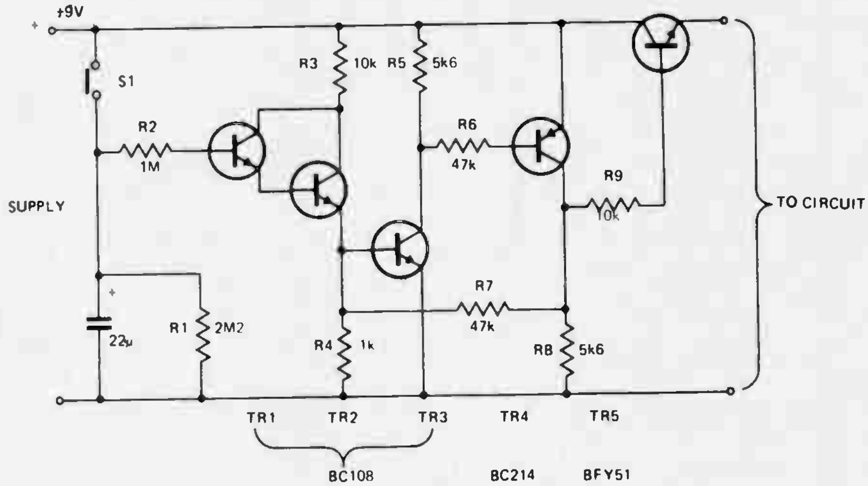
This circuit provides a meter deflection that is strictly proportional to the frequency of the input signal over the range 10Hz–300Hz. The first 555 timer IC is used as a Schmitt trigger, to convert the I/P signal to a fast-edge square wave. This is differentiated by the network C1, R1 and R2, and the resulting spikes used to trigger the second 555, which operates as a monostable, generating constant width pulses. These are used to turn on the constant-current source Q1, so that the average current in the meter movement is proportional to the number of pulses arriving per second. A green LED is used to bias the current source as this gives near-perfect temperature compensation; the 4k7 preset pot gives a fine adjustment for calibration purposes. When the 1mA meter shown is used, fsd is given by 100Hz. To extend the range, reducing C2 to .01μF gives a fsd of 1kHz.

A TIMER WITH LOW STANDBY CURRENT DRAIN

Most timer circuits draw current in both 'on' and 'off' states. For many applications this is no drawback, but for some purposes a timer is required which can remain in 'standby' indefinitely without drawing any supply current. The prototype is used with a toy siren since young "Tech-Tips" features, since young children delight in leaving such devices running indefinitely and a timed cut-out work wonders for parents' mental health! — but more practical uses would include delayed light switches for garages, communal areas in flats, etc, and 'sleep switches' to turn off radios used to lull one to sleep, where it could easily be built into the set.

In operation all transistors are off until push-switch S1 is operated momentarily, charging C1, after which the darlington pair TR1 and TR2 start to conduct. A voltage then appears across R4, and this used to operate the switch proper, TR3 and TR4. These two transistors are used in a positive feedback arrangement to give a sharp switching action as the voltage across C1 gradually falls below a threshold value. Some of the collector voltage from TR4 is fed back to TR3 base through R7 and helps hold it switched on; as TR3 starts to switch off so TR4 turns off and this holding current falls, causing a sharp regenerative switching action to take place. TR5 is driven from TR4 collector and handles the supply current to the circuit being switched. R1 ensures C1 fully discharges since as long as any charge remains TR1 and TR2 continue to pass a small current (1mA).

The time depends on the value of C1 and R1, as shown around 2–3 minutes is obtained, but delays up to ½ hour or so should be easily obtainable with larger values of capacitor. Substituting a pot for R1 would provide some measure of control. C1 should for preference be a tantalum or low-leakage type. Almost any small silicon transistors should work, but germanium types are not recommended as their leakage currents might cause problems.

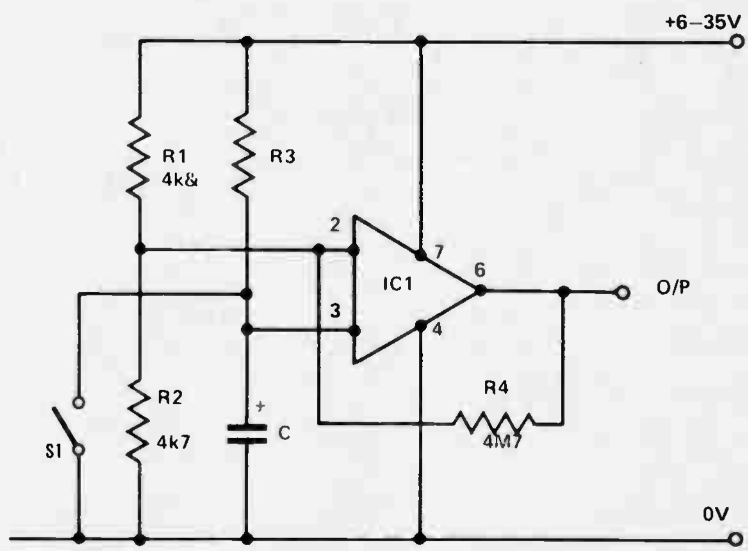


741 TIMER

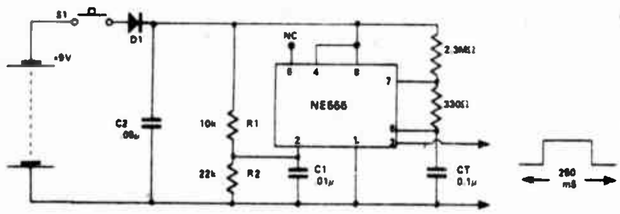
The circuit shows a very simple timer based on a 741 op amp.

R1 and R2 hold the inverting input at half supply voltage. R4 applies some feedback to increase the input impedance at pin 3, but its value is such that negligable damping of pin 2's voltage occurs. Pin 3, the non inverting input, is connected to the junction of R3 and C. After S1 is opened and C charges via R3. When the capacitor has charged up sufficiently for the potential at pin 3 to exceed that at pin 2 the output abruptly changes from 0V to positive line potential. If reverse polarity operation is required simply transpose R3 and C.

R3 and C can be any values and time delays from a fraction of a second to several hours can be obtained by judicious selection. The time delay is 0.7CR seconds where C is in Farads and R in ohms and hence is completly independent of supply voltage.



A SELF-TRIGGERING TIMING CIRCUIT

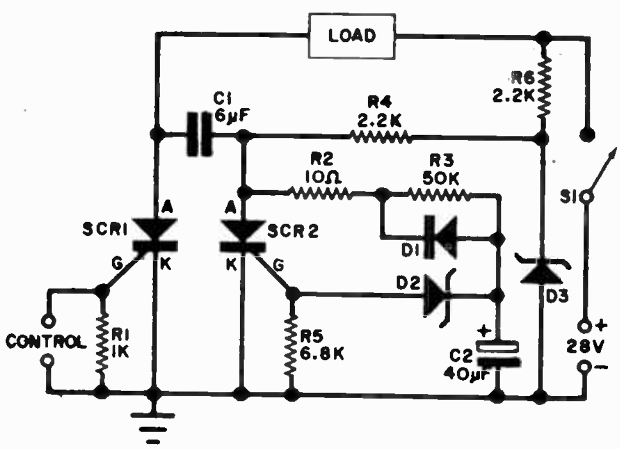


This circuit demonstrates a method of triggering a single time-period from a 555 timer supply by applying power to the circuit, thus eliminating the need for two switches or a two-pole push switch. The original application required that a single 250mSec pulse be produced when a push-button was operated, and the method shown allows the timer circuitry to be switched on and triggered simultaneously by a single MAKE contact. This also economises on battery power as there is no stand-by period.

A 555 timer triggers when the voltage on Pin 2 falls below $1/3V_{CC}$. When S_1 is closed Pin 2 remains briefly below $1/3V_{CC}$ due to the finite time C_1 takes to charge via R_1, and then climbs to $2/3V_{CC}$. Meanwhile the 555 has triggered and the time-period has begun. When S_1 is released, after the end of the period, C_1 discharges through R_2 and the circuit is ready for re-triggering in less than a millisecond. C_2 is essential to prevent supply-line transients resetting the 555 as soon as it triggers, and D_1 provides reverse-polarity protection.

Timing components (C_t and associated resistors) are shown with values for 250mSec pulses, though a wide range of different values can be used.

PULSE TIMER



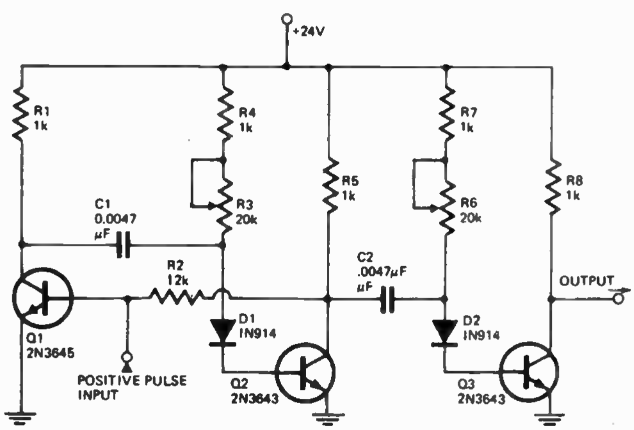
Originally developed by the Unitrode Corporation, this circuit will supply up to four Amps for a period of one second when triggered by an external positive going pulse.

The timing cycle is initiated by applying a positive going pulse of a voltage exceeding SCR1's gate voltage. This causes SCR1 to lock on and apply power to the load. In the meantime SCR2 is switched off by the commutating action of C1 and timing capacitor C2 is charged via R2, R3 and R4 and the constant voltage source D3 and R6.

When C2's charge voltage exceeds the rating of Zener diode D2 by an amount sufficient to exceed the gate voltage of SCR2, this SCR conducts, discharging both C1 and C2 and thus switching SCR1 back into its former non-conducting state.

Zener diode D2 should be rated at 6.8 Volts, and Zener D3 at 18 Volts.

ECONOMICAL PULSE DELAY

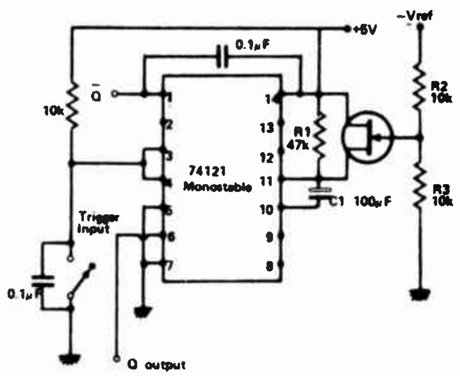


The circuit shown is of an economical pulse delay which utilizes only three transistors instead of the four normally required for a pair of one shots.

Initially Q1 is off, and Q2, Q3 are on. When an input pulse arrives it turns Q1 on, and Q2 off via C1. Transistor Q2 remains off for a time determined by the time constant of C1, R3 and R4, and when it reverts to the 'on' condition it also turns off Q1. Transistor Q3 which is triggered by the output of Q2 stays off for a time determined by R6, R7 and C2, and thus produces a delayed pulse whose duration is approximately 10µsecs, and is adjustable by R6.

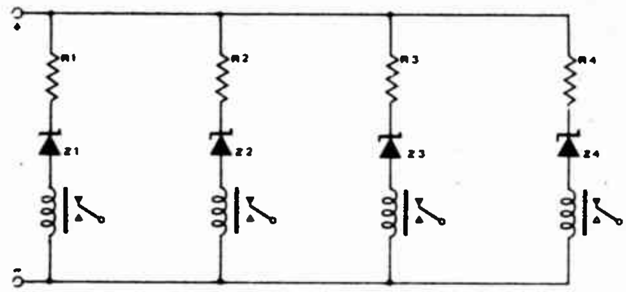
Delay time is adjustable by R3 and for the values shown, is about 10µsecs.

VOLTAGE CONTROLLED MONOSTABLE



This circuit was used to switch a motor on for a variable period as part of a position control system. Using the components shown a range of 20mS to 2 seconds may be obtained, when Vref, is varied between 0 and –5V. The maximum period is governed by the value of R1 (≈0.7R1C1). The minimum by the drain-source resistance of the FET with no gate voltage applied. The FET acts as a voltage controlled resistor in the charging circuit of a 74121 monostable. The FET used was an N-channel 2N3819. If a P-channel device is used, R2 must be taken to + Vref.

SEQUENTIAL CLOSING OF RELAY SERIES



This circuit provides sequential closing of a relay series by appropriately increasing an applied voltage.

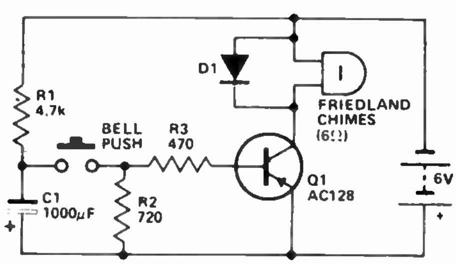
All relays have similar coil resistances but are in series with zener diodes of different voltages.

As the applied voltage exceeds the zener's breakdown voltage, that zener will conduct, thus energizing its associated relay.

The exact voltage at which each relay is energized is determined solely by the zener diode, and not by individual relay characteristics.

By using 5% tolerance zeners from 3.9 volts to 30 volts, 22 selective steps can be obtained.

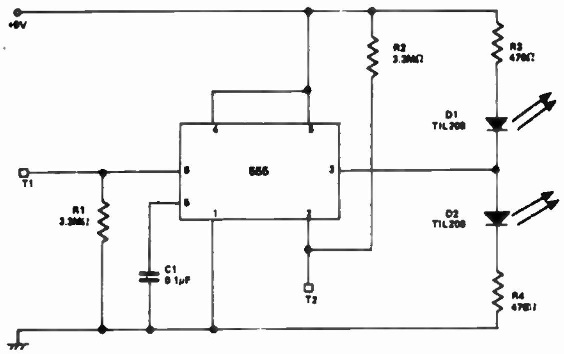
DOORCHIMES DELAY



Ever get tired of people who repeatedly press your doorbell?

With values shown, this simple circuit will permit one operation every 10 seconds or so. Capacitor C1 charges through R1 when the button is released. Making R1 larger will increase the delay.

TOUCH TRIGGERED BISTABLE



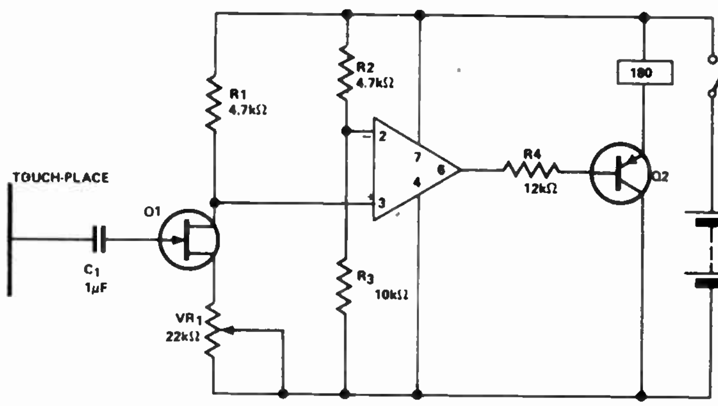
This circuit was devised as part of a touch controlled lighting system. It uses a 555 timer operated in the bistable mode.

Due to the high input impedance presented by the threshold and trigger terminals the 555 can be set and reset by the touch of a finger. Touching T2 causes the output to go high; D2 conducts and D1 extinguishes. Touching T1 causes the output to go low; D1 conducts and D2 is cut off.

The output from pin 3 can also be used to operate other circuits e.g. a triac controlled lamp. In this case the LEDs are useful for finding the touch terminals in the dark.

The capacitor C1 is not absolutely necessary but helps to prevent the circuit triggering from spurious pulses.

TOUCH-SENSITIVE SWITCH



The circuit illustrated can be set to energise the relay when the plate is lightly touched. Under certain circumstances the proximity only of the body is sufficient to operate the switch.

A high impedance input is provided by Q1, a general purpose field effect transistor such as 2N3819. A general purpose 741 op-amp is used as a sensitive voltage level switch and this in turn operates the current buffer Q2, a medium current pnp bipolar transistor, thereby energising the relay which can be used to control equipment, alarms etc.

In the quiescent state, the voltage at pin 3 of the op-amp is set higher than the voltage at pin 2 by adjustment of VR1. This ensures that the voltage at pin 6 is high and Q2 and the relay are off. Upon lightly touching the touch-plate, a decreasing reverse bias V_{GS} increases the drain current flowing through Q1 and the resultant voltage drop across R1 lowers the voltage at pin 3 below that at pin 2. The voltage at pin 6 falls and switches on the relay via Q2. Resistor R4 may need to be selected to ensure that the relay is held off since a small positive voltage at the output remains even though the voltage at pin 3 is lower than that at pin 2 in the quiescent state. This problem can be overcome by using dual power supply for the op-amp in the more usual mode of operation of this device. Component values are not critical and there is considerable scope for experimentation.

The sensitivity of the circuit to the proximity of the body depends upon the nature and strength of the surrounding electromagnetic fields produced by mains wiring and equipment in the vicinity, for it is the pick-up of this energy which the body couples to the circuit.

ELECTRONIC SWITCH

The switch in this circuit uses an N channel FET to present either a high or low impedance path to ground for any incoming signal.

The main advantage of such a switch is that the actual switching of an audio or RF signal can be done in-situ on the board rather than bringing the signal along a cable to and from a mechanical switch.

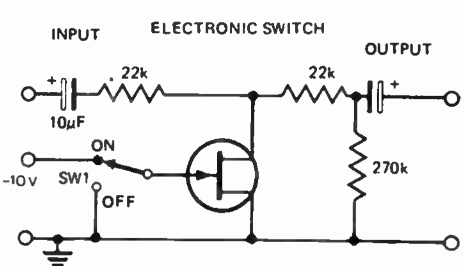
This eliminates hum pick up and other stray problems.

The mechanical switch simply switches dc to the FET gate.

Another feature of the circuit is that one mechanical switch is sufficient to key a number of FET switches with no crosstalk between channels.

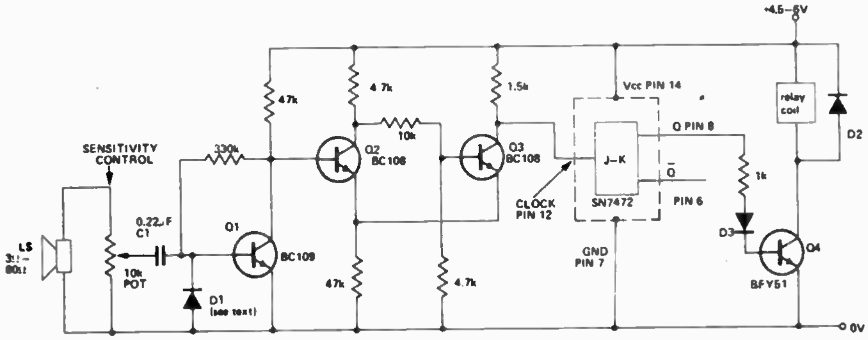
The operation is that when the switch is in the "off" state the FET is biased hard on. Any incoming signal is effectively shorted to ground. In the "on" position the FET is biased to the non-conducting region thus presenting a high impedance to ground. This allows the incoming signal to appear at the output terminals unattenuated.

The output impedance of the circuit



is high and the following stage impedance should be in excess of 50 k if excessive loading is to be avoided.

SOUND OPERATED TWO-WAY SWITCH

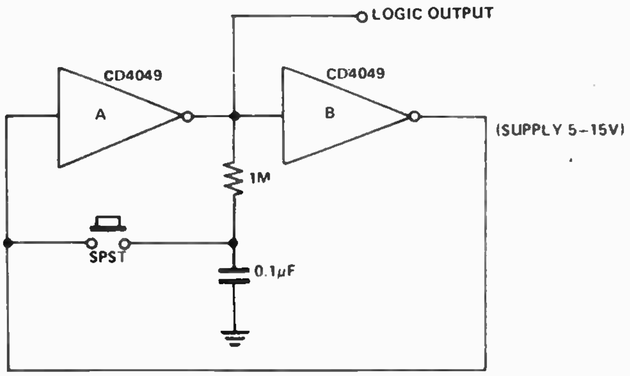


The circuit operates a relay each time a sound of sufficient intensity is made, thus one clap of the hands will switch it one way, a second clap will revert the circuit to the original condition. Q2 and Q3 form a Schmitt trigger. The JK flip-flop is used as a bistable whose output changes state every time a pulse is applied to the clock input (pin 12). Q4 allows the output to drive a relay.

Under quiescent conditions Q1 is on, holding the base of Q2 low and keeping the output of the Schmitt trigger low (Q3 collector). If a sharp noise is made (e.g. a clap) it will generate a pulse in the loudspeaker which is fed through C1 and switches Q1 off. D1 prevents any large pulses damaging Q1. As Q1 switches off, its output goes high causing the output of the schmitt trigger to go high. When the clap is finished Q1 again conducts, causing the output of the schmitt trigger to go low. Therefore each clap causes a high pulse at the Schmitt trigger's output which is fed to the clock of the JK flip-flop causing it's output to change state. This is used to turn a relay on and off. Because the circuit is only sensitive to sharp noises it is generally unaffected by talking or sounds caused by movement. (The sensitivity control can be adjusted to prevent such noises triggering the circuit if this does arise). A moving coil loudspeaker is used as a microphone as it can respond to sounds from any direction. It was found that any loudspeaker from 3–80Ω worked in the circuit. The Q̄ output of the JK flip-flop could be used as well, allowing two relays to be switched on and off complementarily.

The circuit has limitless applications like turning on a radio or controlling motorised toys by clapping. The diodes can be any general purpose silicon types (1N914 etc) and the relay a 5–6V type with minimum resistance of 50 ohms.

SPST SWITCH FLIP FLOP



The circuit gives latching on-off action with a single SPST switch, utilizing the high input impedance property of CMOS logic.

It can be seen that C will go to the logic state opposite to that existing at the input to inverter A. On closing the contact, the input to inverter A is taken to the opposite logic state momentarily and the latch flips.

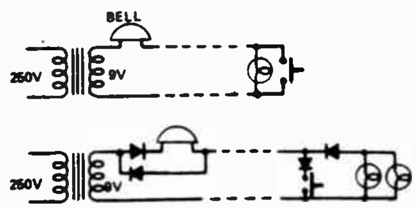
The RC time constant of about 100 milliseconds provides sufficient protection from switch bounce yet gives quick recovery for the next operation. The output may drive other logic gates or external loads through buffer circuits.

TWO CONTROL SIGNALS DOWN ONE WIRE

Many houses have a door-bell push lit up by a small bulb as in Fig. 1. A second bulb was required to light up the house number - but the two bulbs together used so much current that the bell trembler was continually buzzing. Running a separate feed from the transformer to the front door was inconvenient.

The solution was to use four small cheap diodes as in Fig. 2.

Almost any wire ended diode will work - (0.5A, 50V if in doubt). This idea has many other extensions - wherever it is required to send two control signals or two power supplies down a single pair of wires.



INDICATORS

LEVEL

ECONOMY 'LINE-O-LIGHT' DECODER

The decoder works in the following fashion: one NAND gate of IC2 (a 7413) is wired as a clock pulse generator, running at approximately 10kHz. These pulses are used to clock IC4, a 7490 4-bit BCD counter, which in turn drives a 7441 BCD-to-one of ten decoder. The line of ten LEDs is arranged with the anodes linked together and coupled 50 +5V via TR1. The ten LED cathodes can be pulled down to 0V by the 7442 outputs. Thus, if TR1 is turned on, the counter will scan through 0-9 and the LEDs will turn on in sequence. If the clock frequency is high enough, all ten LEDs will appear to be on all the time.

In Fig. 1, the second half of IC2 is used to detect when the counter outputs are at 9, and this signal is used to trigger the 555 timer, IC1, which is connected as a monostable. The output of the mono drives the base of TR1. Thus, every time IC4 reaches 9, IC1 will fire and switch the LED array on.

The voltage input to the decoder, Vin, is fed to the control voltage input of IC1, altering the threshold at which the internal comparators of the 555 will switch. This in turn decides (in conjunction with the values of R3 and C2) the length of pulse available at the mono output. IC1 is, in effect, used as a pulse-width modulator.

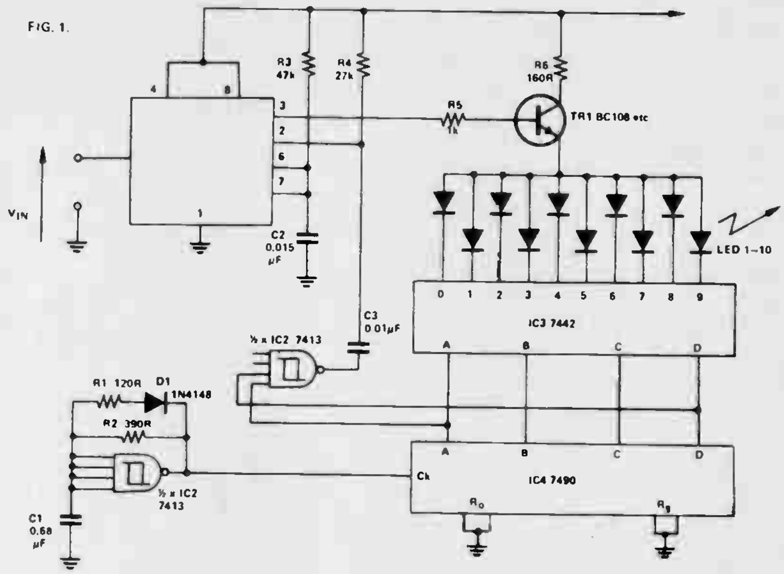
Fig. 2 shows the relative timing of the clock, 555 trigger and 555 output waveforms, together with the 'ON' times of the individual LEDs.

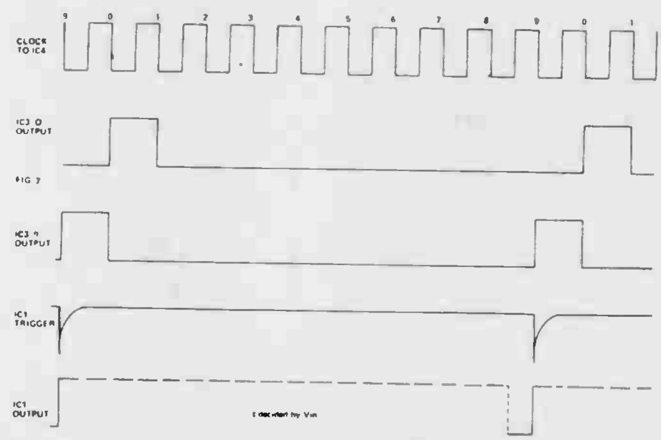
If the value of Vin is such that the 555 pulse length is approximately equal to nine clock pulses, then TR1 will be on almost continuously and the LEDs will all appear to be on. As Vin is reduced, TR1 ON time will become less than nine clock pulses; consequently, the +ve supply to the LEDs will be removed at the same point in every count cycle. This will have the effect of 'shortening' the row of illuminated LEDs. Because the mono pulse length does not decrease in definate steps of one clock pulse, the row of LEDs will decrease in length by the highest gradually dimming, then the next highest, etc. — a rather pleasing effect.

Some experimenting with the value of C2 may be required to achieve a realistic rane for Vin, and to prevent the mono pulse length being greater than ten clock pulses at maximum Vin.

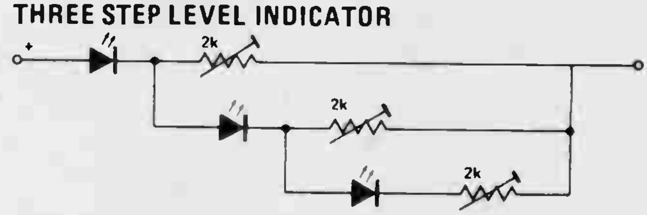
The values for C1, R3 and C2 given are those used in the prototype.

FIG. 1.





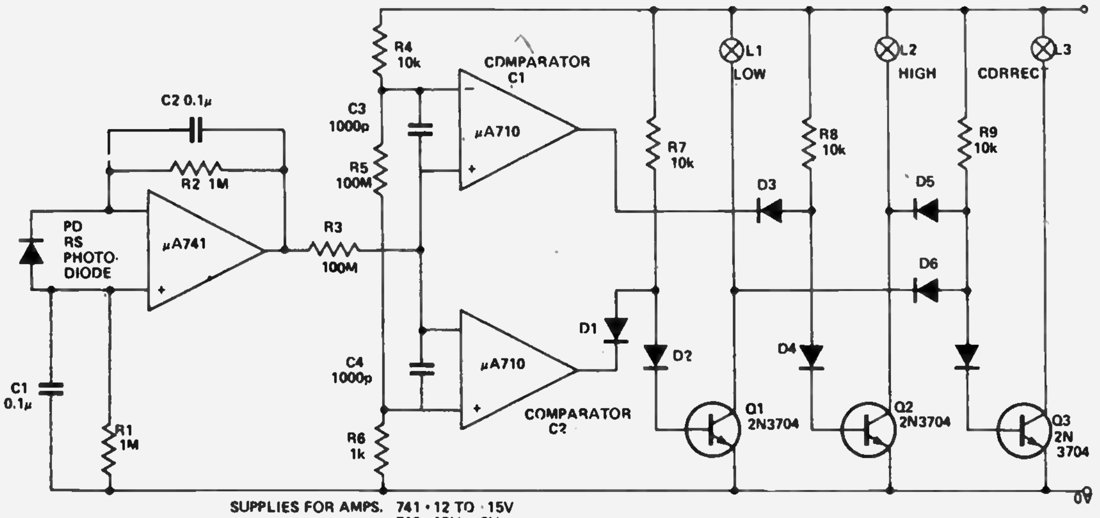
THREE STEP LEVEL INDICATOR



This device makes a very compact and robust level indicator where a meter would be impractical due to lack of space, or not justified due to cost.

Resistor values will depend on type of LED used. In the prototype, the LED's were MV50's and the resistors were 2kΩ ½watt. This gave steps of approx 2V and the current drain with all three LED's on was 5mA. The chain can be extended but current drain increases rapidly and the first LED carries all the current drawn from the supply.

LIGHT LEVEL INDICATOR



SUPPLIES FOR AMPS. 741 · 12 TO · 15V
710 · 12V, −6V
SIGNAL DIODES 1N4148 OR SIMILAR

When conducting optical experiments, or calibrating photocells, it may be necessary to set a known light level each time before the experiment is performed. The circuit provides a simple means of setting a light level to a particular value.

A silicon planar photodiode, strategically placed in the optical system, generates a photocurrent proportional to the incident illumination which is fed to the input of an op amp connected as a current amplifier. The output is thus the equivalent photocurrent developed across a 2Mohm resistor.

Two comparators are used to compare the output voltage with a fixed reference set by a potential divider chain. Comparator 2 is set at nominally 1V and Comparator 1 at 1.1V.

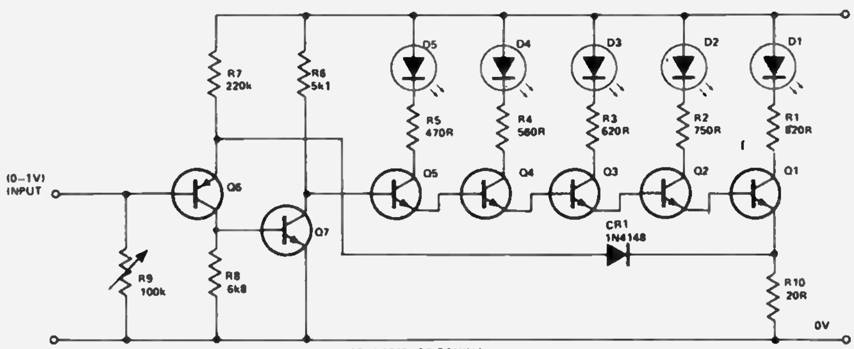
The amplifier output is fed via R3 to the inverting input of comparator 2. When the output is below 1V, the output of comparator 2 is positive which enables the current in R7 to turn on Q1, lighting lamp 1 indicating "Too Low". When the output of the amplifier is above 1.1V the output of comparator C1 will be positive, enabling current in R8 to turn on Q2 and lighting lamp L2 indicating "Too High". If the amplifier output is between the two thresholds, both comparator outputs will be low, both lamps will be off, and the current in R9 will be enabled to Q3 and L3 will light giving the green indication "Correct"

Changing the values of R1 and R2 alters the basic sensitivity of the system, C1 and C2 provide decoupling of noise pick up for remote direction or small content of AC lighting and R3,C3, and C4 minimise instability in the comparators as they pass through their linear region.

Values in the diagram shown give an acceptance band of 10%. Reducing the value of R4 to 50ohms reduces the pass band to 5%. For closer bands, higher gain comparators may be used (eg. µA734 or LM311), but light levels closer than this are rarely necessary.

BARGRAPH DISPLAY

A bargraph display is a useful medium for seeing a monitored variable. Where low resolution (5 to 10 segments) is sufficient the display can be built with LED's and a few transistors.



Q1 Q5 2N3568 (OR EQUIV.) Q6 2N4250 (OR EQUIV.) Q7 2N3565 (OR EQUIV.)

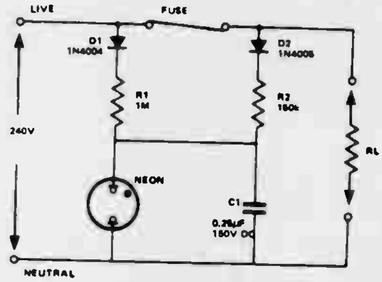
With the 5 segment system shown, transistors Q1 to Q5 saturate successively as the input signal increases from zero. The resulting currents drive LEDs D1 to D5. As each transistor turns on, its emitter current flows through R10. Transistors Q6 and Q7 as well as CR1 and associated resistors, comprise a feedback amplifier that forces the voltage across R10 to equal the inputs voltage. This causes the display to 'deflect' linearly.

For R10 = 20R and a current of 10mA per LED, the resolution is 200mV and the full scale input equals 1V (for five LED's). Diode CR1 cancels the V_{BE} offset of Q6. Resistors R1 through R5 control the LED currents. The voltage across R3 for example is 10V minus 1.5V (two transistors V_{BE}'s) minus 0.6V (30mA – R10). Since V_{CE} (SAT) of Q3 is negligable at 10mA, 6.4V must be dropped.

i.e. $R3 = \frac{6.4V}{.010A} = 640R$. 620R being the nearest standard value.

WARNING

FUSE FAILURE INDICATOR

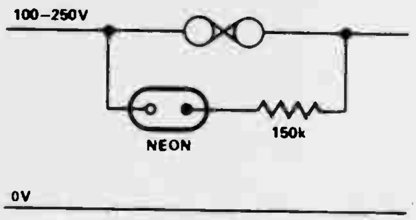


The circuit is built around the neon indicator which is normally used to show that power is being supplied to mains-driven equipment. When the fuse is intact, the neon is lit steadily as normal. However, should the fuse blow or be removed, the indicator flashes at a suitable rate, drawing the attention of the operator.

Effectively, the circuit is a simple modification to a neon relaxation oscillator. Under normal conditions, the time-constant of the RC network is such that the flash-rate of the neon is not detectable by the eye. The removal of the link between the anodes of D1 and D2, however, increases the time-constant and the neon flashes can be clearly seen. The specified component value give a frequency of approximately 2Hz.

An advantage of the circuit is that it will operate regardless of load impedance. Points to note: component values are not critical, but because the capacitor charges up to the striking voltage of the neon, diode D2 must have a PIV rating greater than peak mains plus this voltage. D1 can be rated at peak mains or greater. The types specified are suitable easily-obtained devices. Also, built-in resistors in certain types of neon indicator supplied for use at common supply voltages must either be removed or 'shorted'. The resistors should be ½W.

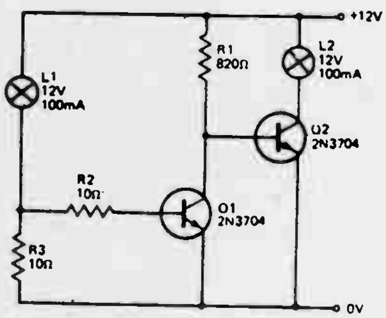
BLOWN FUSE INDICATOR



Here is a very simple method of identifying a blown fuse. This is of course more advantageous on systems employing several fuses.

Across the fuse holder is wired a neon in series with a resistor. When the short circuit, or whatever, blows the fuse, the neon will light indicating immediately the area of the fault. Neons with built-in resistors need not of course have an extra 150k as shown.

BACK-UP FAILURE LAMP



A signal lamp used to indicate failure of a vital piece of equipment is subject to failure itself.

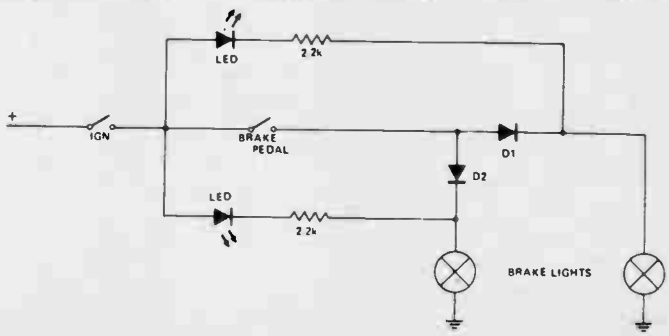
This circuit provides a back up lamp which only comes on if the first lamp fails. As long as L1 is intact Q1 is saturated and Q2 is OFF. If L1 fails Q1 turns OFF Q2 turns on, and lamp L2 is illuminated thus indicating the fault and the failure of L1. The values shown are for 12 volt 100 mA lamps.

If other lamps are to be used the supply should be the same as the lamp nominal voltaoe and the following values should be used:–

LAMP	R1	R3	
6V	50 mA	22 ohm	820 ohm
6V	450 mA	2.7 ohm	220 ohm
12V	250 mA	3.9 ohm	330 ohm
24V	50 mA	22 ohm	820 ohm

DC LAMP FAILURE INDICATOR

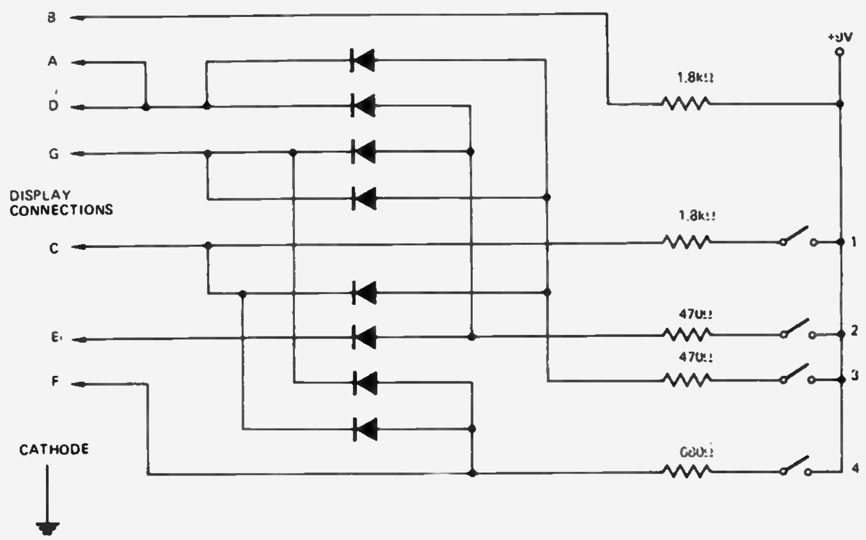
The very simple circuit here provides an on-if-good function for a lamp. D1 and D2 should be generously rated as they are outside the warning loop. On a car type DO4 is recommended for mechanical support. If all the lights on a car are to be monitored the diodes can be mounted in blocks behind the lamp housing that the wiring harness reaches first. A 'line of light' type LED makes a convenient display.

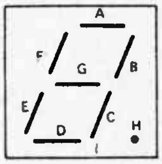


STATION INDICATOR FOR FM TUNERS

Modern FM equipment using varicap tuners generally have four stations pre-tuned and selected by push buttons in addition to a continuously variable control. Most users of such tuners would presumably have the pre-set pushbuttons set to BBC Radios 1,2,3,4. It is, then, quite a simple matter to utilise the unused switch positions which, more often than not, exist in the pushbuttons to set up a display of 1,2,3 pr 4 on a 7 segment LED using blocking diodes and limiting resistors as shown in the diagram. Any general purpose diode capable of carrying 5mA is suitable. A suitable display would be the inexpensive MAN3M which is quite large enough for such an application.

Some users may have a local radio station pre-tuned, this can be indicated with the character "L" using the F,E and D segments of the display. Other alpha characters that can be formed with a 7 segment display are A C E F H L O P S U Y b d g h.





This circuit was designed for use with the ETI Ni-cad Battery Charger. It gives a positive indication by means of an LED that the battery is receiving current. The current I (at any applied voltage from 3–45V) causes a more-or-less constant p.d. of 1.8–2.2V across the 3 silicon diodes. This causes the LED to light up. The circuit is very sensitive and the LED starts to emit at 1.5mA, growing to its maximum brightness at about 10mA. This brightness is maintained over the full current range up to 1A (I_{max} for the diodes used). No current limiting resistor was found necessary for the LED.

The indicator is very cheap – about 30p – and reliable. Although 2V is 'lost', in many cases this is not important. In practice the unit is connected in series with the load (Fig. 2). It is important to note that no indication of the magnitude of the current is given; the whole idea of the circuit is to give a purely qualitative signal.

CURRENT-FLOW INDICATOR

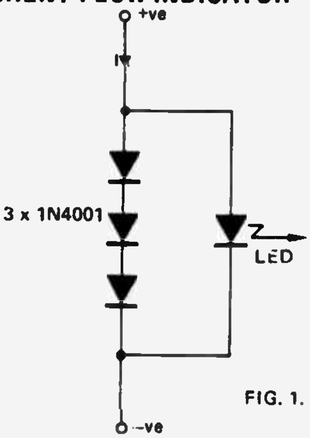


FIG. 1.

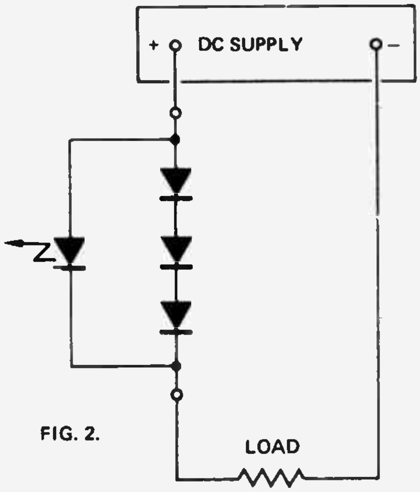
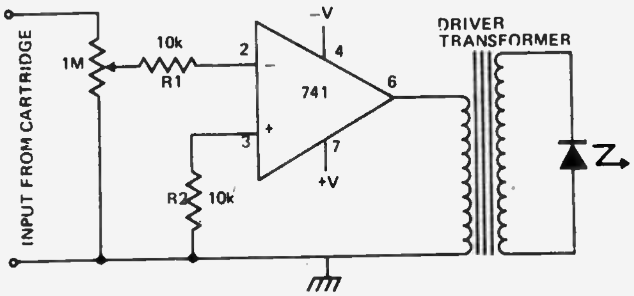


FIG. 2.



A cheap and effective alternative to cueing disco decks by PFL with headphones can be made by using LEDs to give a visual indication of the audio signal using this circuit.

R2 is used to ground the non-inverting input. The 1M pot and inverting input are used to ensure negligible output drain on the cartridge.

The ourpur of the 741 drives into the low impedance winding of an ordinary subminiature driver transformer. The secondary of this gives sufficient drive voltage to light the LED.

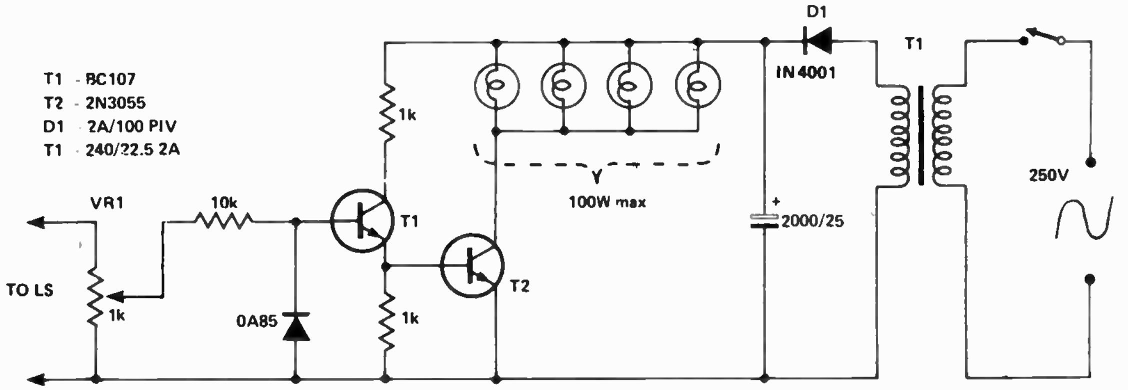
As the current drain for the whole circuit is small it can be supplied either from the main dual-voltage supply lines of the existing equipment or alternatively two PP3s. (Should last for several months).

The 1M pot is set to give no signal at the LED when the pick-up is on the lead-in track of the record.

DANCING LIGHTS

This device will produce a shifting light display in time to the signal from a loudspeaker. Setting will vary according to the volume at which the music is played. When VR1 is at maximum the lights remain lit most of the time. At minimum the lights may not come on at all. A suitable position can be established in between these two extremes.

The bulbs used can be any number at 25V each and the total should not be more than 100W. A heatsink should be used for the power transistor.

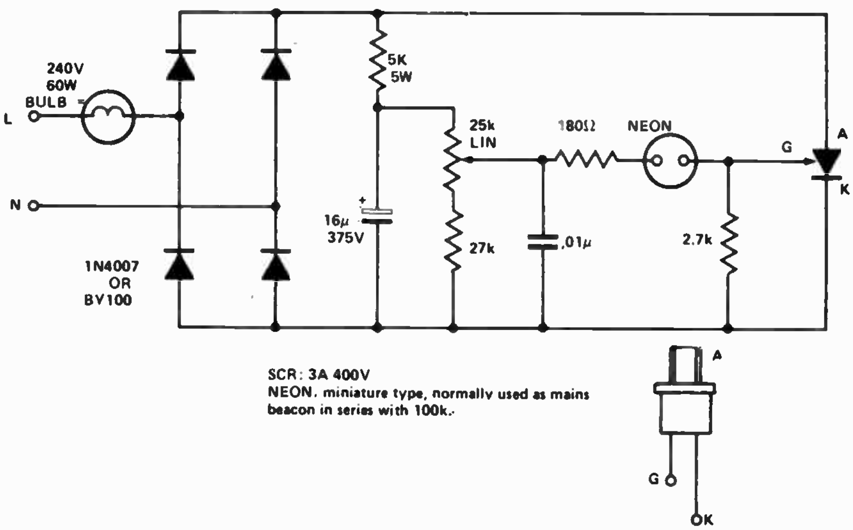


LOW FREQUENCY STROBE

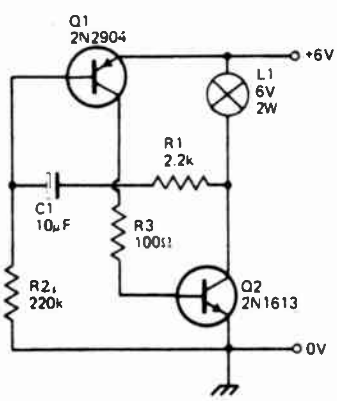
The circuit will flash the bulb at a rate between 0 and 10 Hz. Points to note are:

(i) Because all components are connected directly to the mains, do not touch whilst the unit is on.
(ii) Use a television type 25k pot with insulated spindle.
(iii) Mount in an insulated box with ventilation holes.
(iv) The 5k resistor gets hot, hence the wattage rating.
(v) The 27k may be altered to obtain as full range of control by the pot.

There is a risk of inducing convulsive seizures in people suffering from epilepsy if this unit is operated in their presence. Such people should avoid areas where strobe lights are used. A rate of nine flashes per second is considered the most dangerous and most people will find this unpleasant.

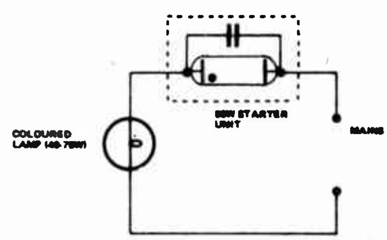


TRANSISTORISED FLASHER



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.

ULTRA-SIMPLE LIGHT FLASHER

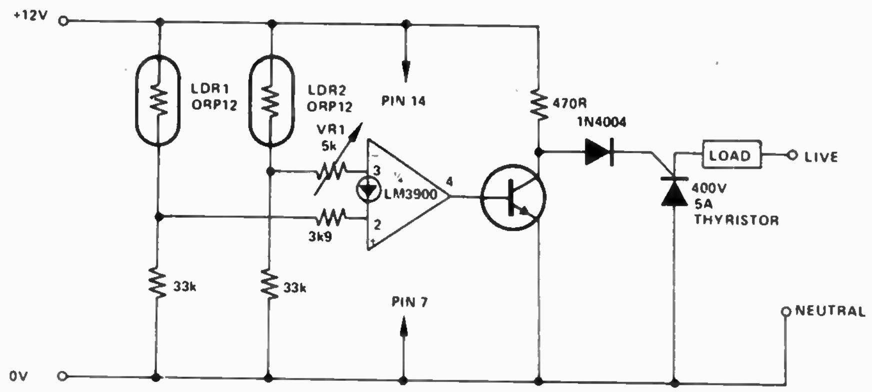


A cheap but effective way of flickering mains lamps suitable for discos etc.

It comprises a fluorescent lamp starter in series with a mains light bulb. The effect is improved with two or more units operating coloured lights. No problem is experienced with radio or TV interference and no supression is needed since the starter has a capacitor in it.

Providing the units are not left on for long lengths of time, the starters will last quite a while.

LDR MAINS CONTROL



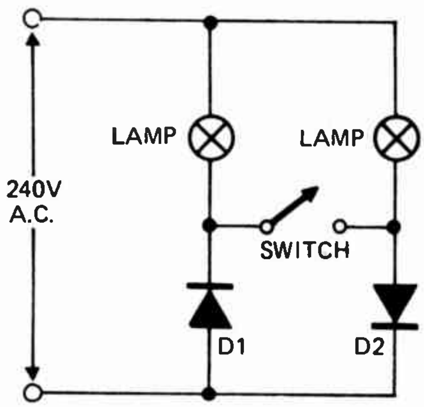
This circuit is used to turn off and on a light of up to 100W. When LDR1 is shaded from any incident light it will cause the output of the amplifier to fall, thus switching of TR1 and causing the thyristor to be turned on. The light will remain on because of the introduction of LDR2, which feeds the inverting input of the amplifier.

To turn the light off, LDR2 must be shielded from the light of the bulb.

To set the circuit up it is necessary to adjust VR1 by trial and error until LDR1 turns the light on and LDR2 switches it back off, and the normal level of background illumination has no effect.

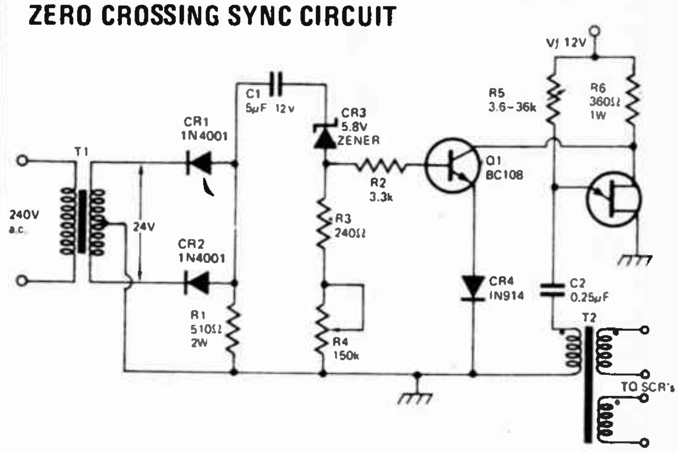
LDR1 and LDR2 should be placed in full view of the light they are switching.

FLOODLAMP POWER CONTROL



When setting up photographic floodlamps it is sometimes desirable to operate the lamps at lower power levels until actually ready to take the photograph. The circuit shown allows the lamps to operate on half cycle power when the switch is open, and full power, when the switch is closed. The diodes D1 and D2 should have a 400 volt PIV rating at 5 amps.

ZERO CROSSING SYNC CIRCUIT



Zero crossing control of SCRs or Triacs is preferable to phase control because less RFI is generated. The circuit shown was developed for a temperature control system and effectively maintained temperature at any set point from ambient to 100°C. Resistor R5 may be a potentiometer, a thermistor or any type of sensing device. R4 is adjusted so that the breakover point of CR3 is at the peak of the reference voltage (zero crossing point of ac wave).

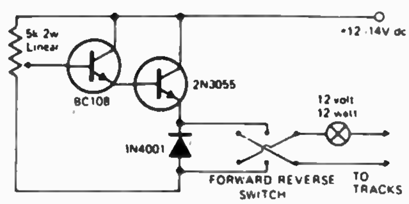
SIMPLE MODEL TRAIN SPEED CONTROL

Two transistors, a diode and a potentiometer can be used in place of the large and expensive rheostat usually provided in model train controllers.

Virtually any npn small signal transistor may be used in place of the BC 108 shown, likewise any suitable npn power transistor can be used in place of the 2N 3055.

The output transistor must be mounted on a suitable heatsink.

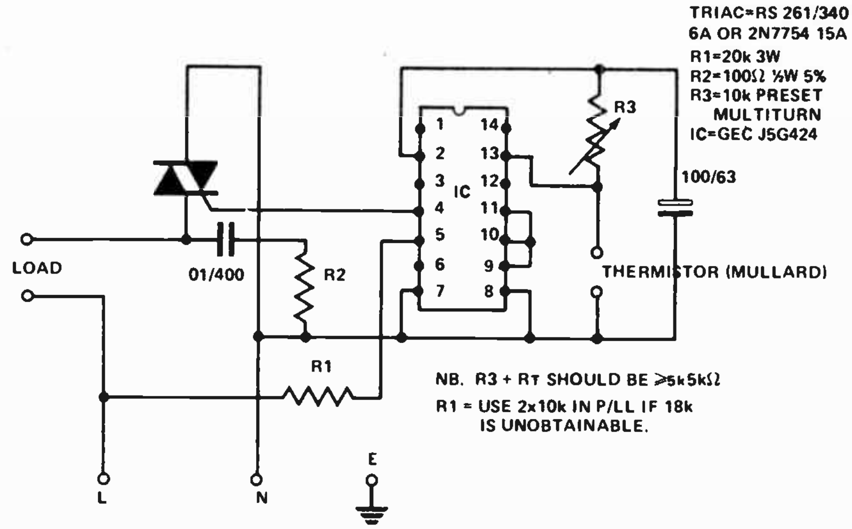
Short circuit protection may be provided by wiring a 12 volt 12 watt globe in series with the output. This will glow in event of a short circuit and thus effectively current-limit the output, it also acts as a visual short-circuit alarm.



LOW DIFFERENTIAL THERMOSTAT

This circuit evolved as a result of the need for a more satisfactory method of controlling the temperature in our paint heaters which operate at 170°F. The differential of conventional mechanical thermostats was too wide, both in actual rating and in % accuracy, so that severe overheating occurred when the demand for paint momentarily lapsed. The result was poor finish and in a number of cases the destruction of the thermometer (at approx £10 a time).

The introduction of the new thermostat completely eliminated the problem. The circuit consists of a GEC J5G424 Zero Voltage Switch in IC form together with a Mullard 2322 640 90004 which is plastic encapsulated, giving it both mechanical and electrical protection. It is available

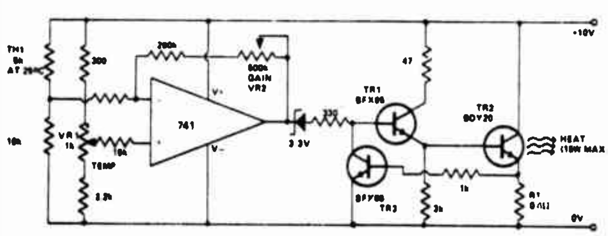


in four sizes with a temperature coverage from −30 to +200°C.

The RC network, 0.1mF + 100ohms, prevents self latching of the Triac.

The J5G424 is, by nature of its design free of RFI. The type of Triac employed will depend on the loading. We were using 6 and 15 amp loads.

A SIMPLE TEMPERATURE-CONTROL SYSTEM



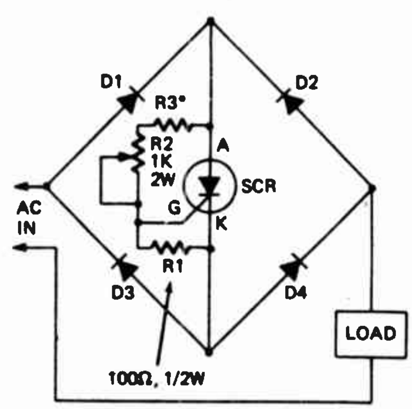
In electronics the need often arises to stabilise the temperature of critical sections of circuitry, such as master oscillators, log converters, and reference supplies. This circuit will control the temperature of a small mass of metal, such as a heat sink onto which critical components can be mounted, simply and efficiently.

The difference between a reference voltage set up on the temperature setting control VR1, and a voltage derived via thermistor TH1, is amplified by the op-amp, gain being set via VR2. This output voltage is applied to heater transistor TR2 via current amplifier TR1. ZD1 is essential for voltage-shifting since without it even the negative saturation voltage of the 741 would leave TR2 turned on. The current flowing in TR2 is limited to 1.5Amp by TR3, which shunts current from TR1 base if the voltage developed across R1 exceeds 0.6V. This arrangement leaves most of the supply voltage across TR2 and hence it is the only component dissipating significant heat.

TR2 is bolted to the mass of metal to be temperature-stabilised, and TH1 mounted as close to its flange as possible, using silicon grease for good thermal contact. The less thermal time-lag between them, the higher the gain that can be used without instability, and hence the lower the steady-state error of the system. Instability can be easily checked for by monitoring the op-amp output; if this is stable at the non-saturated value, then the system is probably operating correctly. TH1 must be the type of thermistor that has a small head of sensitive material at the extreme end of the glass encapsulation; other types have too much thermal inertia.

The prototype held a small (3°C/W) heatsink within ½°C for temperature settings between 30°C and 70°C, when shielded from draughts. The unit must be powered from a 10V PSU having good voltage regulation, to attain this performance.

FULL-WAVE SCR CONTROL



This circuit enables a single SCR to provide fullwave control of resistive loads.

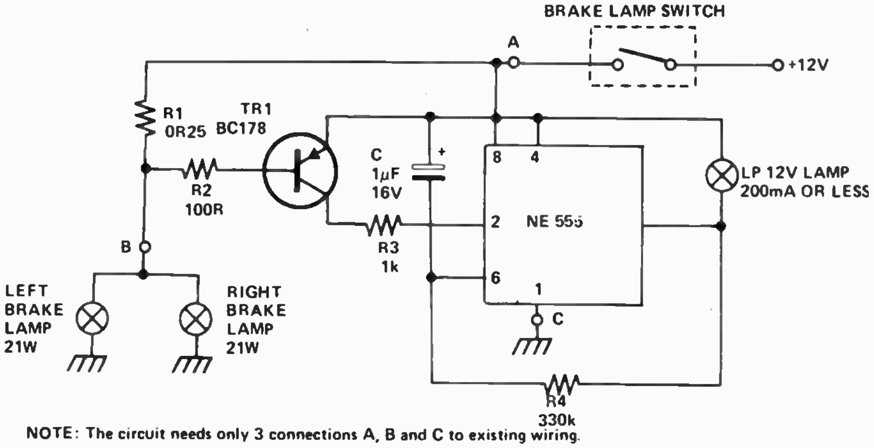
Resistor R3 should be chosen so that when potentiometer R2 is at its minimum setting, the current in the load is at the required minimum level. Diodes should have same current and voltage rating as the SCR.

BRAKE LAMP FAILURE INDICATOR

Here is yet another application for the NE555 timer.

If both brake lamps are working the lamp LP lights but if one or both are open circuit the lamp will flash at 2Hz, alerting the driver.

When both lamps are good the current through R1 turns on TR1 preventing C from charging, and keeping pins 2 and 6 at rail potential. Under these circumstances pin 3 is low and LP is on, however if one or both lamps are faulty TR is not turned on and the NE555 time oscillates freely at 2Hz, flashing LP.



NOTE: The circuit needs only 3 connections A, B and C to existing wiring.

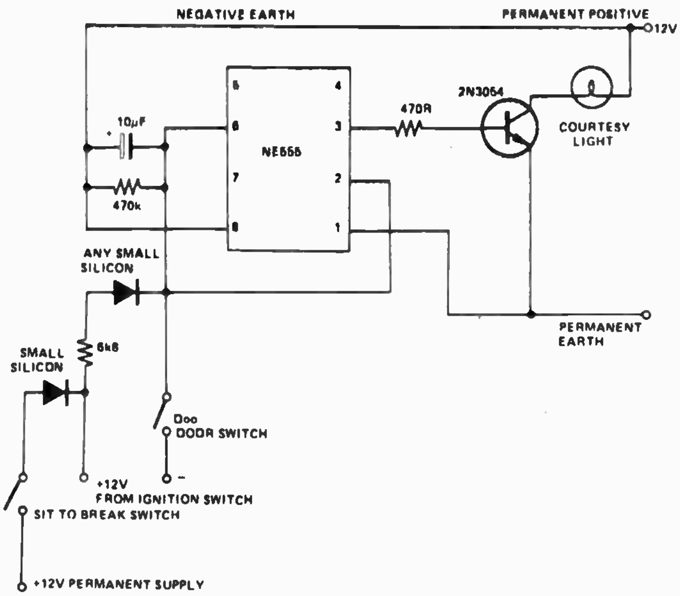
R1 is calculated on the basis of two 21W brake lamps (42W total). If a different total wattage is used, use the formula:

R1=10.5/P where P is the total wattage of two lamps.

COURTESY LIGHT DELAY

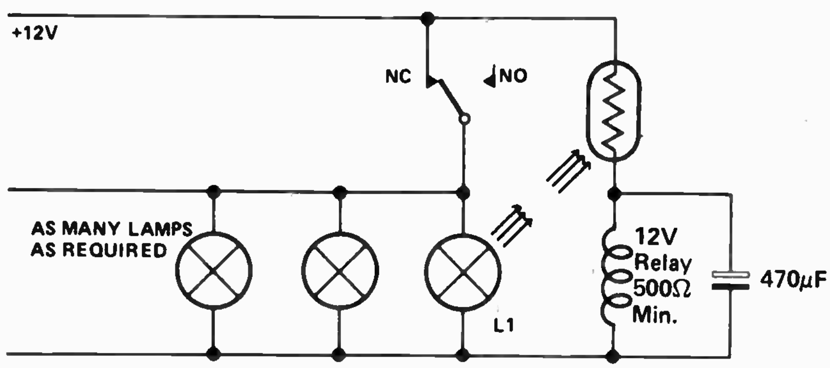
There are two problems with courtesy light delay switches, the first is that having the courtesy light stay on in a dark car park albeit only for a few seconds, can attract unwanted attention when you are leaving the car. The other problem is that most delay switches are set for about 10 seconds, which is too long when you wish to drive off straight away and is too short if you wish to go into the glove compartment or perhaps put on gloves.

This circuit defeats both these problems, the light stays on for a maximum of 1 minute or goes out as soon as the ignition is turned on. Though a switch mounted under the seat which is fitted so the switch breaks when you set on it, the light goes out as soon as you leave the car.



SIMPLE HAZARD LIGHT

When switched on, the lamp will light, at the same time, this light will lower the resistance of the LDR thus operating the relay, which in turn disconnects the supply to the lamp. This causes the LDR resistance to increase thus de-energising the relay. Time delay is introduced by the addition of the capacitor across the coil. Lamp L1 must be positioned close to the LDR. The only limit to the number of other lamps is the current rating of the relay contacts.

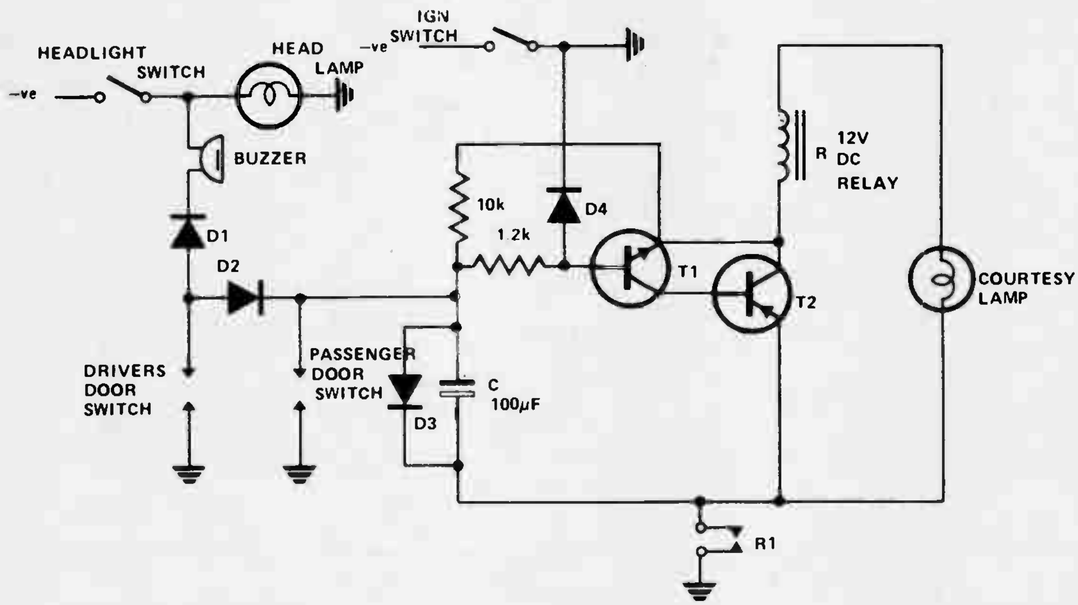


COURTESY LIGHT EXTENDER AND HEADLAMP REMINDER (+VE EARTH)

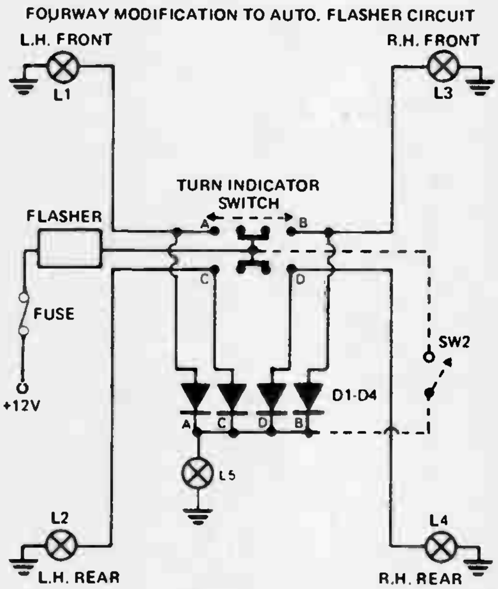
With the ignition switched off, an earth from the passenger or drivers door causes C to discharge, the relay to operate and the courtesy lamp to light. The relay is operated through transistor T2 which is biased on by T1. T1 and T2 remain on once the door is shut until C is recharged, hence giving approximately 15 seconds delay before the courtesy lamp extinguishes. Operation of the ignition inhibits the delay switch by biasing T1 off. i.e. courtesy lamp only alight when door open. D2 and D3 must be capable of carrying full courtesy lamp current.

The headlight reminder operates only when the headlights are left on and the drivers door is operated, thereby allowing departure of passengers without disturbance.

For −ve earth diode polarities and capacitor C should be reversed and transistor types changed.



FOUR-WAY FLASHER ADAPTOR UNIT



Many current model cars now incorporate a turn indicator switch position which causes all four indicator lights to flash simultaneously. This is a valuable safety device if stalled on the road – especially at night.

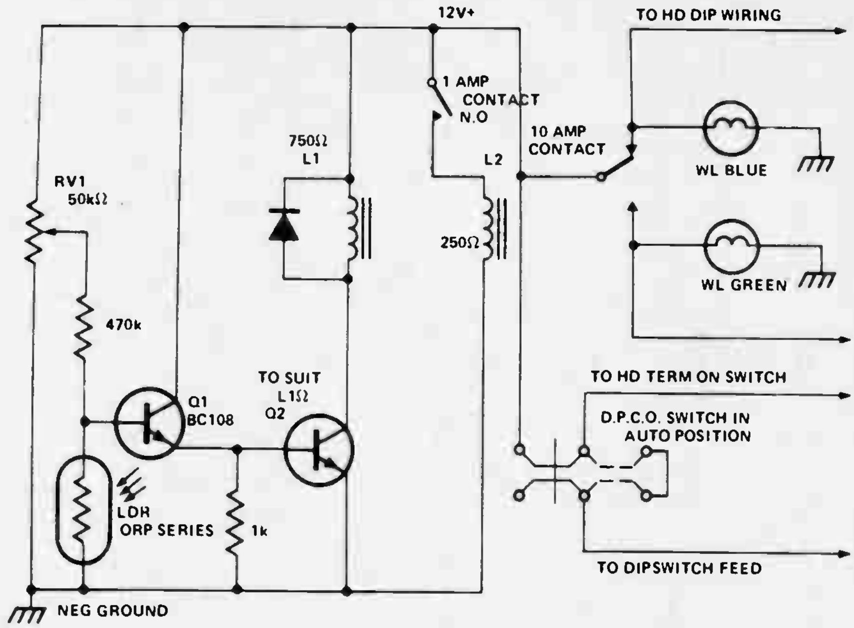
Older model cars fitted with normal winking indicators can be converted to include this facility with the aid of a few diodes, a switch and a heavy duty flasher unit.

Since in the "four" position the flasher must switch twice its normal load it is advisable to substitute the normal flasher unit with a heavy duty one as supplied for use with caravans and trailers.

Diodes D1-D4 are any rectifier types capable of handling about 3 A. Switch 2 is fitted in on the dashboard and L5 is a optional indicator also located to the dashboard.

The circuit as shown will work with both 6 and 12 volt negative earth systems. If the wiring is positive earth, reverse the direction of the diodes.

AN AUTOMATIC HEADLAMP DIPPER UNIT



The circuit follows an automatic parking light circuit in that when light from an external source falls upon the light dependent resistance LDR causing it to go low, the transistors in the circuit are not triggered, but when the external light fades, the resistance of the LDR goes high, allowing Q1 base to go positive and conducting so that Q1 emitter and Q2 base also go positive. Q2 collector current rises, energising the relay L1, this being 'normally open' contact arrangement, the contacts close and energise relay L12 which livens up the headlamp bright filaments. When approaching rays from street lamps or oncoming cars, the relay L1 drops out and disconnects L2 which drops out and energises the dipped filaments. RV1 controls the sensitivity.

The change over switch when switched to manual allows the dip switch to be used in a normal manner.

The unit can be placed under the forward edge of the dash. Then the headlamp wire is removed from the headlamp switch and taken to the unit and another is run back to the vacated terminal. The LDR should be mounted in a bicycle rear light housing (torpedo shape) and mounted at bumper height on the offside of the car.

UNIVERSAL WIPER DELAY

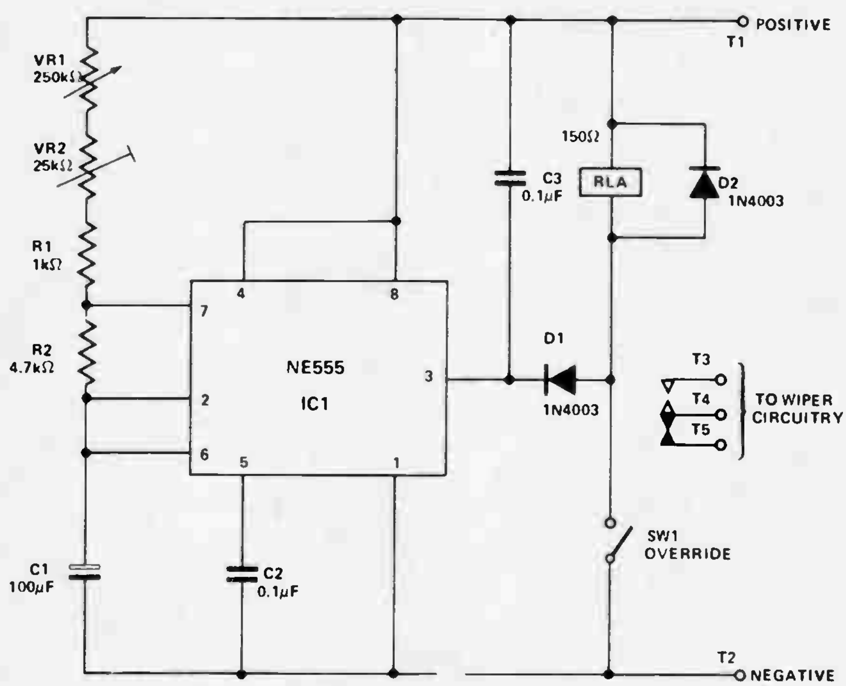


Fig. 1. Wiper Display Unit

Having recently experienced some difficulty in trying to fit a thyristor type wiper delay unit to the car, the trouble was eventually found to be a result of the design of the car's wiper circuit and also by noise spikes which spuriously trigger the thyristor. The following circuit should overcome these problems in both negative and positive earth vehicles.

IC1 is connected in the astable mode, driving RLA. C3, D1 and D2 prevent spikes from the relay coil and the wiper motor from triggering IC1. VR2 is adjusted to give the minimum delay time required. VR1 is the main delay control and provides a range of from about 1 second to 20 seconds. SW1 is an override switch to hold RLA

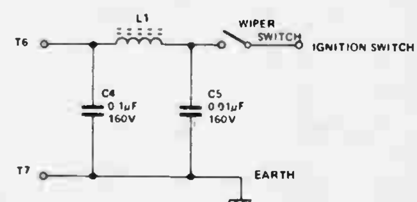


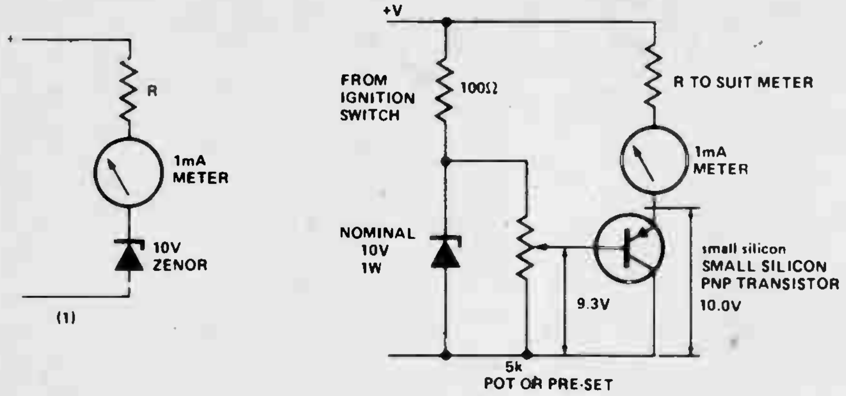
Fig. 2. Suppression Circuit
For Negative Earth Vehicle connect T6 to T1 and T7 to T2
For Positive Earth Vehicle connect T6 to T2 and T7 to T1

permanently on (for normal wiper operation).

The relay should have a resistance of at least 150Ω and have heavy duty contacts. A set of change-over contacts, as shown in Fig 1, are necessary if the circuit is to be used on a car whose wipers are wired as on the Anglia or Cortina (inspection of the car's wiring diagram will confirm this).

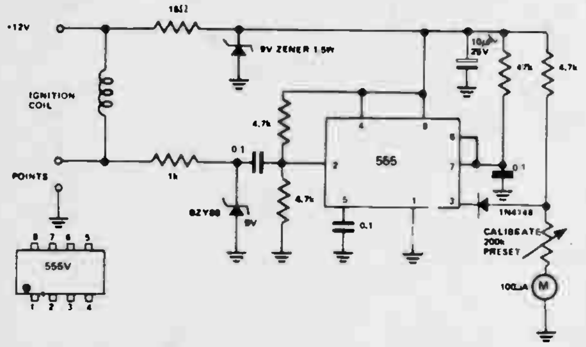
The suppression circuit shown in Fig 2 was found to be necessary for the protection of IC1.

SUPPRESSED ZERO VOLTMETER FOR THE CAR



To make a meter cover the range 10 to 15V or 10 to 20V over its whole scale, then circuit (1) is often used. The zener must be exactly 10.0V and may not be available. In this case use the arrangement shown in (2).

AUTOMOBILE REV COUNTER/TACHOMETER



The Signetics NE555V Timer/Monostable Multivibrator is probably the most important standard IC to appear since the 741 op amp. Its versatility is obvious, but it can also perform a variety of tasks, with a 4½ to 18V supply and its ability to sink 200mA.

The Tachometer is an obvious high volume application and many working variations are possible on the above circuit.

Pulses from the points are fed to the 1k resistor and 5V zener for clamping and then trigger pin 2, which causes the output to go high on pin 3 for a duration set by the R/C ratio on pins 6 and 7. During this time the 1N4148 on pin 3 is reversed biased, and the 4.7k resistor and the preset supply a constant current to the meter, which is calibrated in Rev/Min. The meter is giving an analogue representation. When the time duration elapses pin 3 goes low, shunting all current around the meter. The ratio of current flow to the time it is shunted gives a representation of RPM which is integrated, (or smoothed) by the meters mechanical movement to give a very accurate indication, when calibrated, of the RPM. Accuracy is nominally to 2%. The 9V zener, 15 ohm resistor and 10µF capacitor are to stablize the current supplied to the meter.

Calibration can be made using 50Hz 12V pulses derived from the domestic mains if it is remembered the points operate at 2 times the engine RPM on a 4-cylinder, engine, i.e. 50Hz = 3000 cycles/min equivalent to 1500 Rev/Min for a 4-cylinder 4-stroke. On an 8-cylinder engine it would equal 750 RPM. On a 6-cylinder 50Hz equals 1125 Rev/Min.

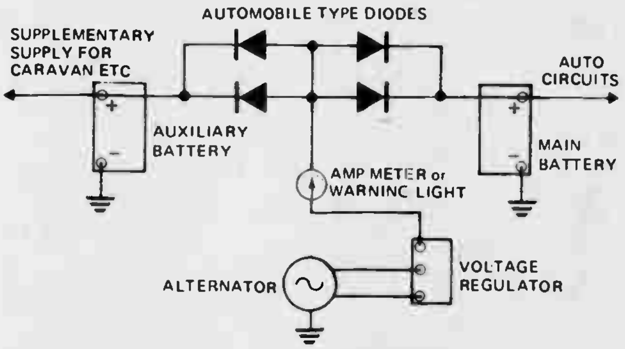
The circuit can easily be built using 0.1 pitch vero board. The 555 IC gives temperature stability and solid state reliability.

AUXILIARY BATTERY ADAPTOR

When towing a caravan or using the automobile battery supply for other heavier duty purposes the drain on the battery may be excessive.

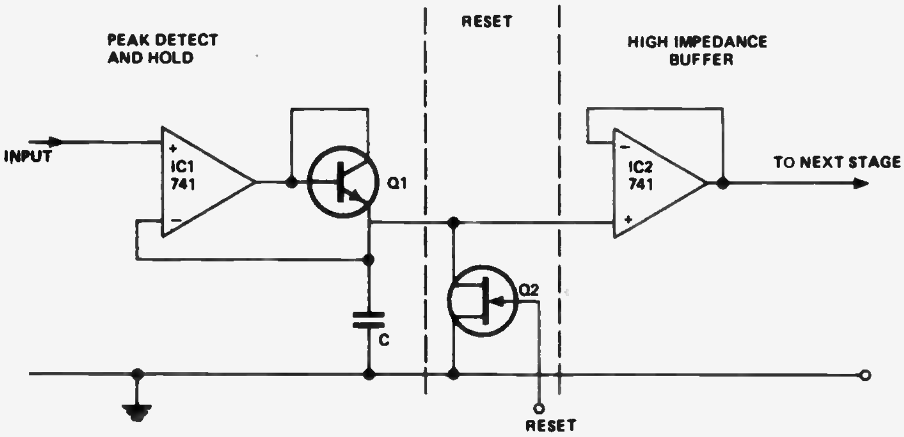
Here is a method of hooking up an auxiliary battery to the auto's charging circuit without upsetting the existing battery, and limiting discharge to external circuitry, to the auxiliary unit only.

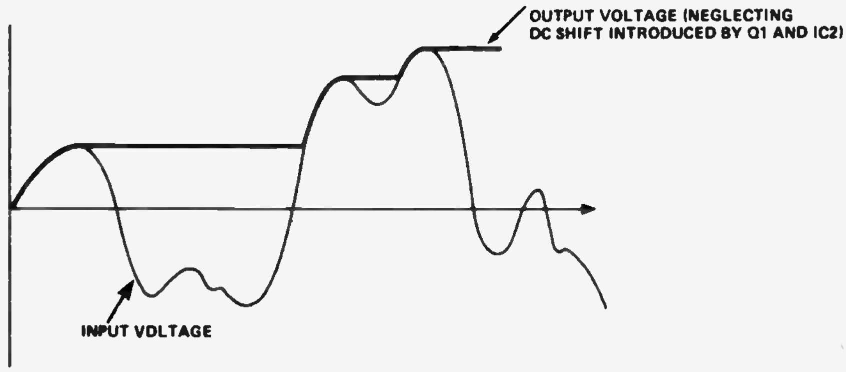
The four isolating diodes are of the automobile type as used in alternators, being capable of carrying up to 25 A they should be mounted on heatsinks.



AUXILIARY BATTERY ADAPTER FOR NEG. EARTH AUTO. SYSTEM

PEAK DETECT AND HOLD CIRCUIT





If the voltage at the input becomes bigger than the voltage on the capacitor, then the output of the 741 goes positive, the diode conducts, and the capacitor is charged up to the input voltage-forward voltage drop of diode. When the voltage at the input is less than that on the capacitor, the output of the 741 goes negative, and the diode cuts off. To prevent the capacitor from discharging through the input resistance of the next stage, a high input impedance buffer stage (IC2) is used. The circuit can be reset by means of a FET or similar high impedance device connected across the capacitor.

WINDOW DECTOR CIRCUIT

This circuit de-energises a normally energised relay if an input voltage goes above or below two individually set voltages.

It consists of one IC and one transistor driving the relay. The transistor is normally turned on by R4, so the relay is normally energised. If the cathode of D1 or D2 is taken negative, Q1 will turn off and the relay will de-energise.

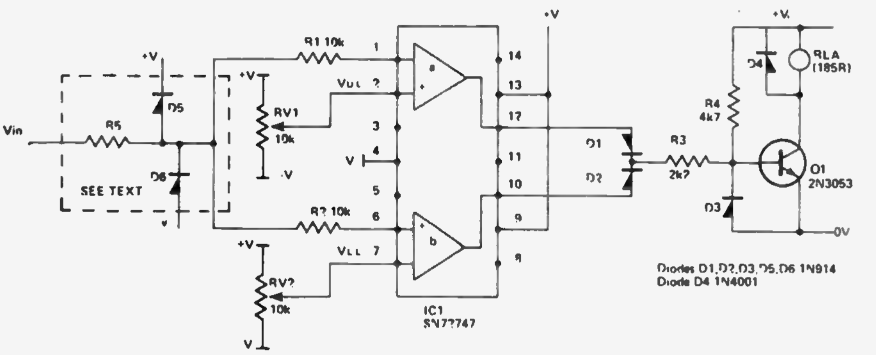
The IC is a 72747 (dual op amp). The op amps are used without feedback, so the full gain (about 100dB) is available. The amplifier output will thus swing from full positive to full negative for a few mV change at the input.

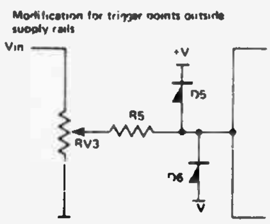
The relay is therefore only energised if Vin is between V_{UL} and V_{LL}. The two limits can be set anywhere between the supply rails, but obviously V_{UL} must be more positive than V_{LL}.

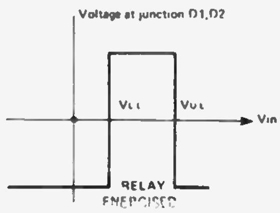
If Vin can go outside the supply rails D5, D6 and R5 should be added to prevent damage to IC1.

If V_{UL} and V_{LL} are required to be outside the supply rails, Vin can be reduced by RV3.

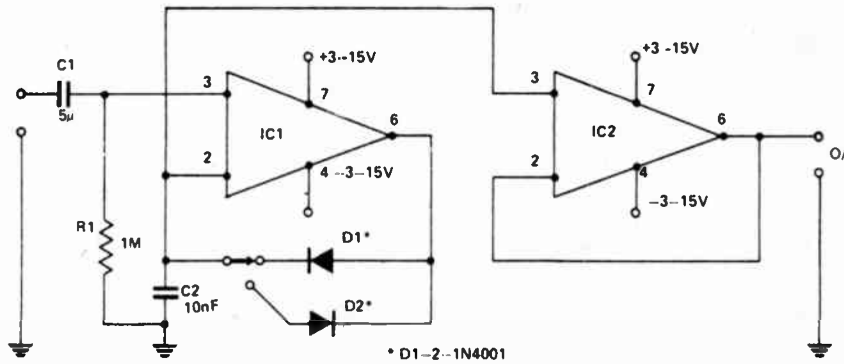
The supplies can be any value providing the voltage across them is not more than 30V.







PEAK PROGRAM DETECTOR



The circuit shown will allow a multimeter to display the positive or negative peaks of an incoming signal.

A 741 is used in the non-inverting mode with R1 defining the input impedance. D1 or D2 will conduct on a positive or negative peak charging C2 until the inverting input is at the same DC level as the incoming peak.

This will maintain the voltage until a higher peak is detected, when this will be stored by C2.

In order to prevent loading by the multimeter another 741, (IC2) is employed. This is also connected in the non-inverting mode as a unity gain buffer, output impedance is less than 1Ω, low enough to feed the most insensitive meter.

As shown the instrument has a useful response from 10Hz–100kHz (±1dB). High linearity is ensured by placing the diodes in the feedback loop of IC1, effectively compensating for the 0.6V bias that these components require.

POSITIVE PEAK DETECTOR

A positive-peak detector having gain may be constructed using two LM301As as shown in Fig. 4.

The output is the peak voltage at the input amplified by the ratio (R1 + R2)/R2. Typical error is 2 (R1 + R2)/R2 millivolts.

If unity gain is required R2 is deleted. The combined resistance of R1 and R2 should be in the range of 10 to 100 k and the minimum load resistance 2.2 k. Where negative peak detection is required reverse the polarity of both IN914 diodes.

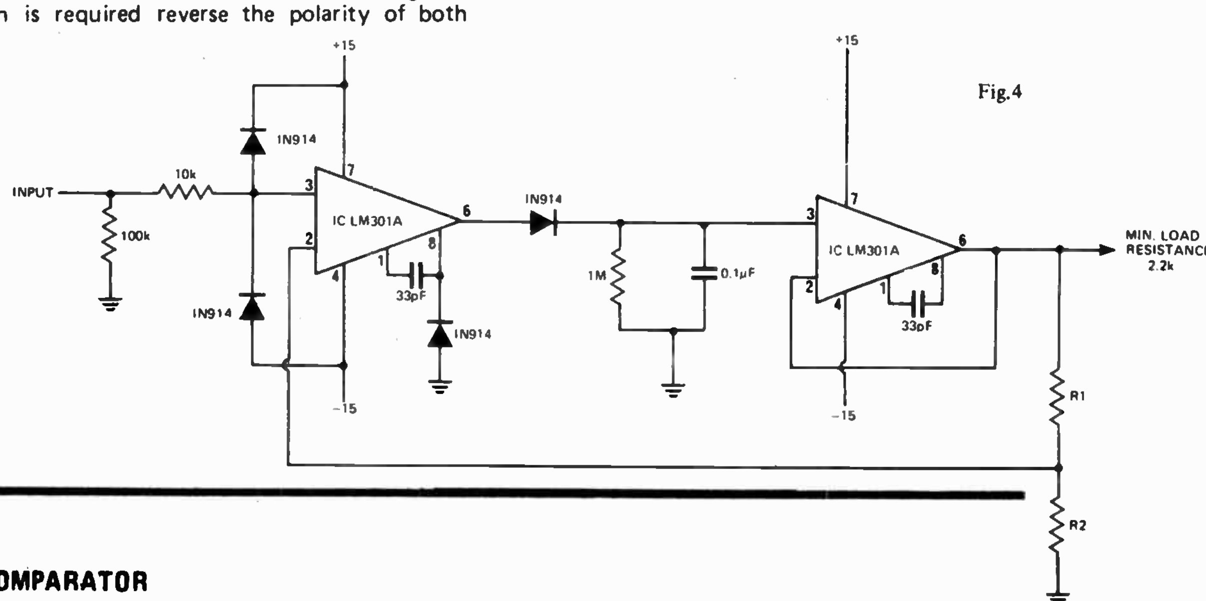
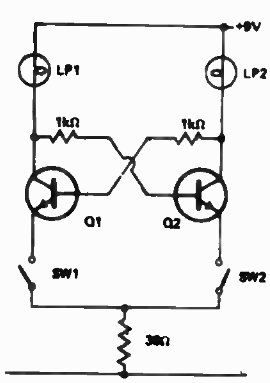


Fig.4

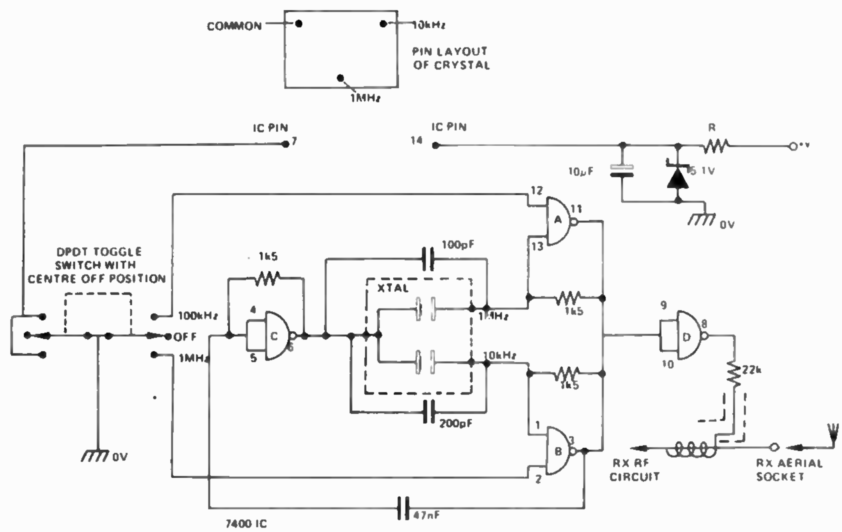
REACTION COMPARATOR



The circuit above can be used as an independent arbiter of which two people can throw a switch first. With both switches off the circuit is completely open and so neither bulb lights. If SW1 is made - even the tiniest fraction of a second before SW2 - Q1 will conduct and LP1, acting as the load, will light up. At the same time the voltage at the collector of Q1 falls meaning that even if SW2 is made there is insufficient bias current to drive Q2 on.

If SW2 is made first the converse is true. The bulb should be 6V, 40mA or 60mA types. The transistors can be almost any silicon NPN type (BC107 etc).

TTL CRYSTAL MARKER



This circuit was designed to make use of an old 100kHz–1000kHz 10X, 3 pin twin crystal.

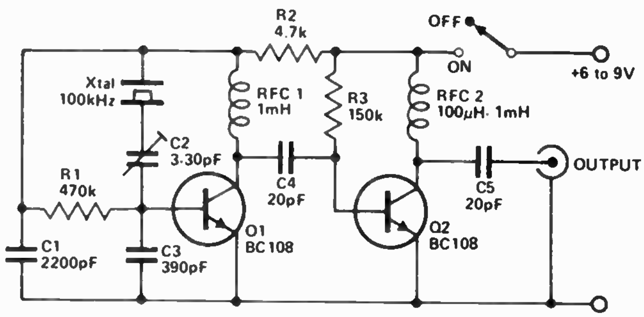
With the switch in the 100kHz position the 0V line is applied to pin 7 of the 7400 (connecting ie. to supply) and also to pin 12 which disables gate A. This allows gates B and C to operate as an oscillator at the crystal frequency of 100kHz.

With the switch in the 1Mz position the 'inhibit' line is fed to pin 2. This disables gate B and allows gates A and C to operate as a 1MHz oscillator.

As DC switching is employed the switch can be remote from the oscillator unit. Gate D is used as a buffer/shaper stage.

The output of the unit is loosely coupled to the aerial input lead providing adequate marker amplitude well above 30MHz.

100 KHZ MARKER GENERATOR



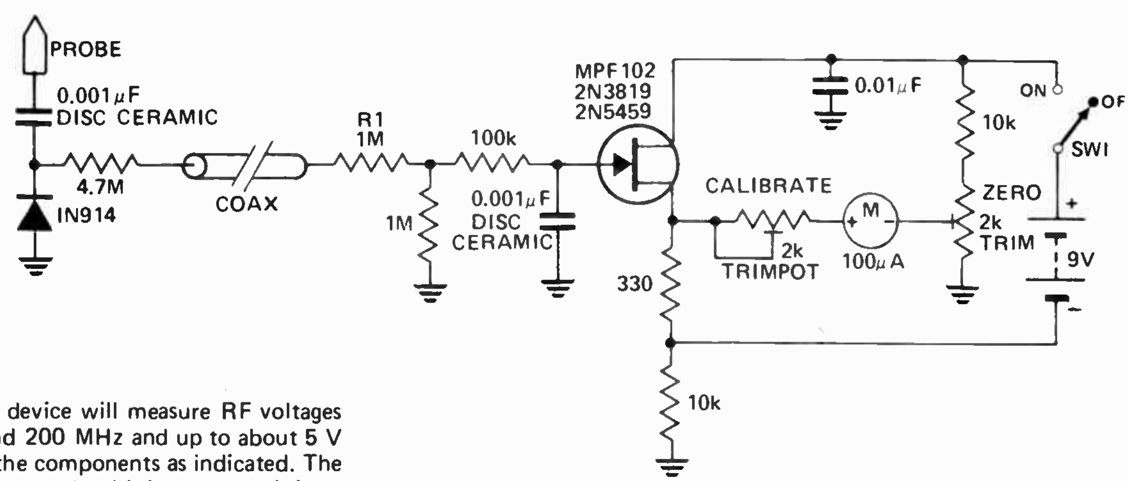
The above marker generator will produce strong signals every 100 kHz from 100 kHz to over 200 MHz. It is very useful for calibrating receivers and for use as a signal generator.

Cheap transistors type BC108 give good results but almost any PNP or NPN transistor having a gain-bandwidth product greater than the desired frequency range will give good results.

The oscillator should be calibrated by adjusting it to zero-beat with WWV at 10 MHz, or 15 MHz, on a communications receiver, or with a digital-frequency meter.

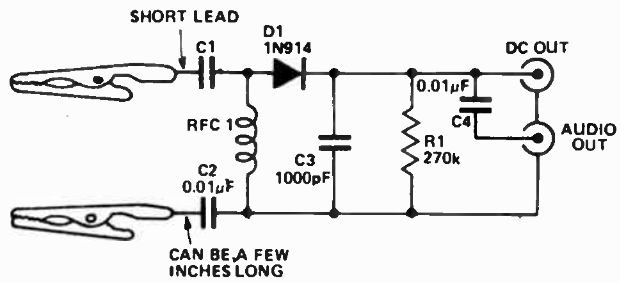
The choke, RFC1, can be any suitable small RFC (i.e. Aegis C13), the same for RFC 2 (i.e. Aegis C13 or UPC 100 to UPC 560).

SENSITIVE RF VOLTMETER



This device will measure RF voltages beyond 200 MHz and up to about 5 V with the components as indicated. The diode etc should be mounted in a remote probe, close to the probe tip. Sensitivity is excellent and voltages less than 1 V peak can be easily measured. The unit can be calibrated by connecting input to a known level of RF voltage, such as a calibrated signal generator and setting the calibrate control. The output indicates in RMS. As it is it reads about 2 V RMS full scale. This can be increased to 20 V or more by increasing R1 to 20 M (two 10 M in series). The 100 μA meter could be a multimeter if desired.

GENERAL PURPOSE RF DETECTOR



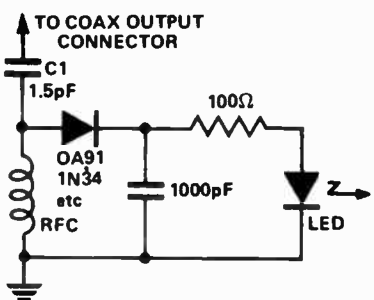
When constructing or developing communications equipment, such as transmitters, receivers etc, a very handy gadget is this general purpose RF detector. It provides dc output to a meter and audio output (if necessary) for checking transmitters or modulated signals.

It can be used also as a field strength meter or transmitter monitor.

The values of C1 and RFC vary depending on the frequency range in use. Below 1 MHz, C1 can be .001 μF and RFC at 2.5 mH or 5 mH RFC (i.e. Aegis C2, C4 or C9). In the HF region to 30 MHz, C1 can be 20 pF or a 5-40 pF trimmer while RFC1 can be a 2.5 mH choke (i.e. Aegis C2 or C4) or any choke down to 470 μH (i.e. Aegis C13, UPC560 or VPC470). In the VHF range C1 can be a 2 to 10 pF capacitor or 0.8 to 7 pF trimmer. RFC1 can be between 47 μH and 150 μH (i.e. Aegis VPC 150, UPC120, VPC100, VPC82, VPC68, VPC56 or VPC47).

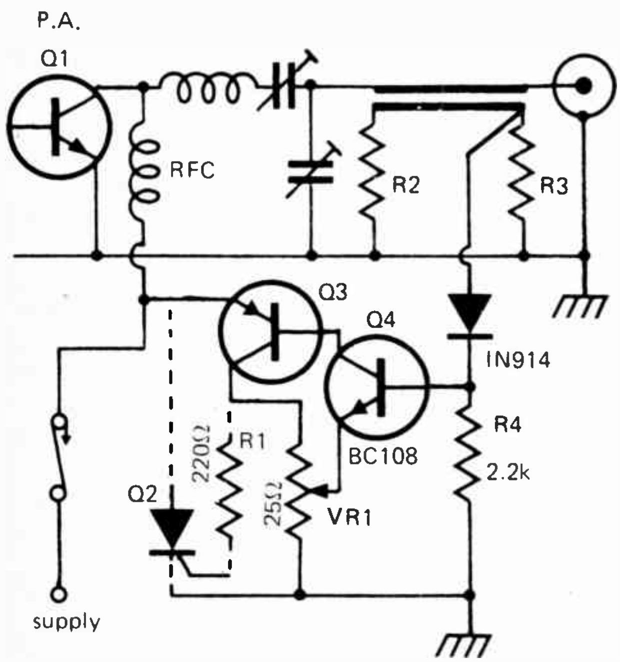
Diode D1 can be almost any germanium diode or a hot-carrier diode. Mixer diodes such as the IN21 and IN23 series are also excellent. Use a diode with a high reverse-voltage rating if working with valve transmitters.

LED RF INDICATOR



An RF output indicator using a LED is very useful for monitoring the output of a transmitter. This circuit will give indication from a 5 W transmitter. The capacitor C1 and the RFC are chosen for the appropriate frequency. The RFC could be replaced by a resistor for wideband use. The sensitivity depends on the value of C1 and the resistor used if the RFC is replaced. For high power transmitters, C1 could be a small 'gimmick" capacitor.

RF AMPLIFIER PROTECTION



RF power amplifier output transistors may be destroyed by high standing wave ratio loads.

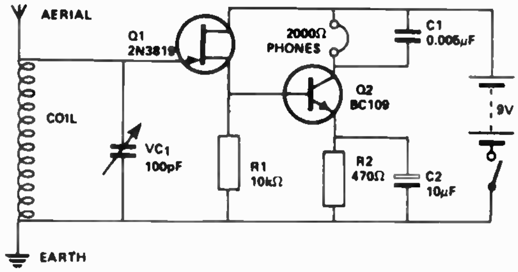
This circuit senses the SWR conditions existing at the transmitter output. If the SWR exceeds a predetermined level, the PA collector voltage will automatically be reduced: in the event of a really high SWR the protection circuit will blow the supply fuse.

The SCR shown in dotted lines – should be included if the circuit is to be used to protect a high power stage. Otherwise Q3 is used to draw an excessive current from the power supply, this transistor must be rated to dissipate the maximum power necessary to cause the fuse to blow. Potentiometer VR1 must also be rated accordingly.

The SWR sensing element employs standard SWR bridge techniques and should present no difficulties.

Make sure that you have the right size fuse though!

FIELD-EFFECT TRANSISTOR RADIO RECEIVER



The circuit shown in the figure provides a simple radio receiver which is both sensitive and selective. A low-cost FET is used - the JUGFET 2N3819.

In order to ensure that the impedance of the parallel tuned circuit is high at resonance, the inductance of the coil should be high and the value of the tuning capacitor should be kept low.

The amplitude modulated carrier wave sets up a varying voltage across the tuned circuit which causes V_{GS} to vary and a changing drain current I_{DS} to flow. A varying voltage is developed across R1 which is amplified by the npn bipolar transistor Q2. Capacitor C2 decouples the emitter of the bipolar transistor to ground for AC signals and capacitor C1 decouples the radio frequency component of the signal from the phones.

Detection of the amplitude modulated carrier wave is achieved by operating Q2 close to the 'knee' of its transfer characteristic. If the receiver tends to be unstable, the tendency for it to break into oscillation can be reduced by coupling the aerial to the circuit by means of a 47pF capacitor.

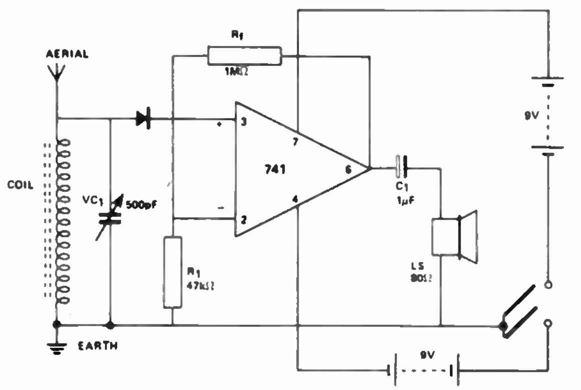
OP-AMP RADIO RECEIVER

The figure shows how to wire an op-amp so that it amplifies the voltage generated across a tuned circuit in order for the circuit to operate as a simple radio receiver. The '741' op-amp is suitable.

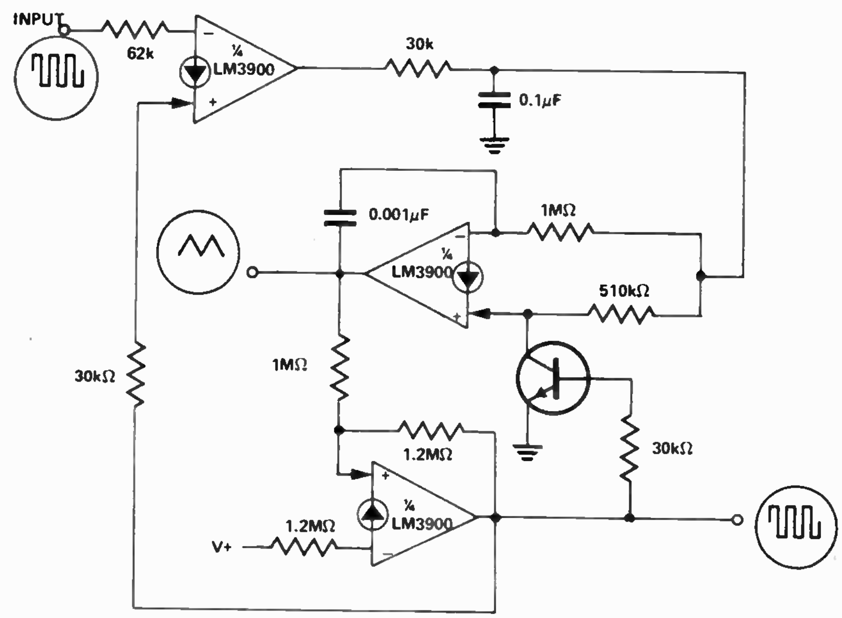
Note that the signal is applied to the non-inverting input of the op-amp so that good selectivity is provided due to the high input impedance of this connection which provides negilible loading of the tuned circuit.

A 2000 ohm earpiece may be used directly at the output of the op-amp but, as shown, an 80 ohm speaker can be driven via a capacitor whose value should be selected for optimum results.

Should the signal suffer from distortion, this may be due to high frequency noise generated by the op-amp and can be cured by connecting a 470pF capacitor across the feed-resistor R_f. The values of the components are not critical.



ECONOMICAL PHASE LOCKED LOOP

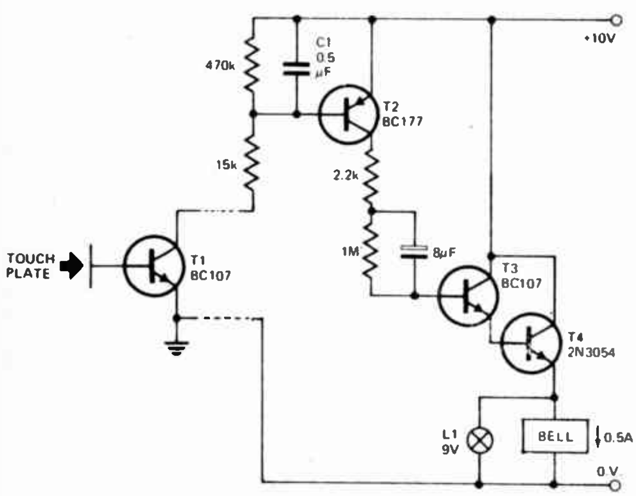


Integrated circuits which have been specially designed as complete phase locked loops are available, but many of them tend to be rather expensive devices. The circuit shows how the economical LM3900N integrated circuit can be used to build a phase locked loop which has a centre frequency of about 3kHz.

The LM3900N contains four current differencing amplifiers in a single 14 pin dual-in-line package. Only three of these amplifiers are used in the circuit shown, so the fourth amplifier is available for other purposes. The price of the LM3900N is about 69p, so it is one of the cheapest linear devices available. The special circuit symbol shown is used for the amplifiers in this device, since they are not a conventional type of operational amplifier.

If desired, the locking range of this phase locked loop may be increased by employing the fourth amplifier in the LM3900N in the input circuit to increased the signal amplitude.

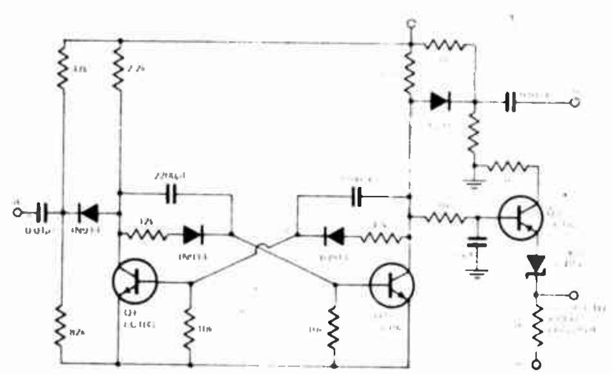
TOUCH CONTROLLED DOORBELL



In the circuit shown, hum is induced into the touch plate when touched with a finger. This hum signal is rectified in envelope detector/amplifier T1 and the resultant dc current amplified by T2, T3 and T4.

Decoupling is provided by C1, and C2 limits the length of doorbell signal to three seconds. The recovery time between signals is five seconds. The lamp L1 may be used to verify the signal, to the caller, if required.

PHASE LOCK CONTROL CIRCUIT



The circuit shown was used to phase lock a UJT oscillator at 10 times the frequency of a 16Hz oscillator which was varying ±25%.

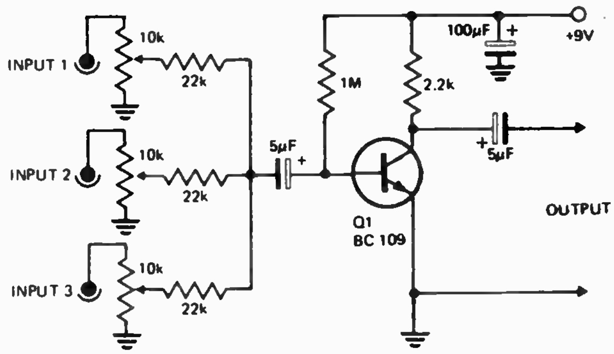
The reference frequency is coupled in at B, and a trigger from the phase locked frequency enters at A. The resultant multivibrater pulse train at Q2 collector is filtered and then dc shifted by Q3 and a Zener diode. The resultant dc error signal may be used to correct the frequency of a voltage tuned multivibrator or UJT oscillator. The sync signal may be generated by counting down the B signal by the ratio of the two signals.

MISCELLANEA

ONE TRANSISTOR AUDIO MIXER

Three inputs are shown on our circuit but more can be added if required.

Each input has its own level control. Sensitivity is 500 mV output for 25 mV input. This is more than adequate to drive most amplifiers.



VIRTUAL EARTH MIXER

This mixer was developed for mixing high quality audio signals prior to recording on a tape recorder.

Q1 is operated as a high gain common emmitter amplifier. Noise is kept low by operating this transistor at the very low collector current of 30μA.

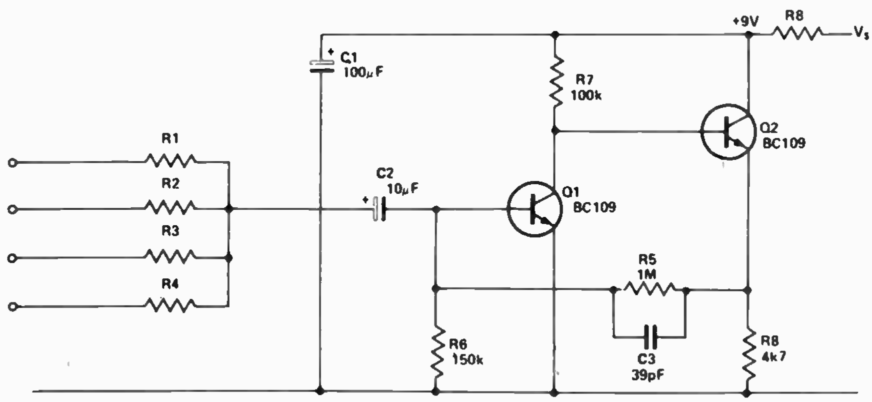
Q2 is connected as an emmitter follower offering a high input impedance at its base to prevent loading on Q1.

Overall feedback, both AC and DC is taken from the emitter of Q2 via R4 and R5. C3 rolls off the response above 40kHz to prevent RF pickup.

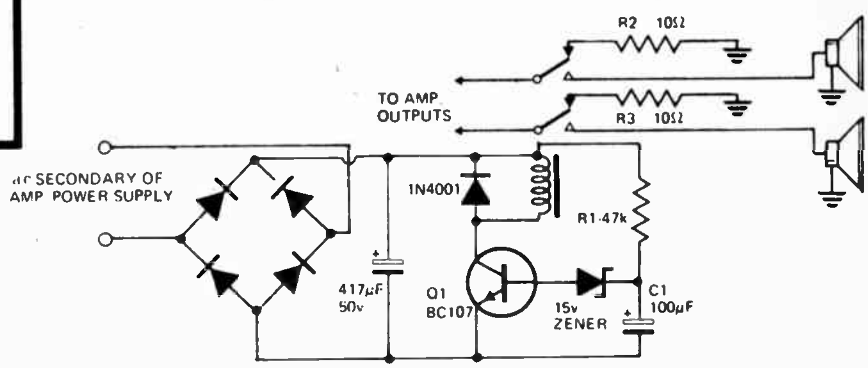
The signals to be mixed are introduced via the input resistors R1–4. C2 isolates unwanted DC from Q1's base whilst coupling the input signal to it.

Overall voltage gain is equal to $\frac{R5}{R1 + Rin}$ 20dB if R1–4 are as shown.

R8 and C1 decouple the mixer from the supply voltage employed. The value of R8 in kΩ is determined from the formula $\left(\frac{Vs - 9}{1.0}\right)$.



PLOP ELIMINATOR



Many hi-fi amplifiers cause an only-too audible 'plop' in the speakers when switched on. The 'plop' is not only disconcerting but can also be damaging to low-power capacity speakers.

The plop is generally caused by the momentarily high inrush current to the series output capacitors.

The circuit shown here brings the speakers into circuit only *after* charge on the output capacitors has been established.

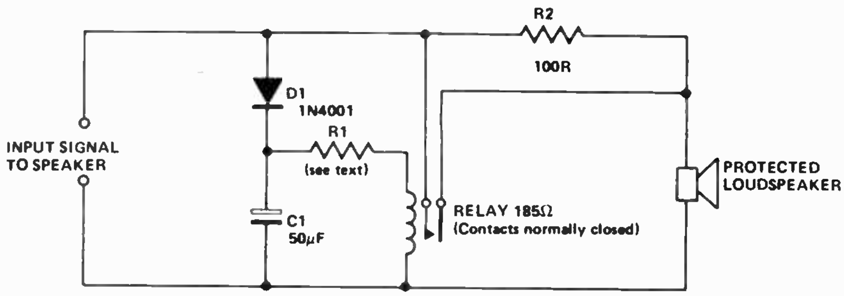
The unit is connected by wiring the rectifier bridge input to the AC secondary winding of the amplifier power supply.

Immediately the amplifier is switched on. C1 charges through R1. When the voltage exceeds the Zener voltage of the diode in series with the base, the transistor conducts and closes the relay.

As soon as the power is shut off, the small smoothing condenser quickly discharges itself through the relay and de-energises it.

The two resistors R2 and R3 provide an alternative path for the onrush current when the amplifier is switchd on.

LOUDSPEAKER PROTECTION UNIT

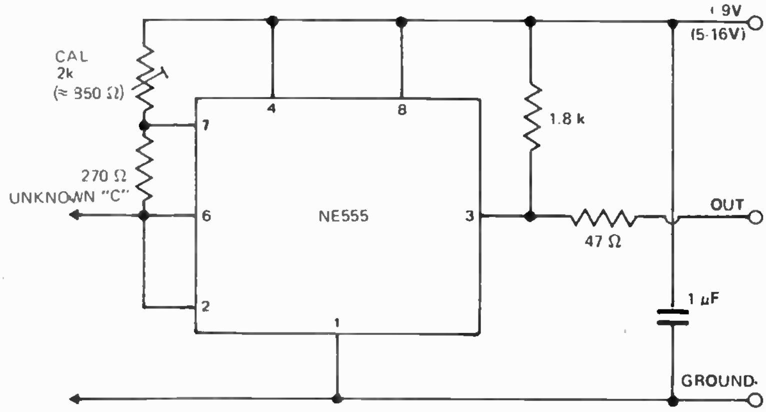


The following circuit will protect loudspeakers against overload if the correct components are used.

Operation of the circuit is quite simple, Diode D1 rectifies the signal across the speaker, which developes a fluctuating DC voltage across C1. When this voltage exceeds a certain level, the relay contacts open, which disconnects the lodspeaker and if required puts a resistor across the signal. In the case of valve amplifiers it is usually necessary to keep a load on the output when there is an input signal present, therefore R2 will have to be included in the design. With most types of transistor amplifiers today, the resistor R2 may be omitted.

R1 is adjusted to give adequate protection at whatever power is being used. Resistor R1 value should be selected according to the power at which the speaker will need to be limited and of course the impedance of the speaker. In my case the resistor R1 was made 220R but this may be too low for very high power applic·ations.

SIMPLE DIGITAL CAPACITANCE PROBE FOR COUNTER



This simple adaptor enables a digital counter to be used to measure capacitance.

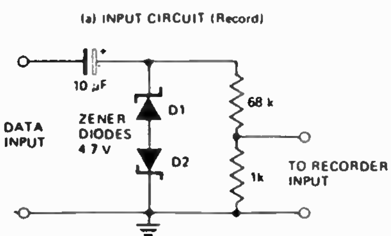
Various ICs may be used but the 555 series is the most practical and readily available.

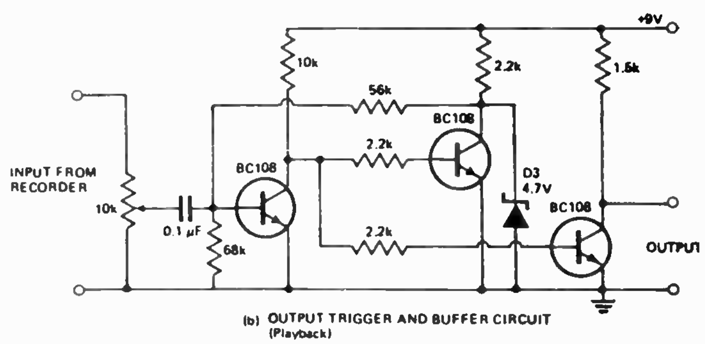
Probe output is coupled to the digital counter via coax. The counter is switched to the 'period' ranges with seconds read as µF, milliseconds as nF and microseconds as pF.

Accuracy depends on the accuracy of the calibration capacitors and of the power supply regulation. A calibration chart could of course be used if great accuracy or small capacitance value is required.

The prototype unit accurately measured a 50 000 µF capacitor and a 2 µF could still be measured accurately with a 47 k resistor paralleled across it. Lowest measurable value was about 500 pF.

DIGITAL TAPE RECORDER ADAPTOR





(b) OUTPUT TRIGGER AND BUFFER CIRCUIT (Playback)

The two circuits shown allow digital data to be recorded and replayed on an ordinary domestic recorder.

The input circuitry differentiates the digital pulses which are then recorded. On playback the recorded pulses are fed into a Schmitt trigger whose output is then amplified restoring the required binary data waveform.

The potentiometer RV 1 is adjusted so that on playback only the peaks will actuate the trigger.

MISCELLANEA

SUBSTITUTE FOR BREAKDOWN DIODE

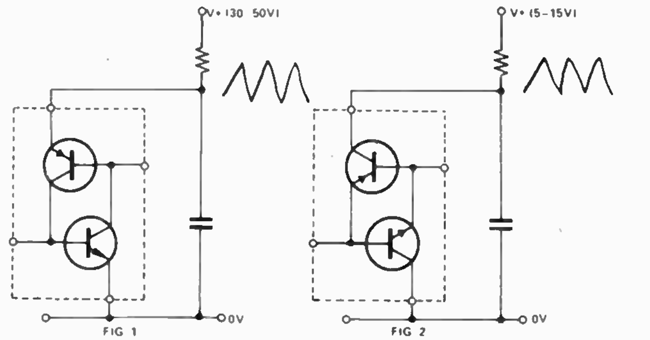
Experimenters may have tried the circuit of Fig.1. as a substitute for a diac. Due to the high V_{cbo} of modern silicon planar transistors, the configuration will snap into conduction only when 20–30V are reached. In addition, the breakdown voltage is ill-defined, varying from sample to sample.

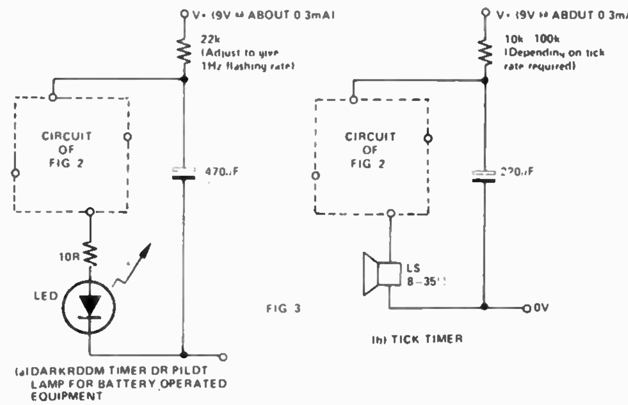
However if the transistors are used in the inverted mode, as shown in Fig. 2, the breakdown voltage now depends upon the V_{cbo} of the transistors used, which is confined to a narrow range for a given transistor type.

Experimental circuits have conducted at voltages from 2 to 6V, depending upon transistor type and quality.

Fig. 3 shows two typical applications. Doubtless other uses like time-base and delay circuits can be conceived.

Note that the circuit can be triggered at either base with an appropriate current pulse.





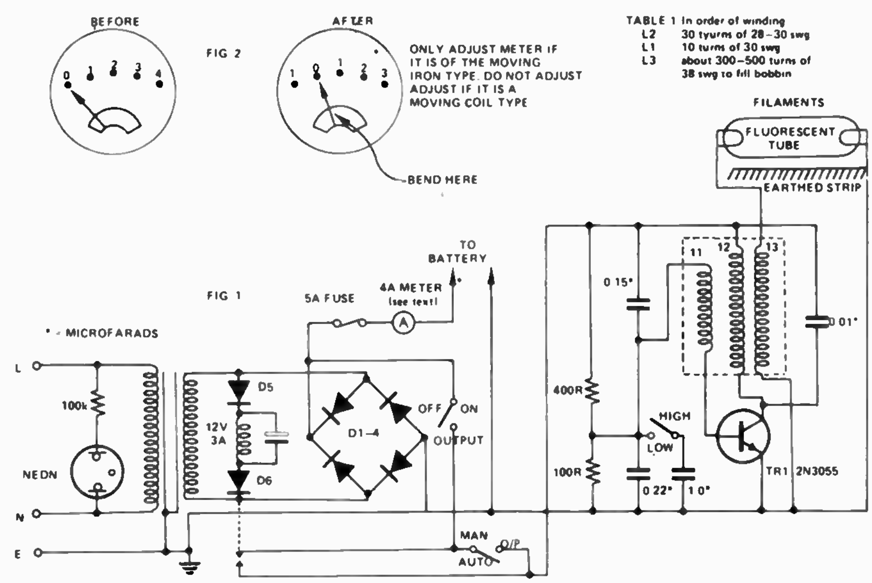
DUAL FUNCTION CHARGER

The charger is quite straightforward; the 13V AC from the transformer is rectified by the power diodes (D1-4) and then goes through the 5A fuse and then goes through a 5A fuse and 4A ammeter to the clips on the battery (it may be better to use lead clips if the clips are to be used for long periods, as steel ones tend to corrode).

However, this 13V is also ½ wave rectified by D5-6 which are connected to a 9V relay (a capacitor may be connected across the relay to smooth out the supply). Most of the time the relay will hold in the contacts and break the circuit – normally closed contacts are used here. But when the power is disconnected (during a power failure) the contacts connect the supply to the inverter and the indicator lights.

(The meter can be made to measure the current drawn by the inverter from the battery by bending the needle slightly as in Fig. 2).

The inverter section is a simple positive feedback oscillator with frequency dependent on the value of C1 and the inductance of T1. This is wound as in Table 1 onto a Mullard LA5 or LA7 pot core. R1, 2 make a voltage divider for the biasing of C2, 3. With C3 in circuit the inverter in fact becomes more efficient, but has a lower output. Using a 2N3055 enables the inverter to be modified to give higher output and also guards against thermal runaway when used for long periods. A heatsink is advisable.

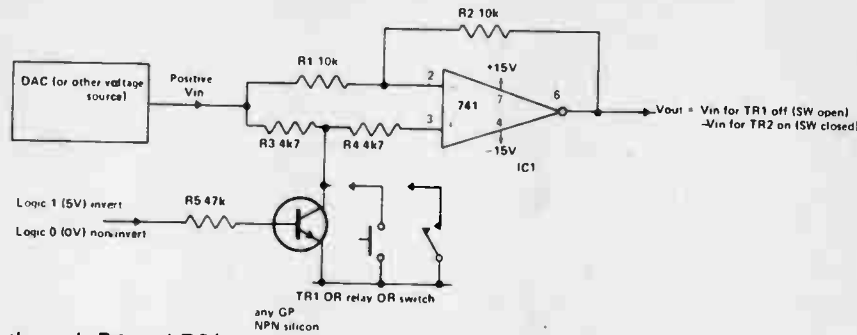


The frequency of the oscillator should be kept as high as possible as the fluorescent light gives out 20 per cent more light if operated at a frequency above 10kHz. The circuit is designed to operate a 12 inch 8W tube but can probably be made to produce up to 30W.

DUAL MODE AMP

This circuit was designed for use as the output of a relatively low accuracy (2%) Digital to Analog Converter (DAC). The DAC was required to have a bi directional output, i.e. both positive and negative. It is quite expensive to provide this normally, so the simple invert/non invert amplifier was used.

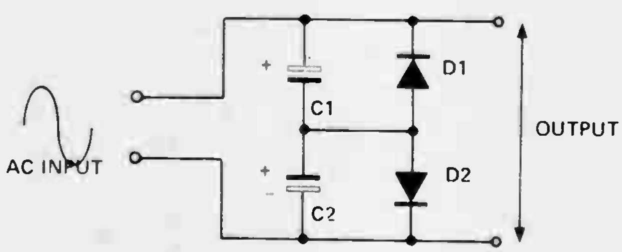
With TR1 turned off the Operational Amplifier IC1 acts as a voltage follower because its inverting and non inverting inputs are at the same potential. In this sense the amplifier has unity gain and little offset. The only errors are caused by the amplifier offset (a few millivolts) which will appear at the output, and the voltage caused by the input currents flowing



through R1 and R3/4.

With TR1 turned on, the non-inverting input of the Op Amp is connected to 0V via R4, and the amplifier now behaves as a conventional unity gain inverting amplifier. The performance in this mode is not as good. The saturation voltage at TR2 collector (typically 0.2V) will appear at the output subtracted from Vin. This limits the performance of the circuit to relatively low accuracy applications. It is possible to switch in a bias offset, but the complexity of this will usually be matched by having a bi-directional DAC.

CAPACITOR SUBSTITUTION

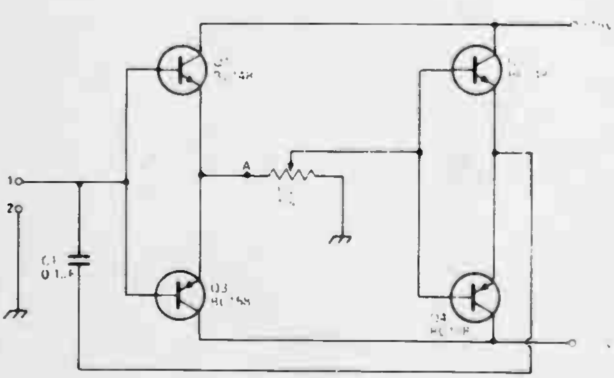


Quite often, especially when constructing operational amplifier circuits, capacitors of several μFarad values are required and they must be of the non polarised type (non electrolytic). One may not have these readily available and as a short term measure I suggest the use of two tantalum capacitors to replace one non-electrolytic, as in circuit diagram.

One can use any diodes, bearing in mind voltages in use, especially maximum reverse voltage that the tantalum will stand. Normal circuit criteria will of course apply to type and value of capacitors chosen. (Two capacitors in series will give a total capacitance of only half the capacitance of one of the capacitors, providing there are of the same value).

The actual operation of this circuit I think is self evident. A negative going voltage would be shorted out across C1 by D1 and applied across C2. A positive going voltage would be applied across C1 and D2 will short C2.

ELECTRONIC CAPACITOR

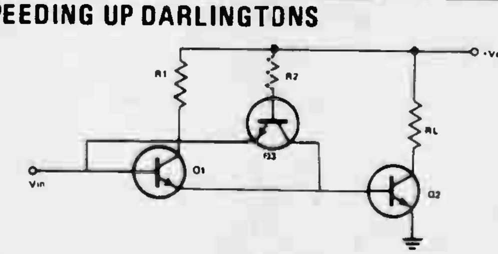


The value of capacitance existing between points 1 and 2 may be varied over a 1000 to 1 range by RV1.

The lower value of capacitance is due to C1, the transistor stages effectively multiply this capacitance, thus the total capacitance available from the circuit, as given, is 100μF.

It is possible to replace RV1 by a NTC or PTC resistor and thus the value of capacitance will depend on ambient temperature.

SPEEDING UP DARLINGTONS



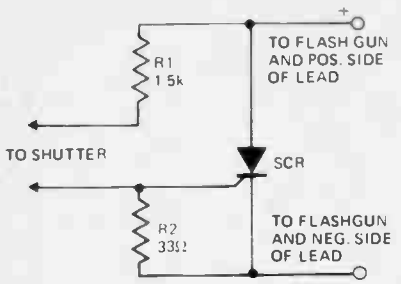
The useful properties of Darlington pairs are somewhat nullified when you need to get any speed out of them. The main drawback of the conventional Darlington in this respect is the long turn-off time, which results from the stored charge at the base of the output transistor.

However, by borrowing a trick from the designers of TTL this situation may be greatly improved, by the familiar looking addition of transistor and resistor R2 to the conventional circuit.

Q3, operating in common - base, draws a relatively steady base current. In consequence, when switching, the base charge of Q3 remains reasonably static, with only the distribution gradients varying. Since the time needed for this redistribution of charge is very small, base drive for the output transistor becomes available within several nanoseconds of positive drive at the input.

At turn-off, Q3 provides a TTL style path for the removal of base excess charges from the output stage resulting in faster turn-off time.

SHUTTER SAVER



This three component device will keep sparks out of your camera shutter, when using a flash gun, by letting an SCR carry the firing current.

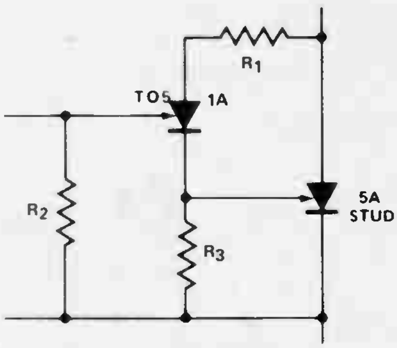
Closing the contacts in the camera shutter applies a triggering voltage, developed across the divider R1, R2, to the gate of the SCR., so firing the SCR and hence the shutter. The value of R1 must be as high as possible, but low enough to carry the needed gate current which may vary from 0.5 mA to a few milliamps. The value of R2 must be sufficiently high to develop necessary gate triggering voltage. R1 and R2 may need to be varied with SCR selected. The SCR should have a rating of 200 volts. If the device is used with a battery-capacitor flash, which operates at about 22 volts, R1 and R2 must be adjusted to suit.

The components can be assembled in a plastic pill tube, fitted in the middle of a flash lead.

INCREASED SENSITIVITY FOR HEAVY CURRENT THYRISTORS

A typical stud thyristor of 5A rating will need 10mA or more for triggering into conduction. This can be reduced to 1mA or less by using an additional 1A thyristor of T05 construction.

The value of R_1 will depend on circuit voltage, ranging from 47Ω at 12V to 1k at 240V R_2 and R_3 are equal, and normally specified in the circuit, being typically 1k or more. The small thyristor should have exactly the same voltage rating as the larger one.



SOUND OPERATED FLASH

The circuit shown enables near instantaneous synchronisation between sound and flash. The latching facility has been incorporated so that the flash is not retriggered.

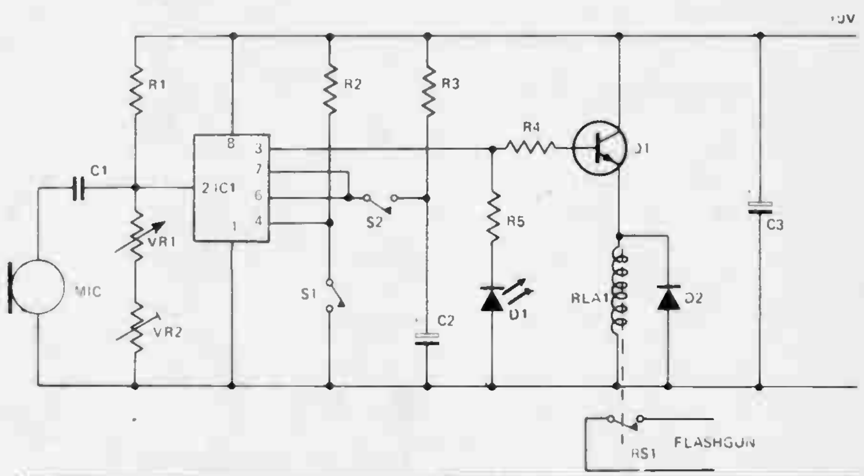
Resetting is by S1. With S2 closed, resetting occurs after a time after a time determined by 1.1xC2xR3. This approximately equals five seconds as shown.

D1 indicates triggering and is used when setting the sensitivity:- set VR1 to zero and increase VR2 until D1 just fails to light. A sharp snap of the fingers causes it to light. S2, R3 and C2 maybe omitted if not required.

The output from IC1 via R4 and Q1 activates the read relay, used for its simplicity and speed of action. The relay itself is connected across the flash sync leads.

The unit maybe battery powered e.g. PP3, as it consumes a mere 15mA or so.

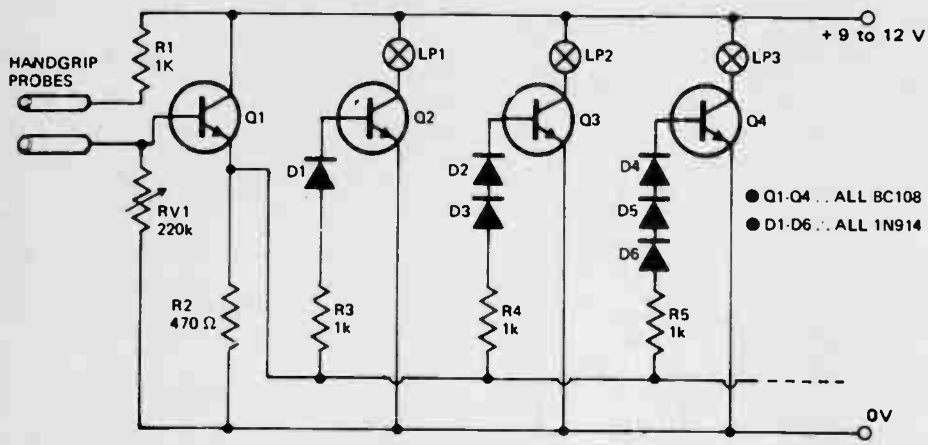
The circuit has been built as the result of many modifications to other circuits. I find it an interesting toy and in its simplicity, should cost less than £2.



COMPONENTS *(see text)

Resistors			**Capacitors**	
R1	330kΩ		C1	10nF
R2	22kΩ		C2	4.7µF, 10V*
R3	1MΩ	* all ¼W, 10%	C3	100µF, 10V
R4	4k7Ω		**MISC.**	
R5	560Ω		mic	crystal mike insert
Potentiometers			S1	push to make switch
VR1	50kΩ	lin	S2	spst switch
VR2	250kΩ	preset	RLA1	reed relay coil
Semiconductors			RS1	normally open reed switch
IC1	NE555			
Q1	BC108			
D1	TIL209 LED			
D2	1N914			

STRENGTH TESTER



Here is a circuit that will quickly tell you and your friends whether it is safe to go onto the beach with no chance of someone kicking sand into your face or conversely that a session with the Charles Atlas academy is advisable!

The idea is to grip the two handgrips which are made out of 25 mm wood dowling (broom handle) covered with aluminium foil.

The stronger the grip, the better the electrical contact made and depending on the strength of grip one two or more lamps will light up.

The circuit operates on the principle that skin contact resistance can be determined to some extent by the amount of pressure applied between the palms and the probes.

The greater the pressure, the lower the resistance and hence the higher the voltage output of the emitter of Q1.

Bases Q2, Q3 . . . are connected to the emitter follower via progressively more series diodes (D1, D2 . . . D6 . . .). Each lamp in the collector circuit will require a progressively higher voltage output from Q1 emitter to ignite. (i.e. a stronger grip).

The number of lamps can be increased as much as one likes, with each stage input having a larger number of series diodes. In the further stages it is not necessary to stack all those diodes since a single Zener will do just as well. For more than four stages reduce the value of R2 to 220 ohms.

RV1 adjusts the sensitivity: Reducing its value lowers the sensitivity.

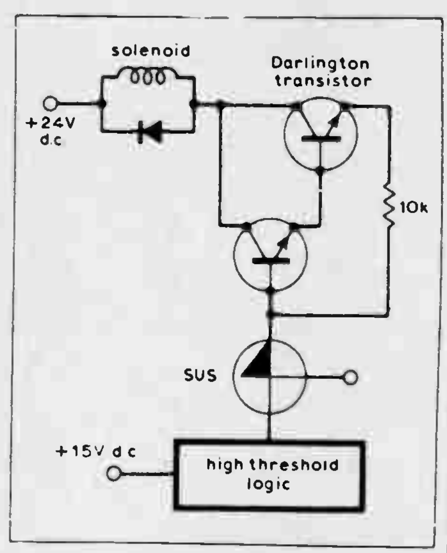
INCREASING THE NOISE IMMUNITY OF LOGIC CONTROL SYSTEMS

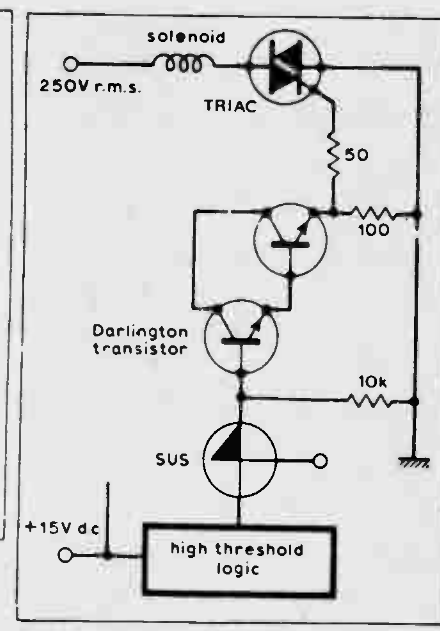
In many industrial control systems, the output from logic circuitry is used for simple on/off control of a solenoid. However, the situation is often complicated by the presence of high levels of electrical noise. Although a high input noise immunity of the logic circuitry can be obtained with the use of high threshold logic ICs, such as Motorola Semi-conductors' MHTL range, the power amplifier feeding the output logic signal to the solenoid must be specially designed for maximum noise immunity in both the on and off conditions.

An extremely simple and low-cost solution involving the use of a silicon unilateral switch (SUS) has been proposed by Motorola Semiconductors.

The SUS is connected between the logic circuitry and the output amplifier as shown in Fig. 1 for a dc solenoid, and for an ac solenoid the connection is shown in Fig. 2. With an SUS with a V_s of 8 V and a V_f of 1.3 V, positive-going noise pulses with the solenoid switched off would have to exceed 8 V for a spurious energisation of the solenoid; negative-going noise pulses occurring while the solenoid is on would have to reduce the input to the output amplifier to below 2 V, from between 12.5 and 15 V, for a spurious de-energisation.

An additional advantage is that the regenerative portion of the SUS switching waveform effectively boosts the drive to the inductive load of the solenoid, speeding up the response to control signals.





High-threshold logic, operating from a 15-V supply, produces a logic '0' output of 0 to 1.5 V and logic '1' output of 12.5 to 15 V. Noise amplitudes which increase the '0' level to 6.5 V and reduce the '1' level to 8.5 V can be tolerated.

IDENTIFYING 74 SERIES SUPPLY PINS

When unmarked IC's are suspected of belonging to the 74 series a simple method of finding the supply pins is possible. If the resistance is checked between any two pins using a multimeter set on the 1 ohm range, all pins with the exception of the supply pins will give readings as open circuit or as a diode. When the supply pins are checked a reading is obtained both ways, the lower reading will be obtained when the positive lead from the multimeter is connected to Vcc.

SOLDERING IC's

As it is no longer an economic proposition to use IC sockets for the cheaper IC's on the market, a method of soldering them without damage can be extremely useful. Cheap commercial soldering heatsinks do not appear to be available, but sprung letter clips could have been made for the job. These are available from most stationers.

These are almost an exact fit for a 14-pin DIL IC. They clamp tightly over the tops of the IC pins, ensuring that heat is rapidly dissipated and that the pins are all at the same potential (preventing damage to CMOS IC's).

These clips could save a small fortune in IC's, they also enable IC's to be unsoldered without damage providing care is taken.

TINNING WITH SOLDER WICK

Do not discard the lengths of solder saturated solder wick. Further use can be made of them to plate printed-circuit boards by pre-tinning the joints, prior to inserting components and soldering.

The simple operation is as follows – place the saturated solder wick on the printed board and apply a heated soldering iron to melt the solder in the wick. At the same time, move the wick and iron along sections or joints requiring tinner. A neat plated copper print will result.

STENCIL FOR PCB's

A childs plastic geometry set-square makes a very useful stencil when using etch-resist pens. The holes should be slightly counter-sunk to avoid smudging. Some suggested configurations are; 8 pin DIL (easily moved for 16 pin), 0.1" edge connector slots, your 'favourite' relay base, preset pot holes, and if you want to be very professional, pairs of holes the correct distance apart for the different sizes of resistors and capacitors.

FINISHING FRONT PANELS

The finish on aluminium panels can be improved by etching them in a caustic soda solution. To get the best effect:

1) Do all marking out on the back of the panel.

2) Drill holes two ways – small pilot hole from front to back. This minimises the problem of getting rid of the 'flash' which arises round the holes while drilling. Removing flash often leaves scratches, and it is better that these be on the back of the panel than the front.

3) Rub the front of the panel with medium grade emery cloth to rid it of all unwanted marks and scratches. The emery should be rubbed only in one direction for the final rubbing. This leaves the aluminium with a bright matt finish. From this point on, avoid touching the front of the panel.

4) Attach a length of thin plastic string or tubing to the panel by tying it through one of the panel holes.

5) Prepare a caustic etching solution. Put about 30 grams of caustic soda in a glass or plastic dish. (The plastic throw-away food containers are ideal.) Carefully pour on about 300 ml of hot water. (1 oz. of caustic soda in half a pint of water, if that's any easier for you.) The strength of the solution is in no way critical. Now, by means of the plastic string, lower the panel into the solution, leaving one end of the string hanging out of the dish. It will fizz fiercely and the solution will get hotter – but all is well.

6) About 3 minutes later remove the panel, rinse it under a cold water tap, and wipe it clean. Rinse it again thoroughly, and if it looks O.K. – dull matt all over, it's finished. Hang it to dry.

A panel finished this way has a satin chrome look to it, and does not retain finger marks the way untreated aluminium does.

DIL DRILLING

Drilling holes in a pcb for 14 and 16 DIL ICs is quite difficult and if the holes are slightly off centre it is tricky to fit the IC. An easy way to get it right use a small piece of 0.1 matrix veroboard as a template.

FLICKER-FREE FLUORESCENT STARTING

Here is an extremely simple, yet effective modification which will eliminate the annoying flickering when a fluorescent lamp is first switched on.

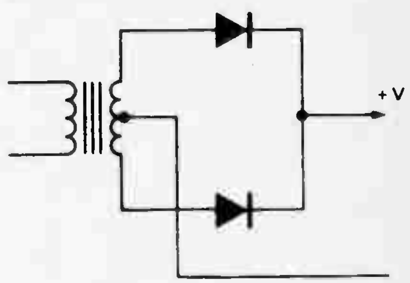
The modification consists of inserting a diode (P.I.V. about 600 V) in series with the starter. This results in a fairly heavy current on initial switch-on, which heats the filaments quickly. When the starter contacts open again, the lamp fires immediately.

NOTE: The effectiveness of the modification, depends largely upon the characteristics of the starter; try and find one that is quick-acting.

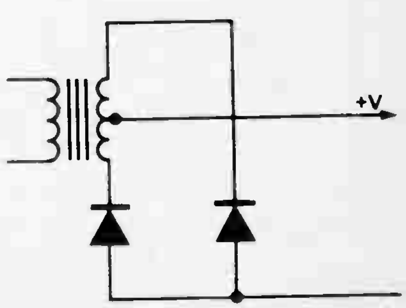
My original unit has been working successfully in my desk lamp for the past three years, and I've had no problems with dc magnetisation of the ballast, or excessive power consumption on switch-on.

AVOIDING INSULATED HEAT SINKS

If a fairly heavy current is to be taken from the type of power supply shown in (1), then the diodes will be of the stud type on insulated heat sinks. By choosing stud anode diodes, and using arrangement (2), the chassis may be the heat sink without the need for insulation.

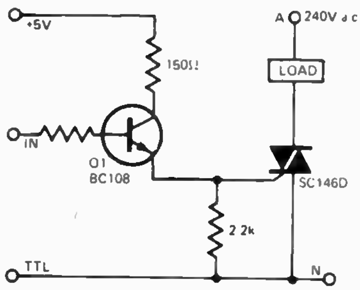


(1)



(2)

TTL-MAINS INTERFACE

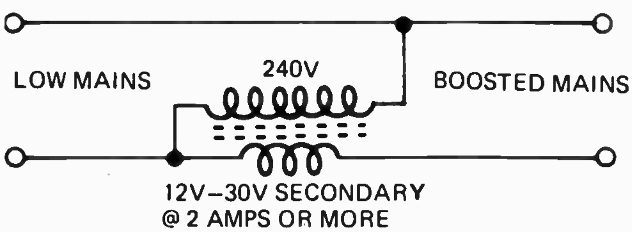


Here is a useful circuit for driven mains operated devices direct from TTL logic circuits. Although it works well, it has the inconvenience that the neutral line is connected to circuit ground.

For inputs other than TTL levels a 10k series resistor may need to be connected between Q1 base and ground to reduce leakage.

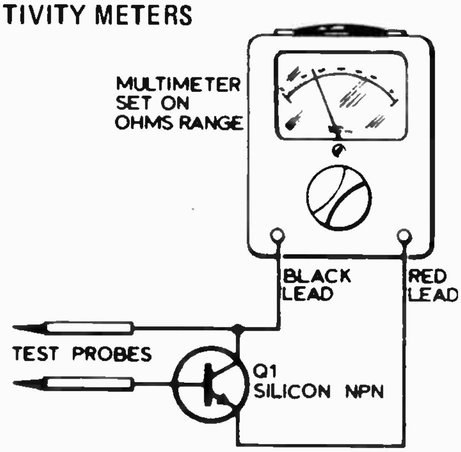
Approximately 1mA at 1.4 volts is required to switch Q1 on. If driving from a low impedance, some means of current limiting will also be required.

BOOST YOUR MAINS



Mains voltage may be boosted by up to 10% by using a standard filament transformer connected as shown above.

MEASURING HIGH RESISTANCE ON LOW SENSITIVITY METERS



Many inexpensive multimeters are unable to give useful readings on the ohms range much above 47k. However, by using almost any silicon NPN transistor (BC107 for example) in the arrangement shown will give considerable deflections for quite high values. The meter's scale will not apply but by noting the readings from high tolerance, high value resistors and some interpolation, fairly accurate measurements can be made. Note that on a multimeter the Black (negative) lead connects to the battery positive on the ohms range.

HIGH VOLTAGE ELECTROLYTICS

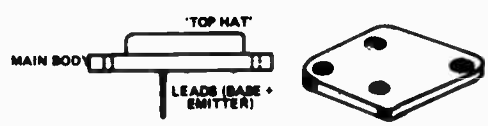
If you have difficulty in getting hold of an electrolytic with a sufficiently high voltage rating for your purpose, you can use two or more in series as long as the combined voltage rating exceeds that of the supply.
This will reduce the capacity of course; for instance if two capacitors with equal values are used, the effective value will be half.
A high value resistor (about 1Mohm) should be wired in parallel with all capacitors used in this way to stabilise the voltage across each.

TRANSISTOR IDENTIFICATION

Those who do a lot of lashups on S-Dec etc and use the same components time after time will find that transistor identifications soon rub off. Why not take a few minutes to colour code your transistors with paint or nail varnish. You can work out your own coding but for instance PNP can be blue, NPN red; high gain types can have yellow, r.f. types green etc. Therefore a PNP, high r.f. transistor would have blue, yellow and green blobs on them.

Woolworths and model shops sell sets of tiny pots of paint for models. This will give you about six colours for only a few pence.

TEMPLATE AND HEAT SINKS FOR POWER TRANSISTORS



Power transistors similar to OC35, 28 etc., can be useful even when there is a complete electrical breakdown. They can be modified and utilized as either a power transistor heat sink, or as a making out and drilling template. Just remove the 'top hat' part of the transistor by squeezing it in the jaws of a vice.

The top hat should fully detach itself from the main body and the main body can be modified further by filing it flat. Then remove the ceramic insulators and base and emitter leads by pulling with a pair of pliers. Little effort is required to do this.

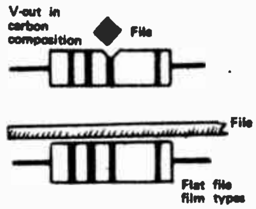
TRANSISTOR SOCKET

If you have ever built a transistor tester and use it a lot, for testing large batches of unmarked devices for instance, you are probably aware of the shortcomings of normal transistor sockets which, in fairness, are not designed for continuous use.
A really hardy socket can be made from a B9A or B7G valve plug (not socket). These consist of pins which are hollowed out and are mounted on an insulated base. The spacing between the pins is ideal for use with most types of transistor.

SOLDER FLOW PROBLEMS

If you solder a lot of p.c. boards and are cursed with the solder flowing across adjacent tracks, chances are that you are either using the wrong type of bit on your iron or the wrong size of solder. There seem to be a lot of people who are unaware that solder is widely available in 22s.w.g. as well as 18s.w.g.

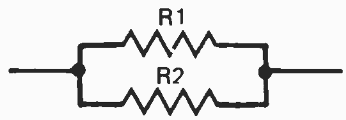
ODD RESISTOR VALUES



If you are faced with finding an odd value resistor, e.g. for multimeter repairs or making shunts for meters, a simple trick is to take a resistor lower in value than that needed and file it until the required value is reached. The resistor should be connected to an accurate multimeter, preferably a digital type, when being filed. Carbon composition or film-types may be used.

The power rating of resistors may be reduced slightly by the filing. When finished, a coating of modeling paint or epoxy resin should be given over the area that has been filed, to prevent moisture changing the resistor value.

RESISTORS IN PARALLEL



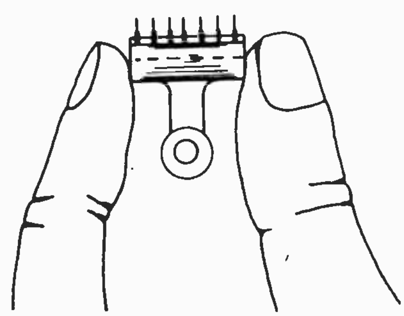
R2 \ R1	10	12	15	18	22	27	33	39	47	56	68	82	100
10	5.00	5.45	6.00	6.43	6.88	7.30	7.67	7.96	8.25	8.48	8.72	8.91	9.09
12	5.45	6.00	6.67	7.20	7.76	8.31	8.80	9.18	9.56	9.88	10.20	10.47	10.71
15	6.00	6.67	7.50	8.18	8.92	9.64	10.31	10.83	11.37	11.83	12.29	12.68	13.04
18	6.43	7.20	8.18	9.00	9.90	10.80	11.65	12.32	13.02	13.62	14.23	14.76	15.25
22	6.88	7.76	8.92	9.90	11.00	12.12	13.20	14.07	14.99	15.79	16.62	17.35	18.03
27	7.30	8.31	9.64	10.80	12.12	13.50	14.85	15.95	17.15	18.22	19.33	20.31	21.26
33	7.67	8.80	10.31	11.65	13.20	14.85	16.50	17.88	19.39	20.76	22.22	23.53	24.81
39	7.96	9.18	10.83	12.32	14.07	15.95	17.88	19.50	21.31	22.99	24.79	26.43	28.06
47	8.25	9.56	11.37	13.02	14.99	17.15	19.39	21.31	23.50	25.55	27.79	29.88	31.97
56	8.48	9.88	11.83	13.62	15.79	18.22	20.76	22.99	25.55	28.00	30.71	33.28	35.90
68	8.72	10.20	12.29	14.23	16.62	19.33	22.22	24.79	27.79	30.71	34.00	37.17	40.48
82	8.91	10.47	12.68	14.76	17.35	20.31	23.53	26.43	29.88	33.28	37.17	41.00	45.05
100	9.09	10.71	13.04	15.25	18.03	21.26	24.81	28.06	31.97	35.90	40.48	45.05	50.00
120	9.23	10.91	13.33	15.65	18.59	22.04	25.88	29.43	33.77	38.18	43.40	48.71	54.55
150	9.38	11.11	13.64	16.07	19.19	22.88	27.05	30.95	35.79	40.78	46.79	53.02	60.00
180	9.47	11.25	13.85	16.36	19.60	23.48	27.89	32.05	37.27	42.71	49.35	56.34	64.29
220	9.57	11.38	14.04	16.64	20.00	24.05	28.70	33.13	38.73	44.64	51.94	59.74	68.75
270	9.64	11.49	14.21	16.88	20.34	24.55	29.41	34.08	40.03	46.38	54.32	62.90	72.97
330	9.71	11.58	14.35	17.07	20.63	24.96	30.00	34.88	41.14	47.88	56.38	65.68	76.74
390	9.75	11.64	14.44	17.21	20.83	25.25	30.43	35.45	41.95	48.97	57.90	67.75	79.59
470	9.79	11.70	14.54	17.34	21.02	25.53	30.83	36.01	42.73	50.04	59.41	69.82	82.46
560	9.82	11.75	14.61	17.44	21.17	25.76	31.16	36.46	43.36	50.91	60.64	71.53	84.85
680	9.86	11.79	14.68	17.54	21.31	25.97	31.47	36.88	43.96	51.74	61.82	73.18	87.18
820	9.88	11.83	14.73	17.61	21.43	26.14	31.72	37.23	44.45	52.42	62.79	74.55	89.13
1000	9.90	11.86	14.78	17.68	21.53	26.29	31.95	37.54	44.89	53.03	63.67	75.79	90.91

CMOS DIL HANDLING DEVICE

For those of us that get into a cold sweat when handling CMOS devices, we can all now sign with relief. A cheap solution in the form of a spring clip can be obtained from a well-known stationery chain-store. The clips are called "letter clips" and cost 30p for a card containing 12 clips.

In the light of experience the following points are worth noting.

1. Before using the clips any internal burrs should be removed, as these will prevent a good contact being made with the IC pins.
2. NEVER remove the IC from the impregnated foam until the clip has been fitted.



3. When fitting the clip, ensure that it 'shorts out' ALL the IC pins.
4. During IC insertion or removal from a PCB or socket, the IC and clip are gripped TOGETHER across their ends, with the thumb and forefinger (see sketch). This proceedure should remove any chance of the clip accidentally releasing the IC.
5. If the IC is to be soldered into a PCB, the clip should be left in place during the soldering process, as it will also act as a heat-sink. If several IC's are to be soldered into a board, it is worth while leaving all the clips attached until the last IC has been soldered in position.

IDENTIFYING SURPLUS IC's

In checking unmarked surplus IC's, a clue can be gained as to the identity of the IC if the ground pin can be located first. In epoxy encapsulated IC's, the truncated part of the lead frame can be seen at both ends, perhaps partially covered by moulding flash. This is generally connected to the substrate. In TTL and most linear IC's, this is the most negative pin (ground). In PMOS (clock and calculator chips) this is Vss, the most positive pin. An ohmmeter can find which pin is connected to the substrate by touching one probe to the frame and the other to each pin in turn.

Another clue to whether the IC is linear or digital is the fact that most digital IC's have diode protection against reverse bias at inputs. Knowing the ground pin, this can be checked rapidly.

EXTENDING BATTERY LIFE

Cell batteries are expensive but it is possible to recharge these to a certain extent. If they are connected to a suitable power supply (positive to positive, negative to negative) before they are too run down, it is possible to extend the life considerably. This can be done several times though each recharging will last a shorter time.

Layer batteries (PP3, PP7, PP9 etc) cannot be recharged in this way.

POWER SUPPLY OR BATTERY?

If you want to work out if it is worth building a power supply rather than using a battery a quick rule of thumb is that the juice from a battery comes to between 1,000 and 100,000 times the expense. The cost of a typical power supply which is comparable to a battery will cost between £1 and £3 so if you are likely to purchase that value of batteries in the lifetime of the equipment and if portability is not a consideration, plump for the power supply.

MUCK REMOVER

When you have had the best from your toothbrush, don't throw it away, add it to your tool box. It's one of the best ways of clearing off dust, dirt and swarf.

LOUDSPEAKER CHECKING

If you suspect that the coil on a loudspeaker is out of alignment and rubbing you can check this easily. Hold the cone of the speaker against your ear and gently thump the back of the magnet with the flat of your hand. If the thump is 'clean' the problem lies elsewhere. If a rasping or scratching is heard the coil is out of alignment.

Pressing in the cone with your fingers should be avoided for unless the pressure is even you can cause the problem you are looking for.

SIGNAL TRACER

There are now plenty of pocket transistor radios around which suffer from some fault on the r.f. or i.f. side. These are ideal for conversion to an audio signal tracer.

The earphone socket can be modified as the input by wiring across the volume control and disconnecting the original connections to this from the r.f. section.

CRYSTAL EARPIECES

One of the cheapest and most useful items of test equipment is a simple crystal earpiece. These are very sensitive and have such a high impedance that they will not load the circuit. Cut off the plug and add a couple of croc clips.

CHEAPIE VARICAPS

Varicap diodes are not all that cheap or easy to get hold of. What is not widely known is that all diodes exhibit a variable capacity effect when reverse biased. Generally speaking the higher the current rating, the larger the capacity and the change when the voltage is varied across it. Even the 1N4000 series with a 1A capacity can be used when small capacity changes (a few picofarads) are required.

BATTERY SNAPS

When a layer type battery (PP3, PP7, PP9 etc) is exhausted, don't throw it away; the connectors can be used as battery snaps for other equipment.

Using a pair of pliers, bend away the metal lip from around the top and the panel holding the connectors will come free. There is usually a wire running to the underside of the layer stack: this should be cut.

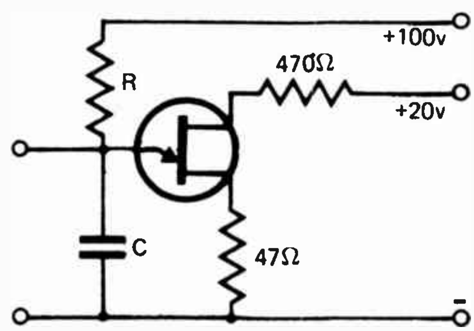
BATTERY CHECKING

Never check a battery off-load simply by using a voltmeter – the readings can be meaningless. Measuring on voltage on-load is o.k. but if this is not practical ensure that you connect a resistor across the meter probes so that a reasonable current is drawn. For a 9V battery a 180 ohm resistor will draw about 5mA.

TRANSFORMERS IN REVERSE

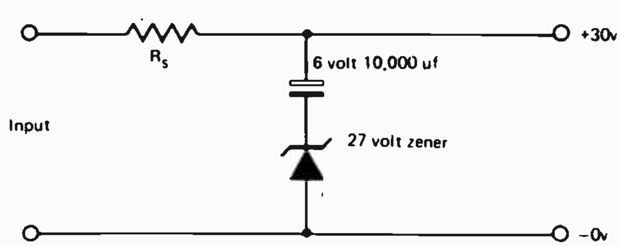
It is frequently overlooked that there is nothing magic about the primary and secondary windings of a transformer: a 250V to 9V will operate equally as well as a 9V to 250V.

IMPROVING UJT LINEARITY



The linearity of a UJT relaxation oscillator may be improved by returning the 'timing' resistor R to a high voltage supply.

ZENER DIODE LIFTS CAPACITOR RATING

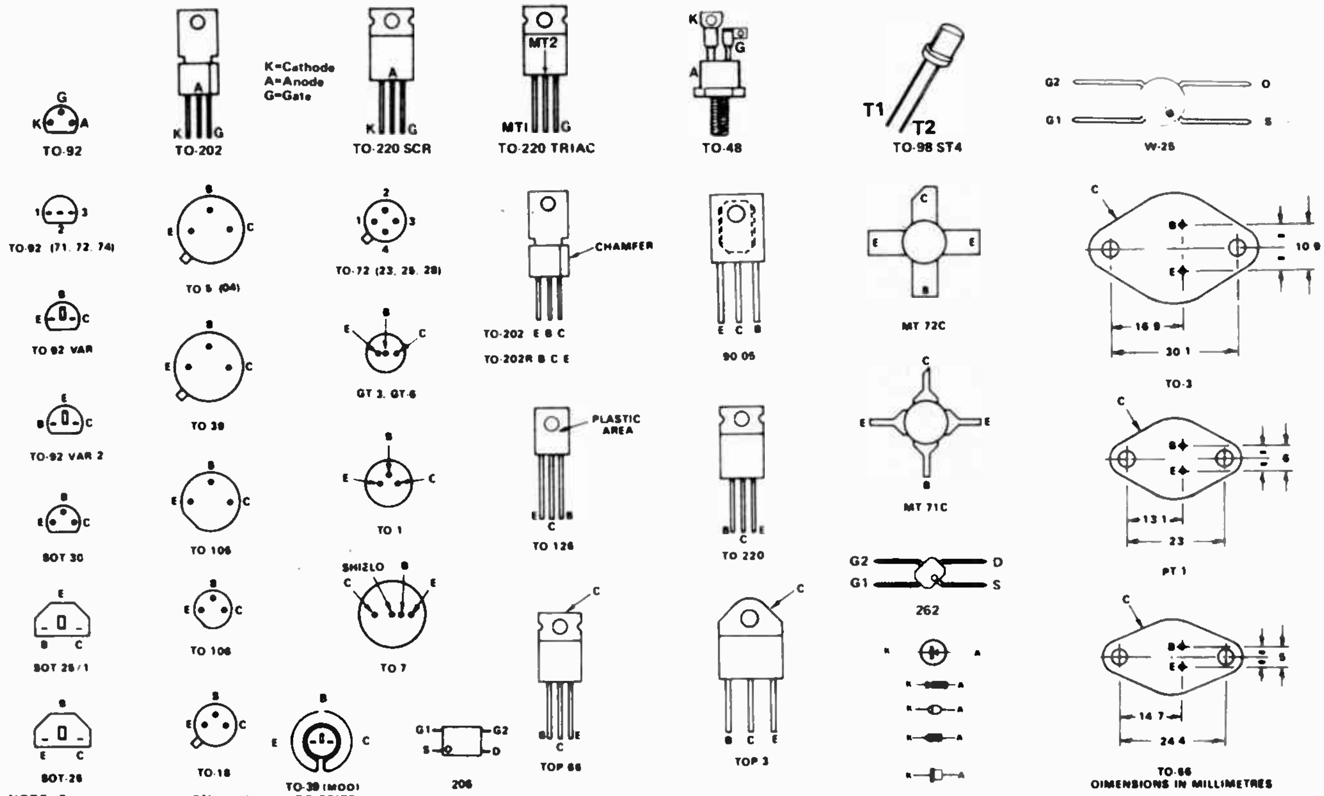


Electrolytics combining large capacity and high working voltage are bulky, expensive, and frequently difficult to obtain.

A drastic reduction in the voltage rating required is achieved by the connection of a series zener diode. In this example a 27 volt zener in series with a 6 volt electrolytic filters a 30 volt line.

MISCELLANEOUS DATA

SEMICONDUCTOR OUTLINES



NOTE For transistors with "PN" prefix, use TO-92(72) pin connections

PIN	72 (STD) T	72 (STD) FET
1	C	G
2	B	S
3	E	D

PIN	71 T	71 FET
1	C	G
2	E	D
3	B	S

PIN	74 T	74 FET
1	B	S
2	C	G
3	E	O

PIN	T (25)	FET (25)
1	E	S
2	B	O
3	C	G or G2
4	case	case or G1

PIN	T (28)	FET P (23)
1	B	S
2	E	G
3	C	O
4	GNO	CASE

BIPOLARS
E EMITTER
B BASE
C COLLECTOR
NG NPN GERMANIUM
PG PNP GERMANIUM
NS NPN SILICON
PS PNP SILICON

FETS
S SOURCE
G GATE
O ORAIN
N/CH N CHANNEL
P/CH P CHANNEL

DIODES
A: ANODE
K: CATHODE

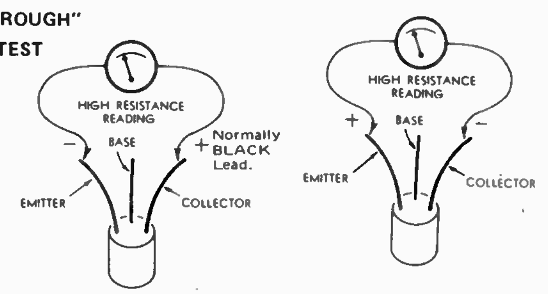
G P GENERAL PURPOSE
S S SMALL SIGNAL
SW SWITCH
O/P OUTPUT
R F RADIO FREQUENCY
H F HIGH FREQUENCY
V H F VERY HIGH FREQUENCY

QUICK TRANSISTOR CHECK

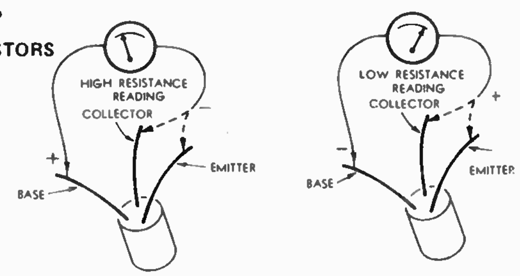
These diagrams illustrate a handy method of checking any transistor with a multimeter switched to a low Ohms range.

NOTE Battery (+) positive on most multimeters is the BLACK lead — NOT the RED lead

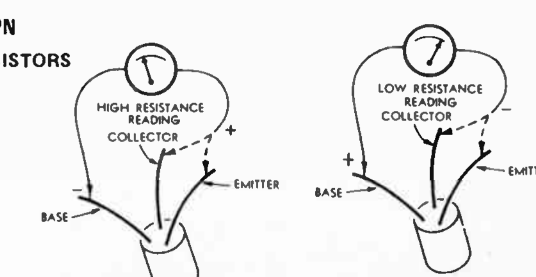
"PUNCH-THROUGH" TEST



PNP TRANSISTORS

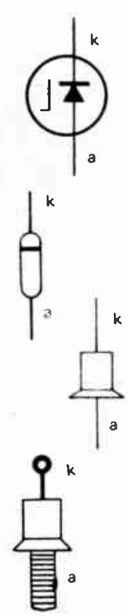


NPN TRANSISTORS



ZENER DATA

VOLTAGE	400mW TYPES*	1 W TYPES**
3.3	IN746	
3.6	IN747	
3.9	IN748	
4.3	IN749	
4.7	IN750	IN4732
5.1	IN751	IN4733
5.6	IN752	IN4734
6.2	IN753	IN4735
6.8	IN754,IN957	IN4736
7.5	IN755,IN958	IN4737
8.2	IN756,IN959	IN4738
9.1	IN757,IN960	IN4739
10	IN758,IN961	IN4740
11	IN962	IN4741
12	IN759,IN963	IN4742
13	IN964	IN4743
15	IN965	IN4744
16	IN966	IN4745
18	IN4112	IN4746
20	IN968	IN4747
22	IN969	IN4748
24	IN970	IN4749
27	IN971	IN4750
30	IN972	IN4751
33	IN973	IN4752



* Dissip. 400mW, Op Temp 50°C, Case DO-7 or DO-35
** Dissip. 1W, Op Temp 50°C, Case DO-41. This range will dissipate up to 3 watts @ 75°C with 10mm lead length as heatsink.