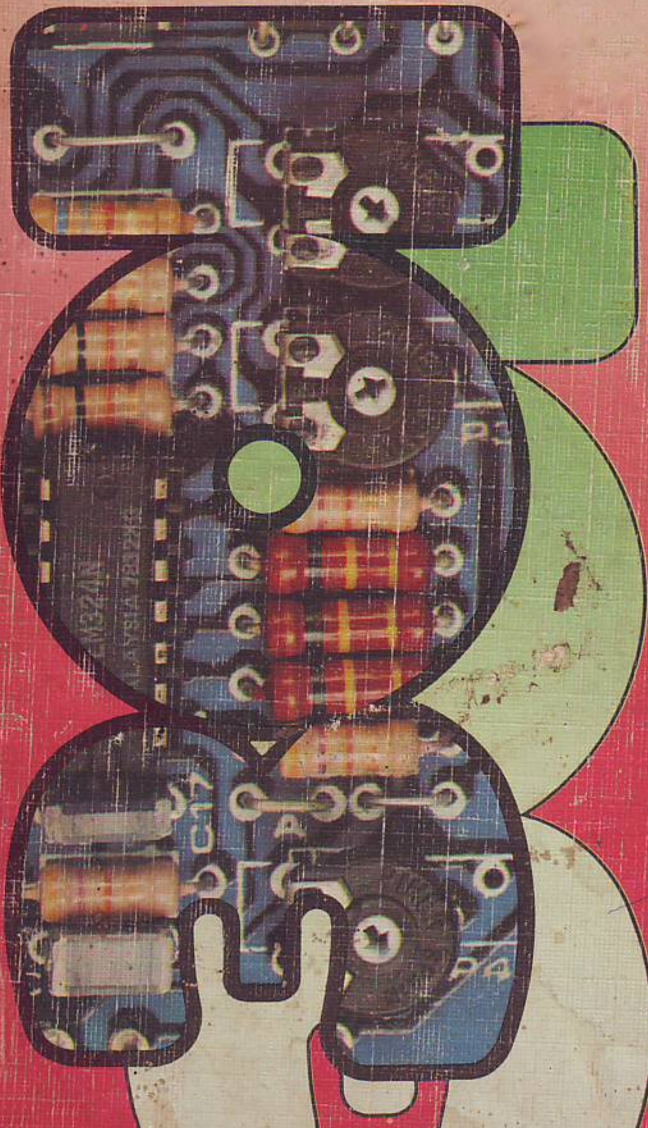
301 circuits

Circuits 180 to 251



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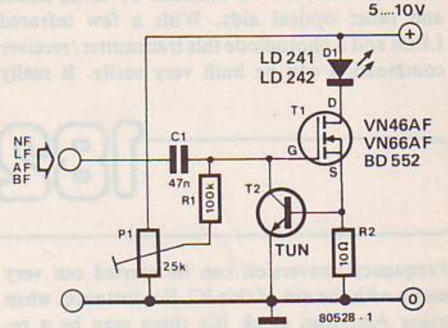
AN elektor PUBLICATION

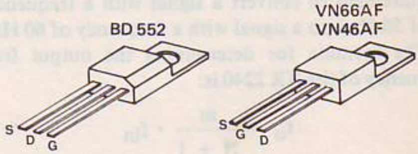
180 Infrared transmitter

This is a very simple circuit and is really intended for hobbyists who like experimenting with infrared or with VFETs, or even both, for that's what this is all about.

The circuit diagram shows an infrared transmitter in its simplest, most elementary form. Its simplicity is achieved with the aid of a VFET. Since FETs unlike other bipolar transistors, reveal a neat linear input to output voltage ratio, it is enough to feed a low frequency signal to the gate and to include an infrared LED in the drain lead. The intensity of the infrared light produced by the LED will then vary according to that of the LF voltage put across it and there you have it – a transmitter.

In order to increase the infrared LED's lifespan, a transistor has been added to achieve current limitation, thereby reducing the FET's maximum drain current to about 60 mA. If the current were to rise, the voltage across R2 would become so high that T2 starts to conduct and the gate of the FET is then short circuited to ground. The low frequency modulation signal which is supplied needs to have a value of around $250\ mV_{eff}$ if the transmitter is to achieve full output. Potentiometer P1 is preset with the input short circuited, so that a voltage of 0.3 volts (drain current 30 mA) is measured across R2.



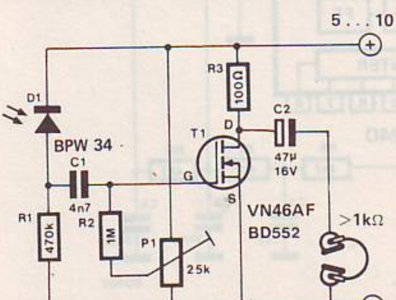


It doesn't matter what type of VFET or what type of infrared LED is used. This is why various types are given in the schematic. In the event of insufficient 'transmission power' several infrared LEDs may be connected in series, as required.

ITT applications

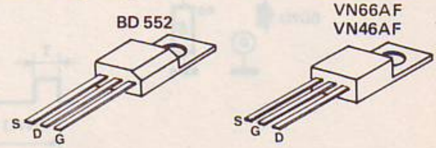
181 Infrared receiver

A transmitter needs a receiver. A receiver will be described here to act as a counterpart to the infrared transmitter. Again, simplicity itself with the aid of a VFET.



The infrared light which falls on the infrared photo diode (here a BPW34 type, but any other will also do) will cause the voltage across R1 to vary. This will of course affect the VFET's gate and the drain current, therefore, will fluctuate according to the modulation of the infrared light received. The modulation can be heard with a set of headphones.

Such simplicity of course has its disadvantages.



For instance, mains light bulbs which happen to be 'on' in the vicinity will be heard as a humming noise. In quiet surroundings, however, fair reception is possible within a range of a few metres, which can be extended by using lenses and other optical aids. With a few infrared LEDs and a photodiode this transmitter/receiver combination can be built very easily. It really works and is ideal for experiments.

One other thing... for it to function properly, P1 will have to be preset so that when the photodiode is completely shielded from light, exactly half the supply voltage will be measured on the FET's drain.

ITT applications

182 Frequency conversion using the XR 2240

Frequency conversion can be carried out very easily with the aid of this IC. For instance, when using American clock ICs there may be a requirement to convert a signal with a frequency of 50 Hz into a signal with a frequency of 60 Hz. The formula for determining the output frequency of the XR 2240 is:

$$f_o = \frac{m}{N + 1} \cdot f_{in}$$

where

f_o = the output frequency

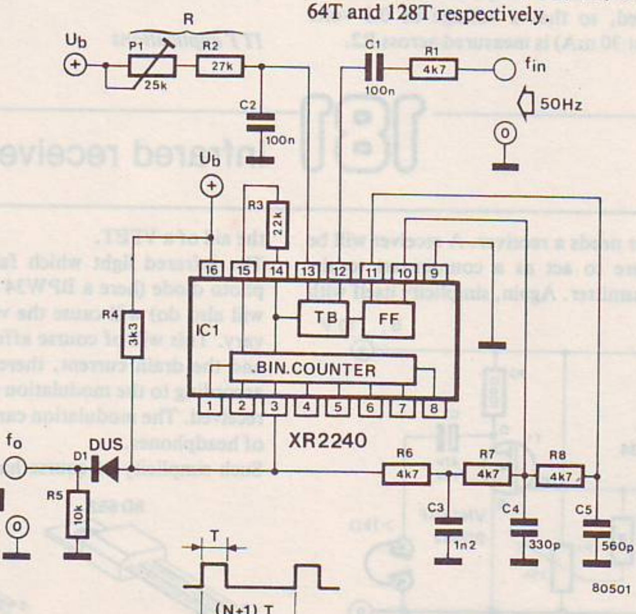
m = the ratio of the input frequency to the time base frequency. This ratio is determined by the position of the potentiometer and is a whole number between 1 and 10.

N = a whole number between 1 and 255 which may be selected by connecting one or more of pins 1...8.

f_{in} = the input frequency

When m = 6 and N = 4 the output frequency will be 60 Hz for an input frequency of 50 Hz. Similarly, when m = 5 and N = 5 the output frequency will be 50 Hz for an input frequency of 60 Hz.

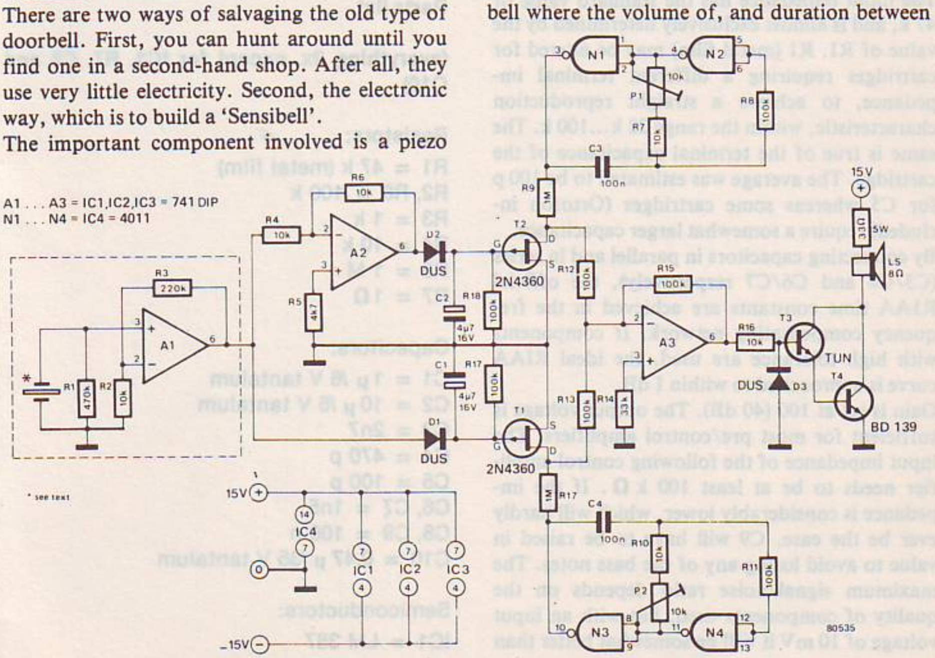
The XR 2240 contains a control flip-flop FF, a time base generator TB and an eight bit binary counter. The period T of the signal produced by the time base generator is determined by the RC product of the components connected to pin 13. Signals are then available at the outputs on pins 1...8 with periods of T, 2T, 4T, 8T, 16T, 32T, 64T and 128T respectively.



If, for example, the outputs T and 4T (pins 1 and 3) are connected to the 3k3 pull up resistor (R4), the generated signal will have a period of T + 4T = 5T. This provides the 'N' factor given in the formula.
A positive going input signal at pin 12 will trigger the time base generator and reset the binary counter. The counter will then operate until the next positive going input pulse occurs.
In the circuit shown the required output frequency is 60 Hz for an input frequency of 50 Hz. Therefore, the frequency of the time base generator needs to be preset to the sixth harmonic of 50 Hz which is 300 Hz (m = 6). This is accomplished with the aid of the potentiometer. With the output from pin 3 (4T) connected, the result is N = 4.
The circuit will operate satisfactorily with a supply voltage of between 4 and 15 volts. With a supply voltage of 9 volts the current comsumption is approximately 8 mA. The synchronisation (input) signal needs to be a square wave with a minimum amplitude of 3 V. The maximum frequency of the time base generator is quoted as being 100 kHz (C2 = 6n8, R = P1 + R2 = 1 k).

183 Sensibell

Remember the old-fashioned copper bell-on-a-chain? In many ways it had considerable advantages over its modern electronic counterparts, advantages which were probably not appreciated at the time. It happened to provide a lot of useful information about the visitor. The way in which he rang – loud, soft, long, short, repeatedly, persistently, etc. said a great deal about him. All this of course is lost with the splendid electronic versions of today.
There are two ways of salvaging the old type of doorbell. First, you can hunt around until you find one in a second-hand shop. After all, they use very little electricity. Second, the electronic way, which is to build a 'Sensibell'.
The important component involved is a piezo element, from an ultrasonic transducer. When a voltage is connected to it it becomes distorted, when distorted it produces a voltage. If we use the element as a bellpush, two voltage peaks are obtained – one when it is pressed and one when released. The height of the voltage peaks corresponds to the pressing force. The interval between both peaks depends on how long the button is pressed.
A simple solution is to construct a ding-dong-bell where the volume of, and duration between



'ding' and 'dong' is determined by the caller. The circuit diagram shows how this is done. The signal originating from the piezo crystal is amplified by A1. Because of the high impedance of piezo elements A1 is best mounted in the doorbell push button. The low impedance output of the amplifier as well as the supply voltage connections can then be connected to the rest of the circuit by a four-core cable. The output of A1 is inverted by A2 so that positive pulses are available both at the beginning and the end of the bell signal. Via T1 and T2 these signals are used to shape the envelopes which modulate the amplitude of two oscillators. The two oscillators produce the 'ding' and 'dong' and are constructed around a single 4011 IC. Their pitch can be preset with P1 and P2 respectively.
A simple output amplifier (T3/T4) completes the circuit.
One word of advice – it is best to power the LF amplifier separately – from the rectified bell transformer voltage, for example. The double 15 V supply then only has to provide a few mA. If after installing the device, the bell sounds like a 'dong-ding' instead of a 'ding-dong', this can be remedied by exchanging the piezo element connections. In any case, the piezo element needs to be removed from its original case with the greatest of care before the connecting wires can be soldered to it.
The element can be protected against outside influences with a coat of epoxy resin or double compound glue.

W. van Dreumel

184 Stereo dynamic preamplifier

The circuit diagram shows how simple it is to use it to build a high quality dynamic preamplifier using just one opamp, in this case the LM 387 from National.
The input impedance has the standard value of 47 k, and is almost exclusively determined by the value of R1. R1 (metal film) may be altered for cartridges requiring a different terminal impedance, to achieve a straight reproduction characteristic, within the range 22 k...100 k. The same is true of the terminal capacitance of the cartridge. The average was estimated to be 100 p for C5 whereas some cartridges (Ortofon included) require a somewhat larger capacitance.
By connecting capacitors in parallel and in series (C3/C4 and C6/C7 respectively), the official R1AA time constants are achieved in the frequency compensating network. If components with high tolerance are used, the ideal RIAA curve is approached to within 1 dB.
Gain is set at 100 (40 dB). The output voltage is sufficient for most pre/control amplifiers. The input impedance of the following control amplifier needs to be at least 100 k Ω. If the impedance is considerably lower, which will hardly ever be the case, C9 will have to be raised in value to avoid losing any of the bass notes. The maximum signal/noise ratio depends on the quality of components used, but with an input voltage of 10 mV it will be somewhat better than 80 dB.
If we double everything shown in the diagram with the exception of the decoupling components.

Parts list

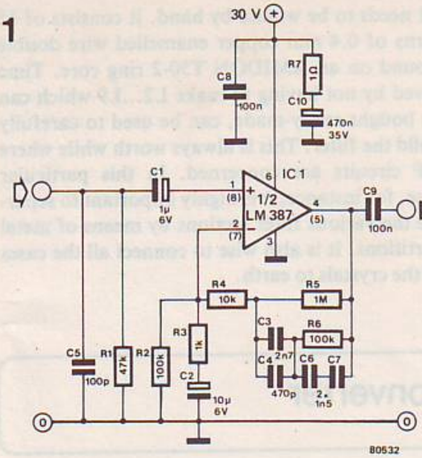
(everything 2x, except for IC1, R7, C8 and C10)

Resistors:
R1 = 47 k (metal film)
R2, R6 = 100 k
R3 = 1 k
R4 = 10 k
R5 = 1 M
R7 = 1 Ω

Capacitors:
C1 = 1 µ /6 V tantalum
C2 = 10 µ /6 V tantalum
C3 = 2n7
C4 = 470 p
C5 = 100 p
C6, C7 = 1n5
C8, C9 = 100 n
C10 = 0,47 µ /35 V tantalum

Semiconductors:
IC1 = LM 387

1



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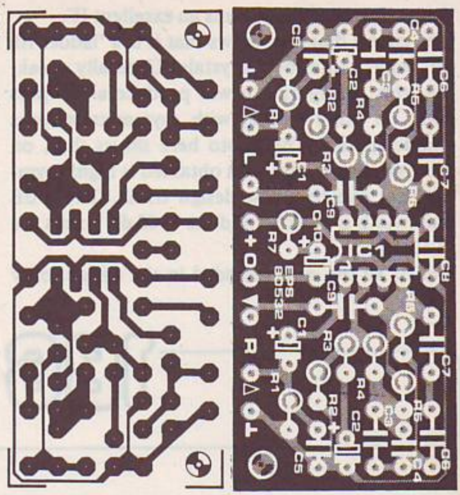
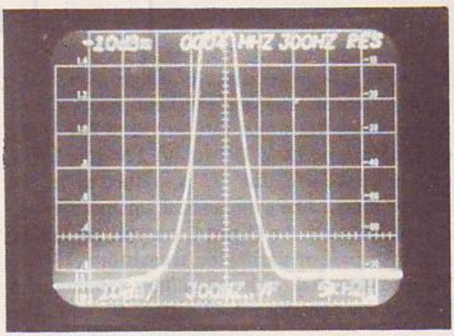
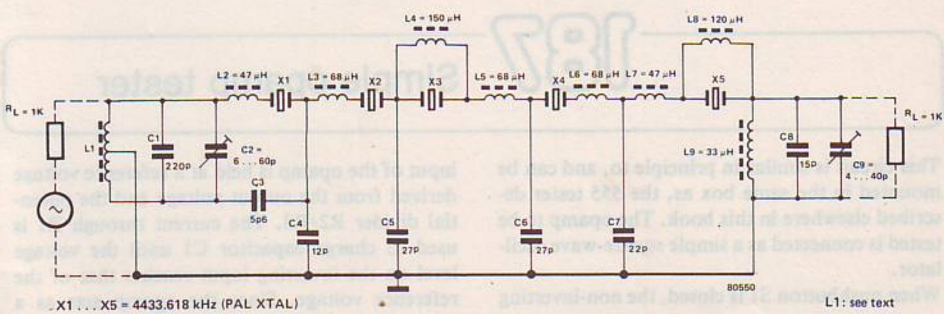


ponents (C8, C10, R7) a stereo dynamic pre-amplifier can be constructed. Because of the small size of the IC and the few components required, a highly compact printed circuit board can be made, for which room can be found in the majority of pre/control amplifiers.

185 4.4 MHz crystal filter

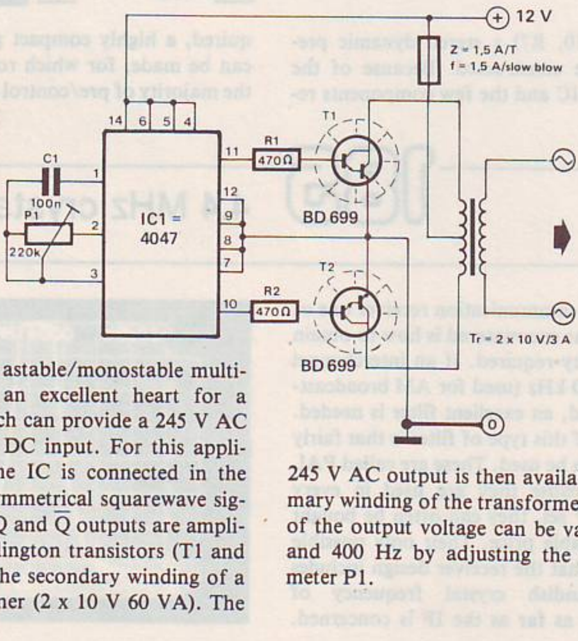
When building a communication receiver one of the main problems encountered is how to obtain the high selectivity required. If an interchannel spacing of 9 to 10 kHz (used for AM broadcasting) is maintained, an excellent filter is needed. One advantage of this type of filter is that fairly cheap crystals can be used. These are called PAL crystals and, because they are used in every (PAL) colour TV set, they can often be bought at a very reasonable price. Their only possible disadvantage is that the receiver design includes a rather outlandish crystal frequency of 4.433 618 MHz, as far as the IF is concerned.





Often, however, this forms an excellent IF.
The circuit diagram shows that it is a 'ladder filter' with a total of 5 crystals. Generally speaking, this configuration will produce a low-pass filter, in other words, with asymmetrical pass characteristics. The photo here shows that, on the contrary, the by-pass obtained is highly symmetrical due to a few design tricks. The 6 dB bandwidth is 5.2 kHz and the −60 dB points are at 12.4 kHz.
Out of all the coils included in the design, only L1 needs to be wound by hand. It consists of 15 turns of 0.4 mm copper enamelled wire double wound on an AMIDON T50-2 ring core. Time saved by not having to make L2...L9 which can be bought ready-made, can be used to carefully build the filter. This is always worth while where HF circuits are concerned. In this particular case, for instance, it is highly important to separate the various filter sections by means of metal partitions. It is also wise to connect all the cases of the crystals to earth.

186 Converter

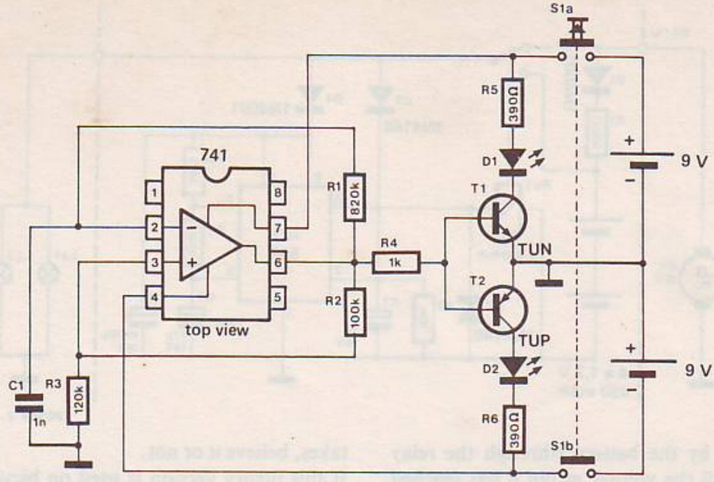


The 4047 low power astable/monostable multivibrator constitutes an excellent heart for a simple converter which can provide a 245 V AC output from a 12 V DC input. For this application, of course, the IC is connected in the astable mode. The symmetrical squarewave signals available at the Q and $\overline{Q}$ outputs are amplified by a pair of darlington transistors (T1 and T2) and then fed to the secondary winding of a low voltage transformer (2 x 10 V 60 VA). The 245 V AC output is then available from the primary winding of the transformer. The frequency of the output voltage can be varied between 50 and 400 Hz by adjusting the preset potentiometer P1.

M. Cafaxe

187 Simple opamp tester

This circuit is similar in principle to, and can be mounted in the same box as, the 555 tester described elsewhere in this book. The opamp to be tested is connected as a simple square-wave oscillator.
When pushbutton S1 is closed, the non-inverting input of the opamp is held at a reference voltage derived from the output voltage and the potential divider R2/R3. The current through R1 is used to charge capacitor C1 until the voltage level on the inverting input reaches that of the reference voltage. Since the opamp acts as a



comparator, its output level will change state thereby producing a reference voltage of opposite polarity. The charge current for C1 will then flow in the opposite direction until the new reference voltage is reached and the whole cycle will be repeated.

When the output is high, transistor T1 will conduct and LED D1 will be on. Conversely, when the output is low T2 will conduct and LED D2 will be on. The transistors are included so that other opamps with the same pin-out as, but less current output than, the 741 can be tested.

The circuit requires a posi.ive and negative power supply and will operate satisfactorily from two 9 volt batteries.

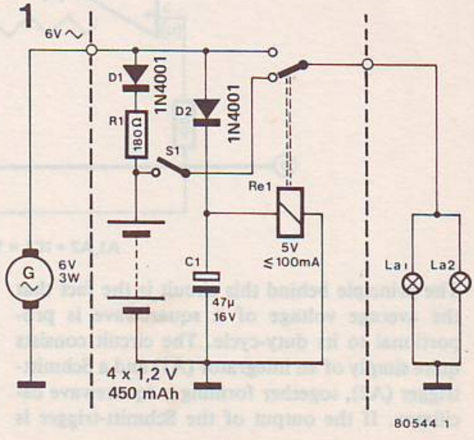
188 Automatic cycle lighting

This simple circuit (figure 1) ensures a great improvement in road safety for nocturnal cyclists. The light remains on when the cyclist stops at traffic lights – a battery supplies the current. During the trip with lights switched on (supplied by the bicycle dynamo) the battery, a parallel connection of four nicads, is charged across D1 and R1 and the relay is operated. When the bicycle stops, the relay drops out and it now connects the bulb to the battery. The one thing to remember of course, is to switch off the lights at the end of the ride, but even this can be electronicised. Figure 2 gives a suitably extended version of the circuit.

Forgetting the lights need now no longer be a problem with this luxury version, which switches the lights off automatically after about 3 minutes. The circuit is of course a little more elaborate than the standard model.

The battery is charged in the same way during the ride with the lights switched on. When the bicycle is halted at a traffic light the voltage is no longer supplied by the dynamo. The trigger input of IC1 (pin 2) then receives a negative pulse and the relay is energised. Now the lights

1



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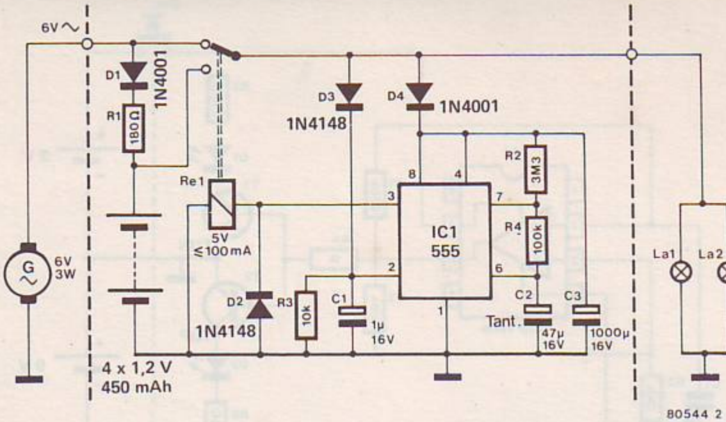
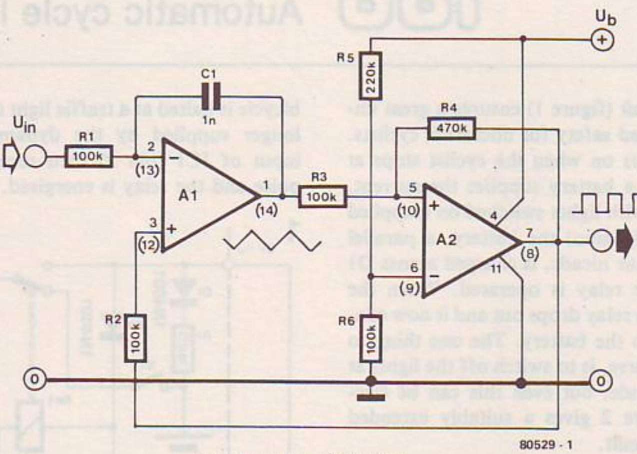


are supplied by the battery (through the relay contacts) until the voltage at pin 6 has reached the level of the internal reference voltage. Then the relay drops out again and the lights and the entire circuit are cut off from the battery. The time is preset by R2 and C2 to approximately 3 minutes. That is more time than a red light takes, believe it or not.

If this luxury version is used on bicycles with a hub dynamo and a switch in the headlamp, it may be useful to mount a switch between the dynamo and the input to the circuit. Not that it draws much power, but the relay clicking in and out could be a nuisance!

189 Voltage controlled duty-cycle



A1, A2 = IC1 = 1/2 LM 324

The principle behind this circuit is the fact that the average voltage of a squarewave is proportional to its duty-cycle. The circuit consists quite simply of an integrator (A1) and a Schmitt-trigger (A2), together forming a squarewave oscillator. If the output of the Schmitt-trigger is low the output of A1 will gradually decrease until the lower threshold of A2 is reached. The output of A2 will then go high (just less than the supply voltage) causing the integrator output to rise until the upper threshold is reached and the Schmitt-trigger output goes low again.

By altering the voltage level on the inverting input of A1 the integrator's characteristics can be altered. As the trigger thresholds of A2 are fixed, the result is a change in duty-cycle. The average voltage of the squarewave output will always be equal to the input voltage, but the frequency will remain constant. In this way the duty-cycle can be varied between 0% and 100%. The control voltage may be anywhere between 0 V and 1.5 V less than the supply voltage.

When the LM 324 is used, the supply voltage may be anywhere between 3 V and 30 V. If another type of opamp is used, the control range may become limited.

190 Exposure meter and development timer

A great deal has been published about exposure meters and development timers in electronic magazines. However, it is very rare to find an article covering both devices at the same time. For this reason a combination design is presented here. As usual an LDR (light-dependent resistor) has been included in the bridge circuit of the exposure meter. The amount of light falling on the LDR determines the degree of imbalance in the bridge network. During measurement, relay Re1 is activated via S2e and the enlarger is turned on. Balance is then restored manually by adjusting potentiometer P1. The final value of P1 will correspond to the exposure time required.

Indication that the bridge is in balance is given by two LEDs (D7 and D8). Of course, this could also be indicated on a centre-zero meter but this could be difficult to read in the dark and would probably be more expensive than two LEDs. The circuit is in balance when both LEDs are extinguished.

Once the above procedure has been carried out switch S2 is changed to its other position. The circuit will now operate as a development timer. The value of P1, together with C3, now determines the pulse duration of the monostable multivibrator IC2. The timer is started when push-button S3 is pressed. The output then goes high activating the relay and lighting the enlarger lamp. LEDs D5 and D6 are optional and are used to illuminate the panel of the case in which the circuit is mounted. If a single pole double-throw relay is used then it is possible to switch the exposure lights off when switching on the enlarger lamp.

The role of potentiometer P2 has not yet been discussed. It enables the characteristics of the bridge amplifier to be altered to suit different kinds of paper, and should be provided with a suitable scale. The usefulness of the circuit will depend on how well it is calibrated.

Readers who wish to carry out spot measurements can simply mount the LDR into a cardboard tube. This is then placed inside a metal film can and covered by a piece of perspex which has been slightly sanded with emery paper. For details see figure 2. When measuring the exposure time the perspex can be removed and a larger piece placed directly under the enlarger lamp. No details of the negative will be visible when the enlarger is switched on. The perspex is, of course, removed during exposure.

The circuit is calibrated by making test strips. Potentiometer P2 is given a linear scale division numbered 1...20. With S2 in the 'time' position,

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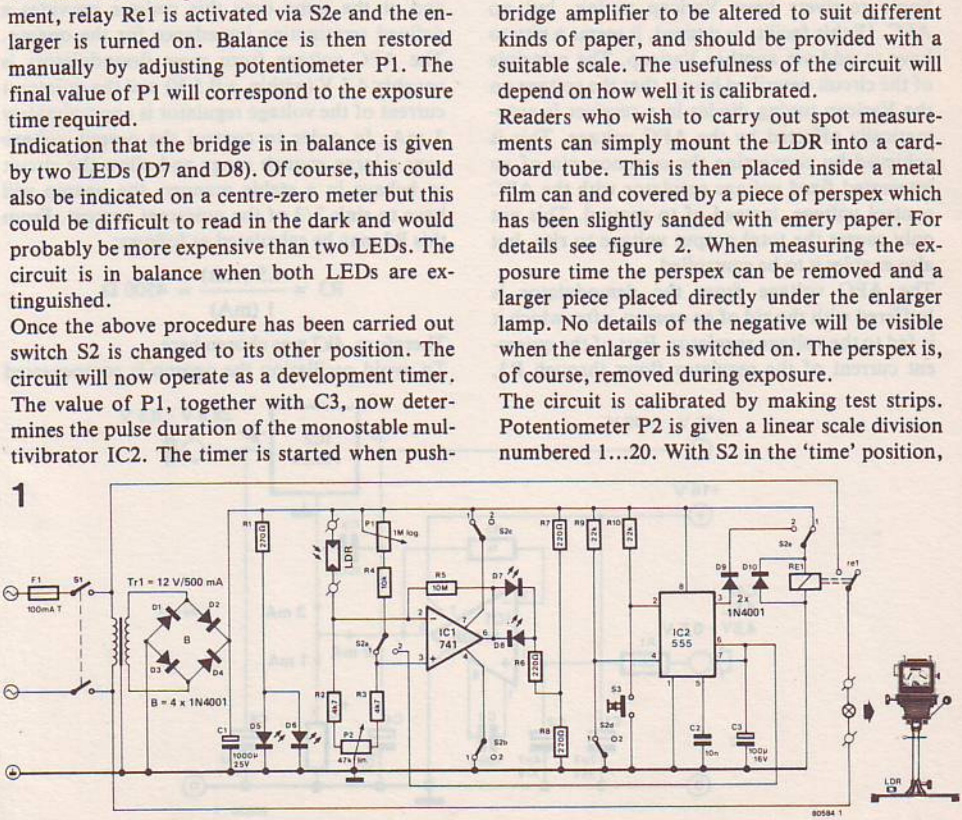

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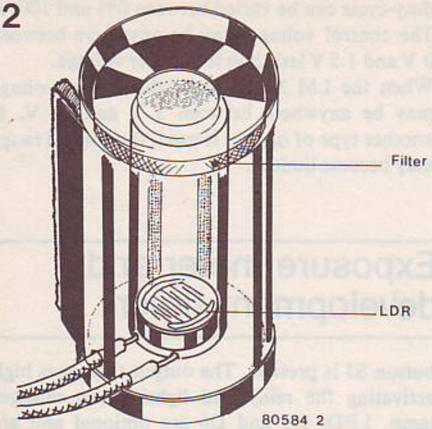


P1 is adjusted until a set of whole numbers is obtained which correspond to the length of time the lamp is on. This will form the scale for P1. Once this has been done a test strip is made using the old method.
The same exposure time is then selected with P1. Test strips are made with P2 in various positions until the best results are obtained. The positions of P1 and P2 are then written on the test strips to provide the correct coding for all types of paper.

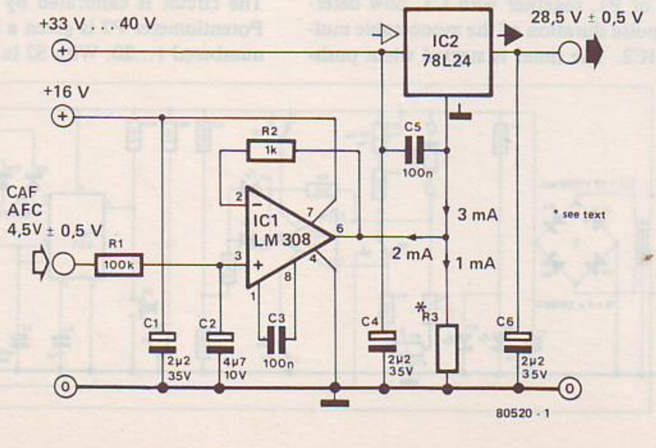
D.S. Barrett

191 AFC via the tuning diode

Some receivers have Varicap tuning, but no AFC. IF this facility is desired, it seems a pity to have to add yet another Varicap. The principle of the circuit described here is that the voltage on the Varicap tuning diodes in a receiver is automatically affected by the AFC voltage. This is achieved by connecting the common pin of an integrated fixed voltage regulator with the AFC control voltage, instead of to ground. This not only causes the total output voltage to rise, but also enables it to be controlled.
The AFC voltage from the demodulator is buffered with the aid of an opamp, after which it is fed to the voltage regulator. Part of the quiescent current of the regulator flows through R3, and at the same time this resistor provides a defined terminating impedance for the opamp. The AFC voltage from most demodulators is roughly 4.5 V (within ± 0.5 V) and the quiescent current of the voltage regulator is approximately 3 mA. In order to control the output voltage over a large enough range and allow the circuit to behave in a stable manner, the opamp will have to sink 2/3 of the quiescent current. From this R3 may be calculated as follows:

$$R3 = \frac{4.5\ (\text{volt})}{1\ (\text{mA})} = 4500\ \Omega$$

Therefore, 4k7 was chosen here.
To avoid oscillation the opamp is compensated



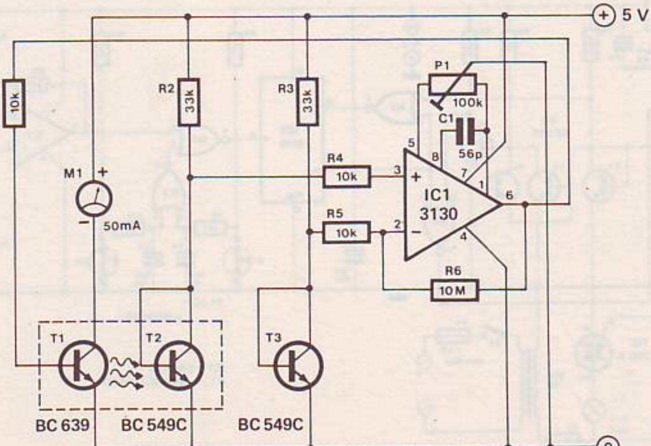
by C3 and the voltage regulator is decoupled with C5. The LM 308 type was chosen as a buffer (IC1) because of its low input current (only 3 nA) and its very low drift.
The circuit's current consumption is approximately 300 μ A.
The AFC voltage is fed to the input by means of a low-pass filter (R1 and C2) which causes any interfering signals to be thoroughly suppressed. It also ensures quiet, stable AFC control. To switch off the AFC the input voltage of the circuit is preset to the average value of the AFC voltage.

S. Hering

192 Wind-o-meter

This device utilises the fact that an air current has a cooling effect on an object which is warmer than its surroundings. The object cooled in this case is a transistor (T2) which is connected as a diode. To make it warmer than its surroundings it has been thermally coupled to a transistor (T1) which has a current flowing through it continuously. The wind's speed is measured by comparing the voltage across the cooling diode with that across a reference diode (T3). These two voltages are fed to the non-inverting and inverting inputs respectively of an opamp.
This amplifier, which is preset for a gain of 1000, passes a current through the heating transistor via resistor R1. When the wind cools the diode, the forward voltage across that diode rises (2 mV/°C) causing the voltage at the non-inverting input of the opamp to increase. As a result, the output voltage of the opamp rises to provide more base drive current for T1 thereby generating more heat in this transistor. The opamp thus tries to compensate for the temperature drop, which leads to an increase in T1's collector current.

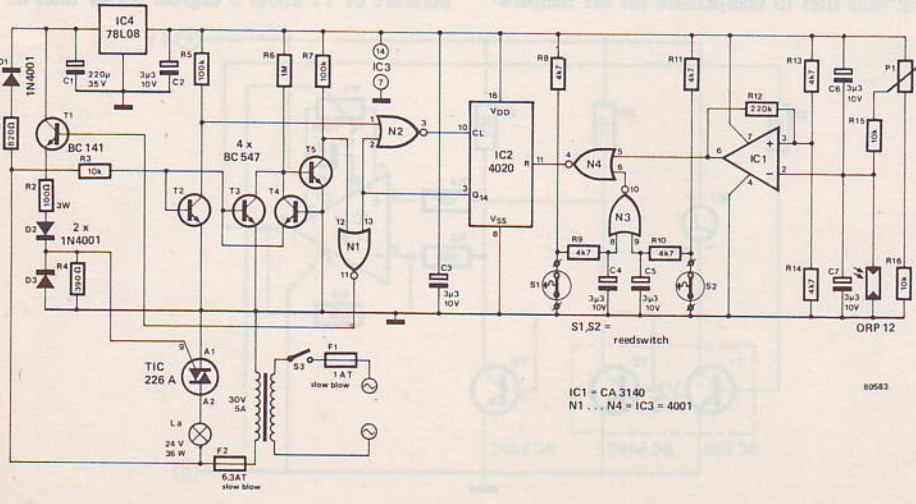
A high sensitivity is obtained by making the temperature of T2 about 5 degrees higher than its



surroundings. This is achieved by presetting the meter to give an offset of about 5 mA when there is no wind blowing. Resistor R1 is selected so that the current through T1 is not excessive.
In the circuit, T1 is shown as a BC 639, but a BC 547 type may also be used: the maximum collector current must then be limited to 100 mA. If the circuit tends to oscillate, the gain of IC1 should be reduced – by increasing the value of R5.
The photograph shows the construction of the wind detector. The two transistors are coupled by glueing their flat sides together with a heat conducting adhesive.

193 Garden path lighting

Plant a new bulb in your garden! This circuit will lead you up the garden path at night in safety. It enables the path to be illuminated whenever required, and it consumes very little energy. The lamp is switched on by reed switches, mounted in the front door and in the garden gate. By using a 24 V lorry bulb, electrical installation remains simple and safe while ensuring sufficient light.
The circuit is powered from the secondary of a 24...30 V/5 A transformer. Where long leads through the garden are a necessity the higher rating is preferable due to the voltage drop caused by the resistance of the leads.
The 50 Hz signal from the transformer is fed to the base of T2 which converts it into a squarewave. This is then gated by N2 to provide the clock signal for the counter IC2, for as long as Q14 remains low. As soon as Q14 goes high the clock signal is blocked. Transistors T3...T5 form a zero-crossing detector which is also controlled by the 50 Hz signal. Each time the transformer voltage crosses zero the collector of T5 is pulled low for 100 s. This pulse arrives at the base of T1 via the buffer N1. This transistor is used to control the triac which in turn switches the lamp on at every zero-crossing. The light will, of course, appear to be on continuously for as long as Q14 remains low. The path will be illuminated for nearly three minutes – long enough for someone to walk up the average path and open the front door.
The circuit is activated when the reset input of the counter is taken high. For this to happen, both inputs of N4 must be low. One of these inputs to N4 is controlled by an opamp whose output is determined by the amount of light falling on an LDR (light-dependent resistor). A certain amount of hysteresis has been incorporated in this part of the circuit. As long as there is sufficient daylight the output of IC1 will be high. The reset input of the counter will remain low thereby inhibiting the system. As darkness draws in the output of IC1 will go low (the actual level



can be preset by means of P1) enabling one of the inputs to N4. The other input to N4 is taken low when one of the two reed switches (S1 or S2) opens and closes, when the garden gate or front door is opened and closed. This fulfils the condition that both inputs of N4 must be low and the inhibit is removed from the counter enabling operation (and illumination).

Thin twin-lead cable can be used for the reed switches, but thicker wire (about 2.5 mm²) is required for the lamp. Current consumption is approximately 100...150 A when the lamp is not on.

B.E. Kerley

194 Hybrid cascode

It is general knowledge that cascoding two or more transistors creates a new transistor with better characteristics than the individual ones (see figure 1). These include a very slight retroaction from point C ('collector') to point B ('base') and a higher collector impedance, thus a much better approach to current source operation at point C.

In the all transistor version of figure 1, the base of T2 will have to be fed a certain voltage with respect to the emitter of T1 −0.6 volts (D1 in figure 1) at least.

If T2 is replaced by an N channel FET, the DC bias of the cascode will be a lot easier to preset – see figure 2. As far as slope is concerned (i.e. the ratio between collector current and base voltage) both versions are equally good.

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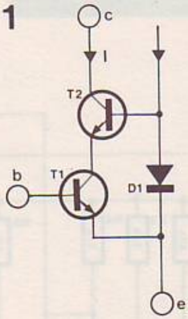

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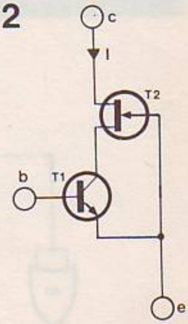
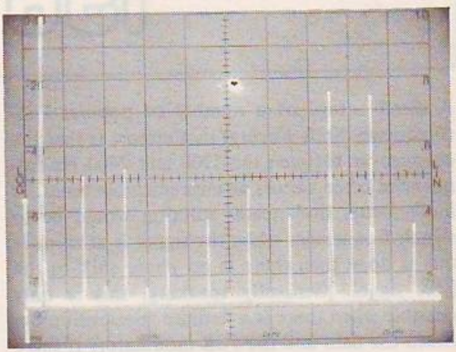


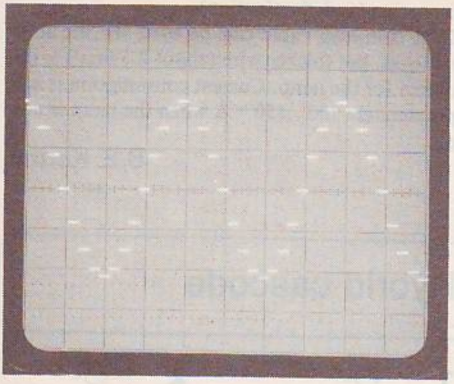
195 Digital sinewave generator

More and more information is being generated by digital means, because of the good frequency and amplitude stability obtained. This particular circuit generates a sinewave, but if R1...R8 were to assume other values, different waveforms would also be possible.

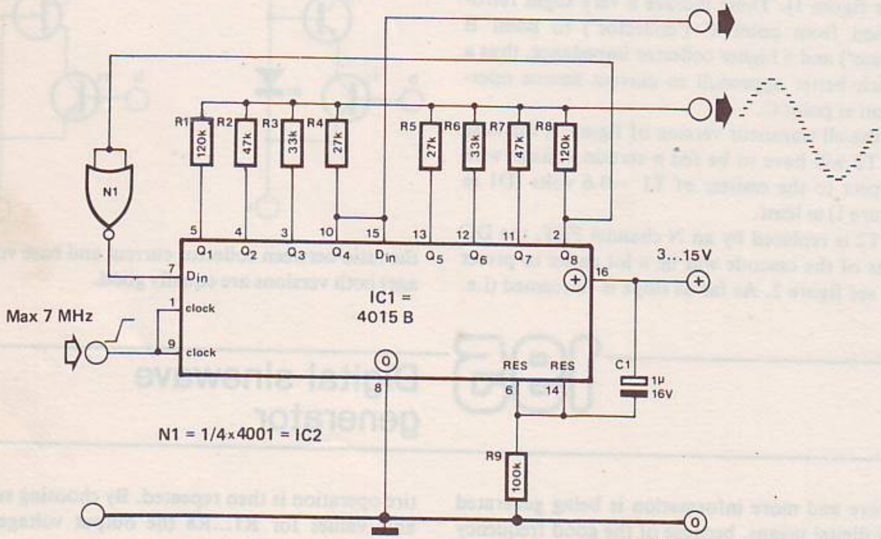
After the supply voltage has been applied, the R9/C1 network ensures a short reset pulse: all outputs become logic zero. Since output 8 is also '0', the inverted level ('1') is offered at input D. With the aid of an external oscillator (not drawn) pulses are fed to the clock inputs. At every positive slope the information in the shift register IC1 moves one place further along. Thus, after the first clock pulse Q1 will be '1' and after the eighth Q8 will also be '1'. As soon as Q8 becomes logic 1 however, the information at input D will change to logic 0. Then zero's will be entered until Q8 also changes to '0'. The entire operation is then repeated. By choosing suitable values for R1...R8 the output voltage is converted into a sinewave.

The output frequency is one sixteenth of the clock frequency. The CMOS IC can process up





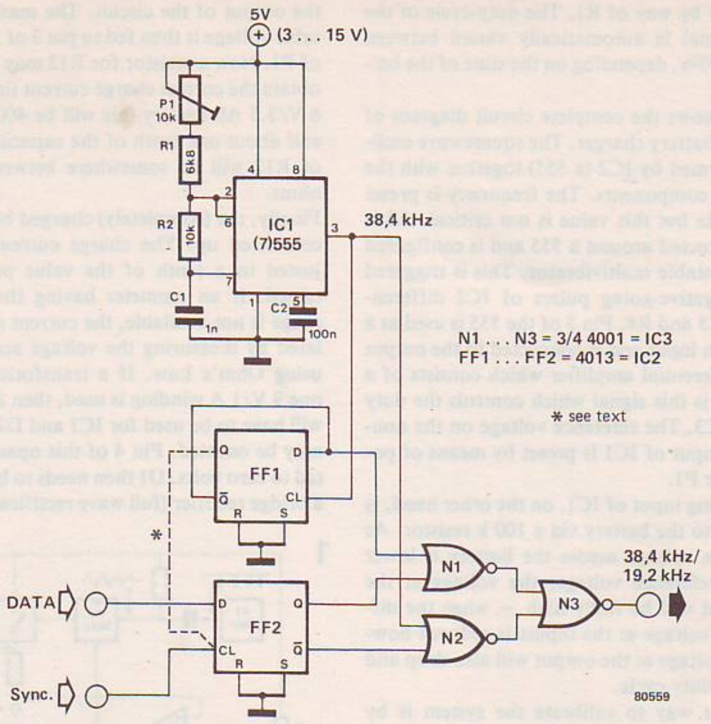
to 7 MHz so that the maximum output frequency is about 0.5 MHz. Gate N1 may be of any type provided that the signal is inverted.
The two photo's show the waveform and the frequency spectrum respectively. The most important harmonics, the third and the fifth, are almost 50 dB below the output level. Although the fifteenth and seventeenth harmonics are much larger, they can be nullified by a simple RC filter because they are further away from the main frequency.
A circuit using a 555 timer as an oscillator may be used for the clock pulses – see the synchronous FSK modulator described below. The sync output provides a squarewave, with the same frequency and phase as the sinewave, which can be used, for instance, to trigger an oscilloscope.



196 Synchronous FSK

A disadvantage of a number of popular FSK (Frequency Shift Keying) modulators is that the changeover between the 1200 and 2400 Hz frequencies often occurs at unpredictable times. A much better and neater solution would be to switch frequencies when the signal is at zero. If this is done then there is no phase shift in the FSK signal. Generally speaking, this is only possible when there is a definite relationship between the data and the FSK modulator. If there isn't then the circuit given here should be of some help.

The actual FSK signal is obtained by means of the digital sinewave generator described previously (see above). At every zero-crossing the sinewave generator produces sync pulses which are used to clock FF2. The data level present at the 'D' input of this flip-flop determines which of the two output frequencies are selected. When a '1' is present at the input, the output frequency



will be 38.4 kHz, and when a '0' is present the output frequency will be 19.2 kHz. This output signal is then divided by 16 by the digital sinewave generator to produce the correct FSK frequencies. The oscillator for the circuit is formed by the well known 555, albeit a CMOS version (that's why it has the slightly different 7555 type number). This IC's characteristics are practically the same as those of the 'ordinary' 555 with the added advantages of a much higher input impedance, less current consumption and its almost total 'spike' suppression during output level switching.

If the circuit is not used in combination with the digital sinewave generator, the sync input must be connected to the Q of FF1.

197 PWM battery charger

This circuit is designed to charge 6 V/3.5 Ah batteries similar to those often used in flash equipment. Of course, there are all kinds of methods to charge lead-acid batteries but what is special about this version is the fact that the charge current is continually corrected according to the state of the battery.

Figure 1 shows the block diagram of the PWM battery charger (PWM stands for Pulse Width Modulation). A1 is a squarewave oscillator which generates a frequency of around 2 kHz. A2 is a monostable multivibrator which is triggered by the negative-going pulses from A1. The pulse width from A2 depends on the control voltage which is derived from the differential amplifier A3. The latter constantly monitors the battery voltage. The output of A3 varies according to the difference in voltage between the preset reference level and the tested battery level. When both are equal the output voltage of A3 will be such that the duty-cycle of A2 will be 10%. This is enough to maintain trickle charge for the battery. The output of A2 controls the electronic switch ES1, so that current is fed to the battery by way of R1. The duty-cycle of the output signal is automatically varied between 10% and 90%, depending on the state of the battery.

Figure 2 shows the complete circuit diagram of the PWM battery charger. The squarewave oscillator is formed by IC2 (a 555) together with the associated components. The frequency is preset at 2.27 kHz but this value is not critical. A2 is also constructed around a 555 and is configured as a monostable multivibrator. This is triggered by the negative-going pulses of IC2 differentiated by C5 and R8. Pin 5 of the 555 is used as a modulation input and is connected to the output of the differential amplifier which consists of a 741 IC. It is this signal which controls the duty cycle of IC3. The reference voltage on the non-inverting input of IC1 is preset by means of potentiometer P1.

The inverting input of IC1, on the other hand, is connected to the battery via a 100 k resistor. As long as the voltage across the battery is lower than the reference voltage, the voltage at the IC's output will be fairly high – when the difference in voltage at the inputs is reduced however, the voltage at the output will also drop and so will the duty-cycle.

The easiest way to calibrate the system is by using a discharged battery (about 2 V per cell) and a fully charged battery (about 2.4 V per cell). The discharged battery is first connected to the output of the circuit. The maximum presettable voltage is then fed to pin 3 of IC1 by means of P1. Now a resistor for R12 may be selected to obtain the correct charge current (in the case of a 6 V/3.5 Ah battery this will be 400 mA; in general about one tenth of the capacity). The value of R12 will be somewhere between 2.5 and 5 ohms.

Finally, the (completely) charged battery may be connected up. The charge current is now adjusted to a tenth of the value previously obtained. If an ammeter having the correct test range is not available, the current may be calculated by measuring the voltage across R12 and using Ohm's Law. If a transformer with only one 9 V/1 A winding is used, then a 3140 opamp will have to be used for IC1 and D2, D3, and C2 may be omitted. Pin 4 of this opamp is connected to zero volts. D1 then needs to be replaced by a bridge rectifier (full wave rectification).

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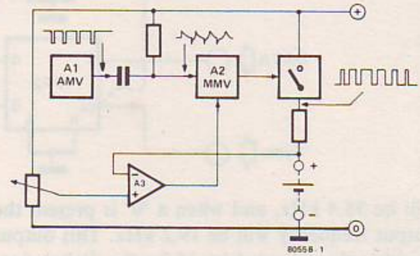

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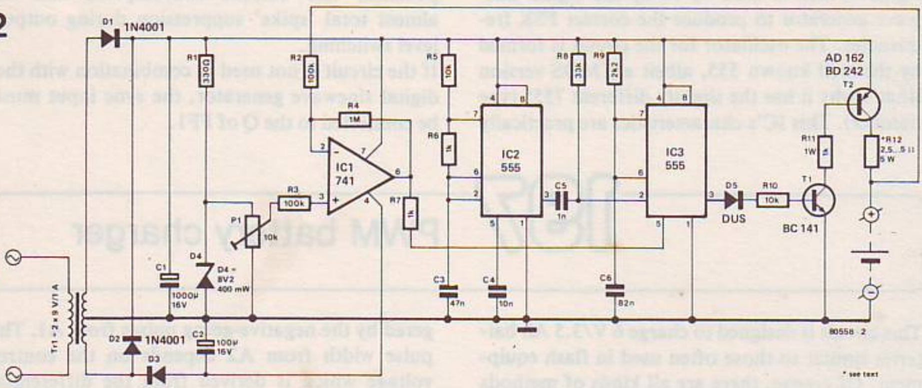
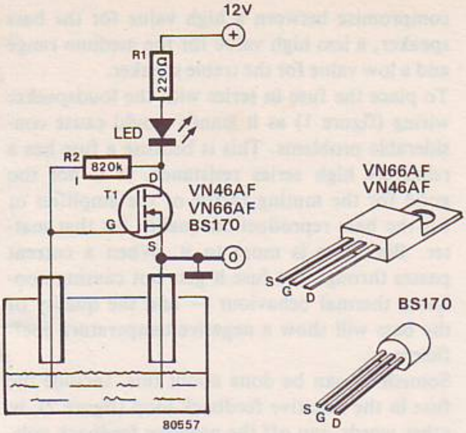


198 Water detector

The level of water in a water tank can be measured in various ways, some of which can of course be more complicated than others. The circuit published here lights a LED whenever the



level of water drops below the electrodes. With a high water level, the FET hardly conducts or not at all, because the gate is then connected to earth and there is no voltage difference between gate and source. When the water level drops, the gate/source connection is interrupted. The gate is then at a positive potential by means of the 820 k resistor, thereby causing the FET to conduct. The LED will now light. If the reverse operation is needed, that is the LED lights when the electrodes are short circuited by the water, just connect the 'earthy' electrode in the circuit to the positive and wire R2 between gate and source.

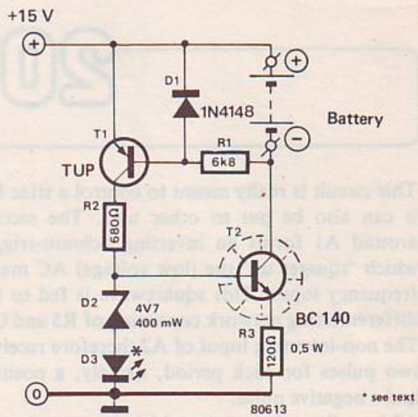
ITT Applications

199 Intelligent NICAD charger

Like all things, NICAD chargers are subject to human error, that is, the batteries can be placed in the holder in two ways, correctly and incorrectly. This charger will refuse to operate unless the cells are fitted correctly. The charger consists of a current source (T2) to maintain the output current at about 50 mA. The zener diode D2 and LED hold the base drive of T2 at a constant level and thus also the voltage across R3. The current through R3 is therefore also constant, providing the correct conditions for charging NICADs at the collector of T2.

The protection circuit includes T1, D1 and R1. The terminal voltage of an incorrectly fitted NICAD will turn off T1 preventing the charger from operating. An indication of this will be given by the LED – it will not light. When the battery is fitted correctly T1 will turn on and the charger will function normally.

Although the charger is capable of charging up to four penlight cells, it will not detect a single cell being the wrong way round if two or three



others are connected correctly at the same time. A small transformer, bridge rectifier and electrolytic capacitor are all that is required for a power supply. The circuit works well providing the NICADs are not completely discharged.

200 Loudspeaker fuse

A loudspeaker protection circuit can be used to protect multipath loudspeaker systems effectively from users with destructive tendencies. It can be done simply by using an old-fashioned glass fuse in series with the loudspeaker wiring. The melting value (in A) of the fuse is based on a

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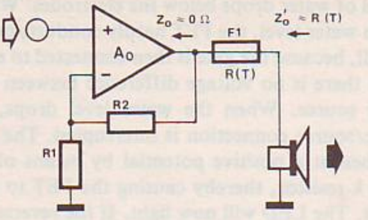

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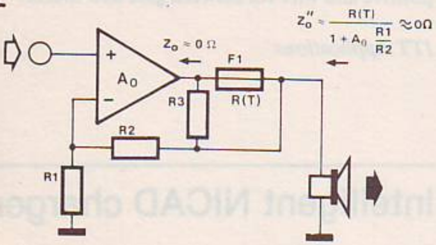


80555 - 2

compromise between a high value for the bass speaker, a less high value for the medium range and a low value for the treble speaker.

To place the fuse in series with the loudspeaker wiring (figure 1) as it stands would cause considerable problems. This is because a fuse has a relatively high series resistance. It is not too good for the muting factor of the amplifier or for the bass reproduction quality for that matter. But there is more to it. When a current passes through the fuse it gets hot causing non-linear thermal behaviour – and the quality of the bass will show a negative temperature coefficient.

Something can be done about this. Include the fuse in the negative feedback loop (figure 2), in other words, tap off the negative feedback voltage from a point behind the fuse. The fuse is bypassed by means of resistor R3 which is small compared to R2 (slight influence on the DC set-up of the amplifier), but large compared to the 4 Ω or 8 Ω load impedance. A value of 220 Ω (1 W) for R3 is fine.

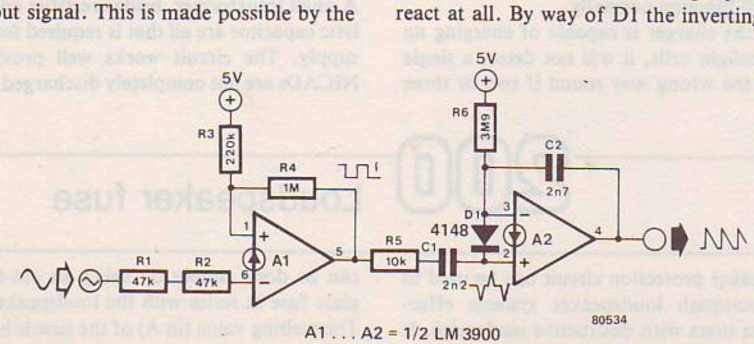
201 Sawtooth synchronous to mains

This circuit is really meant to control a triac but it can also be put to other uses. The section around A1 forms an inverting Schmitt-trigger which 'squares up' the (low voltage) AC mains frequency input. This squarewave is fed to the differentiating network consisting of R5 and C1. The non-inverting input of A2 therefore receives two pulses for each period, namely, a positive and a negative pulse.

A2 is really an integrator which converts the signal into a 'sawtooth' waveform. The mixer reacts to the positive as well as the negative slope of the input signal. This is made possible by the unusual internal architecture of the LM3900.

The opamp reacts normally to positive pulses. As soon as the non-inverting input becomes 'high', the inverting input will also have to become 'high' to maintain the balance. This can only happen if the output voltage is made to rise. The rise in voltage is then passed on through C2.

In order to be able to understand what happens to negative input pulses, it is important to realise that the input circuit of this opamp consists of a transistor whose emitter is connected to earth. For this reason the non-inverting input does not react at all. By way of D1 the inverting input is

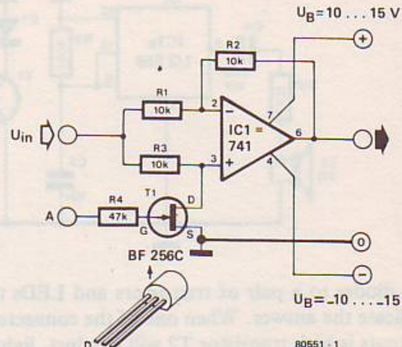


A1 . . . A2 = 1/2 LM 3900

also cut off. When this condition exists the output voltage starts to rise as high as the supply voltage.
Resistors R1 and R2 have been included to limit the input voltage. If the circuit is not connected directly to the mains, it is better to use a single resistor of 100 k. The supply voltage is not critical and may be anything between 4 and 36 volts.

202 Plus or minus unity gain

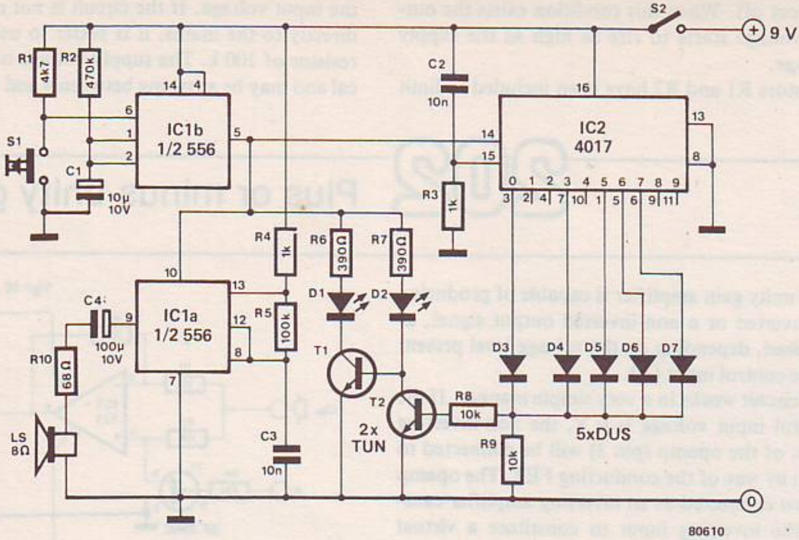
This unity gain amplifier is capable of producing an inverted or a non-inverted output signal, as required, depending on the voltage level present at the control input (A).
The circuit works in a very simple manner. If the control input voltage is 0 V, the non-inverting input of the opamp (pin 3) will be connected to earth by way of the conducting FET. The opamp is now connected as an inverting amplifier causing the inverting input to constitute a virtual earth point (the opamp maintains the voltage level at pin 2 equal to that at pin 3, in this instance, ground). With the given values for R1 and R2 the gain will be -1.
If the control input (A) is connected to $-U_b$ the FET will turn off and will form a high impedance load for the rest of the circuit. Now the output of the opamp will *not* be inverted but the gain will remain the same. The input level must remain within 2 volts of the supply voltage extremes, (thus, $V_b + 2\text{ V} \leq V_{in} \leq + V_b - 2\text{ V}$).



The impedance of the signal source should be as small as possible, since the input impedance depends on whether the FET is conducting or not. A source impedance of 500 Ω is recommended.
The circuit may be used as an automatic polarity inverter for meters etc.

203 Enigma

Less energetic than musical chairs or charades, the use of this circuit can evoke just as much, if not more, humour from populous parties. The idea behind the game is to ask certain questions that can only have a 'yes' or 'no' answer and to produce answers which are hilariously incorrect. For instance, when the vicar is asked if it is true that the drinks fifteen pints of lager before breakfast the answer should be 'yes'. And 'no' must be elicited from the farmer's wife when asked to deny the rumour that she sleeps in the pig-sty. No doubt readers will be able to think of similar questions to put to their particular set of guests, but we cannot print the more obvious ones that spring to mind.
The circuit shown here produces a set series of 'yes' and 'no' answers by lighting one of two LEDs. This set sequence is, of course, only known to the questioner. Once a question has been asked the questionee presses the push-button S1. This triggers the timer contained in one half of IC1 to produce a pulse of around 4 to 5 seconds. The duration of this pulse can be varied by altering the values of R2 and C1 if desired. When the output of this timer goes high it clocks the divide-by-ten counter IC2 and removes the inhibit from the oscillator configured around the second timer IC1b.
This oscillator will produce a tone from the loudspeaker to indicate to all present that the button was in fact pressed.
Certain of the outputs from IC2 are connected



via diodes to a pair of transistors and LEDs to indicate the answer. When one of the connected outputs is high transistor T2 will conduct, lighting LED D2 and turning off T1 thereby indicating a 'yes' answer. The LED will stay lit for the duration of the pulse from IC1a. When one of the unconnected outputs of IC2 is high T2 will turn off turning on T1 and D1 to indicate a 'no' answer.
The sequence of answers for the circuit shown is therefore: 'yes', 'no', 'no', 'yes', 'no', 'yes', 'yes', 'yes', 'no', 'no'. Any other sequence of ten answers can be selected by connecting fewer or more diodes as required. Once the sequence of ten is complete the cycle will be repeated.
Note: Capacitor C2 and resistor R3 are included to reset the counter on initial switch-on. This means that the first sequence will consist of *eleven* answers and the first answer will be 'no' for the circuit as shown – although it doesn't count, since no question has been asked yet!

204 Universal warning alarm

A device to attract one's attention when a certain condition is not being fulfilled would have a wide range of applications. The circuit presented here is versatile enough to provide an alarm for any application whether requiring immediate attention, or just reminding one that a certain function is not being properly accomplished.
The simplicity of the main circuit is immediately apparent (figure 1), consisting of only two CMOS squarewave oscillators and an output buffer. The circuit operates as follows: N1 and N2 form one of the CMOS oscillators. This oscillator is used to pulse a second oscillator (N5 and N6), which is set at an audio frequency. The mark space ratio of the first oscillator can be varied by means of P1 and P2. It can then be seen that P1 determines the length of time that the second oscillator is enabled. Conversely, P2 determines the time that the second oscillator is *disabled*. The frequency of the second oscillator can be varied by means of P3 over the range of 40 Hz...15 kHz.
The output of the circuit will be a pulsed tone of the frequency determined by P3, the 'on' time determined by P1, and the 'off' time determined by P2. The circuit can be enabled in two ways:
1) A logic 0 on the A input
2) A logic 1 on the B input (provided N4 is omitted!)
Thus the circuit can be triggered by either logic state.
The output of N6 is fed to two inverters; N7 and

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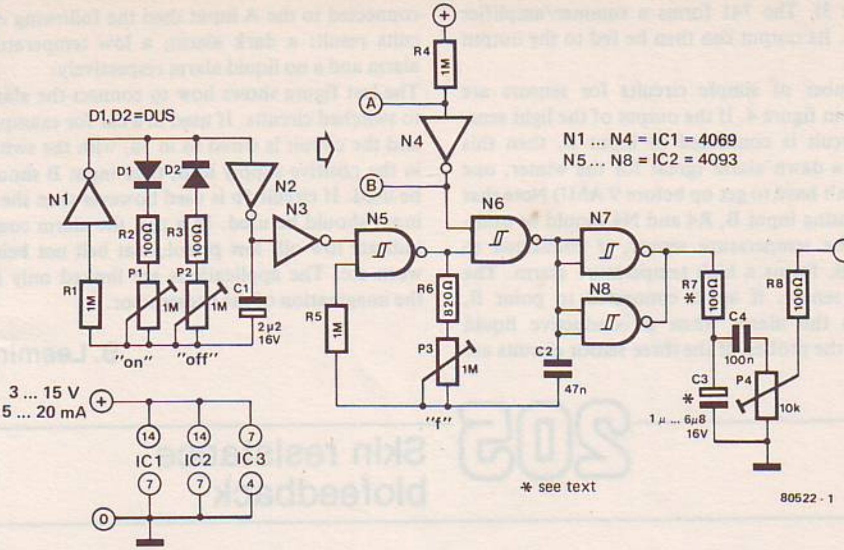

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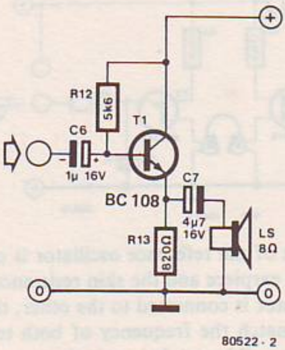

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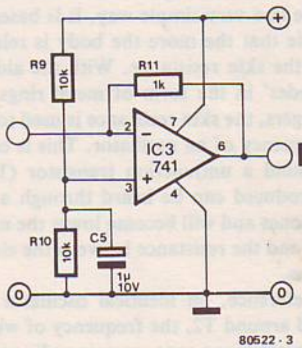

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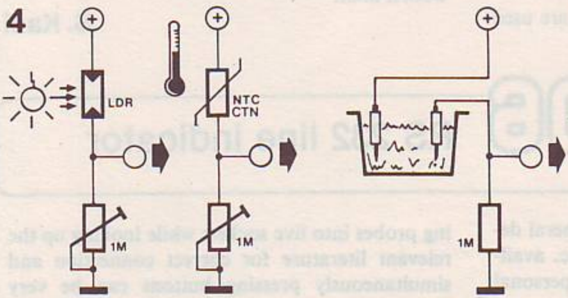

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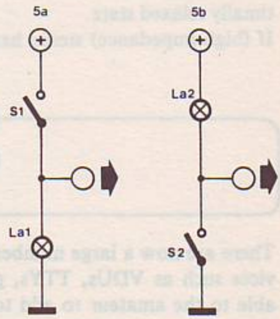


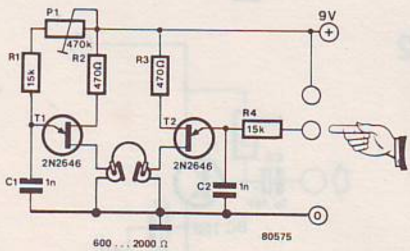
N8. These feed a low pass filter formed by R7 and C3. These reduce high frequency harmonics and give the output a more pleasing sound. Potentiometer P4 is used as a volume control, and therefore the output is taken from the wiper. If the circuit is to be used on its own, then the output can be fed directly to the positive side of C6 and to the output driver formed around T1 (figure 2). If more than one of these circuits is to be connected together, then all the 'out' terminals of the alarm circuits can be commoned and fed to the non-inverting input of IC3 (741)

(figure 3). The 741 forms a summer/amplifier circuit. Its output can then be fed to the output driver.
A number of simple circuits for sensors are shown in figure 4. If the output of the light sensing circuit is connected to input B, then this forms a dawn alarm (great for the winter, one wouldn't have to get up before 9 AM!) Note that when using input B, R4 and N4 should be omitted. The temperature sensor, if connected to point B, forms a high temperature alarm. The liquid sensor, if again connected to point B, sounds the alarm when a conductive liquid covers the probes. If the three sensor circuits are connected to the A input then the following circuits result: a dark alarm, a low temperature alarm and a no liquid alarm respectively.
The last figure shows how to connect the alarm to switched circuits. If used in a car for example, and the circuit is wired as in 5a, with the switch in the positive supply lead, then input B should be used. If circuit 5b is used however then the A input should be used. In a car, the alarm could indicate low oil, low petrol, seat belt not being worn etc. The applications are limited only by the imagination of the constructor.

B. Leeming

205 Skin resistance biofeedback

This device makes a special form of biofeedback possible in a very simple way. It is based on the principle that the more the body is relaxed the higher the skin resistance. With the aid of two 'electrodes' in the form of metal rings around two fingers, the skin resistance is used to control the frequency of an oscillator. This is constructed around a unijunction transistor (Ta). The tone produced can be heard through a pair of headphones and will become lower the more one relaxes and the resistance between the electrodes increases.
As a reference, an identical oscillator is constructed around T2, the frequency of which can be set to produce a tone, corresponding to an optimally relaxed state.
If (high impedance) stereo headphones are used

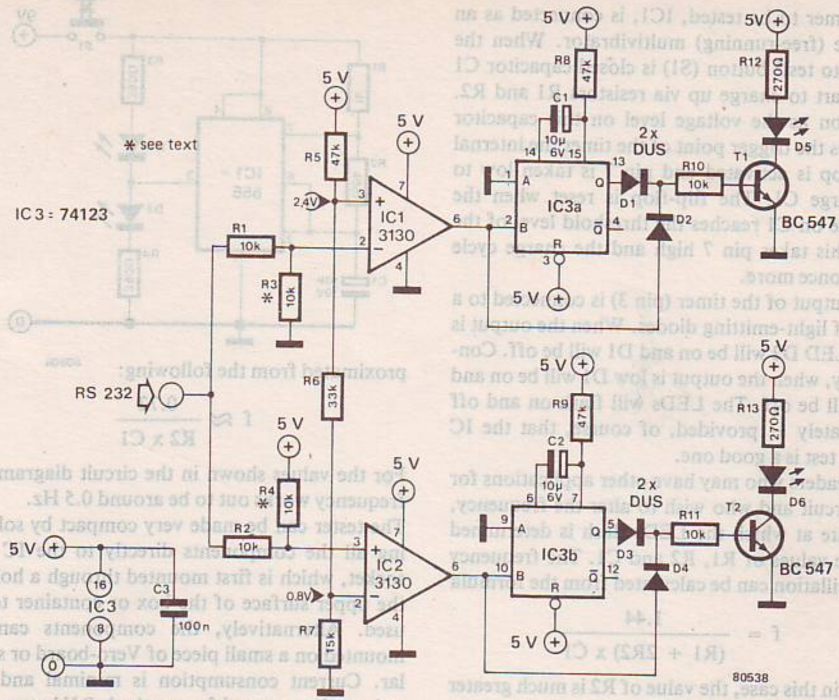


the output of the reference oscillator is connected to one earpiece and the skin resistance sensitive oscillator is connected to the other, the idea being to match the frequency of both tones as closely as possible merely by relaxing.
Cool it man!

S. Kaul

206 RS 232 line indicator

There are now a large number of peripheral devices such as VDUs, TTYs, printers etc. available to the amateur to add to his/her personal computer system. When things fail to operate correctly one of the first checks to be carried out is to make sure that the correct voltage levels are present on the RS 232 line interconnections. As many personal computer operators will testify, this is not a particularly easy task when the only test instrument available is a multimeter. Inserting probes into live sockets while looking up the relevant literature for correct connection and simultaneously pressing buttons can be very frustrating – especially when power supply lines are shorted out accidentally! It would be desirable, therefore, to have a device that will check whether or not the signals are present and at the correct levels. The accompanying circuit is just such a device and can be built into the computer to provide a constant monitor of the RS 232 line.



The circuit is very simple and consists mainly of two comparators and two pulse stretchers. Resistors R5...R7 form a potential divider chain to set the voltages on the non-inverting input of IC1 and the inverting input of IC2 to 2.4 and 0.8 volts respectively. The RS 232 input signal is attenuated by resistors R1...R4 and passed to the inverting input of IC1 and the non-inverting input of IC2. If the circuit is to be used with systems operating on TTL levels only then resistors R3 and R4 can be omitted. The output of IC1 will go high when there is a voltage level greater than 2.4 volts present at the input. Similarly, the output of IC2 will go high when the input level is less than 0.8 volts.

The outputs from the two comparators are fed to a pair of retriggerable monostable multivibrators contained in IC3 (the pulse stretchers). These provide a pulse with a fixed duration to turn on transistors T1 and T2 and light the respective LEDs. These pulse stretchers are included so that no matter how short the input pulse is it will still be seen by the human eye. The pulse duration is determined by the values of R8/C1 and R9/C2.

If the input signal is relatively long then the associated bypass diode takes over to ensure that the LED is lit for the length of the input pulse. In the circuit diagram, the upper LED (D5) indicates the presence of a positive pulse while the lower one (D6) indicates the presence of a negative pulse. Capacitor C3 is included to ensure correct power supply decoupling for IC3.

207 Simple 555 tester

The versatile 555 timer IC has the habit of turning up in wide variety of circuits. As it is such a useful little device it has become very popular over recent years. Although the 555 is generally very reliable, there are occasions when malfunction does occur. The circuit shown here will provide a simple and effective method of testing suspect devices.

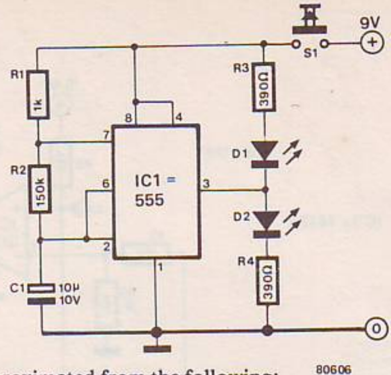
The timer to be tested, IC1, is connected as an astable (free-running) multivibrator. When the 'push to test' button (S1) is closed capacitor C1 will start to charge up via resistors R1 and R2. As soon as the voltage level on this capacitor reaches the trigger point of the timer the internal flip-flop is activated and pin 7 is taken low to discharge C1. The flip-flop is reset when the voltage on C1 reaches the threshold level of the IC. This takes pin 7 high and the charge cycle starts once more.

The output of the timer (pin 3) is connected to a pair of light-emitting diodes. When the output is high LED D2 will be on and D1 will be off. Conversely, when the output is low D1 will be on and D2 will be off. The LEDs will flash on and off alternately – provided, of course, that the IC under test is a good one.

For readers who may have other applications for the circuit and who wish to alter the frequency, the rate at which the LEDs flash is determined by the values of R1, R2 and C1. The frequency of oscillation can be calculated from the formula

$$f = \frac{1.44}{(R1 + 2R2) \times C1}$$

If, as in this case, the value of R2 is much greater than the value of R1, the frequency can be approximated from the following:



$$f \approx \frac{0.72}{R2 \times C1}$$

For the values shown in the circuit diagram the frequency works out to be around 0.5 Hz.

The tester can be made very compact by soldering all the components directly to the IC test socket, which is first mounted through a hole in the upper surface of the box or container to be used. Alternatively, the components can be mounted on a small piece of Vero-board or similar. Current consumption is minimal and the unit can be powered from a single 9 V battery.

208 State variable filter

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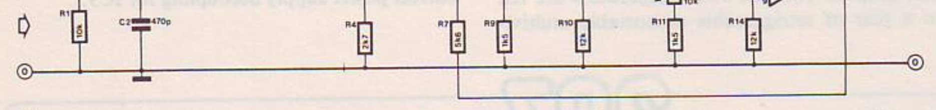

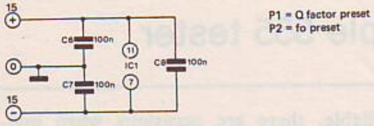

Many multi-band receivers and cheap communications receivers have a bandwidth compatible with the popular broadcasting stations which is too wide for amateur radio purposes. Narrow band receivers or – even better – receivers with an adjustable bandwidth are usually only found in more expensive systems.

In order to be able to listen to amateur radio transmissions (SSB and CW) without too much interference, on a broad band receiver, this ingenious audio band filter may provide a sol-

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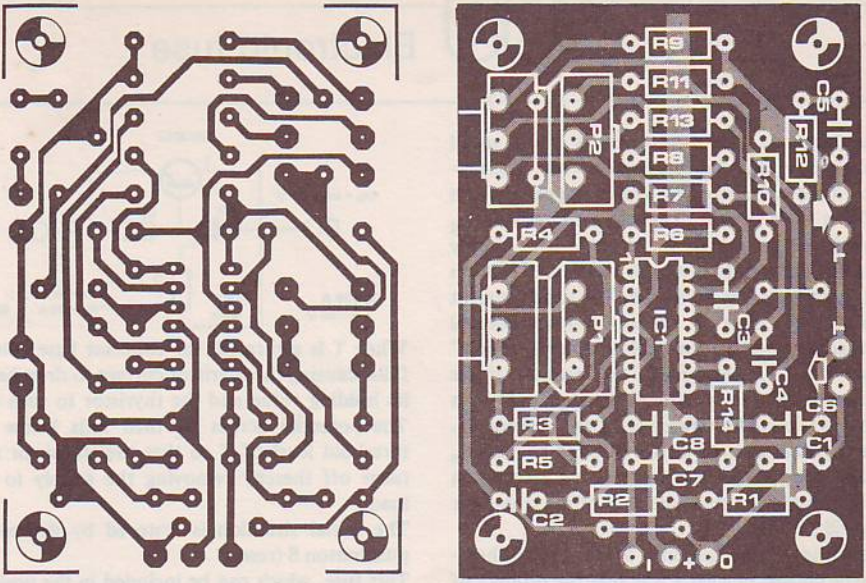


ution. It is a state variable filter, that is one in which both the centre frequency and the bandwidth can be varied. If the filter is placed before the existing AF amplifier, it will be possible to attenuate any interference signals by tuning the filter as accurately as possible to the frequency of the received audio signal. It is of course no comparison to a 'real' narrow band receiver, but the result is usually very satisfactory.

The input filter consisting of C1, C2, R1 and R2 is adequate to reduce the range of the available audio spectrum. The 6 dB points of this filter network are at 500 Hz and 3400 Hz. Opamp A1 acts as a buffer between the input filter and the state variable filter proper. The latter is constructed around the remaining opamps A2...A4. The Q of the filter and thus the bandwidth can be adjusted by means of P1 (control range: $1 \leq Q \leq 50$). The center frequency of the filter can be varied between 200 Hz and 2 kHz by P2. By manipulating these two potentiometers a very small area can be 'extracted' from the total audio spectrum.

Since wide band receivers and cheap communications receivers appear to be very popular at the moment, it seemed worthwhile to produce a printed circuit board for the state variable filter. Fortunately it is very compact as all the opamps in the circuit are contained in one IC (4136). The circuit requires a dual supply voltage (+ and − 15 V), but as current consumption is very low the supply need only produce a few milliamps.

Parts list

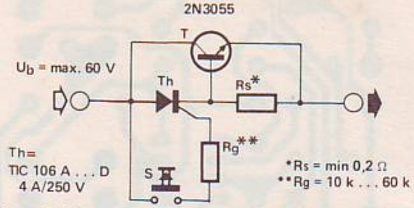
Resistors:
R1 = 10 k
R2 = 100 k
R3, R5 = 2k2
R4 = 2k7
R6, R7 = 5k6
R8, R13 = 8k2
R9, R11 = 1k5
R10, R14 = 12 k
R12 = 3k9

Capacitors:
C1 = 33 n
C2 = 470 p
C3, C4 = 10 n
C5 = 270 n
C6, C7, C8 = 100 n

Miscellaneous:
IC1 = A1...A4 = 4136
P1 = 2 x 100 k log.
P2 = 2 x 10 k lin.

209 Electronic fuse

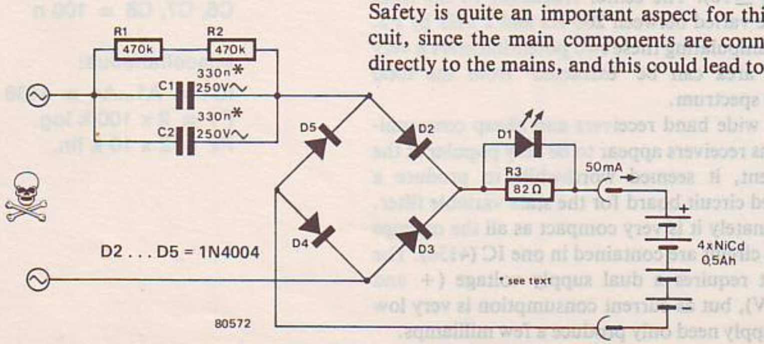
The electronic fuse shown here is a high speed and easily repaired direct current fuse. The thyristor (Th) is triggered by depressing pushbutton S for a brief period. The value of resistor R_g must be about 1 kΩ for every volt of supply voltage. The pushbutton may be released as soon as the thyristor has turned on. The anode current continues to flow – without further control voltage – until it drops below a certain 'hold' value. This will take place, for instance, if the current is conducted around the thyristor! This purpose is served by transistor T and resistor R_S. The thyristor current passes through resistor R_S and as soon as the drop in voltage across it is greater than the threshold level of the transistor this will conduct.
The value of R (minimum: 0.2Ω) must therefore be chosen in such a way that the product of the maximum current and R_S is greater than the transistor threshold level (approximately 0.7 V).



When T is saturated, the collector base voltage falls, causing the thyristor current to drop below its holding value and the thyristor to turn off. The potential across R_S then falls below the threshold level of T so that the transistor also turns off thereby removing the supply to the load.
The initial situation is restored by depressing pushbutton S (reset).
This fuse, which can be included in the positive supply lead of most systems without any difficulty, causes no more than a drop of 1 volt.

210 Nicad charger

Now that the price of small nicad cells is relatively low the cost of the necessary charger is disproportionally high. It is hardly possible to find a less expensive method of charging 4 penlight cells than the one described here. Furthermore, the circuit offers low dissipation and provides a constant current for the cells to be charged.
The circuit uses two capacitors connected in parallel, instead of the usual transformer, to obtain sufficient current (a tenth of the batteries' capacity or 50 mA) from the mains supply. The voltage appearing at the 'cool' end of the capacitors is then rectified by means of four diodes.
The LED has been included to provide an indication that the circuit is actually charging. The resistors, R1 and R2, have been added as a measure of safety. This is because when the battery charger is switched off the capacitors may still be fully charged unless some form of discharge path is provided.
Safety is quite an important aspect for this circuit, since the main components are connected directly to the mains, and this could lead to acci-



dents! For this reason a great deal of care must be taken during its construction.
The entire circuit can be mounted in the hinged part of a cassette case in such a way that it is impossible to touch the parts where a high voltage is present. The section containing the nicad cells has two solder pins which disappear into two holes when the box is closed and which don't make electrical contact until the box is almost completely shut. This eliminates any danger while the cells are being charged. Obviously, adding a 1 A fuse and a double-pole mains switch wouldn't be a bad idea.
Note that capacitors C1 and C2 must be suitable for use at AC mains voltages − 250 V at least. Be warned: the DC working voltage gives no guarantee. It is no exception for a capacitor rated at 400 V DC to have an AC working voltage of only 200 V, or even less!

C.W. Brederode

211 High-frequency optocoupler

It is often necessary to ensure a 'safe' transfer of signals from one circuit to another. The desired AC signal must be passed, but even quite high DC voltages must be completely blocked − even AC *energy* transfer from one circuit to the other is often highly undesirable. Typically, this sort of situation occurs where mains voltages or high DC voltages occur in one circuit, whereas the other must be 'safe to touch'. A standard solution, nowadays, is to use a so-called optocoupler − the desired signal is passed as light.
In the circuit shown, the input signal is applied to T1. This transistor is biased to 20 mA by means of R1...R3. R3 is selected so that I_F (the current through the photodiode) varies from 15 mA to 25 mA as the input voltage swings 1 V peak-to-peak. Linearity can be improved at the expense of signal-to-noise ratio by reducing the I_F swing. This is accomplished by increasing R3 and adding a resistor from the collector of T1 to ground to obtain the desired quiescent current through the photodiode (20 mA).
The output transistor in the IC, T2, is connected in a cascade circuit with T3. Feedback is applied through R4 and R6. R6 is selected for maximum gain/bandwidth product of T3. R7 determines the output swing; obviously, it should be selected to obtain maximum output without clipping. The closed-loop gain ($\Delta U_{out}/\Delta U_{in}$) is determined by R4, as follows:

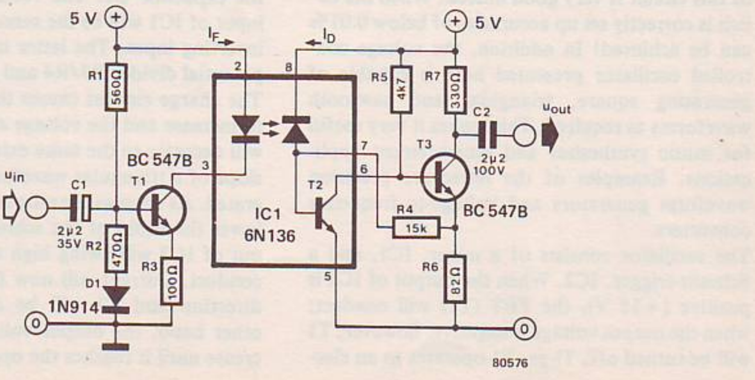
$$\frac{U_{out}}{U_{in}} = \frac{I_d}{I_F} \cdot \frac{1}{R3} \cdot \frac{R4 \cdot R7}{R6}$$

In the unlikely event that the output amplifier (T2/T3) decides to operate as an oscillator, a capacitor of 27...100 p between collector and base of T3 should bring it back into line.

Typical data:
2% linearity over 1 V_{pp} dynamic range

Bandwidth:	10 MHz
Gain drift:	− 0.6%/°C
Common mode rejection:	22 dB at 1 MHz
DC insulation:	3000 V

HP application note

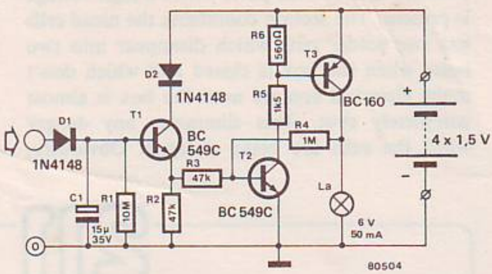


212 Security rear light

Living in a temperate climate involves having to cycle with lights on every now and again. The trouble is that a bicycle dynamo can only produce energy for lighting while a steady rate of pedalling is kept up. As soon as you slow down before traffic lights, zebra crossings etc. the intensity of the rear light drops considerably, thus making it very difficult for vehicles coming up behind to see you. The safety rear light described here is an indispensable addition to the normal reflector. As soon as the dynamo is switched on and producing a voltage, the rear light will come on. The light will have a constant intensity no matter what cycling speed is kept up. Furthermore, it will continue to burn for about four minutes after the bicycle comes to rest, which is ample time to cross the busiest of crossroads.

Unfortunately, this circuit has a drawback – it relies on batteries and when these are dead the system will not work. One consolation is, however, that when 4 or 5 alkaline penlight cells are used there is enough energy available for 35 hours use.

The dynamo lead which is normally wired to the rear light is connected to the input of the circuit.



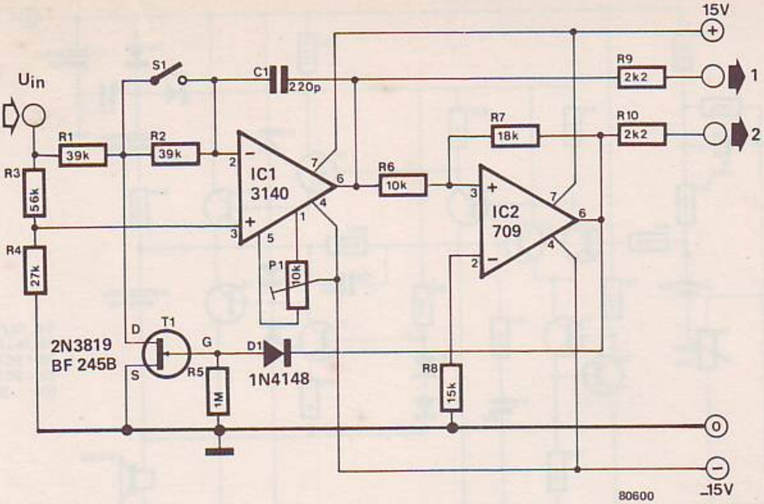
When a voltage is present, T1 is switched on which in turn provides base drive current for T2 and T3. The lamp will now light. When the cyclist stops and the dynamo no longer produces a voltage, T1 continues to conduct for a few minutes until capacitor C1 is discharged through R1 to such an extent that the Schmitt-trigger formed by T2/T3 turns out the light. The entire circuit is now inoperative and no current is drawn from the batteries.

If the system is to be used for long periods at a time, it is advisable to include 5 nicads (also the penlight size). With a capacity of 0.5 Ah and a rear light of 6 V/50 mA the batteries should last for about ten hours.

213 Precision VCO

According to the author, the linearity and synchronisation (when several VCOs are used) of this circuit is very good indeed. When the circuit is correctly set up accuracies of below 0.01% can be achieved! In addition, the voltage controlled oscillator presented here is capable of generating square, triangular and sawtooth waveforms as required. This makes it very useful for music synthesiser and measurement applications. Examples of the latter are precision waveform generators and voltage-to-frequency converters.

The oscillator consists of a mixer, IC1, and a Schmitt-trigger, IC2. When the output of IC2 is positive (+15 V), the FET (T1) will conduct; when the output voltage is negative, however, T1 will be turned off. Thus, T1 operates as an electronic switch. When T1 is conducting, a charge current flows through resistors R1 and R2 into the capacitor C1. The voltage at the inverting input of IC1 will be the same as that at the non-inverting input. The latter is determined by the potential divider R3/R4 and is equal to 1/3 U_{in}. The charge current causes the capacitor voltage to increase and the voltage at the output of IC1 will decrease to the same extent. The descending slope of a triangular waveform is therefore generated. As soon as the output voltage reaches the lower threshold of the schmitt-trigger, the output of IC2 will swing high and T1 will start to conduct. Current will now flow in the opposite direction and C1 will be discharged. On the other hand, the output voltage of IC1 will increase until it reaches the upper threshold of the



schmitt-trigger and the complete cycle can start afresh.

The triangular signal is available at output 1, while a symmetrical squarewave is available at 2. If the switch S1 is closed, the capacitor will be discharged very quickly. Then a descending saw-tooth signal will be generated at twice the frequency of the original triangular wave. The second output will now provide 'needle' pulses. The amplitude of the triangular waveform will be ± 8.3 V, while that of the squarewave will be ± 15 V.

For maximum precision, 1% metal foil resistors should be used (except for R5, R9 and R10) and a ceramic capacitor is recommended for C1.

The frequency of operation can be calculated from:

$$f = \frac{U_{in} \cdot R6}{180 \cdot R7 \cdot R2 \cdot C1} \text{ Hz}$$

With the values shown, a conversion factor of 357 Hz/V is obtained. The circuit is set up by connecting both inputs of IC1 to ground then adjusting P1 to give 0 V out (pin 6).

A. van Ginneken

214 Ulp amp

At first sight, this amplifier design looks like any other. However, it boasts a number of highly interesting characteristics. To start with, the abbreviation 'ulp' stands for 'ultra low power'. This does not refer to the maximum power produced by the amp (100 mW), but rather to the quiescent current consumption which is approximately 1.5 mA. This makes this amplifier particularly suitable for use in a receiver which derives its supply from solar cells.

A further advantage is the wide choice of supply voltages. Apart from the maximum output power, of course, the other characteristics remain unchanged, regardless of whether a 3 V or a 12 V supply is used. The voltage gain is not affected either.

When the circuit in figure 1 is considered, an awful lot of components appear to be involved. However, they are all without exception 'normal' parts.

To enable the amp to work well at all supply voltages, a differential amplifier (T1, T2) has been used as an input stage with a current source (T3, T4) in the emitter lead. The input signal passes from T1's collector to the discrete darlington driver, T5 and T6. To ensure as much gain as possible, there is also a current source (T7) in the collector of T6.

In spite of the fact that the quiescent current of the output transistors is preset (there is only one diode, between the base of T8 and T9), cross-over distortion is kept to a minimum by this type

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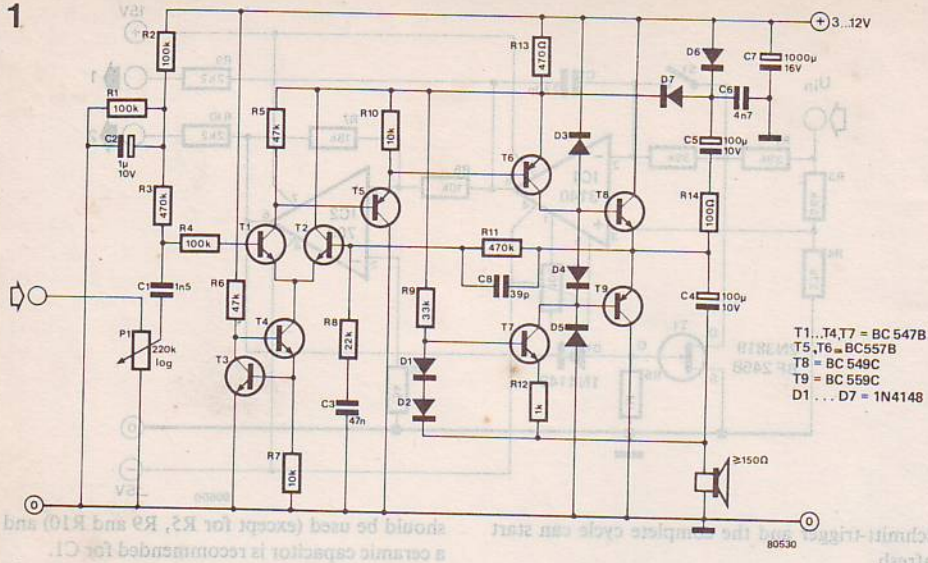

of current control. Negative feedback of course also helps. The feedback network is formed by R11 and C8 and is connected between the emitters of the output transistors and the base of T2. The voltage gain of the ulp amp is therefore determined by the ratio of R11 to R8 which is a factor of 22 in this case.

In order to obtain maximum output amplitude (almost supply voltage – a very rare phenomenon!), bootstrapping has been applied in two ways. As far as the negative half-signal is concerned the 'base' of the current source (junction

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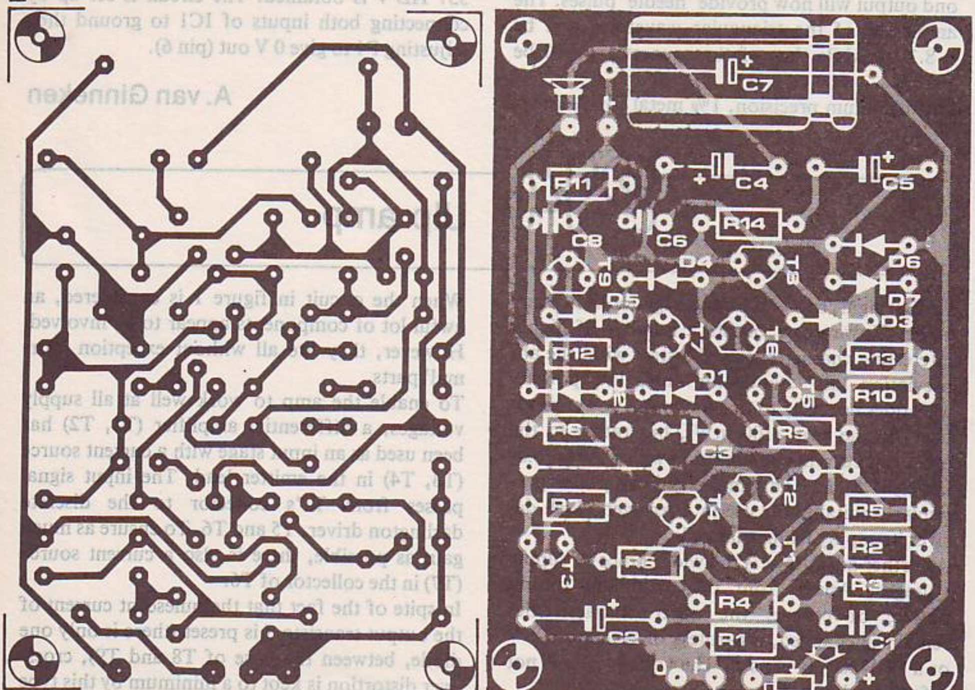


Parts list

Resistors:

R1, R2, R4 = 100 k
R3, R11 = 470 k
R5, R6 = 47 k
R7, R10 = 10 k
R8 = 22 k
R9 = 33 k
R12 = 1 k
R13 = 470 Ω
R14 = 100 Ω
P1 = 220 k log.

Capacitors:

C1 = 1n5
C2 = 1 µ /10 V
C3 = 47 n
C4, C5 = 100 µ /10 V
C6 = 4n7 (cer.)
C7 = 1000 µ /16 V
C8 = 39 p

Semiconductors:

T1...T4, T7 = BC 547B
T5, T6 = BC 557B
T8 = BC 549C
T9 = BC 559C
D1...D7 = 1N4148

D2/R12) is connected to the output capacitor instead of to earth. The peak-to-peak output range will now be greater because the output signal voltage is added to the supply voltage of the driver. Similar use has been made of the positive half-signal and, as far as we know, for the first time in history. The output signal is passed through R14 and C5 to diodes D6 and D7. After it has been rectified it is added to the positive supply voltage, the signal at the R13/D7 junction will therefore rise above the supply voltage. And of course this will increase the peak-to-peak output range of the driver during the positive half-signal. To prevent T8 and T9 from being over-driven, the driver output is limited by two diodes (D3 and D5).

Figure 2 shows the printed circuit board for the amplifier. It may not be the smallest but its low current consumption will certainly take a lot of beating.

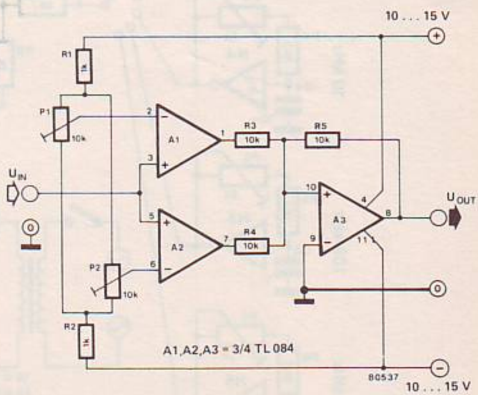
215 Trigger with presettable thresholds

Most triggers with switch hysteresis (the schmitt-trigger included) have switching thresholds which cannot easily be preset (if at all) because the levels affect each other or the switching behaviour of the trigger. This particular trigger is an exception and consists of three opamps.

The switching thresholds can be preset between 0 and 83% of the positive as well as the negative supply voltage, independently of each other, by means of P1 and P2. It makes no difference which pot is used as the upper or lower threshold.

Only when the input voltage is higher than the highest preset threshold voltage, will outputs A1 and A2 both be 'low'. The voltage on pin 10 of A3 will then be lower than that on pin 9, causing the output voltage to become negative.

The trigger can operate with DC or AC signals.



The peak values of the input signal must of course remain within the supply voltage's limits.

216 Simple L/C meter

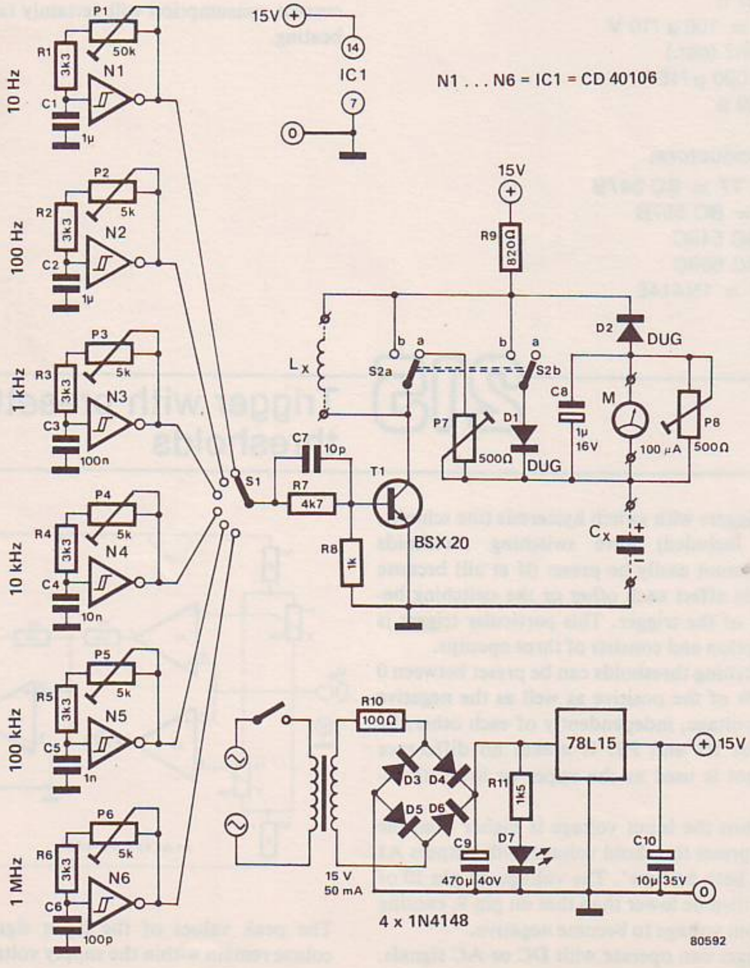
With the aid of this device it becomes a simple matter to check the value of capacitors and inductors. When measuring an inductor (switch S2 in position a) the current flowing through the coil is interrupted periodically so that the self-induced voltage can be monitored. This is achieved by feeding one of the six squarewave signals (N1...N6) to the base of transistor T1. The base drive current to T1 is therefore constant in all cases which means that the collector current is also modulated to the same maximum value. The self-induced voltage U can be found from the formula:

$$U = \frac{-L\Delta I}{\Delta t}$$

where L = inductance
ΔI = change in current
Δt = time during which the voltage change takes place

The self-induced voltage will only alter when a different inductor is placed in the circuit. The average value of the voltage will then be:

$U_{ave} = L \cdot I_c \cdot f$ where
Ic = average collector current
f = control voltage frequency



The average voltage is a measurement of the self-induction. From the proportional relationship between the voltage measured U_{meas} and the inductance L it follows that the scale will be linear. Similarly, it can be proved that the average discharge current of the capacitor C_x (S2 in position b) in this circuit will be:

$I_{meas} = C \cdot U_c \cdot f$ where
U_c = the charge voltage on the capacitor
f = the control voltage frequency

Again the scale division for the capacitance meter will be linear.

The corresponding parameters are given in the table.

In order to calibrate the unit, the squarewave

f in Hz	1 M	100 k	10 k	1 k	100	10
L in H	10 µ	100 µ	1 m	10 m	100 m	1
C in F	100 p	1 n	10 n	100 n	1µ	10

generators must first be adjusted to produce the correct frequencies. A capacitor with a known value is then connected into the circuit and P1 trimmed to give the right reading on the meter. After this, P2 is adjusted with an inductor of known value in the circuit. Accuracy will suffer if a supply voltage of less than 15 V is used. A suitable power supply is shown in figure 2.

Based on an idea by P. Herlitz

217 Sound effects generator

This 'black box' should prove quite popular with (electronic) guitar players. It offers all kinds of possibilities for enriching the sound. There are three controls; the effect of the most important of these (P3) is illustrated in figure 1.

The original input signal is shown at the top. With P3 set to zero, this signal is simply clipped as shown; as P3 is turned up, all kinds of other wave-shapes are obtained – including frequency doubling, even. As will become apparent when we come to the circuit, P3 determines the basic waveform; a further control (P2) determines the 'degree' of the effect; and a final control (P1) sets the sensitivity. As with most circuits of this type, the final effect depends on the input signal level (surprisingly, musicians seem to like it that way!), so that a sensitivity control is both necessary and useful. The overall gain of the circuit depends on the various control settings; it can be anything between x3 and x30 (10 dB...30 dB). Note that it is not the intention to use P1 as a volume control – guitar amplifiers have one of those already.

The circuit is shown in figure 2. A1 is an input buffer amplifier; its gain is determined by P1. The output from A1 is fed to a x10 amplifier (A2) and to a variable-gain amplifier stage (A3), the gain of which is determined by P2a. At this point, things get a bit complicated... Two diodes, D1 and D2, are connected between the outputs from A2 and A3. If the gains of both stages are identical, their outputs will also be identical and the diodes will never conduct. However, as the gain of A3 is reduced, two things start to happen: the output from A2 is clipped at the input to A4, and the output from A3 is *boosted* at the peaks of the signal. The latter signal is inverted by A5 and, at the same

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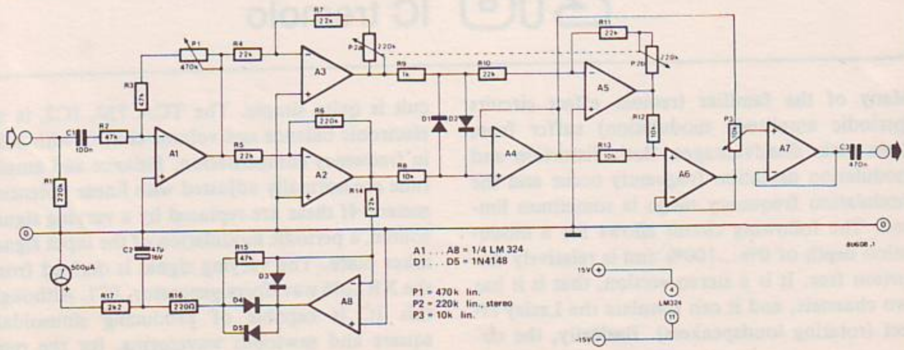

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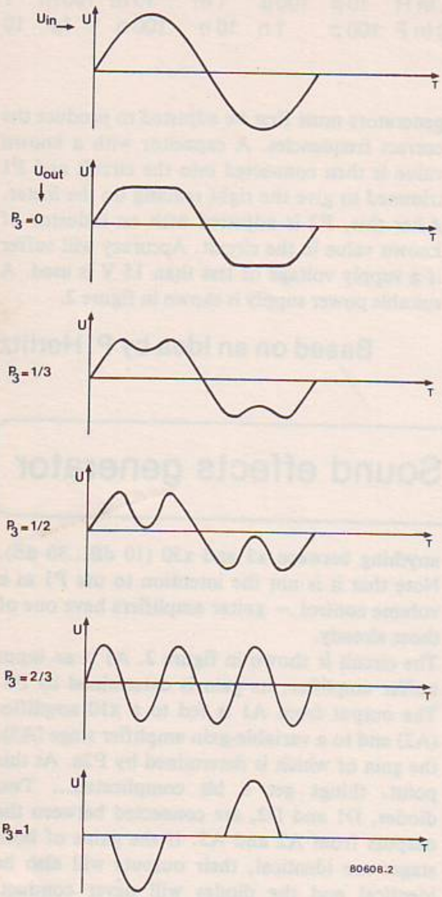


time, the gain in this stage is set by P2b to compensate for the gain difference between the two signal paths that was introduced by P2a. To achieve this, P2a and P2b are connected 'in opposite sense': as the value of one increases, the value of the other must decrease.

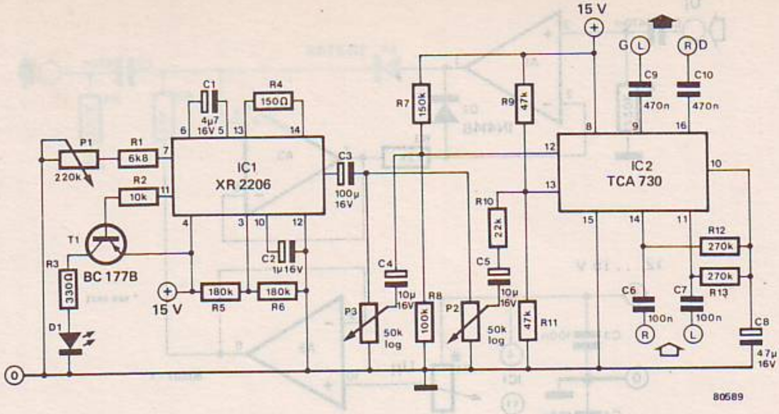
We now have two signals at the same basic level, but in antiphase. Furthermore, where one is 'flattened' at the signal peaks, the other is boosted at those points. These two signals are summed in A6. So what do you get? The basic, undistorted component in the two signals is identical and in antiphase, so it cancels. The distortion components, however, *add*: where the output of A4 is 'down' due to clipping, the output of A5 is at a negative peak level, because this stage inverts. The result of all this is that the output from A6 contains nothing but short peaks, that correspond to the peaks of the incoming signal when D1 and D2 conduct. Or, to be more accurate, short dips that correspond to the peaks and vice versa: when the output of A4 goes high, the output of A6 goes low. P3 can therefore be used to select any desired 'blend' of these two signals, producing the intriguing waveforms shown in figure 1.

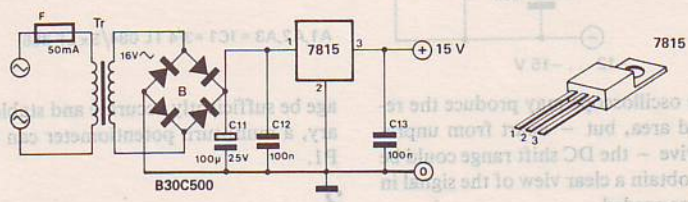
A7 is used as an output buffer. Which leaves one unused opamp, if you're using quad opamp ICs. Pity... may as well do something useful with it: make a simple VU meter circuit (A8).

Using this level meter, it is a simple matter to set up the sound effects generator. The sensitivity control, P1, is set so that hitting one string causes the meter to reach approximately mid scale (40...70%). The basic distortion can now be set between 0 and 100% with P2, and the 'blend' is set with P3. By ear, of course – according to personal taste.

218 IC tremolo

Many of the familiar tremolo effect circuits (periodic amplitude modulation) suffer from three main disadvantages. Both distortion and modulation deviation frequently occur and the modulation frequency range is sometimes limited. The following circuit allows for a modulation depth of 0%...100% and is relatively distortion free. It is a stereo version, that is it has two channels, and it can simulate the Lesley effect (rotating loudspeakers). Basically, the circuit is quite simple. The TCA 730, IC2, is an electronic balance and volume control with built in frequency compensation. Balance and amplitude are normally adjusted with linear potentiometers. If these are replaced by a varying signal source, a periodic modulation of the input signal takes place. The varying signal is derived from the XR2206 waveform generator, IC1. Although this IC is capable of producing sinusoidal, square and sawtooth waveforms, for the pur-





poses of this circuit only the sinewave signal is of interest.

The modulation frequency can be varied with potentiometer P1 as required from 1 Hz to 25 Hz. The squarewave output of the XR 2206 is used to drive a PNP transistor, T1, so that an LED can provide an optical indication of the modulation frequency.

The internal frequency compensation of the TCA 730 (pins 1...7) remains unused. The sinewave amplitude level can be adjusted by P2, thus controlling modulation depth. The degree of balance can be adjusted with P3 (to produce the Lesley effect).

Little needs to be said about the power supply. The 7815 voltage regulator solves all the problems. It is not advisable to use an unstabilised supply as the modulation in the circuit would cause current fluctuations on the supply line. This will then cause deterioration of the modulating sinewave. The supply transformer should have a secondary winding of 15...18 volts at about 120 mA. The voltage regulator requires a heatsink in the form of an aluminium plate of about 10 cm².

T. Stöhr

219 Electronic magnifying glass

Rectification is usually a question of removing the negative half-cycle (during positive rectification) or the positive half-cycle (during negative rectification) of the alternating voltage. The reference for the resultant voltage then becomes 0 V. However, the reference level can be any other positive or negative voltage as required. This is done by removing everything above or below the reference level. An example of this is the circuit in figure 1. It is a precision rectifier which allows all of the input signal, U_i, through unchanged provided it is above the reference voltage U_R (figure 2a). Negative rectification is also possible (figure 2b). All that is required is to change the polarity of diodes D1 and D2. The reference voltage can be preset by potentiometer P1. The circuit works accurately enough for frequencies up to 20 kHz.

What can you do with it? You can make an electronic magnifying glass. Supposing a relatively small portion of the alternating signal is to be examined in detail on an oscilloscope. Increasing

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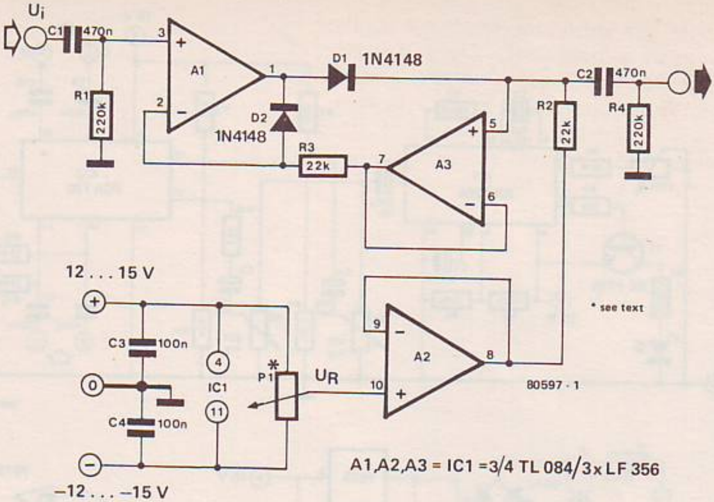

the gain of the oscilloscope may produce the required enlarged area, but – apart from unpredictable overdrive – the DC shift range could be insufficient to obtain a clear view of the signal in as much detail as needed.

Why not then just feed that part of the signal in which we are interested to the oscilloscope. In order to examine the amplitude stability of an oscillator, a positive rectifier is used with a reference voltage preset to a level just below the peak level of the signal. To look at negative extremes a negative rectifier is required. To 'magnify' an area somewhere between the two extremes a positive and a negative rectifier are connected in series.

The value of P1 may be anything between 1 k and 1 M. It is important that the reference voltage be sufficiently accurate and stable. If necessary, a multi-turn potentiometer can be used for P1.

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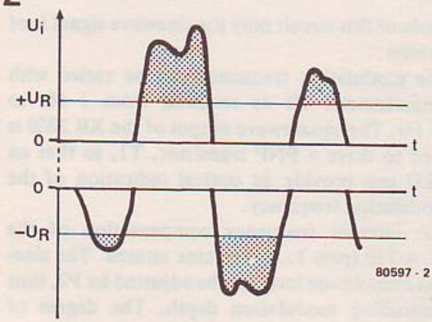

220 Acoustic ohmmeter

There are many instances where it is necessary to test electrical connections, for instance, cable networks, the wiring of connectors, and the various interconnections on a printed circuit board. These tests can be carried out with an ohmmeter, but it is often impossible to keep an eye on everything at once, for while watching the meter dial you have to make sure that the test probes do not short anything out.

One solution is to construct a simple test circuit which produces a sound when the connections

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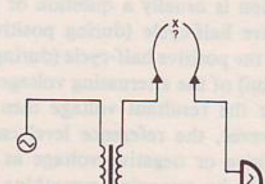

are shorted and which remains silent when the test probes are an open circuit. This permits numerous possibilities. The simplest version which uses a small transformer and a bell is shown in figure 1. The problem with this circuit is that large amounts of current can be produced which could prove too much for the circuit being tested.

By replacing the bell with a small loudspeaker and suitable series resistor the current flow can be limited to below 1 mA. This is shown in the circuit diagram of figure 2. The transformer secondary voltage is full-wave rectified by diodes D1...D4 to provide a 100 Hz signal for the loudspeaker. When the base of T1 is connected to R1 (testing for a short circuit) the two transistors will amplify the signal to produce the required tone.

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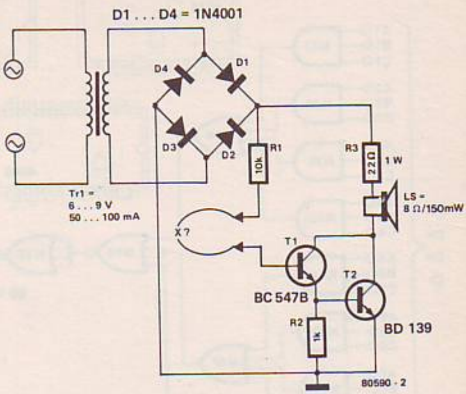

221 Armless bandit

For most of us, giving money away is very easy, all that is required is a car! For many others the odds need to be a little shorter. They use a 'one armed bandit', or 'fruit machine', or 'three in a row', or 'jackpot'. Call it what you will they all do the same job – part you from your money. A

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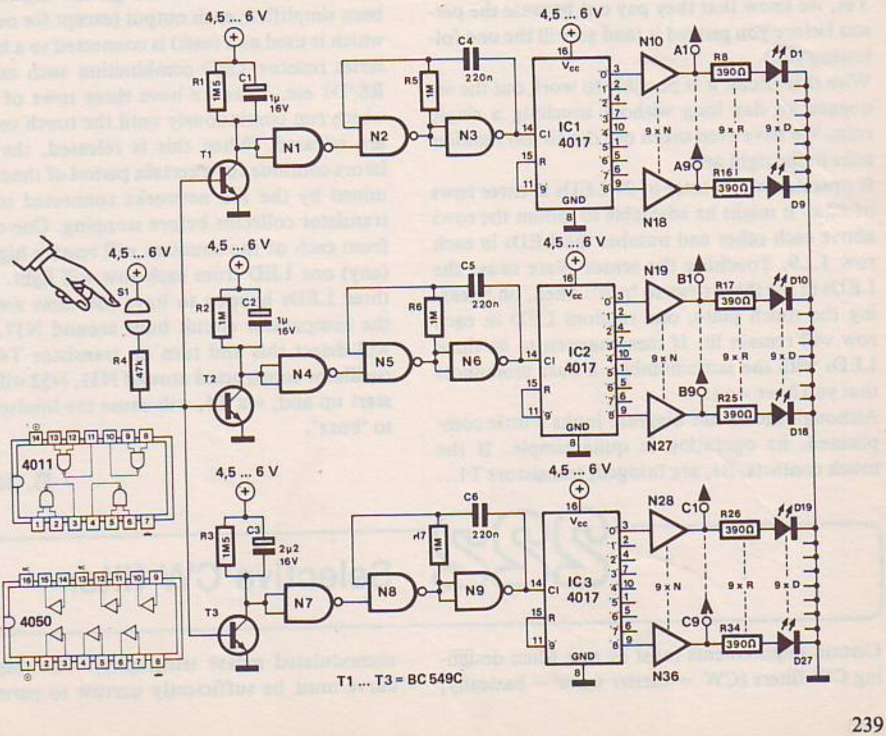
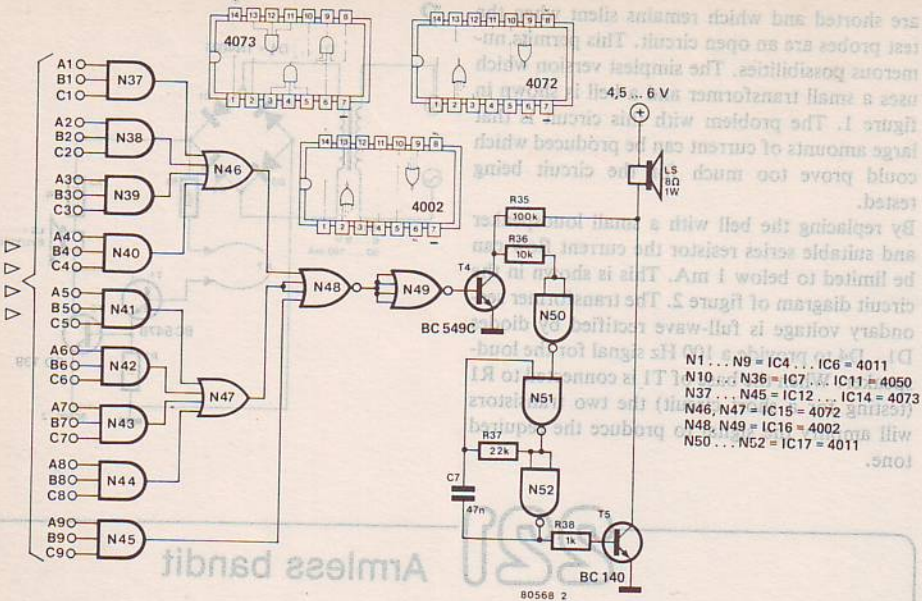




number of rolling drums induce a hypnotic trance in the victim reducing his number of movements to two, inserting coins with the left hand and pulling a lever with the right with only one known surefire cure – atmospheric pockets! Yes, we know that they pay out because the person before you proved it (and so will the one following you).

With this circuit it is possible to work out the sequence all day long without touching a single coin. We have even taken out the all too familiar ache in the right arm.

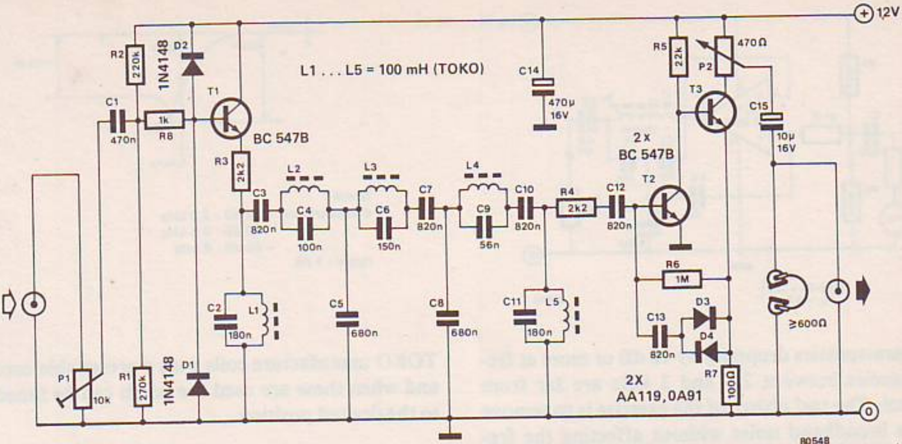
It operates with a total of 27 LEDs in three rows of nine. It might be advisable to mount the rows above each other and number the LEDs in each row 1...9. Touching the sensor plate cause the LEDs in the three rows to 'run'. Then, on releasing the touch plate, one random LED in each row will remain lit. If these happen to be three LEDs with the same number, a buzz announces that you have won.

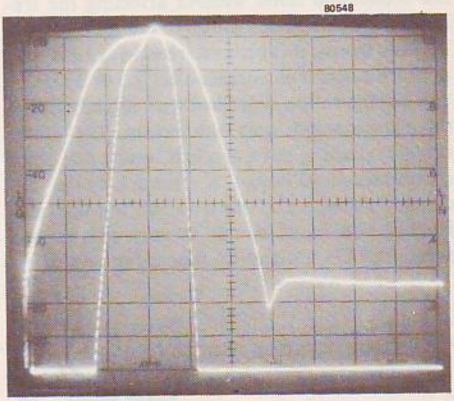
Although the circuit diagram looks a little complicated, its operation is quite simple. If the touch contacts, S1, are bridged, transistors T1... T3 start the three oscillators constructed around N1...N9 which provide the three decimal counters (IC1, IC2 and IC3) with clock signals. The outputs 0...9 of these counters will then go high one after the other. Although the drawing has been simplified, each output (except for output 9 which is used as a reset) is connected to a buffer/series resistor/LED combination such as N10/R8/D1 etc. Thus we have three rows of LEDs which run continuously until the touch contacts are released. When this is released, the oscillators continue for a certain period of time determined by the RC networks connected to each transistor collector before stopping. One output from each of the counters will remain high and (any) one LED from each row will light. If the three LEDs happen to have the same number, the comparator circuit built around N37...N49 will detect this and turn on transistor T4. The oscillator constructed around N51, N52 will than start up and, via T5, will cause the loudspeaker to 'buzz'.

B. Jouët

222 Selective CW filter

Certain requirements must be met when designing CW filters (CW = carrier wave – basically, unmodulated morse telegraphy). The response curve must be sufficiently narrow to permit as





little QRM as possible, but must be wide enough to overcome receiver drift. It must also be phase linear within the bandwidth. Readily available filters usually have a bandwidth in the order of 500 Hz (6 dB points) and give approximately 60 dB attenuation at around 1 to 2 kHz.

In many instances CW signals are insufficiently filtered, so that ringing can often occur in the receiver. A filter with a very high Q allows for maximum selectivity, but is, of course, totally unsuitable for this application. The filter must also be easily reproducable, without any trimming if possible, and the cost must be kept to a minimum.

The circuit shown here fulfils all these requirements and has a comparatively low centre frequency thereby reducing any effect that component tolerances would have at higher frequencies. It was decided to incorporate LC filters in the design as they are superior in operation to RC networks. The coils used need not be of any special quality, virtually any 100 mH choke coils will suffice.

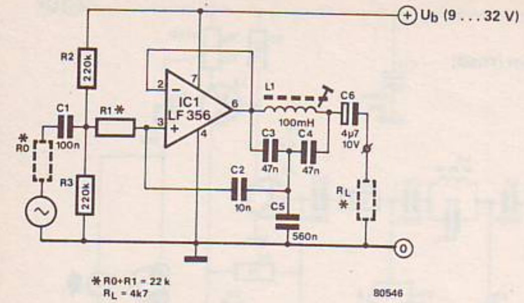
Differences in the signal amplitude of normal receiver filters can be considerable as they are usually more than one CW station 'wide' and the automatic gain control (AGC) may cause the filter to 'pump'. For this reason a limiting circuit (D1 and D2) is included at the input of the filter and a further logarithmic limiter (D3 and D4) is included at the output stage. The audio image signal still needs to be eliminated and this is only possible after the IF signal has had special treatment.

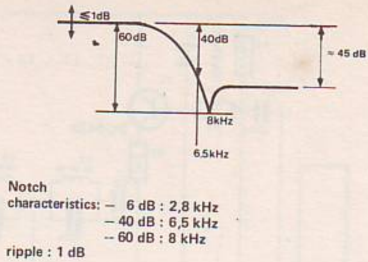
This circuit is virtually indispensable when it comes to displaying CW signals with the aid of a microprocessor and a TV. The filter then only needs a simple interface circuit.

The photo clearly shows the frequency characteristics of the filter circuit. The horizontal scale is 200 Hz per division for both curves. The centre frequency of the filter is around 600 Hz. The narrow curve has a vertical scale of 1 dB per division while the broader curve has a scale of 10 dB per division.

223 Low-pass filter

Most communication receiver designers assume that 'anything above 3 kHz can be cut out' and sometimes a series of four RC filters in a row are given in the receiver. It is obvious that frequency





characteristics dropping by 12 dB or more at frequencies between 2.5 and 3 kHz are far from ideal. The real object of the exercise is to remove the broadband noise without affecting the frequency characteristics within the range required. In other words, it is better to remain at 1 dB with as sharp a cutoff as possible after 2.5 or 3 kHz.
The configuration described here already fulfils this function with only one section. The circuit contains nothing but a single IC and a few other components. Although it appears to be decoupled through C5, this is in fact not the case, because the input circuit is part of the star formed by C3, C4 and C5 (bootstrap principle).
TOKO manufacture coils with a presettable core and when these are used the notch can be tuned to the desired position.

224 The STAMP

Super Tiny AMPlifier

It has nothing to do with the mail, but it is small enough to go through it. This circuit is a versatile, miniature amplifier and loudspeaker combination that can be used in the tightest of electronic corners. The Super Tiny AMPlifier uses only one IC, a loudspeaker and eight other components. It measures two square inches and has an output of 200 mW or more. As if this is not versatile enough, the gain can also be preset (or switched).
Frequently, small projects require an external amplifier of equally small proportions. Finding a suitably sized circuit can present some difficulty. This problem can now be 'STAMPed' out.
The circuit is so simple that it hardly needs any explanation. It is based on the LM 386, IC1. This is produced in a variety of specifications and table 1 lists the major differences, output power and supply voltage being the most significant factors.
The gain of the amplifier is set by the components between pins 1 and 8 of the IC. With

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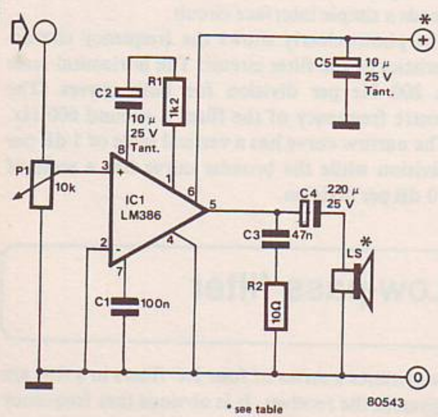

Table 1

Technical data of the LM 386

Operating supply voltage			
LM 386N			4...12 V
LM 386N-4			5...18 V
Quiescent current (U_B = 6 V)			typ. 4 mA
Absolute maximum input voltage			± 0.4 V
Input resistance			typ. 50 k
Output power (THD = 10%)			
LM 386N-1	U_B = 6 V		325 mW
LM 386N-2	U_B = 7.5 V	R_L = 8 Ω	500 mW
LM 386N-3	U_B = 9 V		700 mW
LM 386N-4	U_B = 16 V	R_L = 32 Ω	1 W
Absolute maximum package dissipation (at 25°C)			
LM 386	660 mW		
LM 386A	1.25 W		

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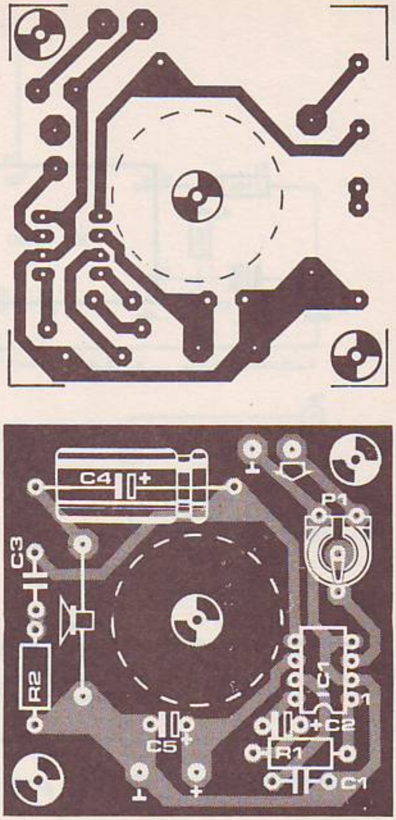


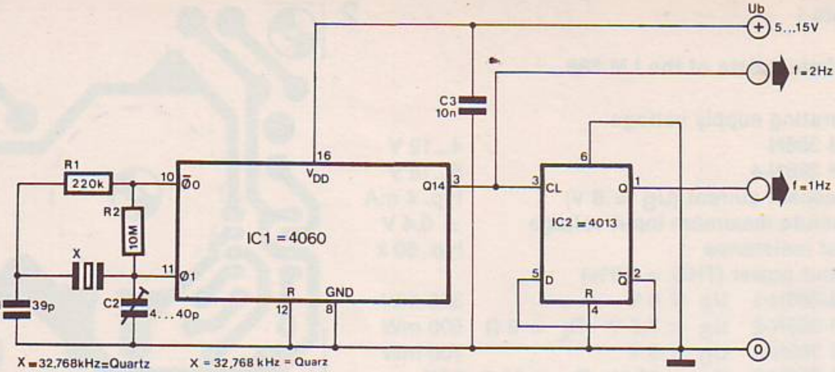
Parts list

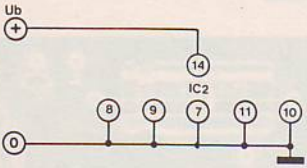
R1 = 1k2 (see text)
R2 = 10 Ω
P1 = 10 k preset potentiometer
C1 = 100 n
C2, C5 = 10 μ /25 V Tantalum (see text)
C3 = 47 n
C4 = 220 μ /16 V
LS = loudspeaker 8 μ /0.2...1 W

both R1 and C2 included (in series) the gain is set at 50. Excluding these two components sets the gain at 20. For the maximum gain of 200, C2 is included and R1 is replaced by a wire link.
The loudspeaker is the limiting factor of the output power if size is the major consideration. The printed circuit board was designed so that, after cutting the hole out of the centre, the board can be mounted over the magnet of the speaker. This of course, means that only the smallest of speakers would be suitable limiting the power output to 200 mW. However, there is no reason why a larger loudspeaker can not be used and the STAMP stuck (with double-sided tape) on or near it. Table 1 should be referred to in this case.

225 Cheap seconds

Miniature crystals for watches are almost invariably tuned to 32.768 kHz, and for a good reason – it is quite easy to derive a 1 Hz signal from this frequency.
Add to this knowledge the fact that the IC type 4060 contains a 14-bit divider stage and oscillator section, and you can put two and two together... The watch crystal is eminently suitable as frequency-determining element for the oscillator. If the maximum division ratio of the divider stage is used (2^{14}), the result is a 2 Hz output. Getting close... To bring this down to the mother of all timing intervals one-pulse-per-second, a single flipflop must be added; half of a





4013, say. The output from this flipflop is a 1 Hz signal, swinging between 0 V and positive supply.

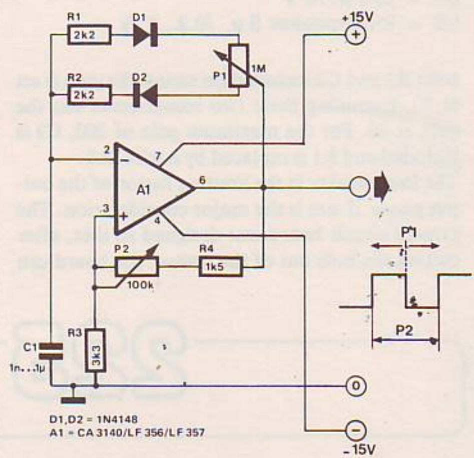
Life would be beautiful, if it weren't for component tolerances. Crystals need trimming – which is where C2 comes in. For absolute accuracy, a frequency meter is required. Connected to pin 9 of IC1, it should read 32.768 kHz.

226 Variable pulse width generator

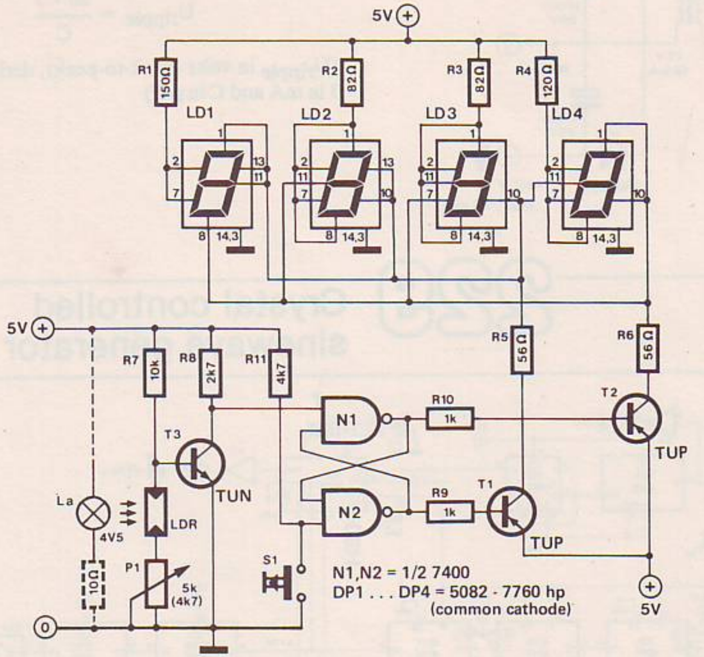
This simple little circuit will find a great many uses where pulse width is critical.

Opamp A1 is connected as a squarewave generator whose frequency can be adjusted by potentiometer P2. Let us assume that the output of the opamp goes high when the system is first switched on. Part of that output voltage is fed to the non-inverting input via the voltage divider R4, P2, R3. As long as C1 is not yet sufficiently charged, the voltage at the inverting input will be lower than that at the non-inverting input and the output will remain high. The moment that the capacitor is charged to such an extent that the voltage at the inverting input becomes greater than that at the non-inverting input, the output of the opamp will swing low. Capacitor C1 then starts to discharge until the voltage at the inverting input becomes lower than that at the other input, whereupon the opamp output will swing high once more.

The mark-space ratio (duty-cycle) may be varied with potentiometer P1 without affecting the frequency. This is done by making sure that the charge time is different (either greater or smaller) than the discharge time. Capacitor C1 is charged via part of P1, diode D2 and resistor R2, whereas it is discharged via resistor R1, diode D1 and the other part of P1. The sum of these two time constants will remain the same (as will the frequency) when P1 is used to alter the mark-space ratio.



227 Post indicator



Has the postman been yet or isn't there any mail today? This question is asked daily by millions of people. As a rule, the answer is to go to the letterbox to have a look. The farther you have to go to find out and the emptier the letterbox, the greater your disappointment.

The post indicator shows on four seven-segment displays whether it is worth the walk or not. Initially, the flipflop formed by N1 and N2 is reset and transistor T2 conducts causing the word 'NONE' to appear on the display. When the light beam to the LDR is interrupted (when a letter falls through the letterbox) T3 will conduct briefly and trigger the flipflop. As a result, T2 will turn off and T1 will turn on. The display will then show the word 'POST'.

The circuit will remain in this state until the reset switch S1 is pressed whereupon it will revert to its initial state. For reliable operation it is advisable to mount the lamp and the LDR as close as possible to the actual aperture of the letterbox.

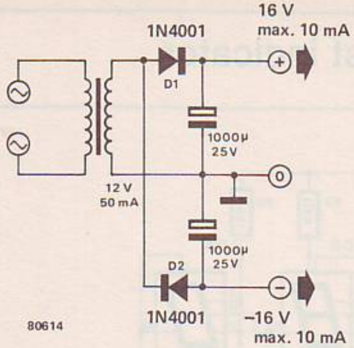
Further to requests by frenzied members of the staff we are now designing a 'bill detector' which will automatically eject unwanted mail.

W. Korell

228 Simple symmetrical supply

Normally, a centre-tapped transformer and a bridge rectifier is used to construct a symmetrical power supply. This seems such a natural solution, that people forget it can be done in a much simpler way. The accompanying circuit diagram shows the simpler version. A disadvantage is the single-sided rectification which makes it necessary to use a larger smoothing capacitor to prevent mains hum.

With the values shown, a maximum of 10 mA



can be supplied at a ripple voltage of about 0.2 V_{p-p}. By using the formula below, values for other currents and ripple voltages can be calculated.

$$U_{ripple} = \frac{20 \cdot I}{C}$$

(U_{ripple} in volts (peak-to-peak), derived current I in mA and C in µ F)

229 Crystal controlled sinewave generator

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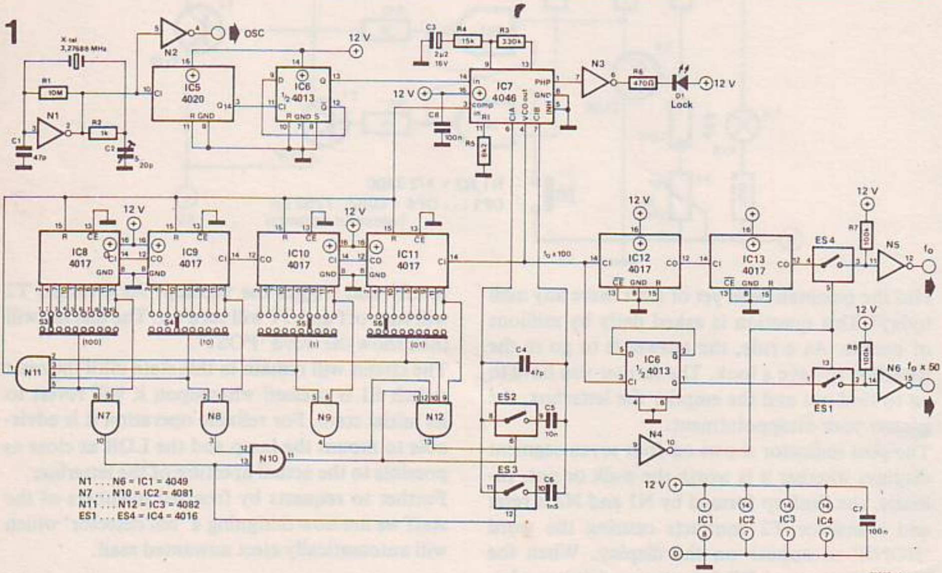

The applications of simple designs are not always restricted to those of a single circuit. A combination of two (or more) circuits can offer new perspectives. For instance, combining a crystal controlled frequency synthesiser and a digital spot sinewave generator produces a very stable sinewave generator. This 'hybrid' uses switches to select the output frequency in 1 Hz steps.

Figure 1 shows the crystal controlled frequency synthesiser section. The heart of the circuit is formed by a phase locked loop (PLL). A very stable frequency is fed to one input of the PLL (IC7) and its output is passed through a variable divider chain before being fed to the other PLL input. The PLL will try to equalise both input frequencies and adjust its output frequency accordingly. Therefore, when the division ratio is set to the figure N, the output frequency will be N times greater than the input frequency. As the input frequency is derived from a crystal source, the output frequency will be very accurate.

The frequency of the crystal oscillator (3.2768 MHz) is divided by a factor of 2^{15} (IC5 and half of IC6) to provide the PLL with an input frequency of 100 Hz. The frequency divider for the

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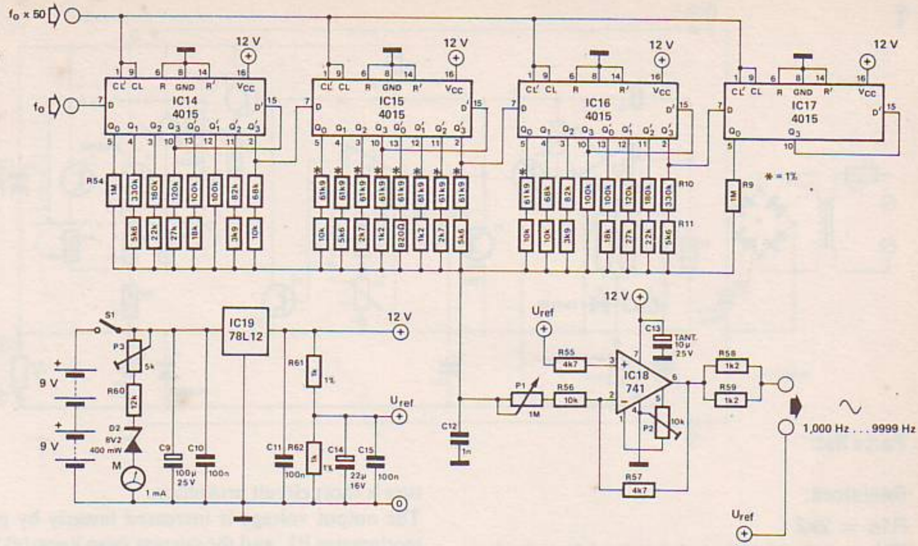

PLL is formed by IC8...IC11 and the desired division ratio (N), and hence the output frequency, is set up on switches S3...S6.

For optimum operation, the value of the capacitor connected between pins 6 and 7 of the PLL will have to be varied with frequency. This is accomplished with the aid of electronic switches ES2 and ES3. The remaining half of IC6 divides the PLL output frequency by two, while IC12 and IC13 form a divide-by-100 counter. This means that two signals are available at the output – one with a frequency fifty times greater than the other.

The circuit of the spot sinewave generator is shown in figure 2. It can be directly connected to the circuit of figure 1. Basically, the circuit consists of a 25 bit shift register and a resistor network.

The fundamental frequency, f_o (output of N5 in
The fundamental frequency, f_o (output of N5 in
figure 1), is fed to the data input of the first shift register (IC14). The higher frequency (output of N6 in figure 1) is fed to the clock input of each shift register. The signals at the outputs of IC14...IC17 are symmetrical squarewaves with a frequency of f_o. The voltages derived from two successive Q outputs are shifted in phase for the duration of one clock period. All 25 output signals are added by means of the resistor network consisting of R10...R54, so that a 50-step sinewave signal is generated across C12. The circuit around IC18 is an amplifier which acts as an output buffer.

The amplitude of the output signal can be varied between 50 mV_{p-p} and 5 V_{p-p} by means of P1. The frequency can be varied in 1 Hz steps between 1000 Hz and 9999 Hz. The sinewave is symmetrical around a reference voltage (U_{ref}) and any offset can be removed by P2. The output impedance of the amplifier is 600 Ω.

Both the 12 V stabilised supply and the reference voltage are derived from a pair of 9 volt batteries (or 4 x 4.5 V). Battery condition is monitored with the aid of the 1 mA moving coil meter, M. Finally, note that the resistors marked with an asterisk (61k9) are listed in the E48 range. If these are not available 62 k 1% resistors from the E24 range will be suitable.

A.G. Hobbs

230 Variable power supply 0-50 V/0-2A

Because of its internal reference source, the LM 10 is eminently suitable for use in power supplies. By using two ICs both the current and the voltage can be made variable. An added fea

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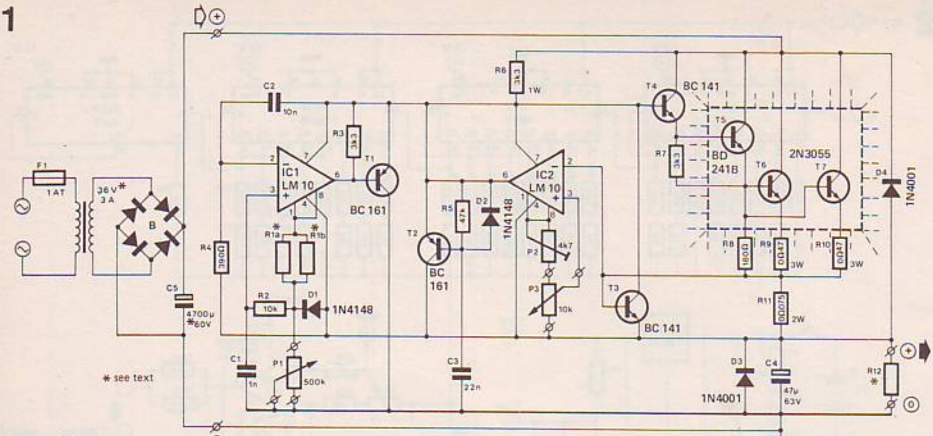

Parts list:

Resistors:

R1a = 2k2
R1b = empirically established (see text)
R2 = 10 k
R3, R7 = 3k3
R4 = 390 Ω
R5 = 47 k
R6 = 3k3/1 W
R8 = 180 Ω
R9, R10 = 0,47 Ω /3 W
R11 = 0,075 Ω /2 W (2 x 0,15 Ω in parallel or resistance wire)
R12 = 470 Ω /5 W
P1 = 500 k lin.
P2 = 4k7 preset
P3 = 10 k lin.

Capacitors:

C1 = 1 n
C2 = 10 n
C3 = 22 n
C4 = 47 µ /63 V
C5 = 4700 µ /80 V (see text)

Semiconductors:

T1, T2 = BC 161
T3, T4 = BC 141
T5 = BD 241
T6, T7 = 2N3055
D1, D2 = 1N4148
D3, D4 = 1N4001
IC1, IC2 = LM 10C

Miscellaneous:

Tr = 42 V (36 V)/3 A transformer
B = B80C2200 (200 V/8 A bridge rectifier)

ture is short circuit protection.

The output voltage is increased linearly by potentiometer P1, and the current (also linearly) by P3. The preset P2 is used to set the peak output current, up to a maximum of 2 A. The maximum output voltage can also be preset by a resistor connected in parallel with R1a. This method ensures better stability and less noise.

The output voltage is stabilised in the following manner. The inverting input of IC1 is connected to the output via R4 with the other input to the junction of P1/R2. The opamp will attempt to prevent any voltage difference by controlling T1. This will either increase or decrease the current through R6 which will consequently vary the voltage to the darlington output stage.

The voltage level at the junction of P1/R2 is generated as follows. Pin 1 of the LM 10 is the reference output. No voltage difference should occur between the two inputs of the opamp, in other words, junction R1/R2 is connected to the same potential as the negative connection (pin 4) of IC1. The reference voltage across R1 will be 200 mV at a current of approximately 100 µ A which will also flow through P1. This means that the potential drop across P1 will be equal to 10^{-4} (100 µ A) times its resistance. Otherwise, there *will* be a difference in voltage at the input of the opamp, so that it will adjust this until the output voltage has reached the exact value.

Current is stabilised by comparing part of the reference voltage (at the wiper of P3) with the voltage dropped across R11 (through which the output current passes). Since the LM 10 is not very fast, conventional current protection has been added with the aid of T3. This limits the current at a fixed threshold value.

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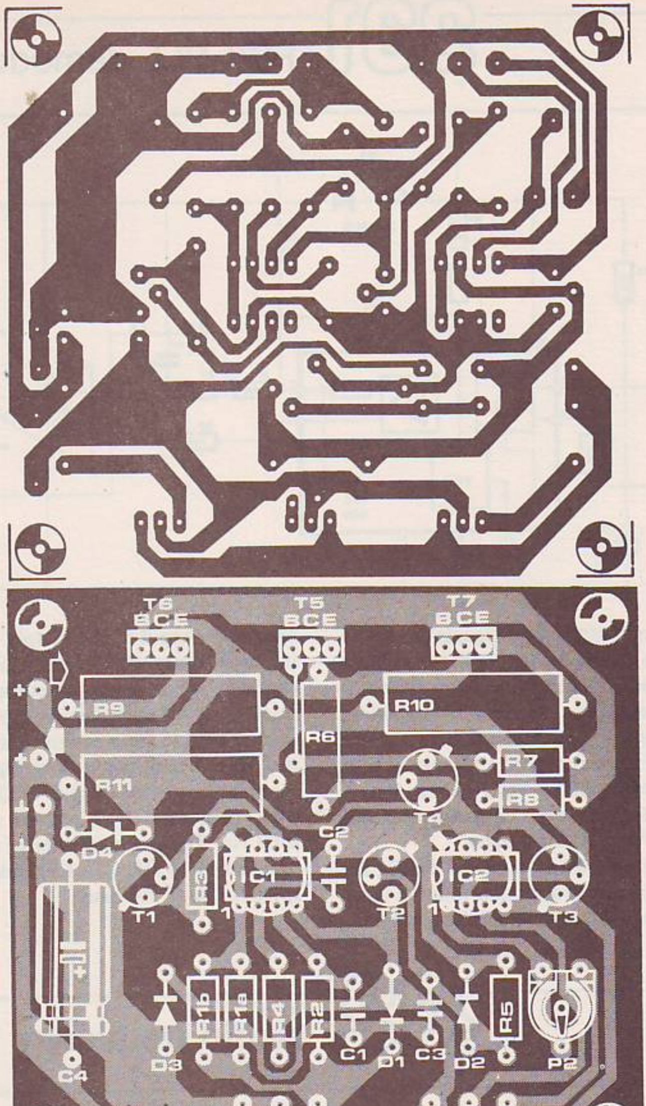


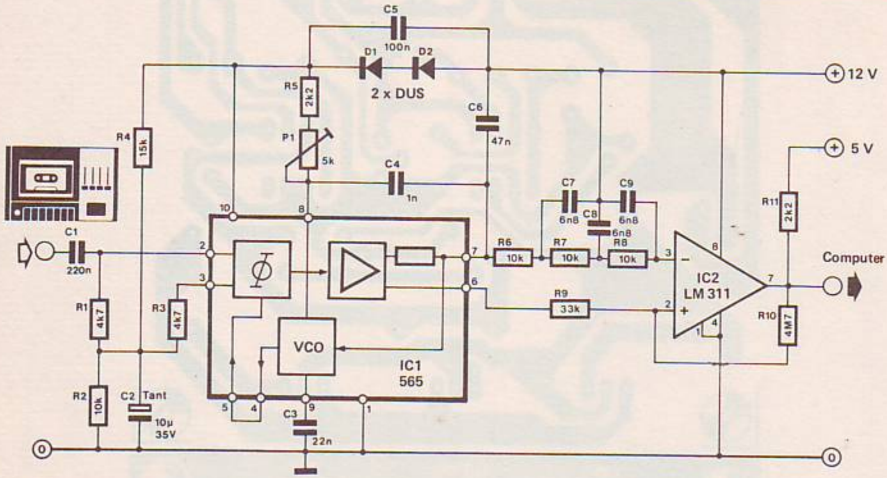
To a certain extent, the minimum output voltage will depend on the load. This is because the (small) supply current of the two opamps passes through the output. It is therefore always advisable to connect a fixed resistor across the output of the supply. With a fixed resistance of 470 Ω (5 W) a minimum output voltage of 0.4 V was measured in the prototype.

The maximum output voltage can be determined with R1b, as mentioned above, and should be no more than 50 V. In many cases however it is better to accept a lower value to work on and use a transformer of 36 V. The 4700 μ electrolytic capacitor may then be the common 63 V type.

Transistors T5, T6 and T7 will have to be mounted on a fairly large heat sink. Figure 2 shows the printed circuit board layout for the supply.

National application note

231 FSK PLL demodulator



FSK (Frequency Shift Keying) signals can be demodulated in a simple manner with the aid of a PLL (Phase Locked Loop). Frequency shift keying is used regularly for data transmission, where a carrier wave is switched between two predetermined frequencies. The frequency shift is obtained by controlling a VCO with the binary data signal, so that the two frequencies are determined by the '0' and '1' logic states.

When a signal is present at the input of IC1 the VCO is locked in synchronisation with the input frequency. This involves an equal change in voltage at the output of the IC (pin 7). The loop filter capacitance (C6) is smaller than usual to eliminate spikes from the output pulse. At the same time, a ladder network of three RC sections is used to filter out the remains of the carrier wave from the output signal. The free-running frequency of the VCO can be preset with potentiometer P1 between about 1900 and 6200 Hz. The characteristics of the circuit (low pass filter R5...R8, C7...C9) make it suitable for speeds of up to 714 Baud.

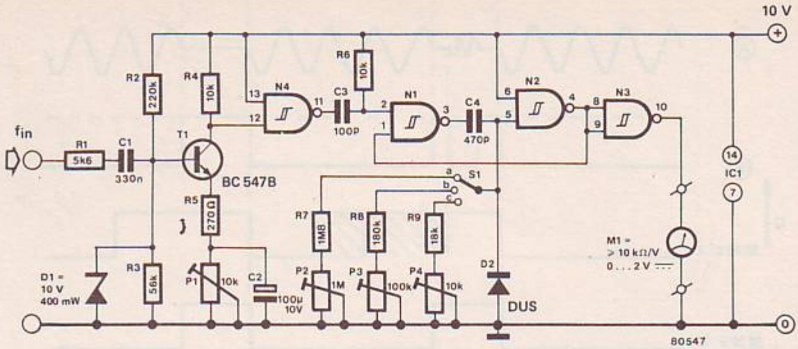
232 Audio frequency meter

If your interest is primarily with audio, a commercial frequency meter, although very nice, is not strictly necessary since most of its range will be redundant. The simple circuit described here is used to convert an ordinary 10 k Ω /volt moving coil volt meter into an audio frequency meter.

The input signal is first amplified by transistor T1 (with a gain of about 40) and then passed through a Schmitt trigger formed by N4. This converts the signal to a square wave and the negative edge of this is used to trigger a monostable multivibrator (N1 and N2). Its output is then inverted by N3 and fed to the multimeter which should be switched to the 2 volts (fsd) range.

The three ranges of the frequency meter are selected by S1. They are 200 Hz, 2 kHz and 20 kHz and are calibrated (with the aid of a frequency generator) by the three potentiometers P2, P3 and P4.

The circuit can be set to maximum sensitivity by P1. This potentiometer varies the DC bias through T1 and therefore the voltage to the input of N4. When this voltage is exactly centred between the two trigger threshold levels the sen-



N1 . . . N4 = IC1 = 4093 B

S1 a = f_{in} max = 200 Hz → 1 V/100 Hz
S1 b = f_{in} max = 2 kHz → 1 V/1 kHz
S1 c = f_{in} max = 20 kHz → 1 V/10 kHz

20 Hz ≤ f_{in} ≤ 20 kHz U_{in} min ≃ 100 mV

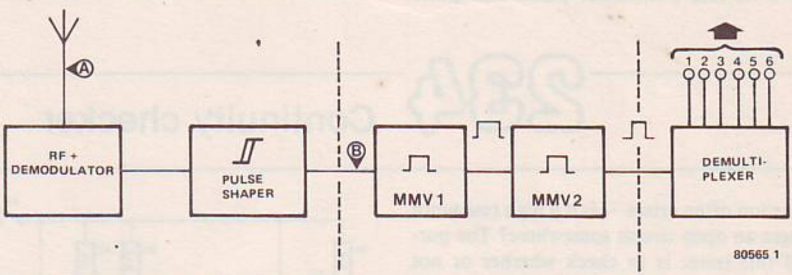
sitivity is then at a maximum.

The input is able to withstand up to 50 volts peak to peak. For low input voltages, less than 14 volts peak to peak, the impedance is about 25 k Ω. At greater input voltages, D1 starts to conduct and the input impedance drops to about 5 k Ω.

The accuracy of the frequency measurement will be determined by the meter used since the accuracy of the circuit itself is better than 2%.

233 Deglitching remote control

1



Interference pulses in remote control receivers are a nuisance. Worse, when the control is for model planes they can be fatal... A quite effective interference rejection circuit can be built, using only two monostable multivibrators. Figure 1 illustrates the basic principle, as a block diagram; figure 2 gives a complete circuit for one common commercial system. The circuit is included in the receiver, between the pulse shaper and the demultiplexer.

In a normal system, interference at a level of only 10...30% of your own signal strength is enough to drive the servo's completely haywire. The 'interference susceptibility' of a given receiver depends, by and large, on its response speed. The faster it reacts, the more it tends to go wild. In general, the end of the 'burst' from the transmitter is the most sensitive period. As illustrated in figure 3, interference spikes after the transmitter shuts down tend to extend the output from a pulse shaper – like MMV1 in figure 1. However, MMV2 in figure 1 is not 'retriggerable': it gives a brief pulse, after which it has

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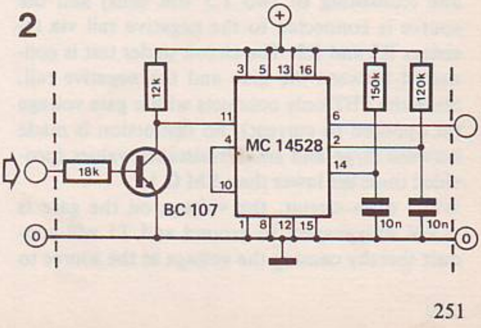

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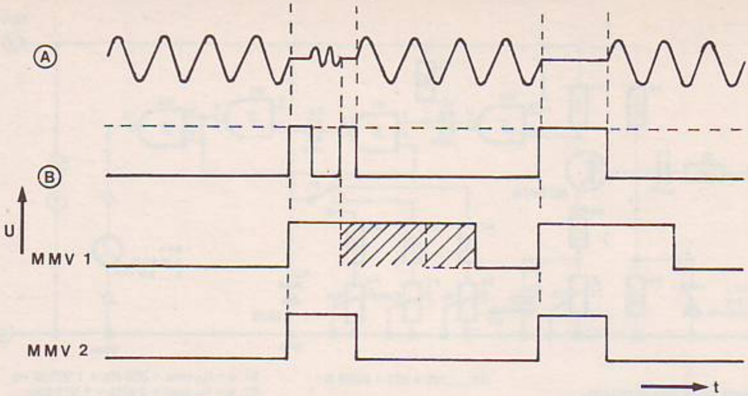


a considerable 'down time'. This second one-shot is triggered at each positive edge from the first MMV, so that it 'reconstructs' the original control signal – ignoring any interference spikes that occur after the transmitter closes down! If the interference is prolonged, MMV1 remains triggered; it doesn't produce any further pulses for MMV2, so that no further output pulses are produced. The servo's remain in their original position. Not that that is necessarily ideal, but it is better than having them run wild! The pulse length of MMV1 should be approximately twice that of a normal transmitter pulse. The pulse length of the second MMV is less critical: anywhere between 0.2 and 0.5 milliseconds should do.

The circuit given in figure 2 is a typical example of the principle. However, component values will vary from one system to another. Until manufacturer's standardise, we can't give one final 'recipe'! The principle is valid, however, for all similar remote control systems that use AM modulation.

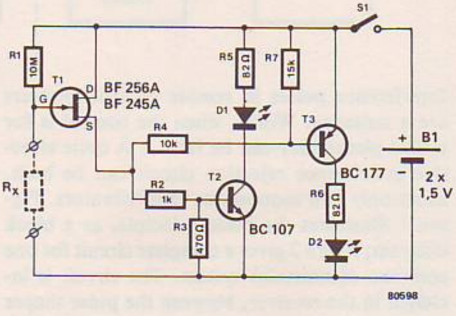
A. Stampfl

234 Continuity checker

The question often arises – is it a high resistance or is there an open circuit somewhere? The purpose of this tester is to check whether or not there is a conductive path with a resistance of less than 5 MΩ between two points. A higher resistance than this is indicated as an open circuit. The results are indicated by two LEDs.

As the circuit diagram shows, the drain of FET T1 is directly connected to the positive supply line (consisting of two 1.5 volt cells) and the source is connected to the negative rail via resistors R2 and R3. The circuit under test is connected between the gate and the negative rail. Since the FET only conducts with a gate voltage (as opposed to current), no distinction is made between large and small resistance values (provided these are lower than 5 MΩ).

When open circuit, the voltage on the gate is +3 V with respect to ground and T1 will conduct thereby causing the voltage at the source to



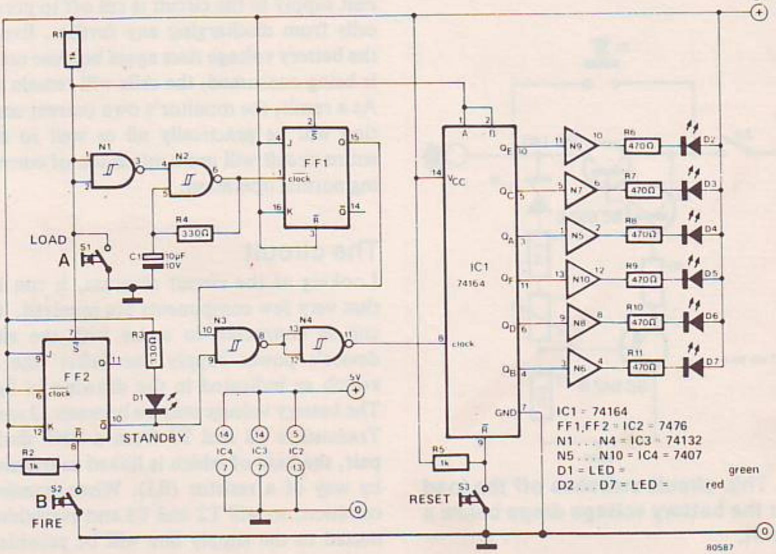
just about reach the supply voltage. This in turn provides transistor T2 with a base drive current and it starts to conduct, with the result that LED D1 will light. If the resistance is lower than approximately 5 MΩ, the gate voltage will drop, so that the FET will behave like a large resistance and the voltage at the source will also drop.

Transistor T2 will then turn off and, of course, so will D1. As far as T3 is concerned, the voltage on its base will also drop causing it to conduct and thereby lighting LED D2.
The value of R1 determines the resistance range which can be tested. With the value given here the highest resistance which can be tested is approximately 5 MΩ.

M.S. Dhingra

235 Pebble game

This circuit is based on the well known video game 'Break Out' where the idea is to shoot down as many bricks as possible in as few attempts as possible. In the version described here the bricks are represented by six LEDs.
Initially the reset key must be depressed for the six 'pebbles' to light up. The device is then activated with the 'LOAD' key whereupon the green LED (D1) will light up. If a point is scored with the 'FIRE' button, the pebble that was hit disappears and the corresponding LED goes out. Scoring a hit is pure coincidence and the system will have to be reloaded after each shot.
There are two methods of playing the game. Either each player has to hit all the pebbles and the winner will be the one to do this in the least number of shots, or the players fire shots alternately and he who hits the last pebble wins.
The operation of the circuit is quite straightforward. The shift register (IC1) is reset by switch S3. All the outputs will then be low and the six pebble LEDs will light via the buffers N5...N10. Operation of the 'LOAD' key, S1, triggers several events. The 'standby' flipflop, FF2, is set so that its $\overline{Q}$ output goes low and lights LED D1. The oscillator formed by N2 is enabled for the length of time the 'LOAD' key is depressed and this will clock flipflop FF1. When the 'LOAD' key is released therefore, the Q output of FF1 will be either high or low depending on the frequency of oscillation and amount of time the key was depressed. No clock pulses will reach the shift register however as they are inhibited by N3.
Once loading is complete the 'FIRE' button can be pressed which will reset FF2 and the standby LED will go out. If the Q output of FF1 was high, the output of N3 will go low and the shift register will receive a clock pulse via N4. As the two serial inputs of IC1 are permanently held high, each of the shift registers outputs will go high (and remain high) in turn every time a clock



pulse is generated. The corresponding LED will, of course, go out to indicate a 'hit'. If the target is missed (the Q output of FF1 is low) the number of pebbles remains unchanged. In both instances the unit is prepared for the next LOAD and FIRE.
The $\overline{Q}$ (or Q) output of FF2 can also be used to control a step counter to indicate exactly the actual number of shots fired. This makes scoring much easier. Power supply requirements for the pebble game are a measly 5 V/100 mA.

H.-J. Walter

236 NiCad battery monitor

keeps the cells 'topped up'

Now that an increasing number of battery-powered devices are being used in the home, it is much more economical to replace 'ordinary' batteries with NiCad cells. If such cells are to lead a long and healthy life, however, they will have to be correctly recharged from time to time. The question is, when is the right moment to recharge them?
More often than not, no indication is given on the electrical device itself and it isn't until the portable radio, the calculator, etc. stops working that its batteries are discovered to have run out, but then, of course, it is already too late... This article describes a small circuit that constitutes a very straightforward and yet highly effective method of keeping NiCads permanently 'topped up'.

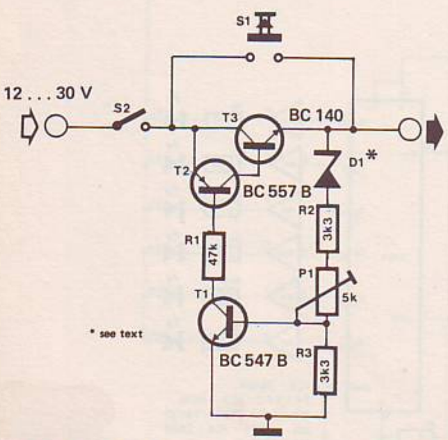


Figure 1. This circuit switches off the load whenever the battery voltage drops below a certain limit.

It would seem that batteries are specifically designed to go flat at the most inoportune moment, during an interesting radio programme or when the calculator is absolutely necessary. In either case, the answer is not simply to replace pen light batteries by NiCad cells, as these need recharging too every now and then. The trouble is, very few devices are equipped with some sort of monitor system, so it is very difficult to know when the cells need boosting. To sit back and wait until they run out won't exactly guarantee the cells a long lifespan – which, remember, was the reason why they were bought in the first place!
The author felt it was high time an end was put to this situation and designed a straightforward circuit to monitor the battery voltage. The circuit operates as follows: when the voltage drops below a certain pre-determined value, the current supply to the circuit is cut off to prevent the cells from discharging any further. Even when the battery voltage rises again because no current is being consumed, the cells will remain cut off. As a result, the monitor's own current consumption will be practically nil as well so that the entire circuit will use a minimum of current during normal operation.

The circuit

Looking at the circuit diagram, it can be seen that very few components are involved. The circuit is connected in series with the electrical device's power supply line 'after' the on/off switch as indicated in the drawing in figure 1. The battery voltage may be between 12 and 30 V. Transistors T2 and T3 from a PNP darlington pair, the base of which is linked to transistor T1 by way of a resistor (R1). When transistor T1 conducts, so will T2 and T3 and everything connected to the supply line will be provided with current. If, on the other hand, T1 stops conducting, T3 will stop too and the cells will no longer supply any current.

The purpose of the circuit is to allow T3 to conduct for the period during which the battery voltage (under load) is higher than 80% of the nominal voltage. This is done by connecting D1, R2, P1 and R3 in series, the junction of P1 and R3 being connected to the base of T1. If the base voltage of T1 drops below 0.6 V this transistor will stop conducting (and so will T3). The values of the zener diode and the resistors are chosen so that the voltage at the base of T1 is greater than 0.6 V when the battery voltage is 0.8 times the size of the nominal voltage. At the same time, the zener diode makes sure that a large share of the change in voltage on the supply line reaches the base of T1. The zener voltage is dependent on the battery voltage and can be calculated as follows:

$U_Z = 0.8 \cdot U_{nominal\ battery} - 1.5.$

D1 may then be the lowest value closest to that result. The zener diode need only be a 400 mW type, as in this particular case the current passing through it will be very low (only about 200 A). Otherwise the true zener voltage will drop way below the level indicated and the calculation will no longer apply.

Pushbutton S1 plays a very important part in the circuit. If we were to construct the circuit without it, or the batteries for that matter, the circuit would never conduct. When the circuit is initially switched on, current is unable to reach the zener diode and the resistor chain, as a result of which the voltage at the base of T1 will prevent the transistors from switching on. If, however, S1 is pressed briefly, current will be able to reach the resistor divider chain via the zener diode. This will enable transistor T1 to conduct and thus switch on the rest of the circuit. It will be apparent that only a momentary operation of S1 is necessary.

The precise moment at which the circuit switches off can be determined with the aid of the preset potentiometer. First of all, the voltage of a fully charged cell that is under no load is measured with an accurate voltmeter. After this, 80% of the measured voltage is fed to the input of the circuit by means of an accurate power supply. P1 is then adjusted very carefully until the point is reached where T3 stops conducting (don't forget to press S1).

The circuit can produce a maximum current level of 1 A. The current consumption is very low. When the circuit is switched on this will be less than 0.5 mA at 12 V and less than 1 mA at 30 V. In the 'off' state, the amount of current consumed will be negligible.

W.-D. Roth

237 RF-test generator

a 'mini test generator' for the 2 metre, 70 cm and 23 cm wave bands

This straightforward little circuit will be an extremely useful tool for high frequency enthusiasts. It is a kind of 'harmonic generator' that can be modulated and which will produce test signals in 9 MHz steps up to the giga hertz range. It can be used both for FM and SSB receivers and is a fairly inexpensive circuit to build.

At one time or another, Ham radio operators who built their own sets are going to need a generator for receiver alignment. A commercially available test transmitter would, of course, be ideal, but they tend to be rather expensive and rather over-sophisticated. In nine cases out of ten a much simpler device will do the job, provided it produces a reliable, stable test signal within the required frequency range.

There is, however, one snag: an absolutely stable generator with an output frequency that is continuously adjustable is almost impossible to obtain. This plus the fact that we are looking for an economic alternative meant that another solution had to be found. A crystal generator was therefore designed that was capable of producing a wide frequency range without having to be tuned. The secret is for it not to be a 'clean' oscillator, but one with an output signal that contains many harmonics. Even though it includes an ordinary transistor, the oscillator produces

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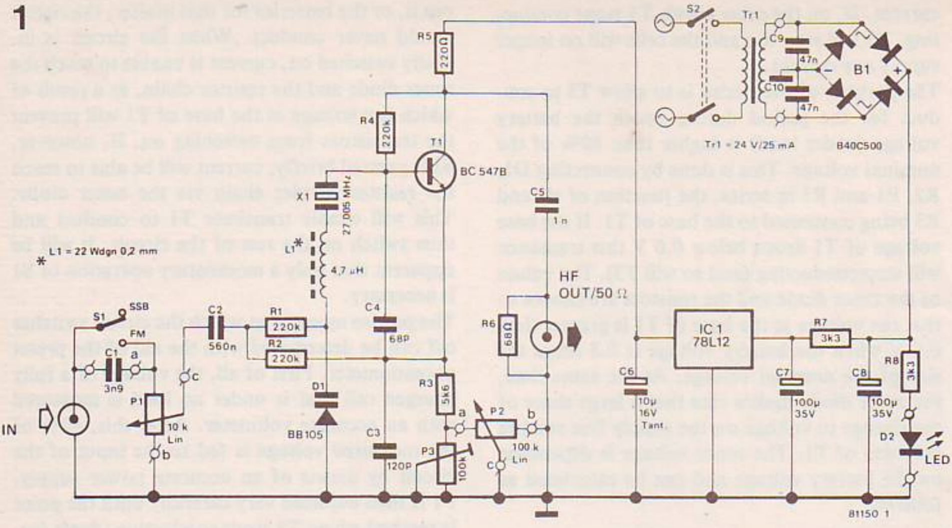

powerful harmonics of gigahertz proportions in addition to its 9 MHz fundamental frequency! This means that the test generator could also be used for reception and transmission on VHF and UHF. The generator's third harmonics are on the 27 MHz wave band (CB), its 16th harmonics

Figure 1. The test generator circuit diagram. It is straightforward and purely functional.

Figure 2. The printed circuit board for the 'mini test generator' is very compact.

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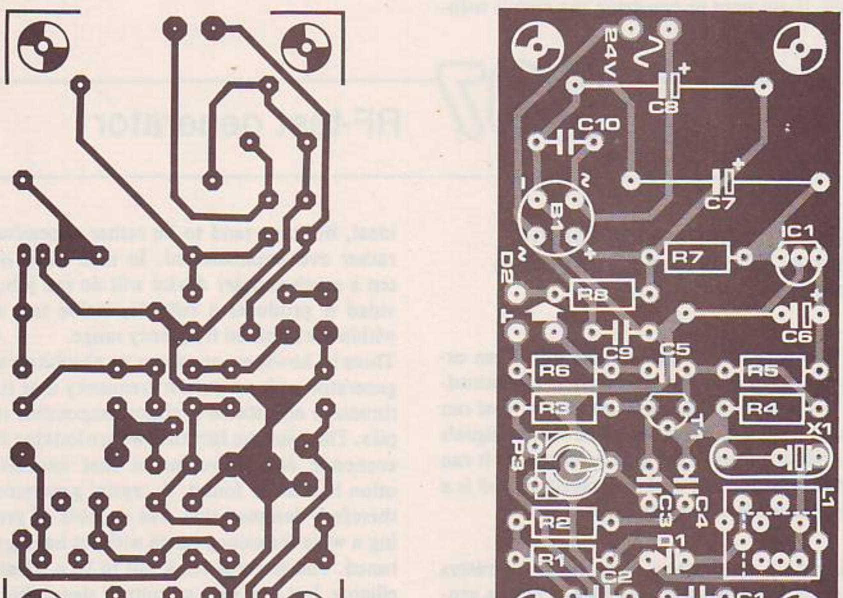


are at 144.08 MHz (2 metre wave band), its 48th harmonics are at 432.24 MHz (70 cm wave band) and its 144th harmonics are at 1296.72 MHz (on the 23 cm wave band).
The circuit is also ideal for testing speech processors.

The circuit

The remarkably simple circuit diagram is shown in figure 1. Around T1 there is a colpitt-like oscillator using a 27 MHz crystal. This does not make use of the third overtone of the crystal but rather the fundamental frequency, this being 9 MHz. This happens to be a very favourable frequency for our present purpose, since its harmonics extend over a range that is very practical for radio amateurs.

Parts list

Resistors:
R1, R2, R4 = 220 k
R3 = 5k6
R5 = 220 Ω
R6 = 68 Ω
R7, R8 = 3k3
P1 = 10 k linear
P2 = 100 k linear
P3 = 100 k preset

Capacitors:
C1 = 3n9
C2 = 560 n
C3 = 120 p
C4 = 68 p
C5 = 1 n (cer.)
C6 = 10 µ /16 V (tantalum)
C7, C8 = 100 µ /35 V
C9, C10 = 47 n

Semiconductors:
T1 = BC 547B
D1 = BB 105
D2 = LED
IC1 = 78L12
B1 = B40L500 round version

Miscellaneous:
X1 = 27.005 MHz crystal
L1 = 4.7 µ H coil (see text)
Tr1 = 24 V/25 mA transformer
S1 = SPDT switch
S2 = DPDT switch

When a crystal is used at the fundamental there is always a considerable difference between the theoretical and the measured frequency. The required frequency (here: 9005.000 kHz) must therefore be tuned precisely with coil L1. With the aid of the varicap diode D1 the oscillator can be frequency modulated. The usable modulation level (presettable with P1) is not particularly high, but high enough to test narrow band FM amateur and other special band receivers.
SSB receivers can be 'whistled through' with the generator. In order to obtain intelligible modulation level for such receivers, the frequency modulation (FM) should merely be converted into a phase modulation (PM). This can be done quite simply by connecting a small capacitor (C1) in series with the modulation input – thus, S1 can now switch between FM and SSB.
In most test generators a separate attenuator is used to measure a receiver's behaviour at very low signal levels. In this particular case this was found to be superfluous since the oscillator continued to be reliable even when barely operating. It is therefore quite a straightforward matter to built an attenuator by making the emitter resistor belonging to T1 adjustable. Pots P2 and P3 have a fairly wide range: at a frequency of 144.08 MHz (2 m wave band) the maximum output signal is around 1 mV and a minimum of around 30 nV (or 0.03 µ V)!

Construction

Obviously, building the board (figure 2) is a simple matter. Even the coil L1 should be no problem; just 22 turns of enamelled 0.2 mm copper wire wound around a Kaschke core, type K3/70/10. If readers happen to dislike this chore, an adjustable 4.7 µ H inductor coil obtainable from Toko will also be suitable.
With the exception of the mains transformer, the simple power supply shown in figure 1 is included on the printed circuit board. Since the circuit consumes very little current (and therefore the transformer can be quite small) the test generator and its power supply can be a highly compact instrument. When putting it into a case, be sure to provide a metal screen between the mains transformer and coil L1, otherwise there will be a lot of hum – a type of modulation that is not always to be desired!

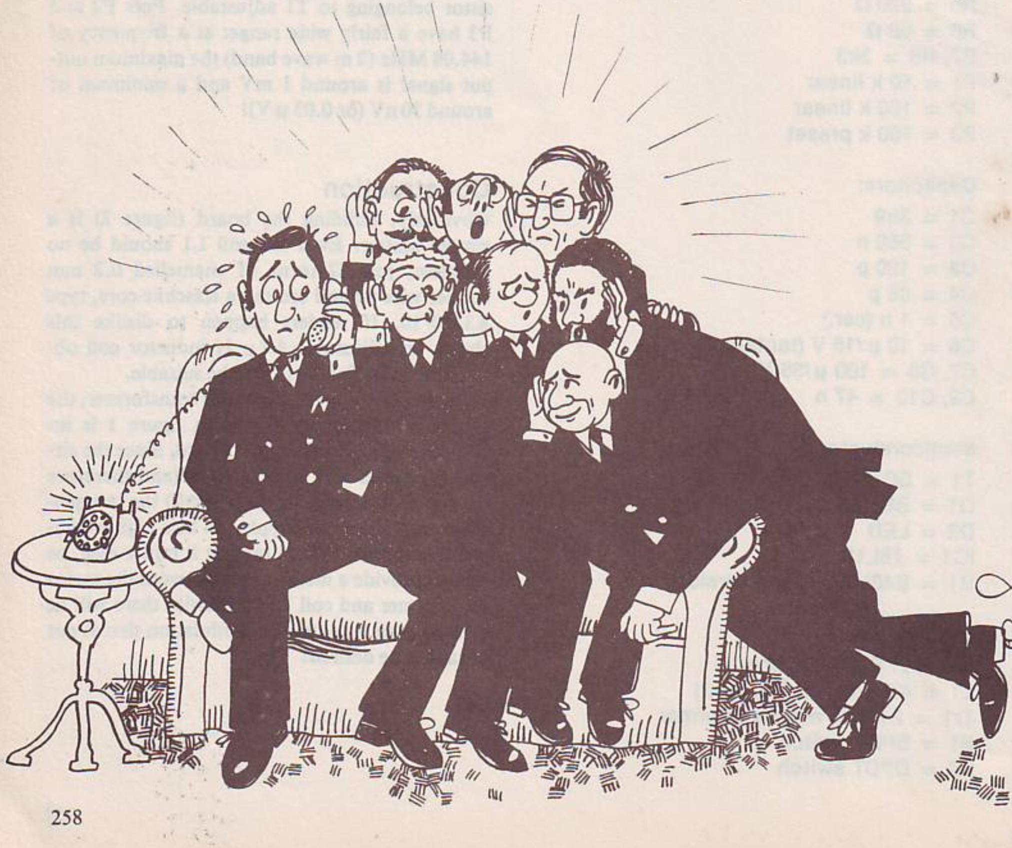
238 Telephone amplifier

makes distant callers loud and clear

'Keeping in touch' is easier said than done, despite the modern telephone networks that stretch to the four corners of the globe. For one thing, a pound for a minute seems a lot of money to hear Granny's faint voice ten thousand miles away and then not understand a word she's saying. The solution is in the form of an amplifier which, when connected to the telephone, enables the whole family to listen in to the conversation.

Some callers, of course, don't need amplifying, as anyone blessed with an old aunt who bellows hearty greetings down one's ear at eight o'clock on a Sunday morning will agree. Here an attenuator would be more appropriate! But then that is an exception. Distant and sometimes even local lines can be very poor indeed, so that an amplifier is really practical. For instance, when relatives ring up from South Africa, say, or Australia, it would be much more economical if the whole family could listen instead of having to 'queue up' to say a few costly words. What's more, the amplifier drowns any interference caused by crossed lines and thousands of 'clicking' relays, so that the once distant voice sounds as loud and clear as if the person were sitting in the same room.

Now that we know what the amplifier is for, we can study the circuit diagram in figure 1. Looking at the drawing from left to right, the circuit starts with a pick-up coil, the centre contains an amplifier and at the other end there is the loudspeaker. The pick-up coil operates according to magnetic principles: any alteration in the mag-



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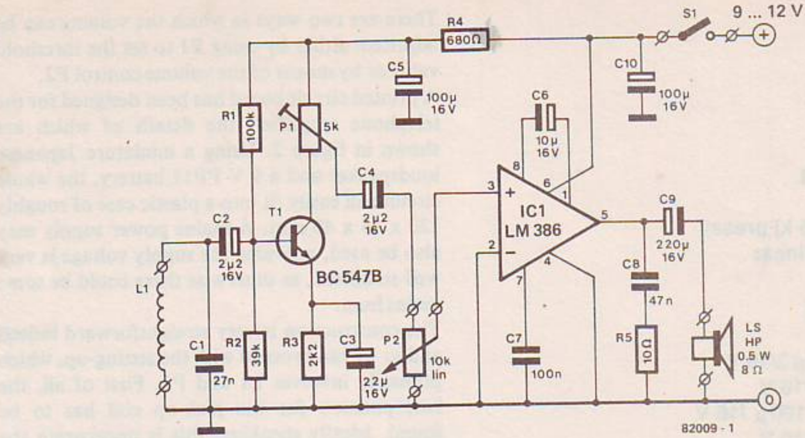

netic field that is radiated by wires in the telephone set or in the receiver will be fed to the amplifier. This slightly roundabout system is necessary, since a direct electrical connection to the interior of a telephone is forbidden.
The rest of the circuit diagram in figure 1 comprises very few components. L1 represents the telephone pick-up coil which is specifically designed for this type of application. A very low AC voltage is induced across the coil and this is amplified by transistor T1 and the amplifier IC1 and then fed to the loudspeaker.

Figure 1. The circuit diagram of the telephone amplifier.

Figure 2. The printed circuit board track pattern and component layout for the telephone amplifier.

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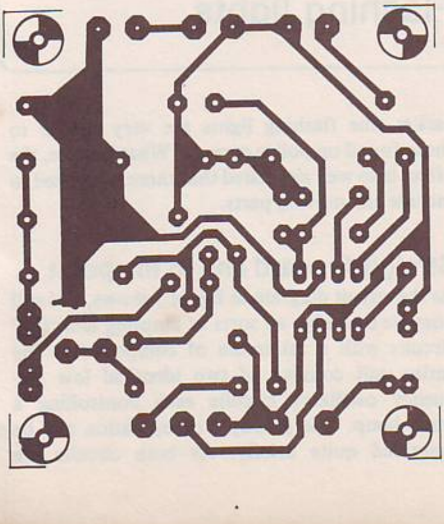
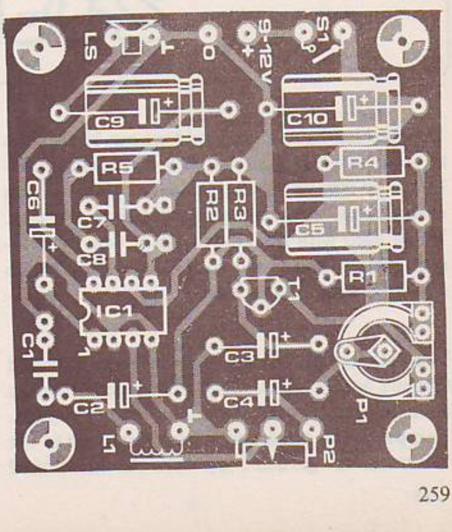




Parts list

Resistors:

R1 = 100 k
R2 = 39 k
R3 = 2k2
R4 = 680 Ω
R5 = 10 Ω
P1 = 4k7 (5 k) preset
P2 = 10 k linear

Capacitors:

C1 = 27 n
C2, C4 = 2μ2/16 V
C3 = 22μ /16 V
C5, C10 = 100μ /16 V
C6 = 10μ /16 V
C7 = 100 n
C8 = 47 n
C9 = 220μ /16 V

Semiconductors:

T1 = BC 547B
IC1 = LM 386

Miscellaneous:

L1 = telephone pick-up coil
LS = 8Ω /½ W miniature loudspeaker
S1 = on/off switch

There are two ways in which the volume can be adjusted: either by using P1 to set the threshold value or by means of the volume control P2.
A printed circuit board has been designed for the telephone amplifier, the details of which are shown in figure 2. Using a miniature Japanese loudspeaker and a 9 V PP11 battery, the whole circuit will easily fit into a plastic case of roughly 120 x 65 x 40 mm. A mains power supply may also be used, provided the supply voltage is very well stabilised, as otherwise there could be some mains hum.
The construction is very straightforward indeed and so we can proceed with the setting-up, which primarily involves L1 and P1. First of all, the best position for the pick-up coil has to be found. Ideally speaking, this is underneath the telephone, but this would mean having to raise the 'phone a little, since the coil is about 3 centimeters high. Another solution is to fit L1 onto the side of the telephone so that it is close to the amplifier. Readers should decide for themselves what the best practical solution is.
Now for the preset P1. This adjusts the maximum volume. Above a certain level, the sound reaching the amplifier input will be so loud that acoustic feedback ('howl round') will occur. This is a kind of echo that has got out of hand and produces a high-pitched tone. After setting P2 to maximum, P1 is adjusted so that this just does not occur. It would of course be feasible to omit all the components to the right of P2 and use HiFi equipment to reproduce the caller's voice, but then, that is up to the reader.

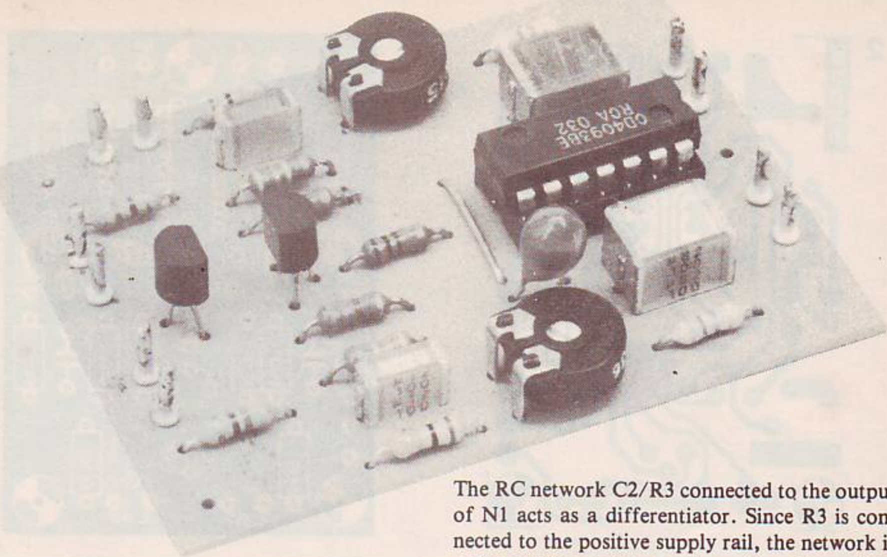
239 Flashing lights

The flashing lights described here can be fitted to an inexpensive (plastic?) toy car to provide an effect very similar to the warning lights seen on ambulances, fire engines and police vehicles. When used with the Hi-Fi siren it will, at minimal cost, add new dimensions to a toy, which any child will find fascinating.

Toy cars are always appreciated and, provided they are not too small, can usually accommodate a small circuit board and a couple of batteries. This particular circuit adds a special touch to the 'common or garden' toy car. As mentioned earlier, the flashing lights are very similar to those found on police cars etc. What is more, the effect is so well simulated that there is no need to include any moving parts.

Straightforward and to the point

As the circuit diagram in figure 1 shows, it is still possible to design all sorts of amusing and 'fun' circuits with a minimum of components. The entire unit consists of two identical low frequency oscillator circuits each controlling a small lamp. The principle of operation can be described quite briefly. As both circuits are



identical, only one need be described. The oscillator (astable multivibrator) is constructed around the Schmitt trigger N1. Capacitor C1 is connected between the inputs of the gate and ground. The output of N1 is fed back to the input via resistor R1 and potentiometer P1. The capacitor is either charged or discharged by way of these resistors, depending on the logic level at the output of N1. Whenever the voltage across the capacitor reaches one of the trigger levels, the output of the gate 'toggles'. Thus, the multivibrator produces a squarewave output signal, the frequency of which is determined by the relationship between the capacitor value and the total resistance of R1 and P1. The frequency can be altered by adjusting P1.

The RC network C2/R3 connected to the output of N1 acts as a differentiator. Since R3 is connected to the positive supply rail, the network is only sensitive to the negative-going edges of the squarewave signal. These short 'spikes' are then converted to usable pulses by gate N2 to drive the darlington transistor T1. In turn, this transistor switches the lamp connected to its collector on for a short period of time. A resistor (R5) has been included across the emitter and collector of the transistor to ensure that the lamp remains at the correct temperature. This has the advantage that the initial current through the lamp is much

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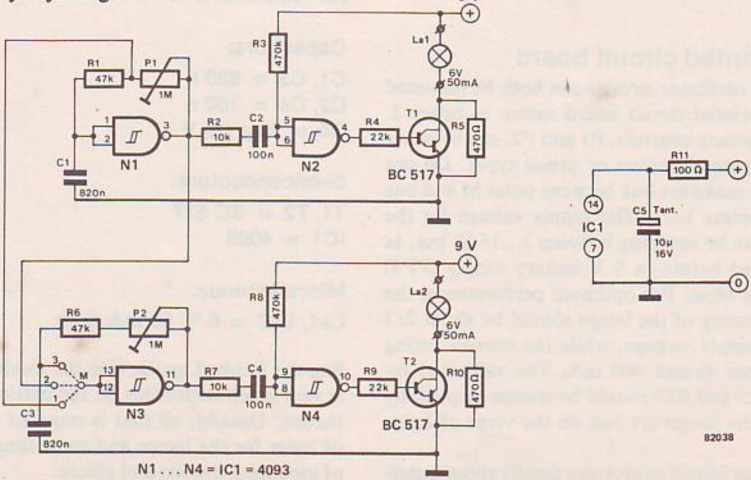

Figure 1. The circuit consists of two identical multivibrators. Depending on the effect required, there are three methods of linking the two circuits.

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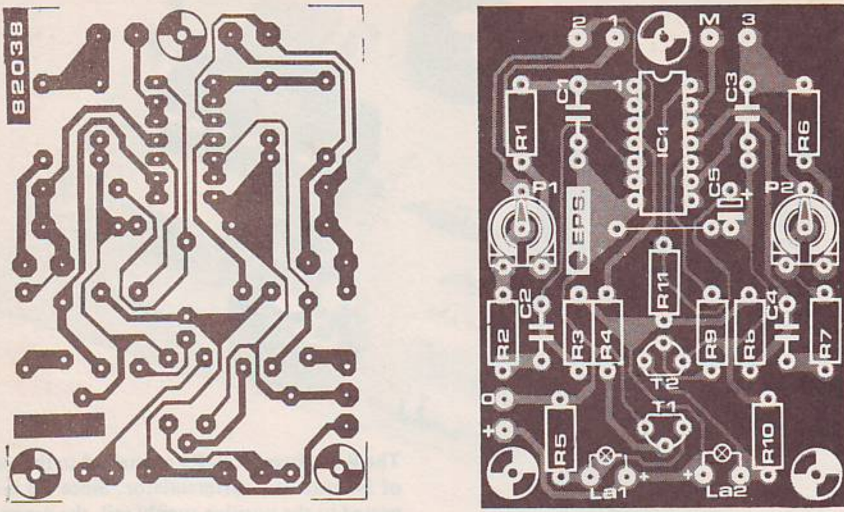


less than normal and therefore the lamp will have a much longer life span. To make the lamp light up brightly, a 6 V type can be used with a (recommended) supply voltage of 9 V.

The only difference between the first and second sections of the circuit is the fact that the second one can be 'programmed' to perform in one of three different ways. This is accomplished with the aid of a wire link on the board. By linking points 3 and M, two completely independent flashing lights are obtained. By linking points 2 and M, the lamps light alternately. The frequency can then be adjusted by means of P1. Finally, if points 1 and M are linked, the two lamps will light simultaneously. Again, the frequency is determined by means of P1.

The printed circuit board

The two oscillator circuits can both be mounted on the printed circuit board shown in figure 2. The frequency controls, P1 and P2, can be either normal potentiometers or preset types. Do not forget to make the link between point M and one of the points 1...3. The supply voltage for the circuit can be anything between 3...15 V, but, as mentioned before, a 9 V battery supply (PP3) would be ideal. For optimum performance, the voltage rating of the lamps should be about 2/3 of the supply voltage, while the current rating should not exceed 400 mA. The values of resistors R5 and R10 should be chosen empirically so that the lamps are just on the verge of lighting.

We do not intend to give any details about installing the finished article into the model car. This is very much dependent on the particular model chosen. Usually, all that is required is a couple of holes for the lamps and some simple method of mounting the bits and pieces.

Figure 2. The printed circuit board and component overlay for the flashing lights circuit. The unit is so small that it can be built into a toy car quite easily.

Parts list:

Resistors:

R1, R6 = 47 k
R2, R7 = 10 k
R3, R8 = 470 k
R4, R9 = 22 k
R5, R10 = 470 Ω (see text)
R11 = 100 Ω
P1, P2 = 1 M preset
All resistors ⅛ W

Capacitors:

C1, C3 = 820 n
C2, C4 = 100 n
C5 = 10 μ /16 V

Semiconductors:

T1, T2 = BC 517
IC1 = 4093

Miscellaneous:

La1, La2 = 6 V, 50 mA bulb

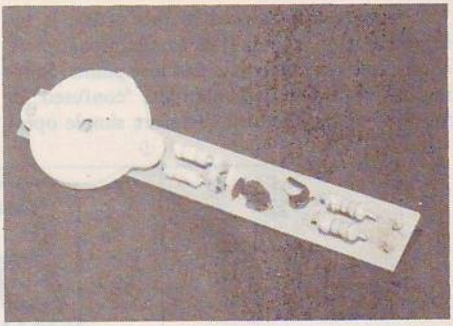
240 Continuity tester

It is possible that after etching a printed circuit there is a break or short circuit in the copper pattern. The chance of this sort of fault occuring increases as the tracks and the insulation between them become narrower. The method of manufacture of printed circuits used by the amateur do not always allow very accurate detailing of the copper layout. A detailed inspection of the results is therefore necessary. This can, of course, be done using a normal resistance meter (a multimeter), but this has the disadvantage that you must keep one eye on the meter. An audible indication can make the testing much quicker and easier leaving both eyes free to check suspect tracks with two test prods. The continuity tester gives a tone when there is a connection, and is silent when there is an open circuit.

The circuit diagram (figure 1) shows that the tester is a very simple design consisting only of a two transistor astable multivibrator. When the two test points are connected, the two transistors conduct alternately causing a square wave voltage to appear across the buzzer, at a frequency of a few kHz. The tone produced by the buzzer indicates the connection.

The circuit operates from a supply of only 1.5 V, and draws no more than 1 mA. Because of the small load, even the smallest 1.5 V battery will have a long life.

The continuity tester is built on the printed circuit board shown in figure 2. This includes space for the buzzer as well. The complete circuit and battery can be fitted into a plastic tube, so that the tester fits easily into the hand.



Parts list

Resistors:
R1, R2 = 2k2
R3, R4 = 470 k

Capacitors:
C1, C2 = 470 p

Semiconductors:
T1, T2 = BC 547B

Miscellaneous:
BZ = buzzer PB-2720 (Toko)
2 test probes
1.5 Volt battery

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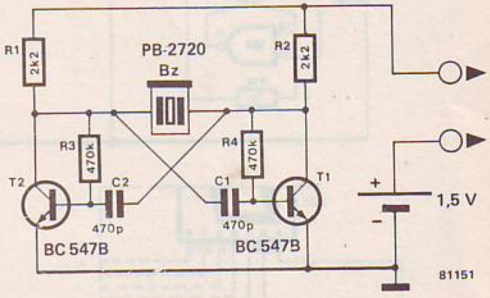

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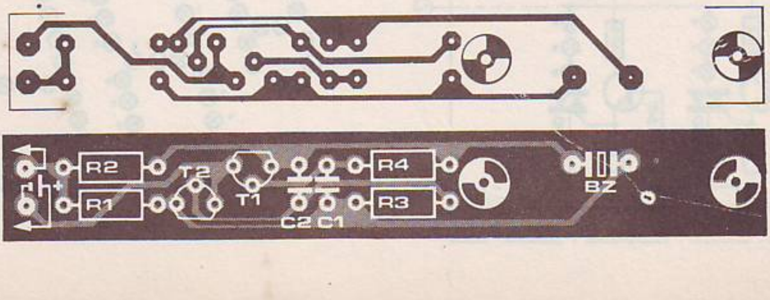
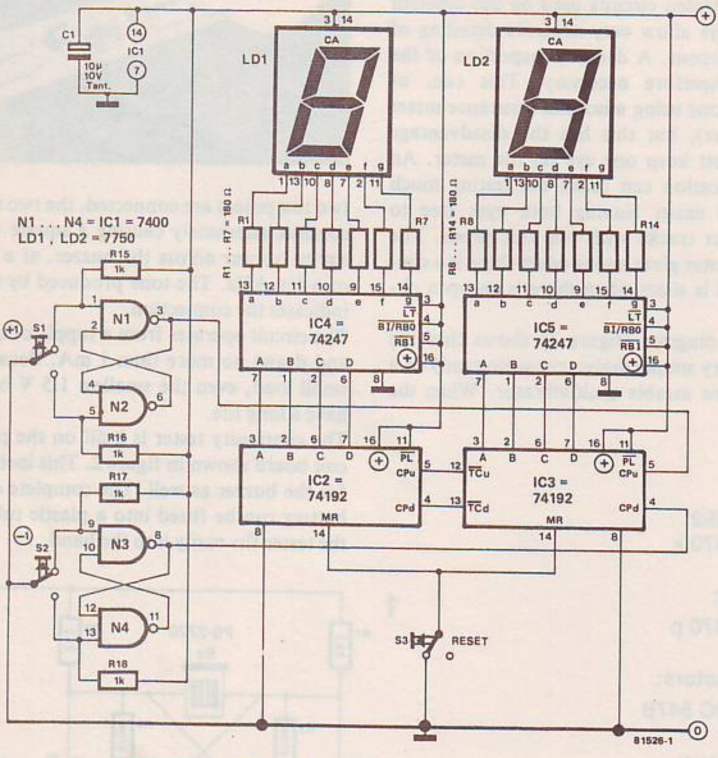
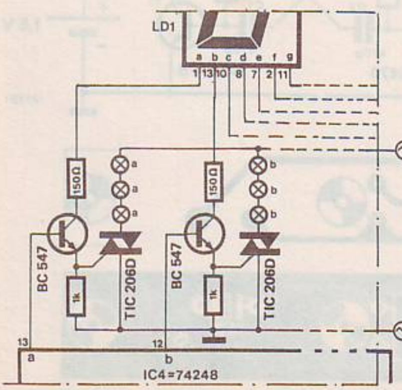


241 Scoreboard

This scoreboard featured in this article is intended for use in quiz type competitions where competitors can both gain and lose points. Scoring can sometimes become quite 'confused' in the heat of the moment therefore simple operation is essential. In this design a point is awarded or deducted by the process of pushing one of two buttons – one push, one point. When for instance the scorer has already awarded points and the referee reverses the de-



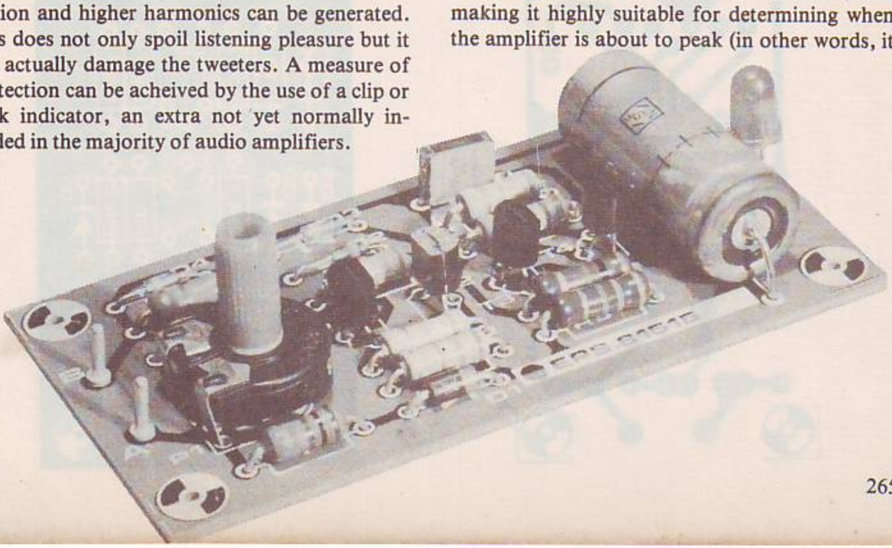


cision the correction can be carried out easily.
The circuit of the scoreboard is shown in figure 1. The counter IC chosen is the well known 74192 decade counter. This has two clock inputs, one for counting up and the other for counting down. The count (or clock) pulses are created by either of the two flipflops formed by gates N1, N2 or N3, N4 which are triggered by switch S1 or S2. The two counters are connected in series to provide a maximum count of 99.
The 74192 presents the information at its output in BCD (Binary Coded Decimal) format and therefore some form of decoding is necessary for the 7 segment displays. For this purpose the 74247 (an updated version of the 7447) BCD to 7 segment decoder/driver is used. This IC carries out all the functions required between the counters and displays which are in fact connected directly to its output via current limiting resistors. Virtually any type of common anode 7 segment display can be used. Switch S3 is included to reset the score and when pressed both displays will revert to zero.
It will be apparent that the LED displays will be altogether too small where a larger audience is concerned and for this purpose the design for a very much larger display is included to the right of figure 1. This uses 240 V bulbs and should be bright enough to be visible from a few hundred yards away. The complete display can be as large as required by carefully spacing the bulbs. The wiring must be as shown in the illustration. Each segment of each display requires a triac and a driver transistor. The bulbs are 15 or 25 Watt and can be obtained in various colours for a more 'professional' appearance. The triac used must have a turn on gate current of 5 mA.
If the mains version is built a 74248 must be used for both IC4 and IC5. The LED display can still be retained for use by the scorekeeper in case he is unable to see the larger one.
LS devices can be used to replace all of the TTL ICs but the two types cannot be mixed. The supply current for LS ICs will be about 350 mA while TTL will require up to 450 mA.
WARNING: Readers who have no wish to renew their aquaintance with the physical properties of 250 V AC when connected with their person should take extra care when constructing the mains display version.

242 Loudspeaker peak indicator

Nowadays, any decent loudspeaker unit is, fortunately, pretty resistant to rough treatment. However, problems can arise in the living room when the volume is turned up high enough for clipping to occur. At that point substantial distortion and higher harmonics can be generated. This does not only spoil listening pleasure but it can actually damage the tweeters. A measure of protection can be acheived by the use of a clip or peak indicator, an extra not yet normally included in the majority of audio amplifiers.

The peak indicator described here can be connected directly to the output of the amplifier or even fitted into the speaker since a separate power supply is not required.
The circuit will respond even to very short peaks making it highly suitable for determining when the amplifier is about to peak (in other words, it



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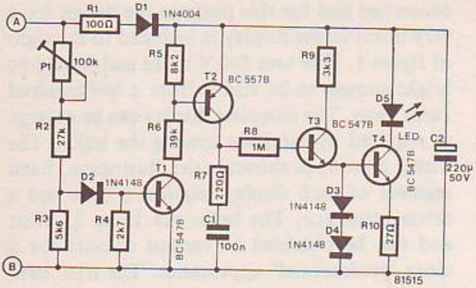

is not just an overload indicator). The peak power level at which the circuit is expected to respond (that is, the peak voltage) is adjustable between 15 and 125 Watts with an 8 ohm speaker (14...45 V). The circuit will light a LED when the amplifier just delivers its peak power enabling the listener to actually see when things begin to go wrong. If the LED only occasionally lights everything is fine. When the LED begins to light continuously then it is time to turn the volume down a little.

The circuit diagram for the indicator is shown in figure 1. Its power supply is derived from capacitor C1 which is charged via R1 and D1 from the speaker output of the amplifier. Half-wave rectification was considered suitable since 'normal' 45 V transistors can be used.

With no signal input all transistors are switched off and therefore current drain from C2 is virtually nil. When the input signal level exceeds a certain value (dependent on the setting of P1), the voltage at the junction of R2 and R3 will reach a point at which T1 will start to conduct.

Components required:

Resistors:

R1 = 100 Ω
R2 = 27 k
R3 = 5k6
R4 = 2k7
R5 = 8k2
R6 = 39 k
R7 = 220 Ω
R8 = 1 M
R9 = 3k3
R10 = 27 Ω
P1 = 100 k adjustable potentiometer

Capacitors:

C1 = 100 n
C2 = 220 µ F/50 V

Semiconductors:

D1 = 1N4004
D2, D3, D4 = 1N4148
D5 = LED
T1, T3, T4 = BC 547B
T2 = BC 557B

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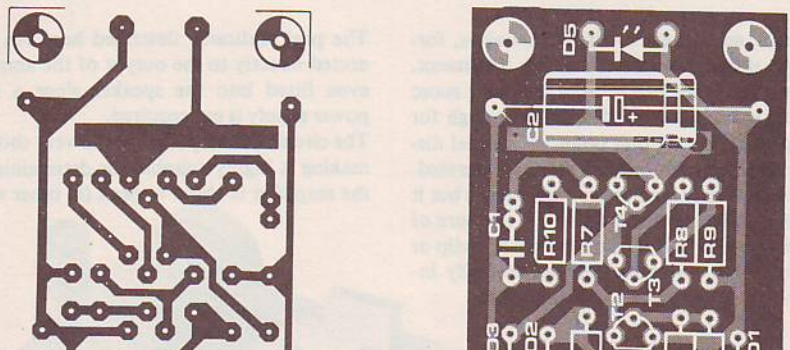

This switches on T2 causing C1 to charge rapidly. Resistor R7 has been included to prevent the maximum permitted collector current of T2 from being exceeded. Both transistors T3 and T4 will now conduct and LED D5 will light. The current through the LED will be maintained at 20 mA by C2, independent of the speaker signal level. When the input voltage then drops below the preset level, T1 and T2 will switch off. However, the LED will remain lit for a few moments longer while C1 discharges via R7 and R8.
Construction should not present any problems if the printed circuit board shown in figure 2 is used. It would probably be advisable to use the larger type of LED for maximum 'visibility'. Calibration is carried out in the following manner. If the peak power of the amplifier is known, its peak voltage can be calculated with the formula:

$$V_{peak} = 2 \times P_{peak} \times R_{speaker}$$

Connect the indicator circuit to a stabilised power supply (positive to point A), and set the DC supply level to the calculated value. P1 should then be turned back until the LED just begins to light. During this operation care should be taken to ensure that the LED does not remain lit for too long because it may cause the dissipation limit of T4 to be exceeded.
Once the clipping level has been set, the circuit may be connected to one of the speaker outputs of the amplifier or, if desired, to one of the speakers. It may be possible to modify the circuit to operate a relay that rings a bell... or fires a cannon perhaps?

243 6 Watt stereo amplifier for a car radio

The TDA 2004 from SGS-Ates contains two balanced class B power amplifiers. The IC was designed especially for use as an in-car stereo amplifier, and for this reason it is housed in a strong package and protected against all kinds of overload. For example, output short-circuits or disconnection of the loudspeaker, overheating of the chip, peaking of the power supply or even briefly reversing the polarity of the supply connections are unable to destroy the device.
With the component values shown and with a supply voltage of 14.4 V (a fully charged car bat-

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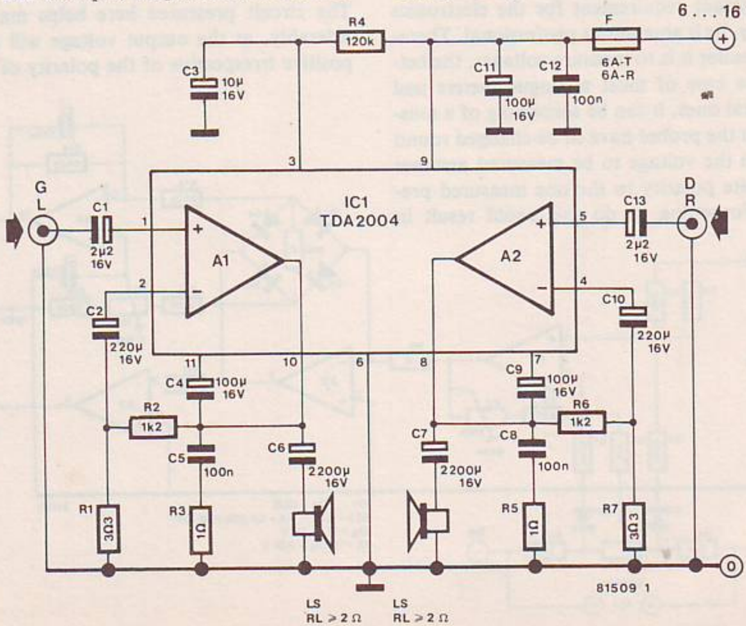

tery), the stereo amplifier is capable of delivering a power output of at least 6 W, typically 6.5 W with a load impedance (RL) of 4 Ω. It can also handle a load impedance of 2 Ω, in which case the output power is a minimum of 9 W, but typically 10 W. Power outputs of this order are subject to about 10% distortion, however, if lower power outputs are acceptable, 4 W with a load impedance of 4 Ω or 6 W with a load impedance of 2 Ω, distortion is only in the order of 0.3%.
The voltage gain of the left-hand channel is determined by the ratio of R2 to R1, and that of the right-hand channel by the ratio of R6 to R7. With the values given, this will be 50 dB. Therefore, a signal of about 50 mV is required at the input to give the maximum output. If this input sensitivity is too great, a 50 kΩ stereo potentiometer can be included at the input. The impedance of the non-inverting amplifier input is minimally 100 kΩ.
The network consisting of resistor R3 and capacitor C5 (and R5/C8) is included to prevent

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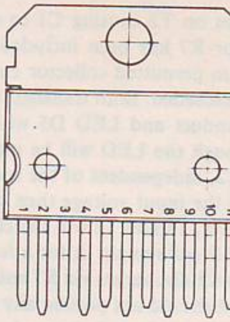
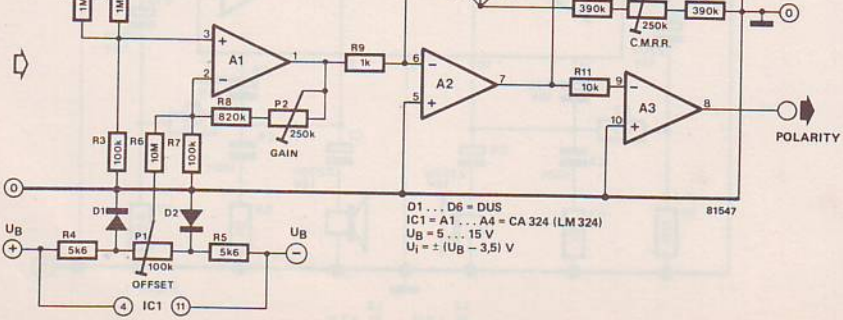


the amplifier oscillating at high input frequencies. The bandwidth of the circuit is more than adequate for use as a car radio amplifier. The frequency response of the amplifier is 40 Hz to 16 kHz (3 dB points).
Obviously, the IC must be kept sufficiently cool. However, the well thought-out design makes it a very simple task to mount the device on an adequate heatsink. The thermal resistance of the heatsink should be at least 4°C/W.

244 Polarity converter

Analogue or digital voltmeters (or both!) are a very important requirement for the electronics laboratory, be it amateur or professional. Therefore, the easier it is to measure voltages, the better. In the case of most analogue meters and some digital ones, it can be something of a nuisance when the probes have to be changed round every time the voltage to be measured assumes the opposite polarity to the one measured previously. Forgetting to do this could result in rather disastrous consequences!!
The circuit presented here helps matters considerably, as the output voltage will always be positive irrespective of the polarity of the input



voltage. The circuit also has a 'polarity' output which will produce an output of $+U_B$ when the input voltage is positive and $-U_B$ when the voltage to be measured is negative. Provided the circuit is calibrated correctly, the overall accuracy is guaranteed to the better than 0.5% of the maximum input voltage (U_i).

The calibration procedure for the circuit is as follows. Resistors R12 and R13 are disconnected from R10 and linked to each other. A level of +1 V is then applied between the junction of R12 and R13 and ground (0 V). The output voltage (A4) is then adjusted to a minimum level by means of preset potentiometer P3. This is called the common mode rejection ratio (CMRR) preset. The polarity of the 1 V test voltage is then reversed (−1 V). A certain voltage (several millivolts) will now be measured at the output and P3 is adjusted once more to reduce this level to about half.

The above procedure is repeated by alternately reversing the polarity of the test voltage and adjusting P3 until the measured output voltage is the same in both cases (at +1 V and at −1 V). The CMRR will then be set to its maximum level. (The low output voltage is due to the offset of A4 and can not be completely eliminated.)

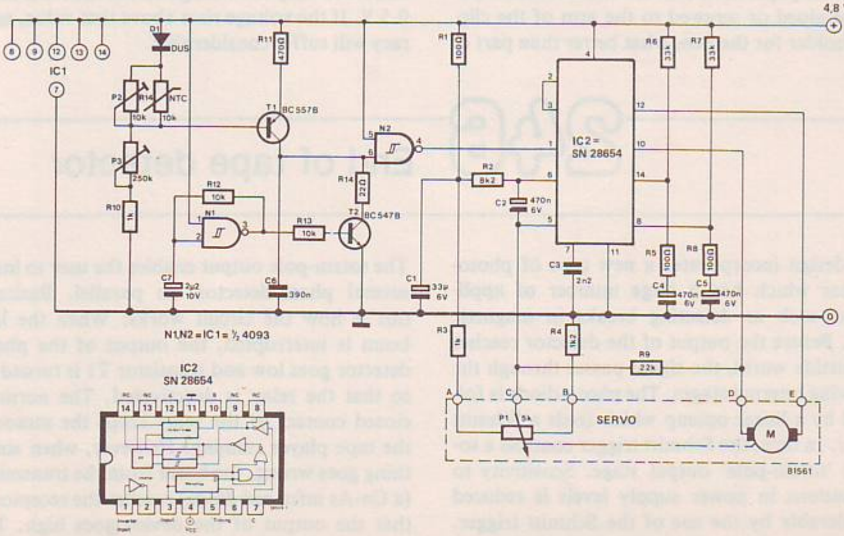
The next step is to connect resistors R12 and R13 as shown in the circuit diagram. The input is then short circuited and the overall offset of the circuit can be reduced to a minimum by means of preset potentiometer P1.

Once this has been accomplished, a known input voltage of, say, +1 V, is applied to the input and the gain of the unit is adjusted by means of P2 so that the output voltage is equal to the input voltage. The circuit will now be correctly calibrated and ready for use.

Last, but by no means least, the circuit should be provided with a stable power supply. This is because fluctuation in supply voltage level would mean having to calibrate the unit all over again.

245 Temperature recorder

This circuit, together with a certain amount of mechanical ingenuity, makes it possible to construct a relatively inexpensive piece of equipment which can be used to record a temperature curve. An ordinary radio control servo is used to operate the pen. It uses a negative temperature coefficient (NTC) resistor as the sensor. The circuitry around N1, N2, T1 and T2 forms an oscillator whose pulse width is determined by the instantaneous value of the NTC resistor. The resultant signal is fed directly to IC2. This IC (the SN 28654) is specifically designed as a servo



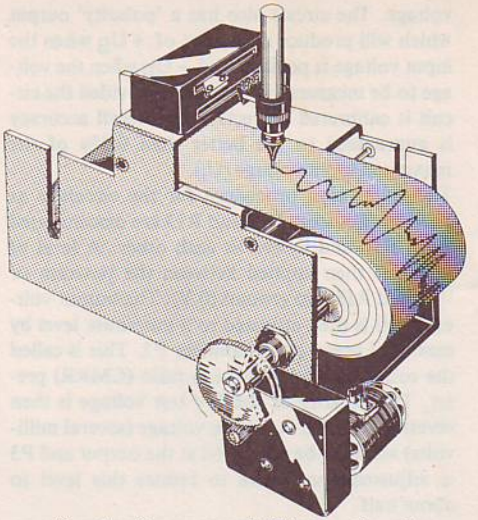
amplifier, which is abundantly clear from its specifications:

- an output current of 400 mA with no external transistors
- change of direction is accomplished with a single supply voltage
- the 'dead space' – the degree of input change required before a change in the output occurs – is dependant on the value of C3
- a maximum dissipation of around 800 mW.

The pulse width modulated signal is fed to pin 1 of IC2. The control output for the servo appears at pins 10 and 12 of the IC. Readers who would like to know more about this particular device than we have room for here are advised to obtain the data sheet from the manufacturers or from one of their distributors.

The non-linear course of the resistance value constitutes a bit of a problem when using NTC devices as temperature sensors. This can be solved, however, by utilising only a small portion of the temperature characteristic. This is accomplished here by the 'sensitivity' potentiometers P2 and P3 which effectively set the amount of deflection of servo per degree of temperature change and the lower limit of the range respectively. This achieves a reasonable degree of accuracy, but of course, we are not trying to construct a piece of laboratory equipment. In reality, the circuit is only intended to record a change in temperature over a period of time rather than measure actual temperatures.

The mechanical parts can be constructed quite simply. The servo can be mounted on a clip over the roll of paper. The holder for the recorder pen can be glued or screwed to the arm of the clip. As a holder for the pen, what better than part of



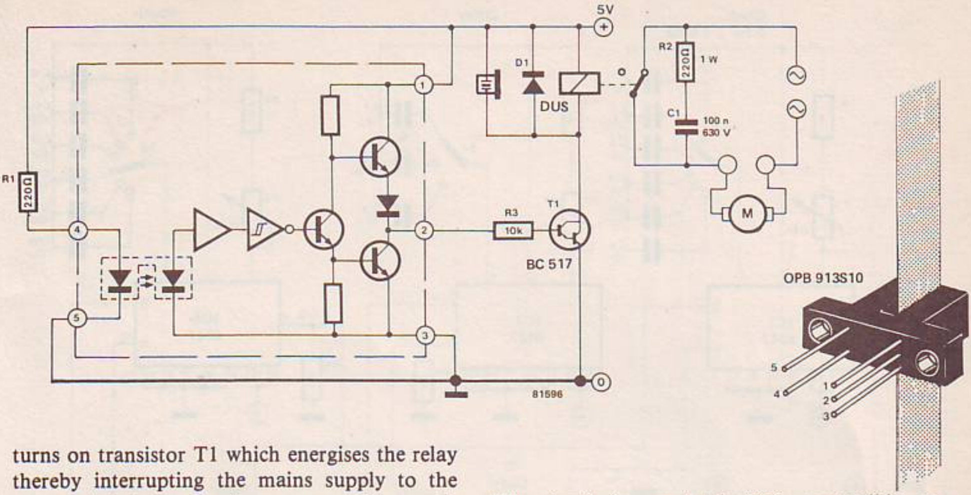
an old pair of compasses? This has the advantage that the pen can be easily removed for cleaning or replacement. The paper can be the type used in printing calculators, but the roll should move at a slow and constant speed when the equipment is in operation. The paper drive can be made with the aid of motors and gears that are normally used by model boat builders, available from any good model shop. Failing this, geared motors, for a variety of voltages, are readily available from surplus stores such as Proops and J. Bull.

To prevent the NTC resistor from heating up on its own, care should be taken when adjusting P2 and P3 to keep the voltage across the NTC below 0.5 V. If the voltage rises above that value, accuracy will suffer considerably.

246 End of tape detector

This design incorporates a new type of photodetector which has a large number of applications such as detecting breaks in magnetic tapes. Before the output of the detector reaches the outside world, the signal passes through the following internal stages. The photo-diode is followed by a linear opamp which feeds a Schmitt trigger. In turn, the Schmitt trigger controls a so-called 'totem-pole' output stage. Sensitivity to fluctuations in power supply levels is reduced considerably by the use of the Schmitt trigger.

The totem-pole output enables the user to install several photo-detectors in parallel. Basically, this is how the circuit works: When the light beam is interrupted, the output of the photodetector goes low and transistor T1 is turned off so that the relay is deactivated. The normally closed contact of the relay keeps the motor of the tape player running. However, when something goes wrong, the beam from the transmitter (a Ga-As infra-red diode) reaches the receptor so that the output of the device goes high. This

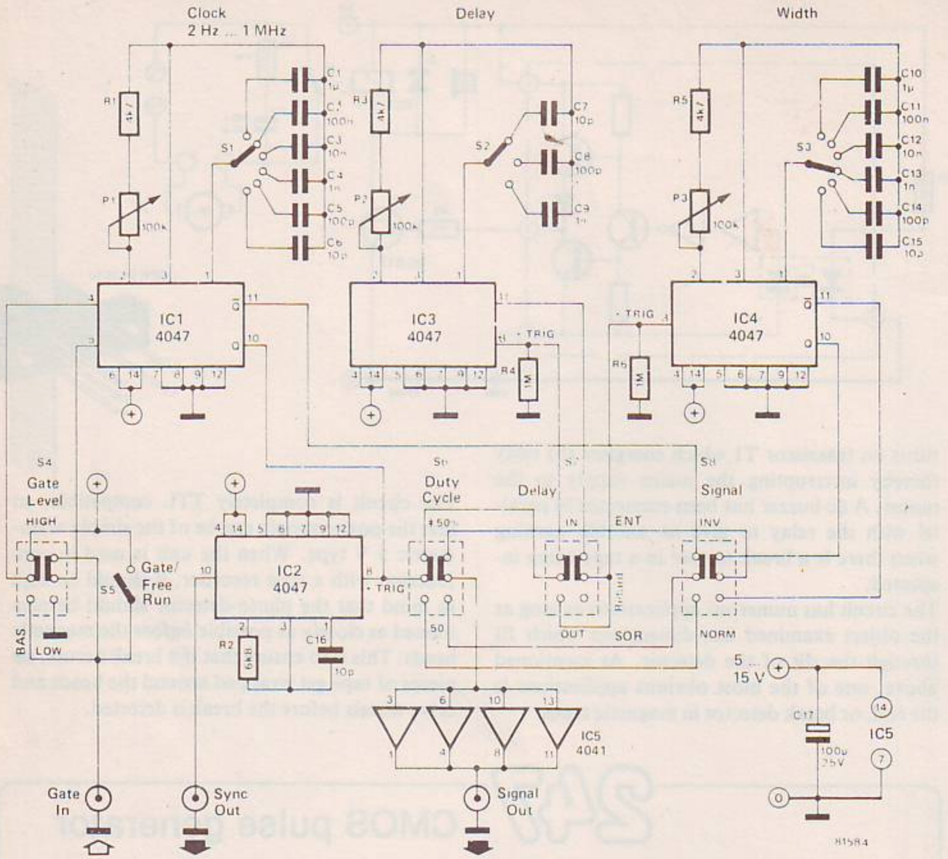


turns on transistor T1 which energises the relay thereby interrupting the mains supply to the motor. A dc buzzer has been connected in parallel with the relay to give an audible warning when there is a break or tear in a tape being inspected.
The circuit has numerous applications as long as the object examined has dimensions which fit through the slit of the detector. As mentioned above, one of the most obvious applications is the end, or break detector in magnetic tapes.

The circuit is completely TTL compatible, so that the power supply can be of the simple asymmetric 5 V type. When the unit is used in conjunction with a tape recorder, it should be kept in mind that the photo-detector should be positioned as closely as possible *before* the magnetic heads. This is to ensure that if a break occurs, no pieces of tape get wrapped around the heads and drive wheels before the break is detected.

247 CMOS pulse generator

A pulse generator can be extremely useful when designing digital circuits. To make the most of its possibilities it must be as flexible as possible. The clock frequency must be variable over a fairly wide range and of course the pulse width must be variable also. An automatic output level control would be a major advantage. All these and a few other features are combined in the circuit given here.
The use of CMOS ICs throughout has two advantages. In the first place it is possible to power the circuit from batteries. Furtnermore, the large supply voltage range afforded by CMOS, from 5 to 15 V, makes it possible to provide the automatic output control mentioned above. This is clear from the fact that if the pulse generator itself is powered from the circuit under test, the supply voltage will equal that of the circuit and the output logic levels must therefore be compatible whether it is CMOS or TTL (output buffers are included). Furthermore, the low current requirement of the generator input results in a very low current drain on the circuit being tested.
The description of the circuit begins at the clock generator, IC1. This IC is wired as an astable multivibrator and its frequency is adjustable between 2 Hz and 1 MHz (depending on supply voltage) by potentiometer P1 and switch S1. With S5 closed and S4 in the 'high' position, IC1 will run continuously. With S5 opened an external signal can be used to trigger IC1 via the 'gate in' socket. Switch S4 can then be used to select the required polarity for pins 4 and 5 of IC1 from the external source.
The output signals of the clock generator appear at pins 10 and 11 of IC1. The Q output (pin 10) is passed to the trigger input of IC2. This IC is used as a pulse shaper to provide a narrow sync pulse output for external trigger purposes. The Q output of IC1 is also fed, via S6 (duty cycle



50%) and S8 (signal normal), to the output buffer stage, IC5.

ICs 3 and 4 are also wired as triggerable monostable multivibrators. With S6 in the 50% position and S7 set to delay out, the Q output of IC1 will be passed to the trigger input of IC4. Any required pulse width can now be achieved by the adjustment of both P3 and S3. This provides a variable duty cycle output at pins 10 and 11 of IC4. Depending on the position of switch S8, either the normal or the inverted signal can be passed via the buffer stage to the signal out socket.

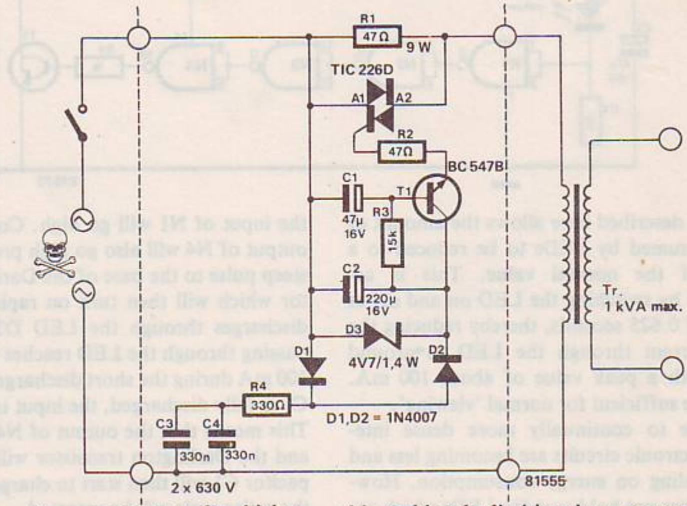
A further modification to the signal can be carried out with IC3 when S7 is switched to the 'delay in' position. IC3 will now be triggered by the clock output signal of IC1 (S6 still in the 50% position). Now, by adjusting P2 and S2, it is possible to delay the output signal from 1.5 s to 250 ms with reference to the sync out trigger pulse. The output of IC3 is now used to trigger IC4. The pulse width can still be modified as required. It should be noted that the delay circuit does not alter the output signal but varies its timing relative to the sink output. By setting the delay to a suitable value, it is possible to move the leading edge of the output signal but varies its timing relative to the sync output. By setting the delay to a suitable value, it is possible to move the leading edge of the output signal pattern to a more central position on the oscilloscope screen enabling the complete waveform to be studied.

The prototype generator was constructed using Veroboard since very little layout work is required. Virtually all of the wiring concerns the controls on the front panel. All of the range capacitors can be mounted on the switches S1, S2 and S3 if two wafers are used for each switch. Mounting resistors R1, R2 and R3 between the potentiometers and switches leaves only five components and the five ICs for the board. However, there are quite a few interconnections to be made between the board and the front

panel so care should be taken. Ribbon cable may prove to be useful for this purpose.
Power for the generator can be derived from the circuit under test or from batteries. If the latter are used, the input and output levels may not be totally compatible.

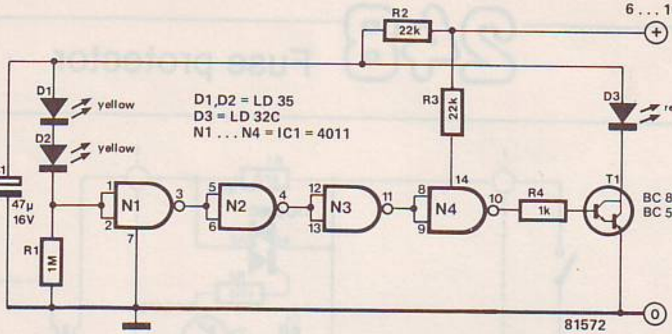
RCA application note ICAN 6230

248 Fuse protector



In recent times, the requirement for higher power hi-fi equipment has grown to such an extent that it has now become necessary to protect even the domestic house fuses from being blown too often. The solution is a 'soft start' circuit – a circuit which maintains the initial surge current to within acceptable limits.
Normal domestics fuses are rated at 13 A and many readers may express some surprise if we venture to suggest that the transformer in their equipment could conceivably draw this seemingly excessive amount of current. The short answer is yes, it can and it does! It should be realised that these large transformers can easily withstand powers of up to 1 kW in some cases.
The 'turn on' current of the transformer goes into a very low impedance, both in the primary and the secondary winding. Furthermore, the smoothing capacitor on the dc side can be so large that when it is still discharged it can virtually have a zero impedance. Effectively, therefore, the fuse in the primary side of the transformer is presented with a short-circuit and it is not at all surprising that the mains fuse does blow on occasion. The fuse protector alleviates this problem by limiting the surge current via resistor R1. Only after approximately 100 ms (two mains periods) will this resistor be shorted out by the triac. The delayed input voltage is obtained by the drive to the triac gate via transistor T1. The mains voltage is reduced by a capacitive series impedance, C3/C4, to the point that after rectification by diode D2, stabilisation by D3 and smoothing by capacitor C2, a dc voltage of 4.7 V appears across the zener diode. Transistor T1 is then turned on via capacitor C1 and remains on. This in turn drives the gate of the triac causing it to turn on and provide a short across resistor R1. The full primary current will then flow through the triac.
The circuit can be constructed with the triac type TIC 226D (as shown) with transformers with a rating of up to 1 kVA. Larger transformers will obviously require larger triacs.
The fuse protector can be used for a variety of applications such as hi-fi equipment (as mentioned), domestic appliance motors (washing machines etc.) and heavy duty lamps - especially ultra-violet and infra-red types.

249 Miser - LED



The circuit described here allows the amount of energy consumed by LEDs to be reduced to a fraction of the normal value. This is accomplished by switching the LED on and off at intervals of 0.625 seconds, thereby reducing the average current through the LED to around 200 µ A with a peak value of about 100 mA. This is quite sufficient for normal 'viewing'.

Mainly due to continually more dense integration, electronic circuits are becoming less and less demanding on energy consumption. However, this does not hold true for LEDs which are used for a variety of indicator functions. Most LEDs consume a minimum current of around 20 mA, which in many instances is several times more than that used by the rest of the circuit. This is an especially unsatisfactory situation where the equipment in question is battery-powered.

The circuit operates as follows: Capacitor C1 is charged via resistor R2. Once the potential across this capacitor is sufficient to overcome the bias presented by (yellow) LEDs D1 and D2, the input of N1 will go high. Consequently the output of N4 will also go high providing a short steep pulse to the base of the Darlington transistor which will then turn on rapidly so that C1 discharges through the LED D3. The current passing through the LED reaches a maximum of 100 mA during the short discharge period. When C1 is fully discharged, the input to N1 goes low. This means that the output of N4 also goes low and the Darlington transistor will turn off. Capacitor C1 will then start to charge up again and the entire cycle will be repeated.

If preferred, several 'ordinary' diodes in series can be used instead of the two yellow LEDs D1 and D2. As the 4011 has a very critical threshold value, it may be necessary to experiment with several different diodes in order to obtain the correct switch-over point. The IC receives its power via resistor R3 which ensures that the current to the IC is restricted to a minimum. The actual physical size of the complete unit is so small that it will cause no problems if it is to be installed into existing equipment.

250 Zero voltage indicator

This zero voltage indicator uses two LEDs to show whether the input voltage lies within a specified small voltage range, which is symmetrical about zero. If the voltage is within the range, the LEDs flash. If it is outside, one of them lights continuously. Within the specified range there is also an indication of whether the voltage is at the edge of the range, or near the centre (i.e. near to zero). At the centre of the range the LEDs flash regularly, but towards the edges they

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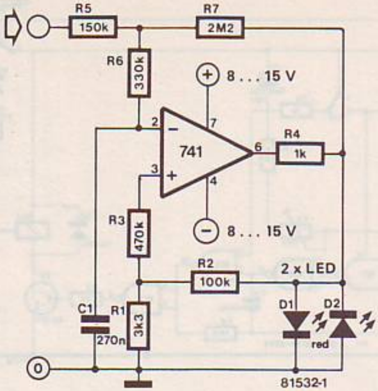


become irregular.

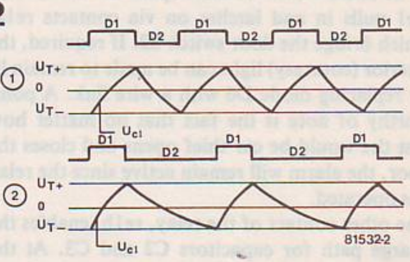
The operation of the circuit is actually quite simple, although it may not at first appear so from the diagram. If you imagine the circuit without some of the components (R3, R4, R5, D1 and D2), you have a normal opamp oscillator. However, by including the potential divider, R7, R5, we ensure that the voltage fed back to C1 is no longer equal to the supply voltage. (It is also limited by D1 and D2.) If R5 is not now connected to earth, but to a DC voltage (the input voltage), the DC level of the feedback voltage will be changed. When this level is so high that the voltage across C1 falls outside the hysteresis loop of the Schmitt trigger, the circuit stops oscillating and one of the LEDs lights continuously.

If the input voltage is exactly 0 volts, then the DC level across C1 is zero, and the LEDs will flash regularly. However, if the input is not exactly 0 volts, (e.g. slightly positive), then one LED (D2) will be on slightly longer than the other.

The sensitivity of the circuit is about 50 mV, i.e. the LEDs change from flashing to a continuous light at plus and minus 50 mV. This can be changed readily by altering the value of R7. A higher resistance increases the sensitivity ($R7_{max}$ 3M3). You must bear in mind that if R7 is reduced, C1 must be increased.

The source impedance of the voltage which is connected should not be greater than 10 k, otherwise a buffer-amp must be interposed.

2



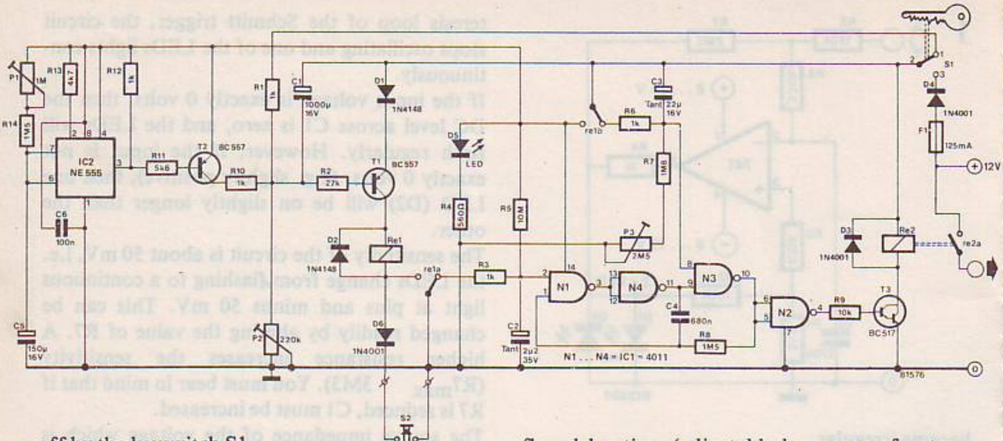
251 Auto theft alarm

Readers who are in the habit of parking their cars close to lampposts at night may presume that the vehicle is considerably safer in a well lit area. However, it could be mentioned that car thieves, as a rule, like to see what they are getting! Further precautions are, therefore, very necessary, or your car could still be found to be gone. Alarm circuits for this purpose must be:
a. reliable; b. easy to operate, and c. failsafe.

After all, if the neighbours have been woken up several times on previous occasions due to false alarms, it may not be easy to maintain good relations with them and they may therefore be less likely to inform yourself or the police when a genuine break-in occurs.

The alarm circuit described here has a number of good features. It uses very little current, it has delayed turn on and delayed alarm, it has repetitive as well as continuous alarm and it will automatically re-arm itself once it has been triggered. All this complexity means that the circuit itself needs to be fairly complex. The actual operation of the unit will become clear as we describe the circuit diagram.

When the alarm circuit is switched on, by means of the (hidden) keyswitch S1, capacitor C1 begins to charge via preset potentiometer P2. This charge time is actually the delay which allows the driver and passengers to leave the car and shut the doors. When the base/emitter voltage of T1 (in series with D1) is sufficient to turn the transistor on, the alarm is activated and is in the 'standby' state. If a door is now opened, the door switch, S2 will make and operate relay Re1. Once the alarm has been triggered, that is, a door has been opened, it can only be switched



off by the keyswitch S1.

Now to the next function performed by the circuit. Once the alarm has been armed by means of S1 and a would-be thief opens the door, relay Re1 pulls in and latches on via contacts re1a, which bridge the door switch S2. If required, the interior (courtesy) light can be made to remain lit by replacing diode D6 with a wire link. A point worthy of note is the fact that no matter how fast the would-be car thief opens and closes the door, the alarm will remain active since the relay has operated.

The other contact of the relay, re1b, enables the charge path for capacitors C2 and C3. At the same time, an indication that the alarm has been triggered is given by LED D5. The charging of capacitor C2 via resistor R5 results in the alarm delay time, in other words, the thief still does not know that an alarm is present in the vehicle. This time delay is long enough, about ten seconds, for the driver of the vehicle to enter and disarm the alarm.

Only when C2 is fully charged will the voltage level at pin 2 of N1 become logic '1'. This level is fed through gates N4, N3 and N2 to the relay driver transistor T3. Relay Re2 will therefore be activated and the alarm will sound continuously. Capacitor C3 will also start to charge at the same time as C2, but this charge time is significantly longer and is adjustable by means of preset potentiometer P3 up to approximately thirty seconds. After this period, pin 8 of N3 will go low, as does the output of N2 and the relay Re2 will drop out again.

The 555 timer, IC2, is connected as a monostable multivibrator and its purpose is to provide the repetitive feature of the alarm circuit. This IC is enabled by relay contacts re1b and is triggered at pin 2 via resistor R13. After the monoflop delay time (adjustable by means of potentiometer P1) has elapsed, transistor T2 is turned on via the output of the 555 at pin 3. Capacitor C1 then discharges through resistor R10 and transistor T1 turns off when the voltage across C1 is reduced to about 1 V. This causes relay Re1 to drop out and capacitors C2 and C3 discharge rapidly via resistors R3 and R6 respectively. At the same time the timer IC is disconnected from the supply. The alarm is now in the original active standby state.

The 4011 (IC1) containing gates N1...N4 also serves another purpose besides that already mentioned. It also functions as a squarewave generator with a frequency of 0.8 Hz. This gives an intermittent alarm signal to the vehicle's horn and/or lights operated by Re2.

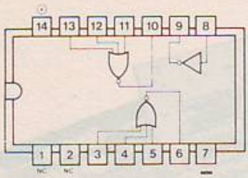
Warning: The horn relay in general use in cars usually has a very low impedance and therefore requires a relatively large amount of current. Transistor T3 must, obviously, be capable of supplying such a current. It may be preferable to incorporate a separate horn for the purposes of the alarm since certain car thieves are aware of the fact that the ordinary car horn is often used for alarms and therefore promptly disconnect it before opening the car door.

It is common sense to conceal the alarm unit and the operating switch as much as is practical for obvious reasons. The current consumption of the circuit is a mere 4 µ A in the standby condition.

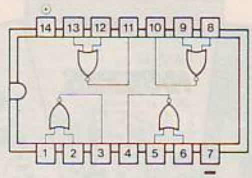
R. Rastetter

BLANK

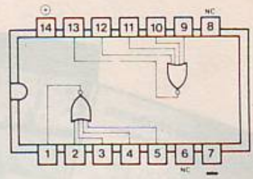
DUAL 3 INPUT NOR GATE PLUS INVERTER
4000



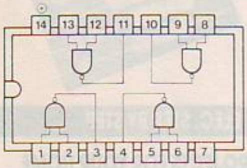
QUADRUPLE 2 INPUT NOR GATE
4001



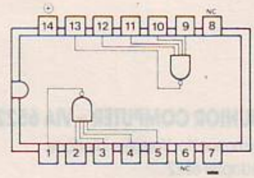
DUAL 4 INPUT NOR GATE
4002



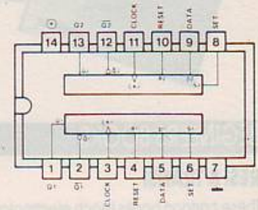
QUADRUPLE 2 INPUT NAND GATE
4011



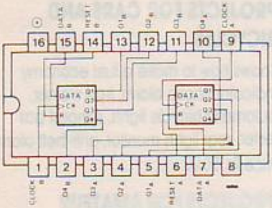
DUAL 4 INPUT NAND GATE
4012



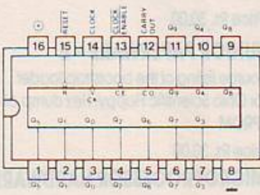
DUAL D FLIP FLOP
4013



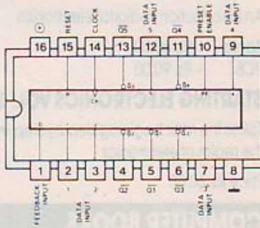
DUAL 4 BIT STATIC SHIFT REGISTER
4015



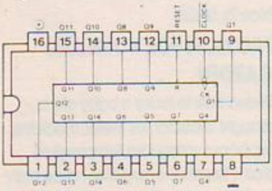
DIVIDE BY 10 SYNCHRONOUS COUNTER
4017



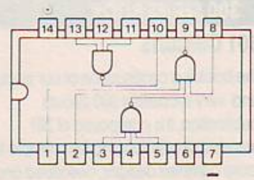
SYNCHRONOUS PRESETTABLE DIVIDE BY 'N' COUNTER
4018



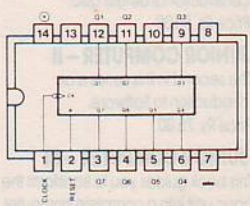
14 BIT BINARY RIPPLE COUNTER
4020



TRIPLE 3 INPUT NAND GATE
4023

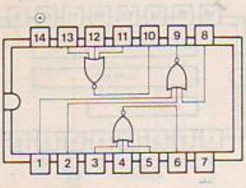


7 STAGE BINARY RIPPLE COUNTER
4024



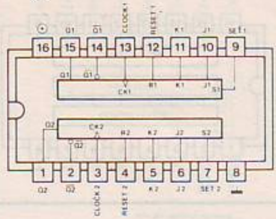
TRIPLE 3 INPUT NOR GATE

4025



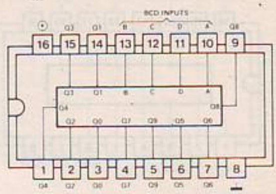
DUAL JK FLIP FLOP

4027



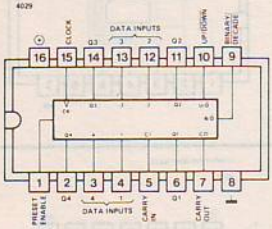
BCD TO DECIMAL DECODER

4028



SYNCHRONOUS PRESET TABLE BINARY/DECADE UP/DOWN COUNTER

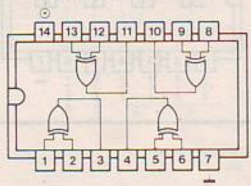
4029



QUADRUPLE 2 INPUT EXCLUSIVE OR GATES

4030

4070 low power TTL compatible (fan out - 2)

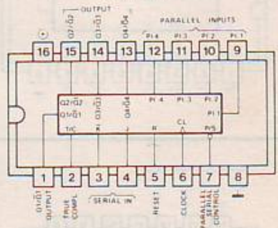


4035

4 BIT

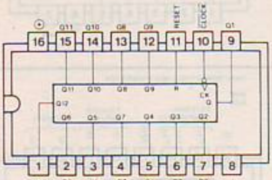
PARALLEL IN/PARALLEL OUT

SHIFT REGISTER



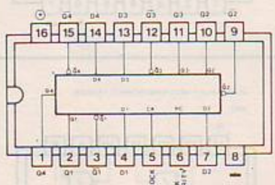
12 BIT BINARY RIPPLE COUNTER

4040



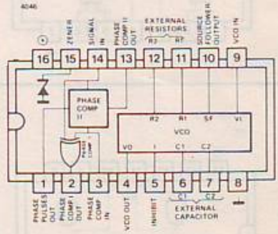
QUAD CLOCKED "D" LATCH

4042



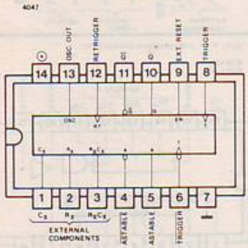
MICROPOWER PLL

4046



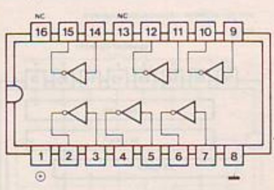
MONOSTABLE/ASTABLE MULTIVIBRATOR

4047



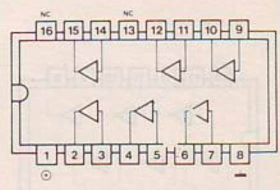
HEX INVERTING BUFFER

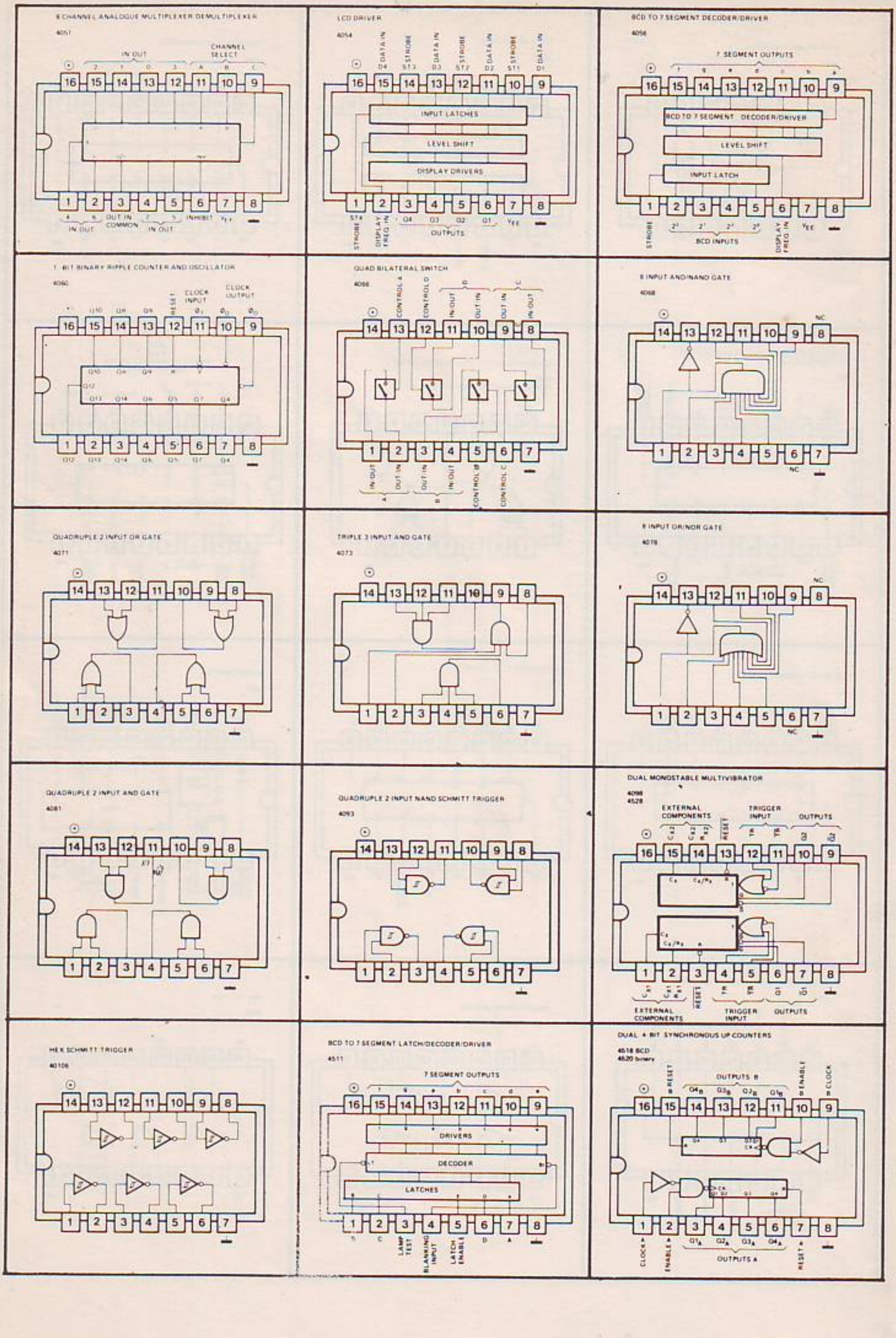
4049



HEX BUFFER

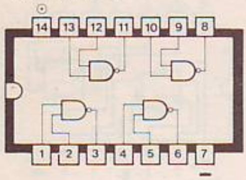
4050




8 CHANNEL ANALOGUE MULTIPLEXER DEMULTIPLEXER
4051
LCD DRIVER
4054
BCD TO 7 SEGMENT DECODER/DRIVER
4056
7 BIT BINARY RIPPLE COUNTER AND OSCILLATOR
4060
QUAD BILATERAL SWITCH
4066
8 INPUT AND/NAND GATE
4068
QUADRUPLE 2 INPUT OR GATE
4071
TRIPLE 3 INPUT AND GATE
4073
8 INPUT OR/NOR GATE
4078
QUADRUPLE 2 INPUT AND GATE
4081
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER
4093
DUAL MONOSTABLE MULTIVIBRATOR
4098
4528
HEX SCHMITT TRIGGER
40106
BCD TO 7 SEGMENT LATCH/DECODER/DRIVER
4511
DUAL 4 BIT SYNCHRONOUS UP COUNTERS
4518 BCD
4520 binary

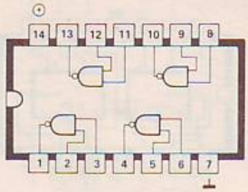
QUADRUPLE 2 INPUT NAND GATES

7400
7403 open collector outputs
7437 power driver (fan out · 30)



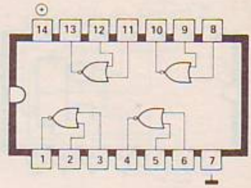
QUADRUPLE 2 INPUT NAND GATE WITH OPEN COLLECTOR OUTPUT

7401



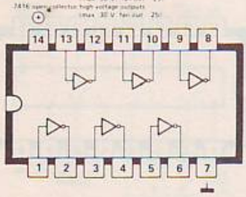
QUADRUPLE 2 INPUT NOR GATES

7402
7428 power driver (fan out · 30)



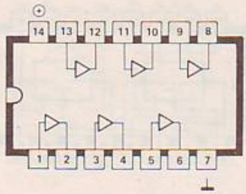
HEX INVERTERS

7404
7405 open collector outputs
7406 open collector high voltage outputs (max. 30 V · fan out · 25)
7416 open collector high voltage outputs (max. 30 V · fan out · 25)



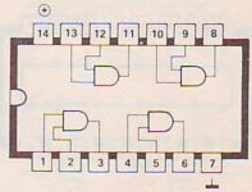
HEX BUFFER/DRIVER WITH OPEN COLLECTOR HIGH VOLTAGE OUTPUTS (max. 30 V · fan out · 25)

7407



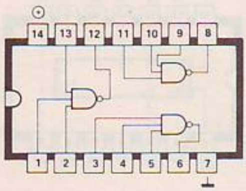
QUADRUPLE 2 INPUT AND GATES

7408
7409 open collector outputs



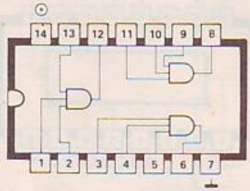
TRIPLE 3 INPUT NAND GATES

7410
7412 open collector outputs



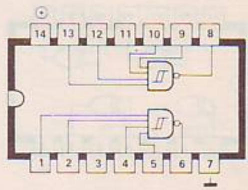
TRIPLE 3 INPUT AND GATE

7411



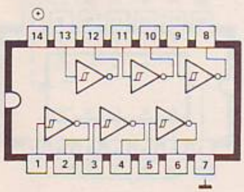
DUAL 4 INPUT NAND SCHMITT TRIGGER

7413



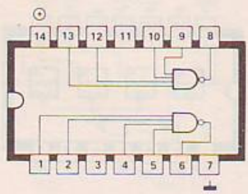
HEX SCHMITT TRIGGER INVERTER

7414



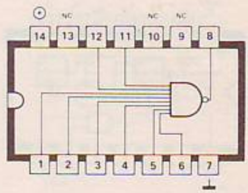
DUAL 4 INPUT NAND GATES

7420
7440 power driver (fan out · 30)



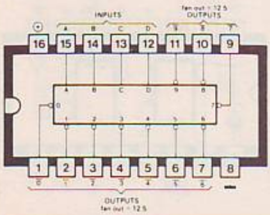
8 INPUT NAND GATE

7430



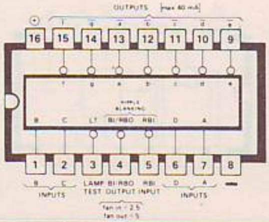
BCD TO DECIMAL DECODER/DRIVER WITH OPEN COLLECTOR OUTPUTS (max. 30 V)

7445



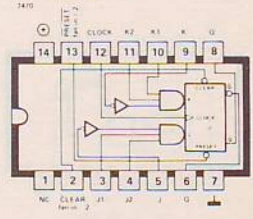
BCD TO 7 SEGMENT DECODER/DRIVER

7447



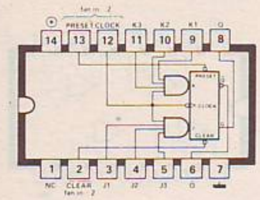
AND GATED J K POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7470



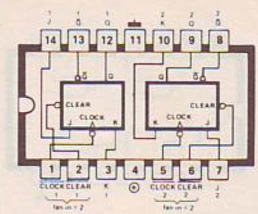
AND GATED J-K FLIP FLOP WITH PRESET AND CLEAR

7472



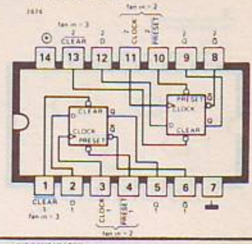
DUAL J-K FLIP FLOP WITH CLEAR

7473



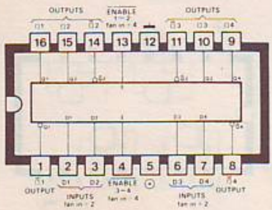
DUAL D TYPE POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7474



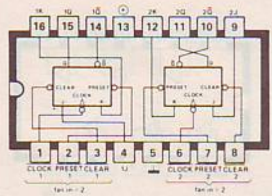
4 BIT BISTABLE LATCH

7475



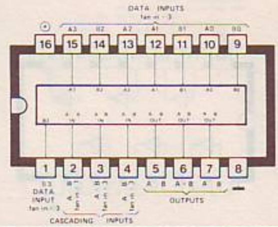
DUAL J-K MASTER SLAVE FLIP FLOP WITH PRESET AND CLEAR

7476



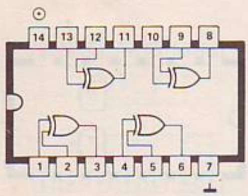
4 BIT COMPARATOR

7485



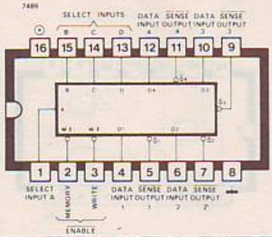
QUADRUPLE 2 INPUT EXCLUSIVE OR GATE

7486



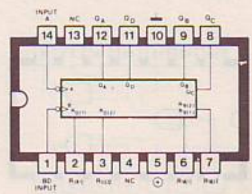
64 BIT READ/WRITE MEMORY (fan out = 7.5)

7489



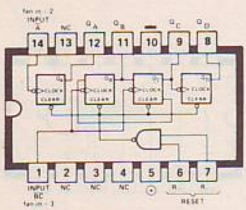
DECADE COUNTER

7490



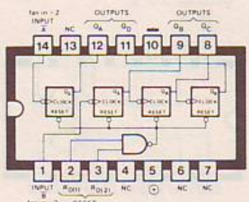
DIVIDE BY TWELVE COUNTER (÷ 2 and ÷ 6)

7492



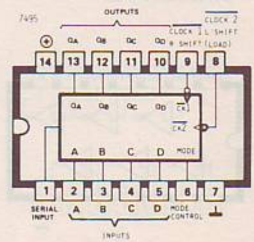
4 BIT BINARY COUNTER

7493



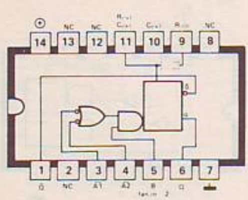
4-BIT PARALLEL-IN PARALLEL-OUT SHIFT REGISTER

7495



MONOSTABLE MULTIVIBRATOR

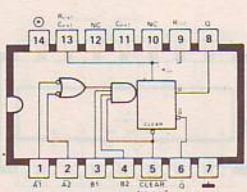
74121



121 R_{int} = 2k NOM
L121 R_{int} = 4k NOM

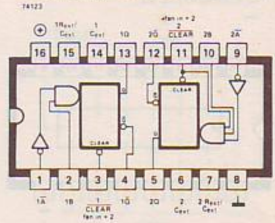
RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74122



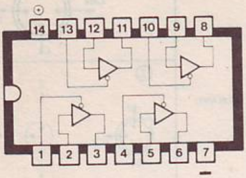
DUAL RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74123



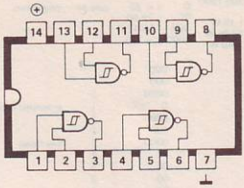
QUAD BUFFER (3 STATE)

74125



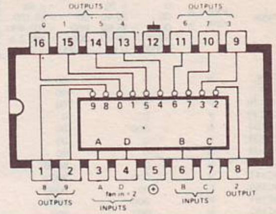
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER

74132



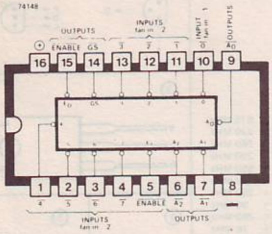
BCD TO DECIMAL DECODER DRIVER
(outputs max 60 V max 7 mA)

74141



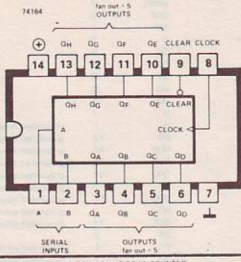
PRIORITY ENCODER

74148



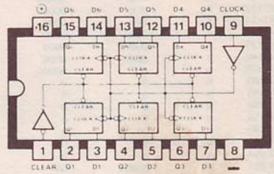
8 BIT SERIAL IN PARALLEL OUT SHIFT REGISTER

74164



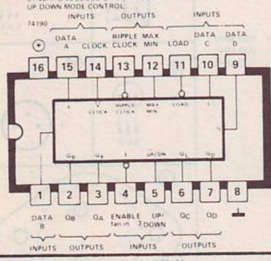
HEX D FLIP FLOP WITH CLEAR

74174



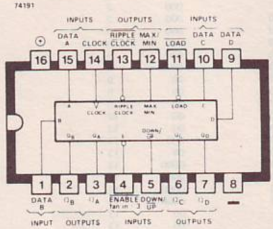
SYNCHRONOUS BCD UP DOWN COUNTER WITH UP DOWN MODE CONTROL

74190



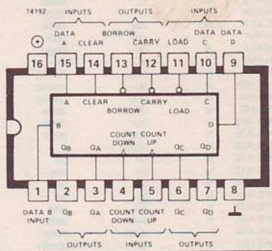
SYNCHRONOUS 4 BIT BINARY UP/DOWN COUNTER

74191



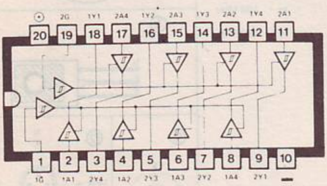
SYNCHRONOUS UP DOWN DECADE COUNTER

74192



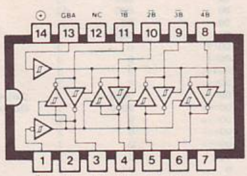
OCTAL BUFFER AND LINE DRIVER (3 STATE)

74LS241



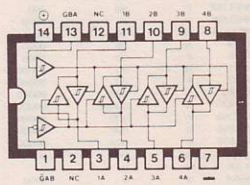
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS242



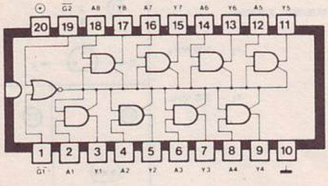
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS243



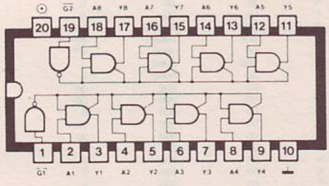
OCTAL BUFFER (3 STATE)

81LS95

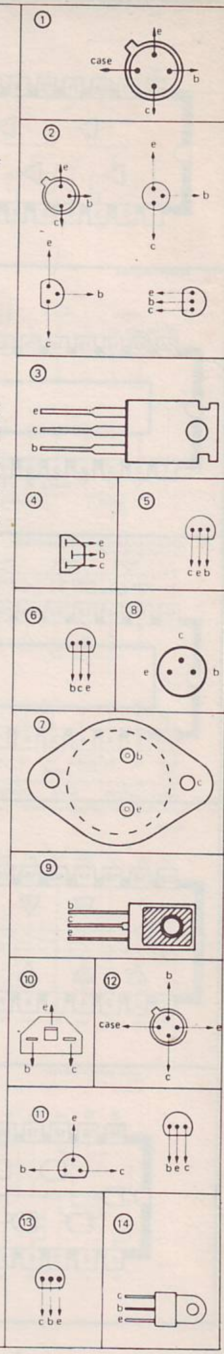


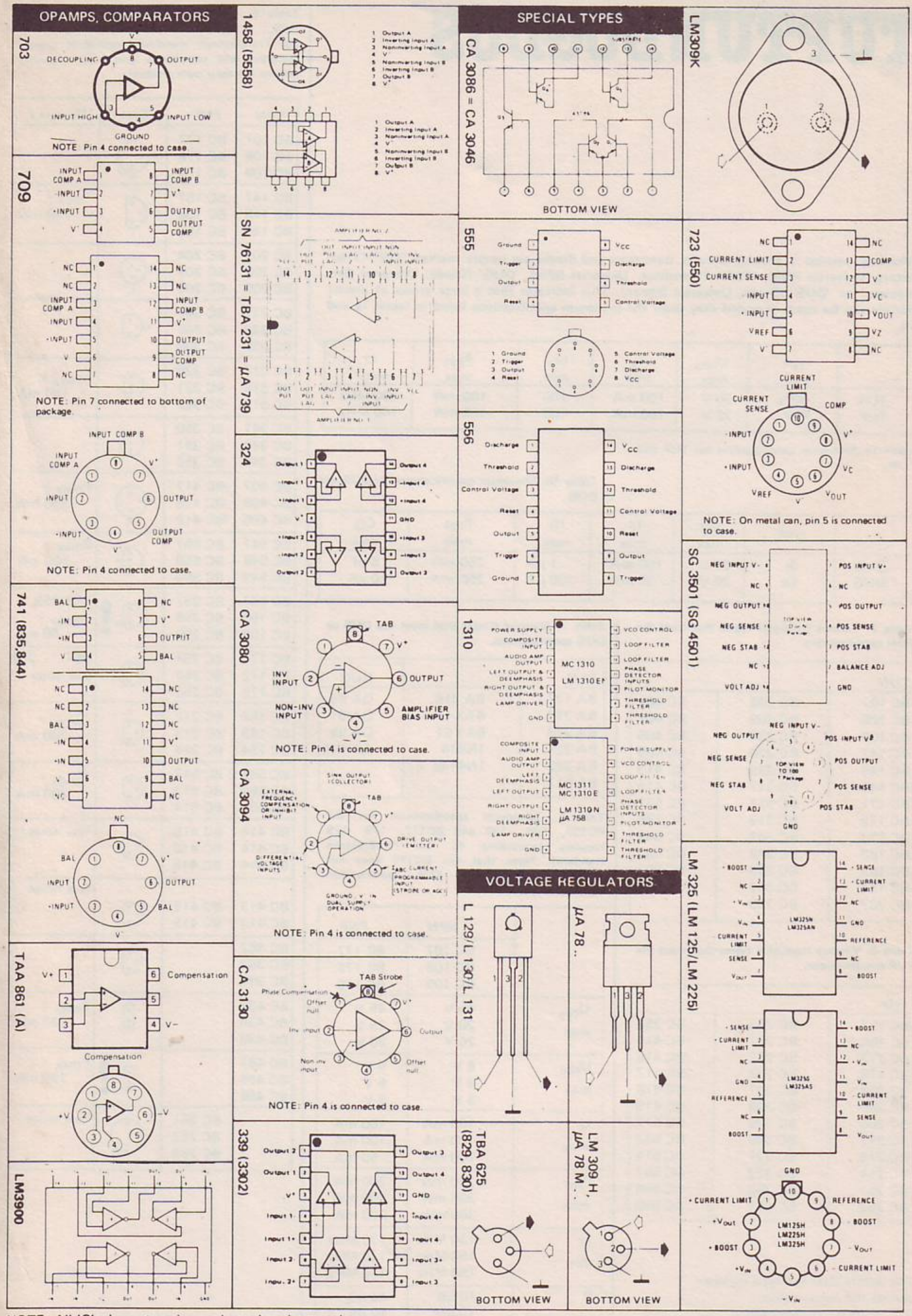
OCTAL BUFFER (3 STATE)

81LS97



Type	PNP = P NPN = N	U_{CEO} (Volt) 0 = ≤ 20 00 = 25-40 000 = 45-60 0000 = 65-80 00000 = ≥ 85	$I_{c(max)}$ (mA) 0 = ≤ 50 00 = 55-100 000 = 105-400 0000 = 405-2 A 00000 = ≥ 2 A	P_{max} (mW) not cooled: 0 = ≤ 300 00 = 305-1000 cooled: 00☆ = 1-10 W 00☆☆ = 10-35 W 00☆☆☆ = ≥ 40 W	$h_{FE(min)}$ 0 = ≤ 20 00 = 25-50 000 = 55-120 0000 = ≥ 125	case nr.	comments
TUN	N	0	00	0	000		
TUP	P	0	00	0	000		
AC126	P	0	00	00	0000	2	
AF239	P	0	0	0	0	1	grounded base: f f_T = 700 MHz
BC107	N	000	00	0	000	2	
BC108	N	0	00	0	000	2	
BC109	N	0	00	0	0000	2	low noise
BC140	N	00	0000	00☆	00	2	
BC141	N	000	0000	00☆	00	2	
BC160	P	00	0000	00☆	00	2	
BC161	P	000	0000	00☆	00	2	
BC182	N	000	000	0	0000	2	
BC212	P	000	000	0	000	2	
BC546	N	0000	00	00	0000	2	
BC556	P	0000	00	00	000	2	
BD106	N	00	00000	00☆☆	00	7	
BD130	N	000	00000	00☆☆☆	0	7	
BD132	P	000	00000	00☆☆	00	9	
BD137	N	000	0000	00☆	00	9	
BD138	P	000	0000	00☆	00	9	
BD139	N	0000	0000	00☆	00	9	
BD140	P	0000	0000	00☆	00	9	
BDY20	N	000	00000	00☆☆☆	0	7	
BF180	N	0	0	0	0	1	grounded base: f_T = 675 MHz
BF185	N	0	0	0	00	12	grounded base: f_T = 220 MHz
BF194	N	0	0	0	000	10	grounded emitter: f_T = 260 MHz
BF195	N	0	0	0	000	10	grounded emitter: f_T = 200 MHz
BF199	N	00	0	00	000	11	grounded emitter: f_T = 550 MHz
BF200	N	0	0	0	00	1	grounded base: f_T = 240 MHz
BF254	N	00	0	0	000	11	grounded emitter: f_T = 260 MHz
BF257	P	00000	00	00	00	2	grounded emitter: f_T = 90 MHz
BF494	N	0	0	0	000	11	grounded emitter: f_T = 260 MHz
BFX34	N	000	00000	00	00	2	f_T = 70 MHz
BFX89	N	0	0	0	00	1	grounded emitter: f_T = 1000 MHz
BFY90	N	0	0	0	00	1	grounded emitter: f_T = 1000 MHz
BSX19	N	0	0000	0	000	2	
BSX20	N	0	0000	0	000	2	
BSX61	N	000	0000	00	000	2	
HEP51	P	00	0000	00	000	1	f_T = 150 MHz
HEP53	N	00	0000	00	000	1	f_T = 200 MHz
HEP56	N	0	00	00	000	5	f_T = 750 MHz
MJE171	P	000	00000	00☆☆	00	9	
MJE180	N	00	00000	00☆☆	00	9	
MJE181	N	000	00000	00☆☆	00	9	
MJE340	N	00000	0000	00☆☆	00	9	
MPS A05	N	000	0000	00	00	13	
MPS A06	N	0000	0000	00	00	13	
MPS A09	N	0000	0	00	000	13	
MPS A10	N	00	00	00	00	13	
MPS A13	N	00	000	00	0000	13	
MPS A16	N	00	00	00	0000	13	
MPS A17	N	00	00	00	0000	13	
MPS A18	N	000	000	00	0000	13	
MPS A55	P	000	0000	0	00	13	
MPS A56	P	0000	0000	0	00	13	
MPS U01	N	00	00000	00☆	00	14	
MPS U05	N	000	00000	00☆	00	14	
MPS U56	P	0000	00000	00☆	00	14	
MPS2926	N	0	00	00	00	13	f_T = 300 MHz
MPS3394	N	00	00	00	000	13	
MPS3702	P	00	000	00	000	13	f_T = 100 MHz
MPS3706	N	0	0000	00	00	13	
MPS6514	N	00	00	0	0000	13	f_T = 480 MHz
TIP29	N	00	0000	00☆☆	0	3	
TIP30	P	00	0000	00☆☆	0	3	
TIP31	N	00	00000	00☆☆☆	0	3	
TIP32	P	00	00000	00☆☆☆	0	3	
TIP140	N	000	00000	00☆☆☆	0000	7	Darlington
TIP142	N	00000	00000	00☆☆☆	0000	7	Darlington
TIP2955	P	000	00000	00☆☆☆	0	3	
TIP3055	N	000	00000	00☆☆☆	0	3	
TIP5530	P	000	00000	00☆☆☆	0	3	
2N696	N	000	0000	00	0	2	
2N706	N	0	0	0	0	2	
2N914	N	0	0000	00	00	2	
2N1613	N	000	0000	00	00	2	
2N1711	N	000	0000	00	000	2	
2N1983	N	00	0000	00	000	2	
2N1984	N	00	0000	00		2	
2N2219	N	00	0000	00	00	2	
2N2222	N	00	0000	00	00	2	
2N2925	N	00	00	0	0000	13	
2N2955	P	00	00	0	0	2	≠ MJE2955, TIP2955!
2N3054	N	000	00000	00☆☆☆	00	7	
2N3055	N	000	00000	00☆☆☆	0	7	
2N3553	N	00	0000	00☆	0	2	f_T = 500 MHz
2N3568	N	000	0000	0	000	13	
2N3638	P	00	0000	0	000	13	
2N3702	P	00	000	00	000	13	
2N3866	N	00	000	00☆	0	2	f_T = 700 MHz
2N3904	N	00	000	0	00	13	
2N3905	P	00	000	00	000	13	
2N3906	P	00	000	00	000	13	
2N3907	N	000	0	0	000	13	
2N4123	N	00	000	0	00	13	
2N4124	N	00	000	0	000	13	
2N4126	P	00	000	0	000	13	
2N4401	N	00	0000	00	0	13	
2N4410	N	0000	000	00	000	13	
2N4427	N	0	000	00☆	0	2	f_T = 700 MHz
2N5183	N	0	0000	00	000	2	





NOTE: All IC's shown top view, unless otherwise stated.

TUPTUNDUGDUS

Wherever possible in Elektor circuits, transistors and diodes are simply marked 'TUP' (Transistors, Universal PNP), 'TUN' (Transistor, Universel NPN), 'DUG' (Diode, Universal Germanium) or 'DUS' (Diode, Universal Silicon). This indicates that a large group of similar devices can be used, provided they meet the minimum specifications listed in tables 1a and 1b.

	type	U_{ceo} max	I_c max	h_{fe} min.	P_{tot} max	f_T min.
TUN	NPN	20 V	100 mA	100	100 mW	100 MHz
TUP	PNP	20 V	100 mA	100	100 mW	100 MHz

Table 1a. Minimum specifications for TUP and TUN.

Table 1b. Minimum specifications for DUS and DUG.

	type	U_R max	I_F max	I_R max	P_{tot} max	C_D max
DUS	Si	25 V	100 mA	1 µA	250 mW	5 pF
DUG	Ge	20 V	35 mA	100 µA	250 mW	10 pF

Table 2. Various transistor types that meet the TUN specifications.

TUN		
BC 107	BC 208	BC 384
BC 108	BC 209	BC 407
BC 109	BC 237	BC 408
BC 147	BC 238	BC 409
BC 148	BC 239	BC 413
BC 149	BC 317	BC 414
BC 171	BC 318	BC 547
BC 172	BC 319	BC 548
BC 173	BC 347	BC 549
BC 182	BC 348	BC 582
BC 183	BC 349	BC 583
BC 184	BC 382	BC 584
BC 207	BC 383	

Table 3. Various transistor types that meet the TUP specifications.

TUP		
BC 157	BC 253	BC 352
BC 158	BC 261	BC 415
BC 177	BC 262	BC 416
BC 178	BC 263	BC 417
BC 204	BC 307	BC 418
BC 205	BC 308	BC 419
BC 206	BC 309	BC 512
BC 212	BC 320	BC 513
BC 213	BC 321	BC 514
BC 214	BC 322	BC 557
BC 251	BC 350	BC 558
BC 252	BC 351	BC 559

The letters after the type number denote the current gain:

A: α' (β, h_{fe}) = 125-260

B: α' = 240-500

C: α' = 450-900.

Table 4. Various diodes that meet the DUS or DUG specifications.

DUS		DUG
BA 127	BA 318	OA 85
BA 217	BAX 13	OA 91
BA 218	BAY 61	OA 95
BA 221	1N914	AA 116
BA 222	1N4148	
BA 317		

Table 5. Minimum specifications for the BC107, -108, -109 and BC177, -178, -179 families (according to the Pro-Electron standard). Note that the BC179 does not necessarily meet the TUP specification ($I_{c,max}$ = 50 mA).

	NPN	PNP
	BC 107 BC 108 BC 109	BC 177 BC 178 BC 179
U_{ceo} max	45 V 20 V 20 V	45 V 25 V 20 V
U_{ebo} max	6 V 5 V 5 V	5 V 5 V 5 V
I_c max	100 mA 100 mA 100 mA	100 mA 100 mA 50 mA
$P_{tot.}$ max	300 mW 300 mW 300 mW	300 mW 300 mW 300 mW
f_T min.	150 MHz 150 MHz 150 MHz	130 MHz 130 MHz 130 MHz
F max	10 dB 10 dB 4 dB	10 dB 10 dB 4 dB

Table 6. Various equivalents for the BC107, -108, . . . families. The data are those given by the Pro-Electron standard; individual manufacturers will sometimes give better specifications for their own products.

NPN	PNP	Case	Remarks
BC 107 BC 108 BC 109	BC 177 BC 178 BC 179		
BC 147 BC 148 BC 149	BC 157 BC 158 BC 159		P_{max} = 250 mW
BC 207 BC 208 BC 209	BC 204 BC 205 BC 206		
BC 237 BC 238 BC 239	BC 307 BC 308 BC 309		
BC 317 BC 318 BC 319	BC 320 BC 321 BC 322		I_{cmax} = 150 mA
BC 347 BC 348 BC 349	BC 350 BC 351 BC 352		
BC 407 BC 408 BC 409	BC 417 BC 418 BC 419		P_{max} = 250 mW
BC 547 BC 548 BC 549	BC 557 BC 558 BC 559		P_{max} = 500 mW
BC 167 BC 168 BC 169	BC 257 BC 258 BC 259	!	169/259 I_{cmax} = 50 mA
BC 171 BC 172 BC 173	BC 251 BC 252 BC 253		251 . . . 253 low noise
BC 182 BC 183 BC 184	BC 212 BC 213 BC 214		I_{cmax} = 200 mA
BC 582 BC 583 BC 584	BC 512 BC 513 BC 514		I_{cmax} = 200 mA
BC 414 BC 414 BC 414	BC 416 BC 416 BC 416		low noise
BC 413 BC 413	BC 415 BC 415		low noise
BC 382 BC 383 BC 384			
BC 437 BC 438 BC 439			P_{max} = 220 mW
BC 467 BC 468 BC 469		!	P_{max} = 220 mW
	BC 261 BC 262 BC 263		low noise