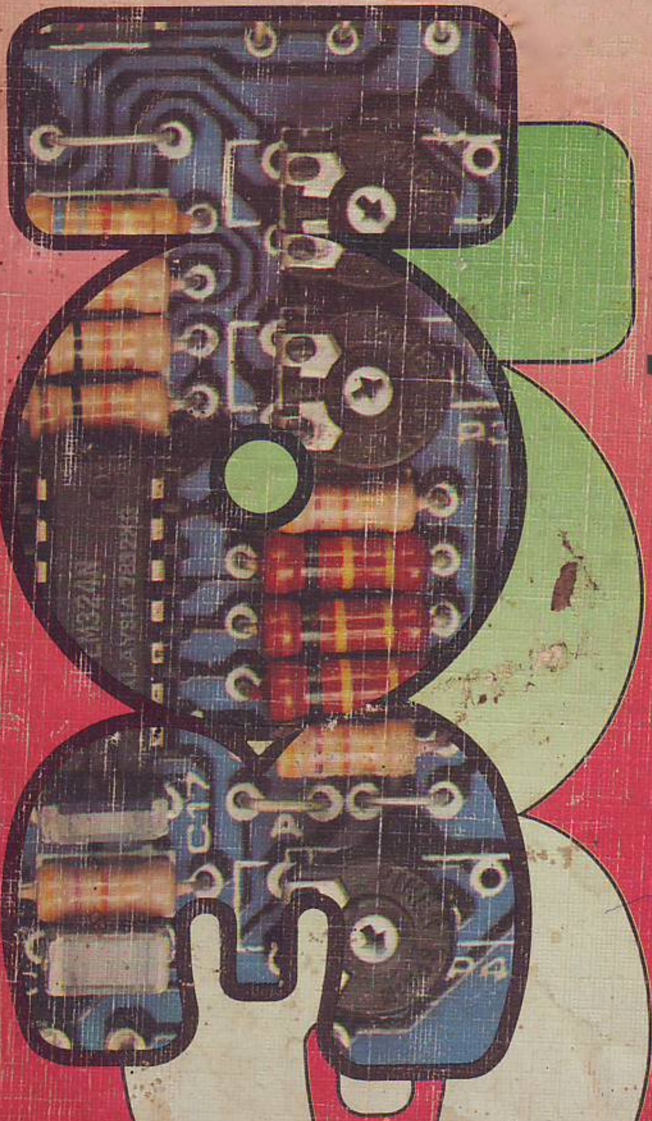
301 circuits

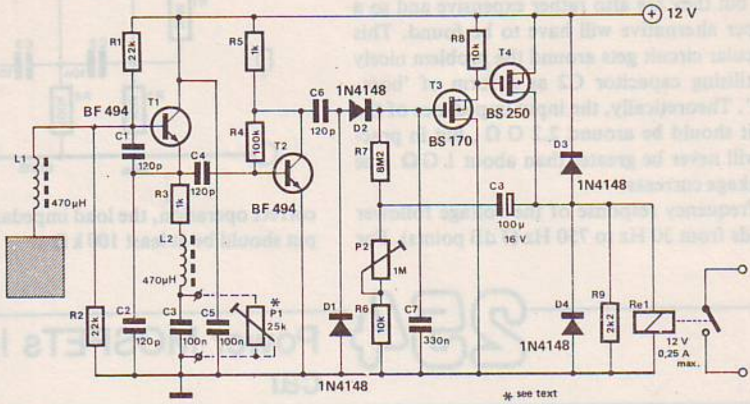
Circuits 252 to 301

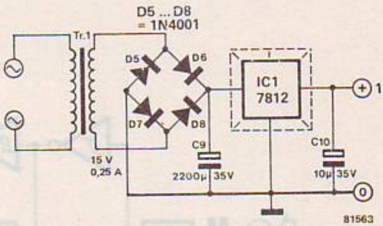


Vol - 4

AN elektor PUBLICATION

252 Objektor





The aim of this circuit is to detect the presence of a conductive object, provided the object is within a certain range. The operation of the circuit is totally independant of the condition of the object (dead, alive, static or moving), as long as it remains within this range. The sensitivity of the circuit can be set remotely, by adjusting the preset potentiometer P1. This is to avoid wearing out shoe leather during the initial (trial an error) alignment procedure – repeatedly walking up and down to obtain the optimum setting for that particular range.

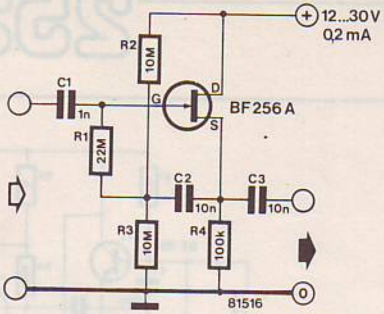
A rather innocent application of the circuit is its use as an invisible doorbell sensor, as the sensor could be located inside the house. The most important part of the circuit is the Clapp-oscillator constructed around transistor T1. The capacitor that should be connected in series with the coil L1 is formed by the sensor plate and the object to be detected. Due to the losses of this capacitor, the output from the oscillator is rather low, therefore a single amplifier stage (T2) has been added. The Schmitt trigger and monostable functions are performed by transistors T3 and T4. VMOS FETs have been chosen for the sake of simplicity and the fact that less components are required with respect to bipolar devices increases the reliability factor ('what you don't fit don't go wrong!'). This does of course mean that the cost of the project will be somewhat higher.

Another application is fluid level measurements in noisy environments. As the sensor does not require physical contact, the fluid could possess agressive qualities (for instance, fuming sulphuric acid).

253 High input impedance voltage follower

It is common knowledge that Field Effect Transistors (FETs) feature a very high impedance – well within the Giga-ohm range, in fact. As shown in the circuit diagram, one FET is quite enough to construct a buffer amplifier with an input impedance of 1 GΩ. The circuit is simply an impedance converter (source follower) with a gain of 1 and can be used for a variety of applications, such as a buffer for high impedance (capacitor) microphones, as an oscilloscope

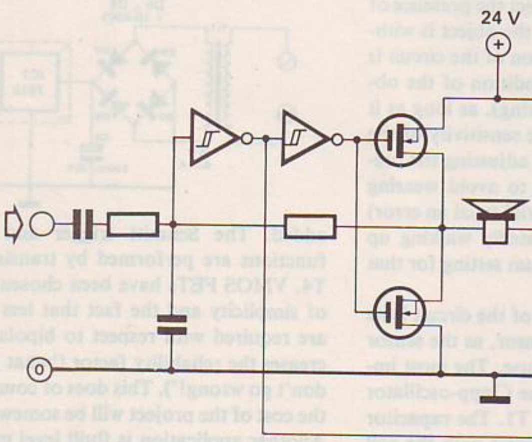
probe etc.
One way to obtain the required input impedance is to connect a 1 G Ω resistor between the gate of the FET and ground. Unfortunately, such high value resistors are not only difficult to obtain, but they are also rather expensive and so a cheaper alternative will have to be found. This particular circuit gets around the problem nicely by utilising capacitor C2 as a form of 'bootstrap'. Theoretically, the input impedance of the circuit should be around 2.2 G Ω, but in practice will never be greater than about 1 G Ω due to leakage currents.
The frequency response of the voltage follower extends from 30 Hz to 750 Hz (3 dB points). For



correct operation, the load impedance at the output should be at least 100 k Ω.

254 Power MOSFETs in the car

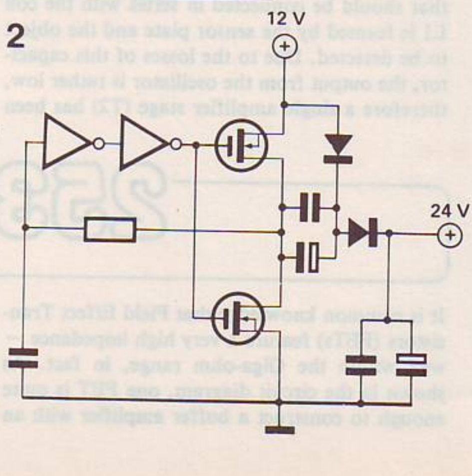
1



Due to the recent breakthrough in power-semiconductor technology it has become relatively simple to construct high power equipment for use in the car. Exemples of two possible applications for power MOSFETs are a 50 Watt booster and a 12 V to 24 V converter as shown in figures 1 and 2 respectively.
A booster amplifier intended for use in vehicles should at least meet the following requirements:

1. The output power should be greater than 10 Watts, thereby providing sufficient audio output to overcome the level of ambient background noise (engine, wind etc.).
2. The amplifier needs to be compact without putting any constraints on cooling.
3. Performance must be acceptable even under

2



conditions of large fluctuations in power supply voltage.

The circuit shown in figure 1 meets these demands quite adequately. It is a bridge version of the self-oscillating PWM amplifier with an output power in the order of 50 Watts.

If such a high power audio amplifier has to be supplied from a 12 V battery, one of the following possibilities may be chosen:

1. The amplifier operates at 12 V in combination with a low impedance load.
2. The amplifier operates at 12 V and a step-up transformer is connected to the output.
3. A voltage converter is used to increase the supply voltage so that the amplifier can deliver the rated output. (The converter may or may not include a transformer).

The advent of the power MOSFET makes the third possibility especially favourable. The simplicity of the design is very apparent from the circuit diagram shown in figure 2. In fact, the entire converter is just a CMOS power astable multivibrator. After rectification and smoothing, the output voltage is added to the battery voltage.

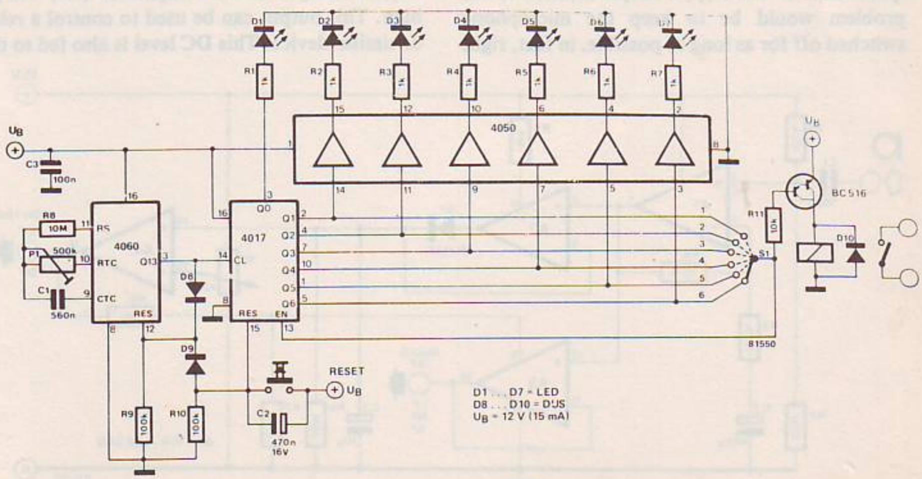
Obviously, the circuit diagrams only reveal the basic principle of operation of the two designs. However, both the 50 Watt booster and the 12 V to 24 V converter are likely to receive further, more extensive treatment in a future issue of Elektor.

255 Six hour timer

This control unit was originally designed to turn off stereo equipment automatically at night, so that music lovers who drift off to sleep in their easy chairs no longer need to worry about the cost of the next electricity bill. As the switching unit controls a relay, the circuit can be used for a multitude of other applications as well.

The heart of the timing circuit is a 4060 CMOS device, which contains an oscillator and a 14 stage divider. The frequency of the oscillator can be adjusted by means of potentiometer P1 so that the output at Q13 is approximately one pulse per hour. The duration of this clock pulse will be very short (about 100 ns), as it also resets the entire 4060 IC via diode D8.

The 'once per hour' clock pulse is fed to the second (divide-by-ten) counter, the 4017 IC. One of the outputs of this counter will be high (logic one) at any one moment. As soon as the 4017 is reset, output Q0 will go high. After an hour, output Q0 will go low and output Q1 will go high, etc. Switch S1 therefore enables the operator to select a time period of from one to six hours. As soon as the selected output goes high, the transistor will stop conducting and the relay will be deactivated (thereby switching off the



radio/record player etc.). As the enable input of the 4017 is also connected to the wiper of S1 any subsequent clock pulses will have no effect on the counter. The unit will therefore remain in the 'off' state until the reset button is depressed.

The 4050 CMOS buffer IC and the seven LEDs have been included to give an indication of the number of hours that have actually passed. These components can, of course, be omitted if an elapsed time display is not required. The supply voltage for the circuit is not critical and may be anywhere between 5 and 15 V. The current consumption of the circuit, not including the relay, is in the order of 15 mA. It is best to choose a supply voltage that is equal to the rating of the relay, so that any trouble is avoided. The BC 516 transistor can pass a current of 400 mA. If desired, two BC 557 (or similar) transistors may be connected as a darlington-pair instead.

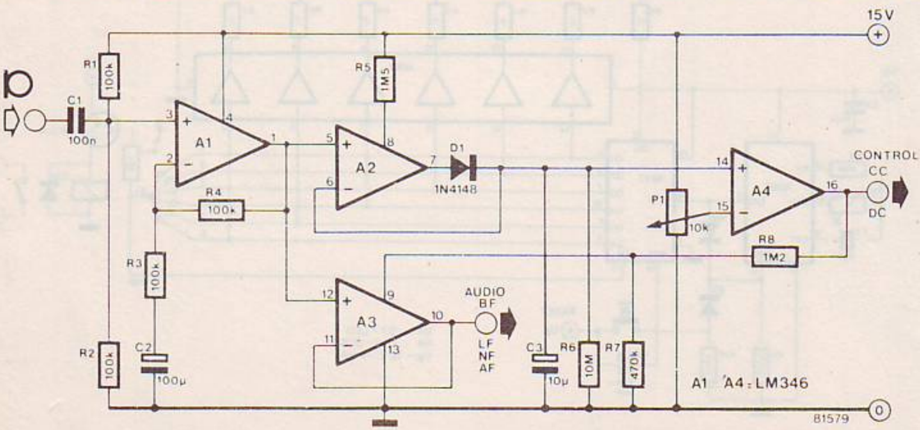
K. Siol

256 VOX for PA systems

An annoying problem with PA systems is their tendency to whistle or 'howl' due to feed back phenomena. There are a number of methods of avoiding the problem, the most obvious being the repositioning of the speakers in relation to the microphone. However, this is not always possible and the object of this article is to provide an answer for those difficult situations where feedback appears to be 'built in'.

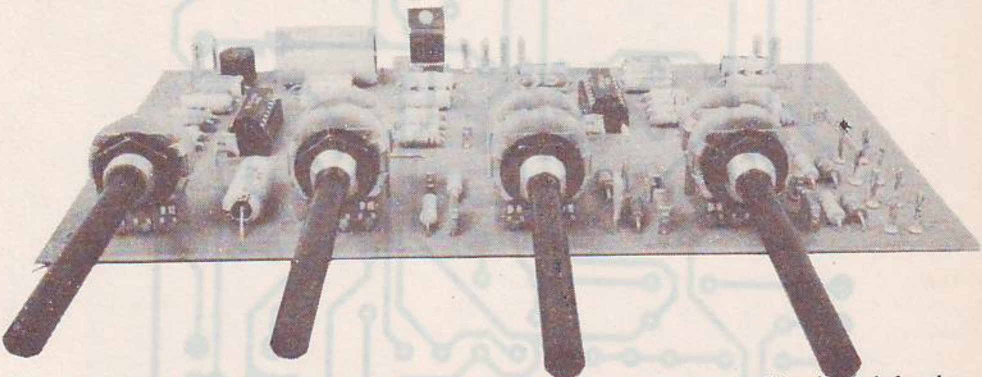
It is well known that lowering the frequency between the microphone and the PA amplifier by about 5 Hz will reduce feedback in many situations when all else seems to fail, but a 'frequency shifter' is an expensive piece of equipment and even then, its effectiveness is not always on a par with its cost. Since feedback requires time to build up, a simple solution to the problem would be to keep the microphone switched off for as long as possible, in fact, right up to the point at which speech begins. In other words, a voice operated switch or VOX.

This design is based on a National IC, the LM 346. This contains four programmable opamps which can be used in a variety of applications. Briefly the circuit works as follows: The speech signal from the microphone is amplified by opamp A1 and then fed to two further opamps, A2 and A3. The latter is simply a unity gain buffer for the PA amplifier. Opamp A2, together with diode D1, is used as a rectifier which converts the amplified microphone signal into a positive DC level. Any ripple voltage is smoothed out by capacitor C3. When the voltage across this capacitor is higher than the level set on the inverting input of A4 (with potentiometer P1), the output of the comparator (A4) will go high. This output can be used to control a relay or similar device. This DC level is also fed to the



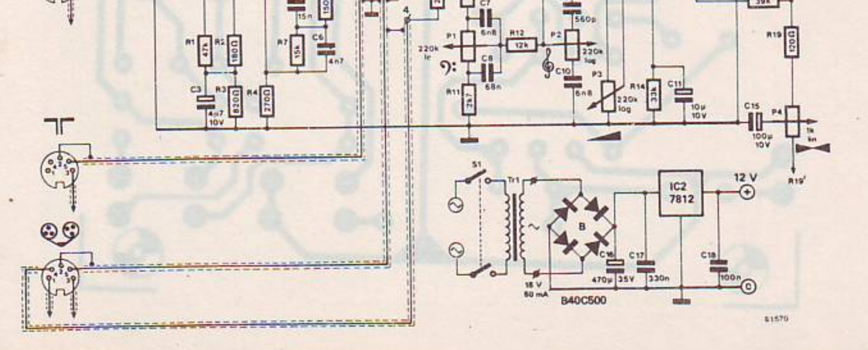
control pin (9) of opamp A3 via resistor R8. This opamp will only operate while this pin is held high. Therefore, when speech into the microphone ceases, A3 no longer operates and the main power amplifier remains inactive.
A high value resistor (R6) has been included in parallel with C3 to ensure that this capacitor discharges very slowly. This is very important because the signal path to the amplifier must remain open when the speaker pauses for a few moments. As mentioned previously, the output of A4 can be used to control a number of devices via a relay etc. This should find a number of applications especially in discos and the like.

257 Hi-fi pre-amplifier

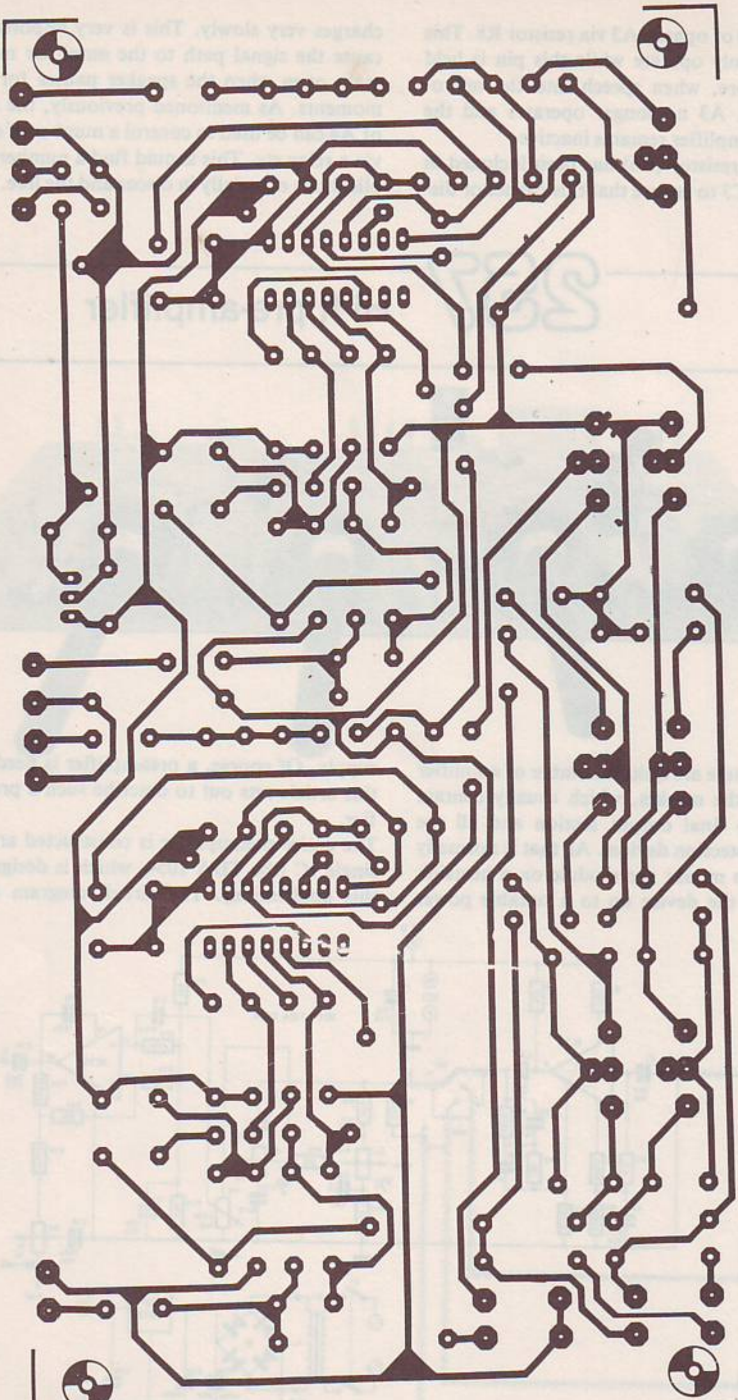


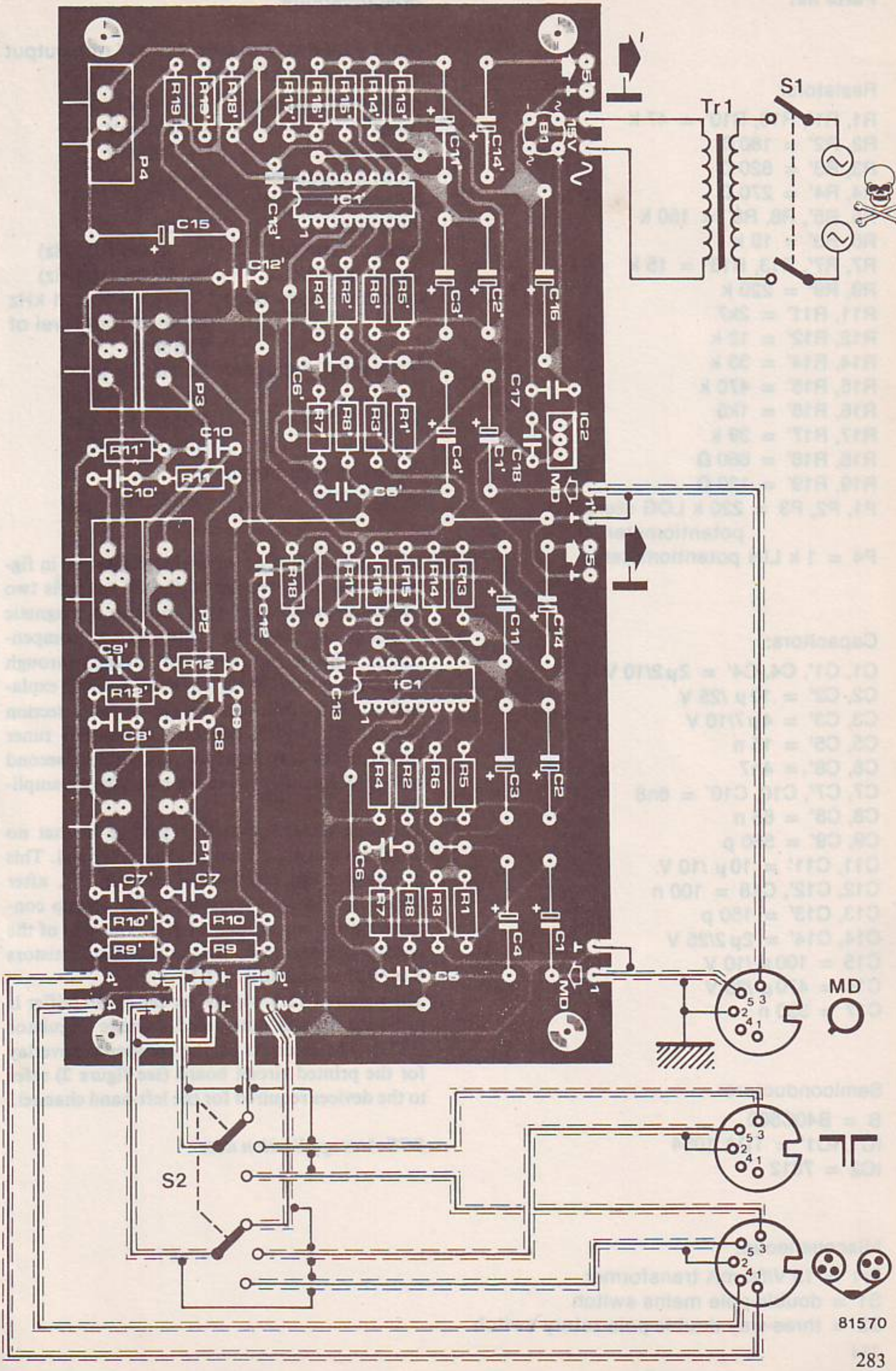
Nowadays, there are quite a number of amplifier modules on the market, which usually contain the complete final output section and all the necessary protection devices. All that is normally required is to mount the module on a heatsink and connect the device up to a suitable power supply. Of course, a preamplifier is needed and this article sets out to describe such a preamplifier.
The entire preamplifier is constructed around a single IC (the TDA 1054, which is designed for this application). The circuit diagram for the

1



2





81570 2

Parts list

Resistors:

R1, R1′, R10, R10′ = 47 k
R2, R2′ = 180 Ω
R3, R3′ = 820 Ω
R4, R4′ = 270 Ω
R5, R5′, R8, R8′ = 150 k
R6, R6′ = 10 k
R7, R7′, R13, R13′ = 15 k
R9, R9′ = 220 k
R11, R11′ = 2k7
R12, R12′ = 12 k
R14, R14′ = 33 k
R15, R15′ = 470 k
R16, R16′ = 1k5
R17, R17′ = 39 k
R18, R18′ = 680 Ω
R19, R19′ = 120 Ω
P1, P2, P3 = 220 k LOG stereo potentiometer
P4 = 1 k LIN potentiometer

Capacitors:

C1, C1′, C4, C4′ = 2μ2/10 V
C2, C2′ = 10 μ /25 V
C3, C3′ = 4μ7/10 V
C5, C5′ = 15 n
C6, C6′ = 4n7
C7, C7′, C10, C10′ = 6n8
C8, C8′ = 68 n
C9, C9′ = 560 p
C11, C11′ = 10 μ /10 V
C12, C12′, C18 = 100 n
C13, C13′ = 150 p
C14, C14′ = 2μ2/25 V
C15 = 100 μ /10 V
C16 = 470 μ /35 V
C17 = 330 n

Semiconductors:

B = B40C500
IC1, IC1′ = TDA 1054
IC2 = 7812

Miscellaneous:

Tr1 = 15 V/50 mA transformer
S1 = double-pole mains switch
S2 = three-way double-pole rotary switch

Specifications:

input sensitivity to give 775 mV rms output at a frequency of 1 kHz:

magnetic cartridge	– 3 mV
tuner	– 220 mV
tape input	– 220 mV
input impedance:	50 kΩ
balance control variation:	12 dB
bass boost/cut:	± 13 dB (100 Hz)
treble boost/cut:	± 13 dB (10 kHz)
harmonic distortion:	< 0.05% (f = 1 kHz at an output level of 775 mV)
frequency response:	20 Hz...24 kHz (± 3 dB, tone controls in the midposition)
signal to noise ratio (at 775 mV):	> 65 dB

left-hand channel of the preamp is shown in figure 1. The first section of the IC contains two transistors which are used to form a magnetic cartridge preamplifier with RIAA compensation. This is the 'standard' straight-through type and is not likely to need further explanation. It is followed by the input selection switch, S2, which connects either the tuner socket or the tape recorder socket to the second half of IC1 in addition to the cartridge preamplifier.

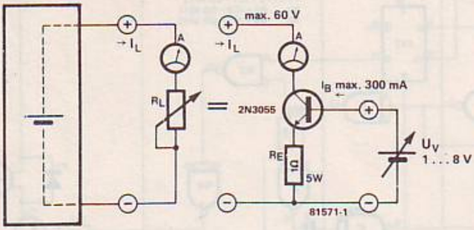
The tone control section is passive so that no problems can arise from too much control. This is followed by the volume control, P3, after which the signal is boosted by the opamp contained in the second half of IC1. The gain of the opamp is determined by the ratios of resistors R16/R17 as well as R18/R19 + P4.

The required 12 V supply for the preamplifier is provided by an integrated voltage regulator (7812). The asterisks on the component overlay for the printed circuit board (see figure 2) refer to the devices required for the left-hand channel.

SGS-Ates application note

258 Variable power 'resistor'

1



A major difficulty encountered when testing power supplies is the availability (or non-availability) of a suitable load. Usually, the problem is solved by a lash-up of resistors which, although not particularly elegant, will enable tests to be carried out. However, resistors with a power rating of 10 Watts and upwards can be rather expensive and certain values difficult to obtain. Furthermore, this type of load will not be variable. The simple circuit described here can overcome these problems effectively and economically.

A 2N3055 transistor with a variable gain controlled by an independant supply will form an infinitely variable load 'resistor'. This circuit will have a power capability of up to 50 Watts if a suitable heat sink is used. The ability to maintain a fixed load current when the power supply output voltage is varied is an added advantage.

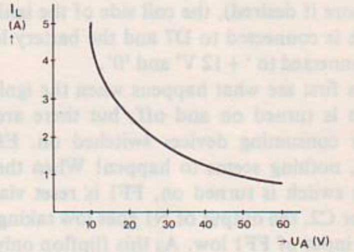
The maximum dissipation of the transistor should be borne in mind. As figure 2 shows, a current of 2.5 A at 20 V is only 50 W but 2.5 A at 50 V is 125 W which may prove to be a little high for the 2N3055.

The base voltage for the transistor can be obtained via a voltage divider across an external power source. If this is not available the circuit shown in figure 3 can be used. In this case, the BD 139 is used as a driver transistor in order to keep the battery supply current drain as low as possible.

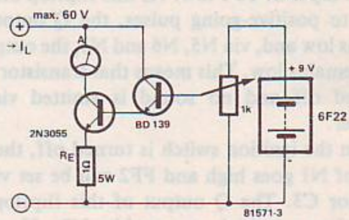
The power level can rise fairly steeply with a rising voltage and for this reason some indication of current and voltage levels in the circuit are essential. By using the graph shown in figure 2 it can be easily determined whether the maximum allowable power is being exceeded.

The circuit can also function as a current limiter. This facility will be useful when charging batteries with a constant current. The battery should be placed in the circuit between the ammeter and the collector of the 2N3055. It should be remembered to ensure that the voltage does not rise to a level high enough to cause damage to the plates of the battery.

2

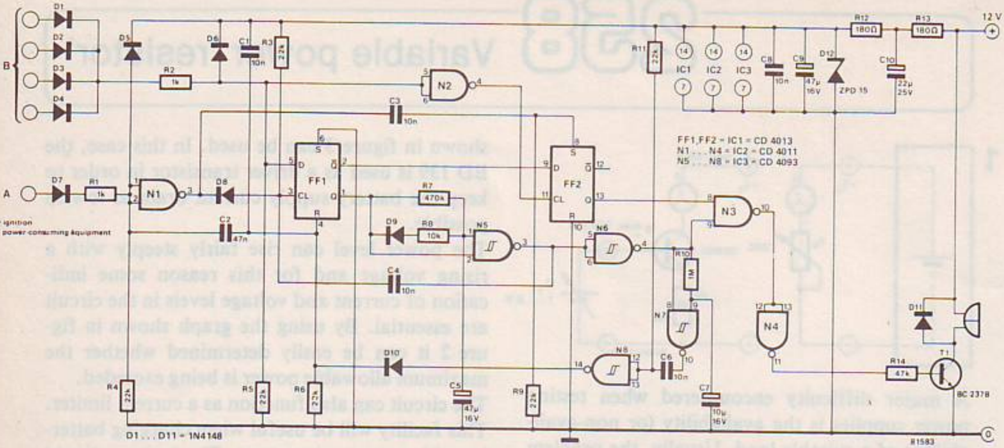


3



259 Auto reminder

The headlights in the majority of motor vehicles are unaffected by the ignition switch. This means that it is possible to leave the headlights turned on after leaving the car. In many instances, this is also true of auxiliary equipment, such as car radios etc., which have been wired directly to the car battery. For the forgetful driver, this can be something of a nuisance when trying to start the car the following morning. This circuit is intended as a warning to the driver that there is still something switched on which is consuming an unacceptable amount of power. The problem can be solved with a bit of logical thinking. After all, what is more obvious than to



apply a little digital technology?

At first glance, the circuit looks a lot more complicated than it actually is. It only requires three ICs. The switched supply rails of the equipment to be monitored are connected to diodes D1...D4 (or more if desired), the coil side of the ignition switch is connected to D7 and the battery leads are connected to '+12 V' and '0'.

Let us first see what happens when the ignition switch is turned on and off, but there are no power consuming devices switched on. Effectively, nothing seems to happen! When the ignition switch is turned on, FF1 is reset via capacitor C2, the output of N1 goes low taking the clock input of FF1 low. As this flipflop only reacts to positive-going pulses, the Q output remains low and, via N5, N6 and N3, the output of N4 remains low. This means that transistor T1 is turned off and no sound is emitted via the buzzer.

When the ignition switch is turned off, the output of N1 goes high and FF2 will be set via capacitor C3. The Q output of this flipflop will therefore go high to enable N3. However, although FF1 receives a clock pulse via R3, the information at the data (D) input is low as there are no items of equipment on and so the outputs of FF1 remain as they were. The end result is that the output of N4 remains low and T1 stays off.

Now let us take the case when one or more auxiliary circuits are turned on. When the ignition switch is turned on the end result is the same as that previously described. When the ignition switch is turned off, however, things start to happen! Flipflop FF1 receives a clock pulse via R3 as before. This time, as the data input is high, the Q output will also go high. Capacitor C5 is charged up via resistor R7 and, when sufficiently charged (charge time = R7 x C5), the output of N5 will go low. This signal is differentiated by C4/R3 to provide a further clock pulse for FF1.

During the period that C5 is charging, the driver still has the opportunity to turn off the relative equipment, thus preventing the warning signal from being sounded. If this is the case, the Q output of FF1 will again go low at the second clock pulse and C5 will discharge via R8 and D9. The output of N5 goes low thereby resuming the original condition. If, on the other hand, the driver neglects to switch off certain items, the output of N6 will go high when C5 is charged. This, in turn, takes the other input of N3 high (N3 was enabled when the ignition was turned off as described previously) so that the output of N3 goes low, the output of N4 goes high and transistor T1 turns on to sound the buzzer. At the same time, capacitor C7 charges up via resistor R10 over a period of about ten seconds. After this time the output of N7 goes low. This pulse is differentiated by C6/R11 to provide a reset pulse for FF1 via N8 and D10. Once reset, the Q output of FF1 goes low and the warning is cancelled.

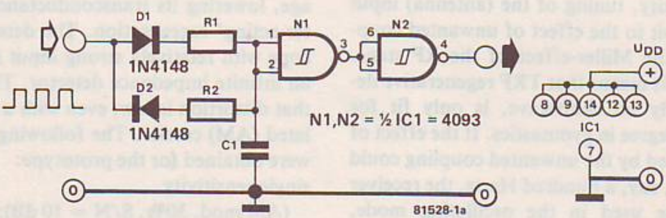
It is possible, of course, to interrupt the alarm by switching the ignition back on. The circuit also allows for the possibility of deliberately leaving a particular item (such as sidelights) on without the alarm sounding. This is accomplished by first turning off the ignition with the desired piece of equipment switched on. This equipment is then switched off and back on again before the alarm sounds. This means that the output of N2 will go high when the item is switched off, providing a clock pulse for FF2. As the $\overline{Q}$ output of FF1 is low at this time, the D input of FF2 is also low therefore the Q output of FF2 will go low, so that N3 is disabled and transistor T1 is turned off. When the desired item is turned back on, the clock input of FF2 goes low, but the outputs remain the same. Although FF1 and gates N5...N8 go through their cycle, the alarm will not be operated.

It must be admitted that simpler alarm systems do exist. However, this circuit incorporates some interesting ideas and is simple to use. Nevertheless, the device does have one drawback. If, for instance, two items of equipment are left on inadvertantly and only one is switched off when the alarm sounds, the unit will not indicate this fact. Care must be taken, therefore, to have a good look round the dashboard when the alarm sounds!

W. Gscheidle

260 Adjustable square-wave edges

1



The ability to delay the leading and trailing edges of a square waveform will find many applications in digital circuits. The diagram in figure 1a shows that only a very few components are required to do this. The circuit makes use of the fact that the output of a Schmitt trigger gate will not change state until the voltage level on the input reaches a certain critical point known as the trigger threshold.

During a rising edge at the input, capacitor C1 is charged via D1 and R1. This increases the time it takes for the voltage level at the input of the gate to reach the trigger threshold point. With a logic 1 at the input to the circuit, the potential across C1 will continue to rise until, for all practical purposes, it reaches the supply voltage level. When the input returns to zero C1 will discharge via D2 and R2, again delaying the time at which the trigger threshold is reached. A clear understanding can be gained from figure 1b which shows the waveforms at various points in the circuit.

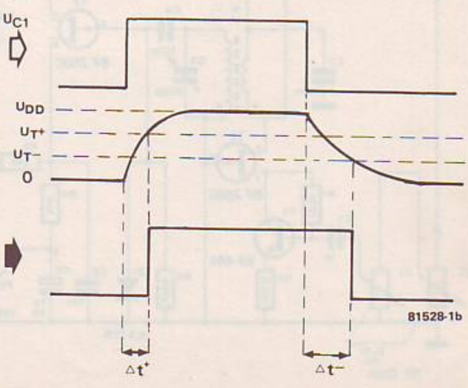
It must be remembered that the trigger threshold point of a Schmitt gate is highly dependant on the supply voltage. The following figures are quoted for the RCA 4093:

U_{DD}	U_T^+	U_T^-
5	3.3	2.3
10	7	5.1
15	9.4	7.3

The leading edge delay can now be calculated with the equation:

$$\Delta t^+ = -R1 \times C1 \times \text{nat.log.}\left(1 - \frac{U_T^+}{U_{DD} - 0.7}\right)$$

2



The delay in the trailing edge amounts to:

$$\Delta t^{-} = -R2 \times C1 \times \text{nat.log.}\left(\frac{U_T^{-}}{U_{DD} - 0.7}\right)$$

Therefore, the choice of values for R1, R2 and C1 will give any required delay to the leading and trailing edges of the square-wave. However, the maximum delay for either edge should not be more than 80% of the pulse duration.

Nothing in this world is perfect and different manufacturers of IC1 quote rather large trigger threshold tolerances. This means that the actual delay times can be quite different from those calculated.

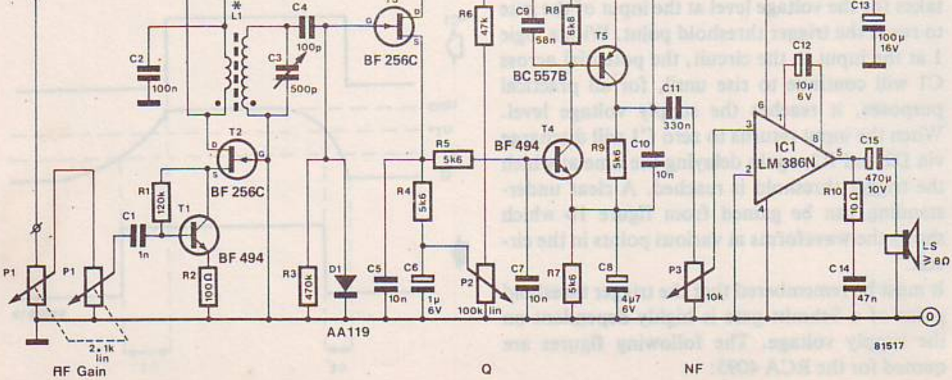
261 Simple short-wave receiver

The most notorious effects of regenerative TRF receivers are radiation and unwanted coupling between the antenna and the LC circuit that is acted upon by the regeneration. Apart from the extra complexity, tuning of the (antenna) input adds quite a bit to the effect of unwanted coupling, due to the Miller-effect of the RF stage. Invariably, this means that TRF regenerative design, especially on shortwave, is only fit for those with a degree in gymnastics. If the effect of detuning caused by the unwanted coupling could be reduced to, say, a hundred Hertz, the receiver could also be used in the oscillating mode, thereby providing the owner with the possibility of product-detector reception modes, such as CW, RTTY and SSB. It should be stated at this point that this feature is positively enhanced when a frequency counter is added.

In order to achieve a minimum of pulling, the RF stage of the circuit shown consists of a bipolar transistor and an FET in cascode. As can be seen, the input is aperiodic. The disadvantage of input overload is more than compensated for by the high sensitivity (even a tiny whip antenna can be used). A smooth control of regeneration is obtained by D1, which starts supplying a negative bias for T3 on reaching the threshold voltage, lowering its transconductance, thus 'counter-acting' regeneration. The detector is able to cope with relatively strong input signals, as it is an infinite impedance detector. This also means that distortion is low, even with a heavily modulated (AM) carrier. The following specifications were obtained for the prototype:

single sensitivity
(AM mod. 30%, S/N = 10 dB): 1 µ V
single signal sensitivity
(SSB, S/N = 10 dB): 0.3 µ V
frequency range with a 500 p
tuning capacitor: 4.4...17 MHz.

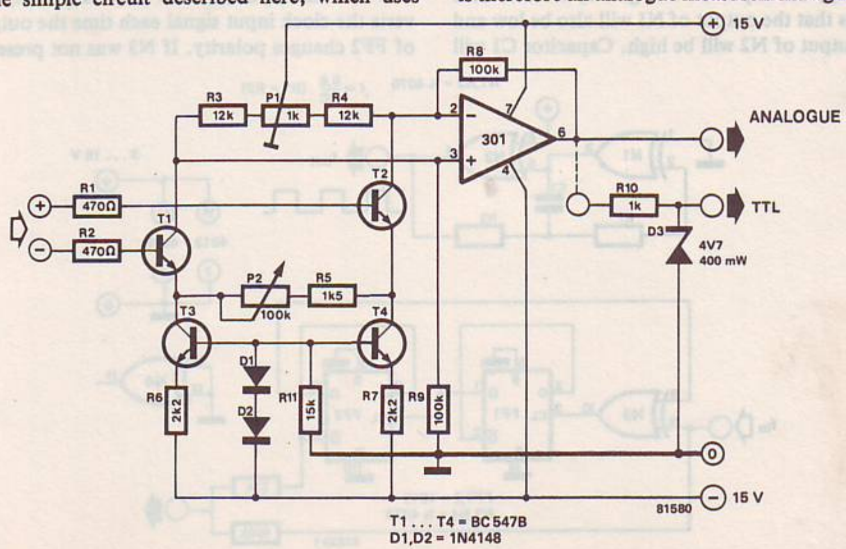
The term single signal sensitivity may need some



explanation. It is the figure obtained by measuring the sensitivity with the aid of a signal generator in the absence of all other signals. Due to input overloading (of the wideband RF stage) and envelope detection taking place on strong adjacent channel signals, the full benefit of this sensitivity will never be obtained, except perhaps in countries like Australia, where the spectrum is not yet polluted by OTHR's and BC jammers.
In the product detector mode, the suppression of AM will be in the order of between 40 and 60 dB, depending on tuning. The lower figure refers to the highest tuning frequency. Improvement can be made by reducing the L/C ratio.
The primary winding of L1 consists of 6 turns of 0.25 mm enamelled copper wire on an Amidon ring core type T94-6. The secondary winding consists of 25 turns of 0.68...0.8 mm enamelled copper wire over the complete length of the core. The primary winding should be situated at the 'cold' end and layed between the turns of the secondary.

262 Universal measuring amplifier

An analogue multimeter is, by now, one of the more 'standard' items of equipment owned by electronics enthusiasts. Even digital multimeters are now becoming quite common. However, it is not uncommon to find that the capabilities of the measuring instruments available are nowhere near as extensive as one would wish. Either the input sensitivity is not high enough (in other words, it can not measure low voltages) or else the input impedance of the instrument is too low. The second disadvantage is the worst, in fact it is often the reason why measurements taken are totally inaccurate. In general, interpretation of incorrect results will lead to incorrect conclusions!
The simple circuit described here, which uses only a few components suffers from none of these disadvantages. The circuit consists of a discrete differential amplifier constructed around transistors T1 and T2. A separate constant current source is connected in series with both emitter leads. The current source for transistor T1 consists of D1, D2, T3 and R6 and that for transistor T2 consists of D1, D2, T4 and R7. The constant emitter currents make the measuring amplifier independant of supply voltage variations.
The differential amplifier (T1 and T2) is followed by an integrated differential amplifier (the LM 301 from National Semiconductor). This opamp is connected to give unity gain. Its output is therefore an analogue measurement signal and



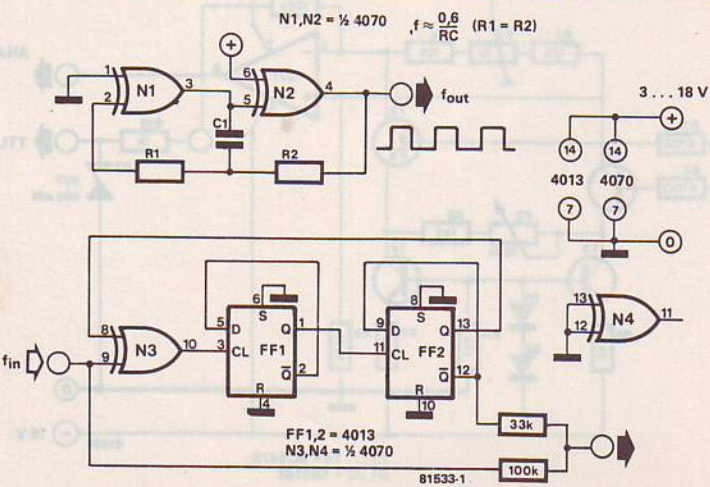
can be used directly as such. Two further components, R10 and D3, convert the analogue output to a TTL-compatible signal.
Now that we have described the basic circuit, what can it be used for? Two possibilities have already been mentioned: It can be used as a preamplifier for a normal (analogue) multimeter and it can also be used as a preamplifier for a digital multimeter. Furthermore, the circuit can be used as an AF preamplifier for frequency counters or similar devices. In this case, potentiometer P2 can be used to set the trigger level. Finally, it is also possible to use the circuit as a preamplifier for an inexpensive oscilloscope.

Regardless of the actual application for the circuit, the only main calibration is the same in all cases. This is the 'zero offset' and is adjusted by means of preset potentiometer P1. This potentiometer must be set so that when the inputs are shorted (in other words, when the left-hand sides of R1 and R2 are connected together) the output from the opamp is exactly zero volts. In normal applications, potentiometer P2 adjusts the input sensitivity. With the aid of this potentiometer, it is possible to adjust the gain of the circuit over a wide range; from a gain of 2 to a gain of 130. It may therefore be useful to provide this potentiometer with a calibrated scale.

263 Digital sinewave oscillator

The circuit consists of two sections, each of which could have many useful applications on its own: an oscillator constructed around a pair of EX(clusive) OR gates and a divide-by-three circuit constructed around two ordinary flipflops.
The oscillator is made up from both a non-inverting gate (N1) and an inverting gate (N2). If only inverting gates had been used, at least three would have been required – as a non-inverting gate can be made up from two inverting gates connected in series.
The circuit works as follows. Let us assume that, initially, the input of N1 (pin 2) is low. This means that the output of N1 will also be low and the output of N2 will be high. Capacitor C1 will then be charged via resistor R2. After a short while, the input of N1 will go high via R1 and the whole procedure will be reversed. Readers interested in this type of oscillator are referred to the National Semiconductor Application Note No. AN-118, which is included in their current CMOS data book.
The divide-by-three section consists of two flipflops which both divide by two, in other words, it would be expected that together they would divide by four. However, another EXOR gate (N3) has been included between the output of FF2 and the input of FF1. This effectively inverts the clock input signal each time the output of FF2 changes polarity. If N3 was not present,

1

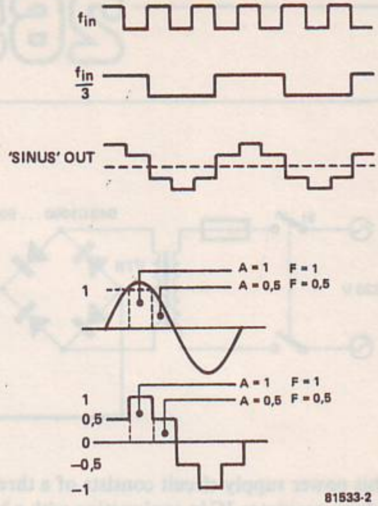


the output state of the flipflop would not change until the end of the current clock period. With the addition of N3, the clock signal is inverted and the positive-going edge triggers the flipflop after every half period. Therefore, the dividing factor here is three, not four.

The sinewave signal is generated via a pair of resistors (R3 and R4). When the input to both resistors is low (logic zero) there will be no output voltage. When the input to both resistors is high (logic one) the output voltage will be high. When one input to the resistors is low and the other high the output voltage will be either ¼ or ¾ of the supply (high) level.

Obviously, this can be proven mathematically, but a simpler method to justify it is to examine a single sinewave period diagrammatically. A small rectangle may be drawn in the centre of the sinewave to represent a logic one level. Two further rectangles of the same size can then be drawn to each side of the first. The area inside the sinewave of the last two rectangles will be half that of the first. The digital simulation technique generates a signal with the same areas as the above.

2



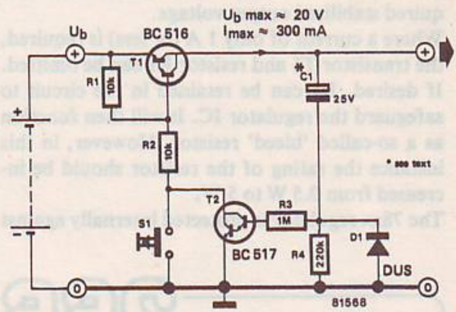
When constructing the circuit it should be noted that CMOS inputs should never be left 'floating'. In other words, pins 12 and 13 of the EXOR chip (N4) should be connected to ground (0 V).

264 Battery supply timer

A need often arises for battery powered equipment to be switched off after a certain period of time. Although timing circuits are legion, it is not such a simple matter to switch a power supply off after some hours of use. The circuit featured here will do this at the expense of a current drain of only a few nano amps.

A look at the circuit diagram will show that not many components are involved. Switch S1 is the 'on' button and, when pressed, supplies a base drive current to the darlington transistor T1 which will then conduct to supply power to the equipment in use. Transistor T2 will now also switch on to act as a latch across S1 maintaining the base current to T1.

Capacitor C1 will now start to charge via R4. When the voltage across R4 drops to about 1.2 V, T2 will switch off. This in turn will switch off T1 and therefore the supply. The only current now flowing will be the leakage current through both transistors but this will only amount to a few nano amps at most. To all intents and purposes, the battery supply will be switched off.

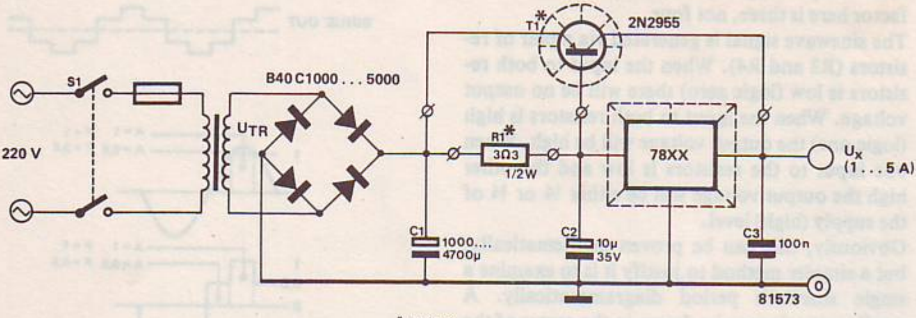


The time period for which the supply will be switched on can be calculated from the rather horendous formula:

$$t = -22 \cdot 10^4 \cdot C1 \cdot \text{nat.log.} \frac{1 \cdot 2}{U_B} \text{ s (C1 in Farads).}$$

For those in doubt the 'try it and see' method may take longer but will work just as well. If required, the two darlingtons can be substituted by discrete transistors.

265 Power stabiliser



*see text

This power supply circuit consists of a three-pin voltage regulator IC in conjunction with a buffer transistor. This combination is a result of the fact that the 78xx series of voltage regulators are only capable of delivering a current of 1 A. In this design, when the output current exceeds about 200 mA, the buffer transistor takes over the task from the voltage regulator IC thereby allowing currents up to 5 A to be drawn.

The 78xx IC is available for a range of different voltages. By selecting the required regulator in the series, the circuit can be adapted for any voltage, provided of course the transformer output voltage is at least 4 V greater than the required stabilised output voltage.

Where a current of only 1 A (or less) is required, the transistor T1 and resistor R1 can be omitted. If desired, R1 can be retained in the circuit to safeguard the regulator IC. It will then function as a so-called 'bleed' resistor. However, in this instance the rating of the resistor should be increased from 0.5 W to 5 W.

The 78xx regulator is protected internally against overheating, but in practice the demand on this protection circuitry is not altogether satisfactory. To obtain a stable design, two measures have been taken in the circuit shown here: The current through the regulator IC can never exceed 300 mA except when there is a short-circuit on the output. The buffer transistor has a more than adequate current rating.

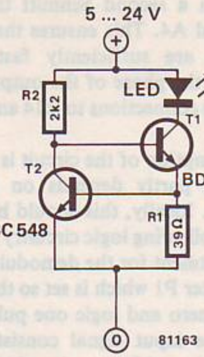
Provided the heatsink used is sufficiently large, both the voltage regulator and the buffer transistor should be able to survive momentary short-circuits, when the peak current may well exceed the maximum output current of 5 A. However, the actual amount of short-circuit current will be limited as the voltage regulator limits the amount of base drive current to transistor T1.

Capacitor C1 will smooth out any AC ripple, but its value should be modified to cope with the maximum flow of current. For a current flow of up to 1 A, a value of 1000 μ F should be sufficient, but for a current of 5 A the value should be increased to 4700 μ F.

266 Constant current LED

It is normal nowadays to use a LED as a panel indicator whenever possible. However, in keeping with all electronic devices, they do have limitations and their operating parameters can make life difficult at times. For instance, if the supply voltage varies by any great degree the brightness of the LED will follow suit. Should the voltage level become too high it will result in the LED giving a permanently off indication! The ingenious circuit here can get around these problems quite effectively.

The maximum current capability of a LED is normally about 50 mA but brightness will not significantly increase above 20 mA. This figure is about the optimum economic current level and the purpose of this circuit is to maintain this value irrespective of fluctuations in supply voltage levels.

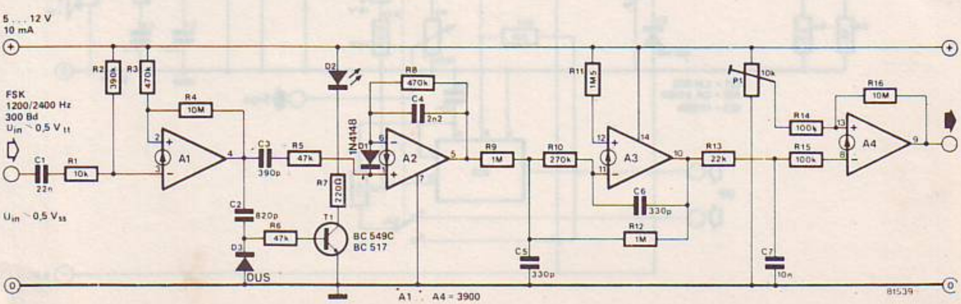


The two transistors, T1 and T2, form a constant current source and will maintain the current level to within the acceptable limits of 15 and 27 mA with variations in voltage level between 5 and 24 V.
Operation is relatively straightforward. A rising supply voltage will cause the collector voltage of T1 to rise. This in turn will increase the base drive current to T2. The subsequent drop in potential at the collector of T2 will reduce the base current to T1 and therefore counteract the rise in current to the LED. The circuit will now be stabilized.
The table below gives an indication of the LED current at various supply voltages.

5 V – 15 mA	15 V – 22 mA
9 V – 18 mA	18 V – 24 mA
12 V – 20 mA	24 V – 27 mA

267 Kansas City demodulator

The most popular form of FSK (frequency-shift keying) modulator used by computers nowadays is the 'Kansas City Standard' where digital information is transmitted sequentially by means of carrier pulses of constant amplitude and two different frequencies (1200 Hz and 2400 Hz). A logic low level generates a frequency of 1200 Hz whilst 2400 Hz is generated by a logic high level. Obviously, for every modulator there has to be a demodulator.
As can be seen from the circuit diagram, the Kansas City demodulator described here is quite straightforward. Opamp A1 acts as a Schmitt trigger and simply converts the output of the tape recorder into a perfectly symmetrical squarewave signal. Opamp A2 is configured as a 'charge pump' and converts the incoming frequency into a low or high logic level. The output of A2 is fed to a low pass filter constructed around opamp A3 and the decoded, filtered signal is then fed to a second Schmitt trigger (A4) to 'clean up' the final digital output.
Transistor T1 and LED D2 are included to indicate whether or not the input (FSK) signal is of sufficient amplitude to drive the system.
The LM 3900 contains four opamps which are slightly unusual in that they react to differences in input *currents* rather than *voltages* (this type of device is commonly called a 'Norton amplifier'). This means that the output of the first Schmitt trigger (A1) is zero under no-signal conditions, because the current flowing through the inverting input (via R2) will be greater than that flowing through the non-inverting input (via

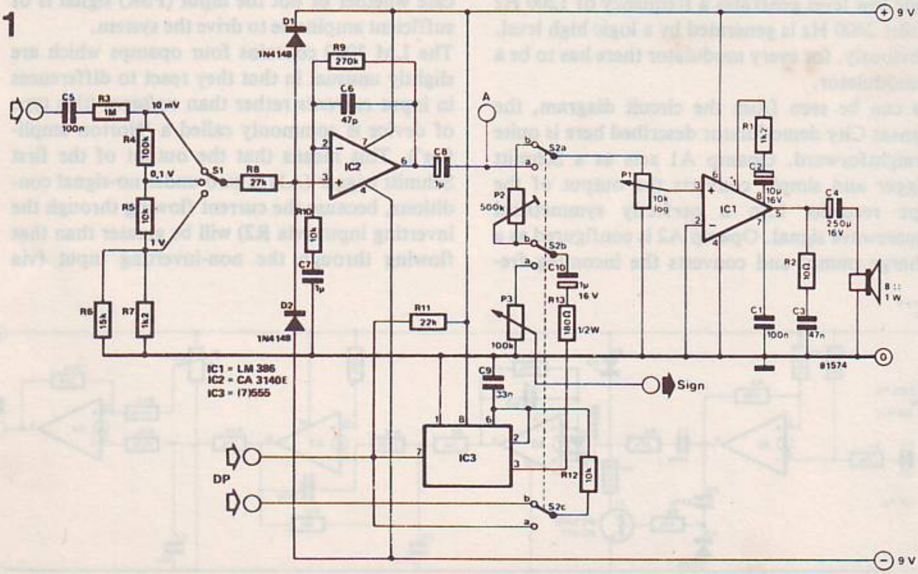


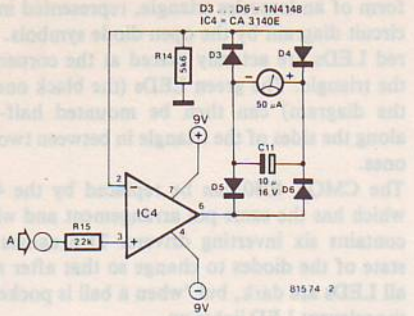
R3).
The charge pump works as follows: When there is no input signal, capacitor C4 is discharged via R8, so that the output of A2 is virtually zero. If at this point a positive-going pulse is received from A1, a brief current pulse will flow into the non-inverting input of A2 via capacitor C3. This means that an identical current must pass through the inverting input for the circuit to remain in equilibrium. This can only be achieved via C4 which is therefore charged a small amount. As a result, the output voltage of A2 will rise whenever a positive-going signal edge is present at the input. Afterwards, capacitor C4 will be discharged via resistor R8 and the output voltage will drop once more. The more pulses at the input, the higher the output voltage.
The circuit around opamp A3 is a conventional low pass filter. The turn-over frequency of the filter depends on the *baud rate* of the incoming signal. At 300 baud the maximum frequency at this point will be 150 Hz therefore the turn-over frequency must be slightly higher.
The output of the low pass filter has rather poor edges and is too low in amplitude to be processed by logic circuitry. For this reason the signal is passed through a second Schmitt trigger constructed around A4. This ensures that the final output pulses are sufficiently fast to drive CMOS ICs. If the phase of the output signal is not correct, the connections to R14 and R15 may be reversed.
Current consumption of the circuit is only a few milliamps and partly depends on the actual supply voltage. Ideally, this should be the same as that of the following logic circuitry.
The only adjustment for the demodulator is preset potentiometer P1 which is set so that the duration of logic zero and logic one pulses are the same when the input signal consists of eight cycles of 2400 Hz and four cycles of 1200 Hz.

268 Signal injector - tracer

A signal injector must certainly be one of the cheapest and most useful pieces of test equipment in the hobby workshop. The design described here doubles as a signal tracer and features an audio output enabling both eyes to be kept on the work in hand.
The circuit is very simple and consists of three main parts; a signal generator (IC3); a pre-amp

1





2

(IC2); and a main amplifier. The signal generator, IC3, is a 555 connected as a 1 kHz oscillator. With S2 in position b, this acts as a simple continuity tester. Since the test points DP are in series with the oscillator's RC network, the circuit will only oscillate when there is continuity between these points. Two test probes connected to the DP points can therefore be used to establish continuity between two points in the circuit under test. The output of the oscillator is fed via P2 to the main amplifier and a tone will be heard from the loudspeaker when there is a short circuit between the probes. Nothing will be heard if the probes are open circuit of course.

With switch S2 in position a, IC3 will oscillate continuously. Its output level is then controlled by P3 and fed directly to probe Ⓑ to be injected into the circuit under test. Capacitor C10 and resistor R13 are included to prevent the oscillator from being loaded by the test circuit. Probe Ⓐ is then used to trace the signal through the stages of the test circuit. A resistance or potentiometer will attenuate the signal and a transistor will usually amplify it.

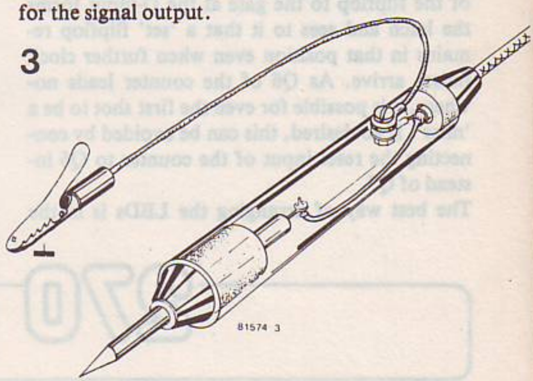
In order to detect changes in amplitude, it is important not to overload the loudspeaker. A simple switched attenuator is therefore included at the input to the pre-amp to provide three different input levels. To avoid loading the circuit under test a high impedance input is ensured by C5 and R3. The probe signal is amplified by IC2 and fed, via C8 and S2a, to P1 which is used to adjust the input level to the pre-amp, IC1.

Some readers may prefer a visual output indication and for this purpose a moving coil meter can be used as shown in figure 2.

Because of the high input impedance of the pre-amp it is necessary to use a screened lead and probe for this input. Failure to do this will result in plenty of noise at the output.

A design for an easily constructed probe is shown in figure 3. Normal test leads can be used for the signal output.

3



269 American billiards

The popularity of American Pool continues to increase and this electronic game is played in a similar manner to the original table billiards. However, the billiard balls are symbolically represented here by six LEDs.

The circuit is based on a random generator. When the reset switch is depressed, all the LED's light up, and when the 'hit' button is pushed a number of random occurrences take place so that either one of the LED's goes out or everything stays as it was. One of the LED's going out, means that one of the balls has been pocketed.

The game can be played in two ways. In the first game, each player must pocket all the balls whereby all the LEDs are extinguished, and the person who can do that with the least number of 'hits' is the winner. In the second variation of the game, the number of players is restricted to only two. One player starts: if the first ball he hits is a red one, then he is obliged to pocket also all the other red ones. The second player then does the same with the green balls. As long as the first player continues to hit his own colour, it remains his turn. Only when he fails to hit a ball or hits one of his opponent's colour, is it the turn of the other player. As soon as three balls of one colour have been pocketed, the player of that colour has won the game. This makes it impossible for the game to end in a draw, and is a good way of deciding the winner of the first type of game that was drawn.

Let us now have a look at the circuit. To start, all six flipflops, FF1...FF6, are set to zero with pushbutton S2 so that all LEDs light up. The multivibrator, constructed with N1 and N2, delivers a clock frequency of about 800 Hz to the Johnson-counter, IC2, of which the outputs deliver '1' in sequence. The gates N5 to N10 are wired as latches and connect the counter outputs to the D-inputs of the flipflops.

Operating the 'hit' button S1 will supply a pulse to all the flipflops. The counter outputs which are high will then, via the latches, set the flipflops that were not already set, and the relevant LED goes out. The feedback from the Q-output of the flipflop to the gate at the D-input forms the latch and sees to it that a 'set' flipflop remains in that position even when further clock pulses arrive. As Q6 of the counter leads nowhere, it is possible for even the first shot to be a 'miss'. If so desired, this can be avoided by connecting the reset input of the counter to Q6 instead of Q7.

The best way of arranging the LEDs is in the form of an isosceles triangle, represented in the circuit diagram by the open diode symbols. The red LEDs are actually placed at the corners of the triangle. The green LEDs (the black ones in the diagram) can then be mounted half-way along the sides of the triangle in between two red ones.

The CMOS 4050 can be replaced by the 4049 which has the same pin arrangement and which contains six inverting drivers. This causes the state of the diodes to change so that after reset all LEDs are dark, but 'when a ball is pocketed' the relevant LED lights up.

The two sets of rules for the games described earlier are not the end. There are other possibilities. One ball can be designated as the 'black' in snooker which means that it is the last one to be pocketed. Another variation consists of, prior to 'hitting a ball', deciding which colour it is going to be. No doubt it will be possible to think up further variations, once the circuit and its possibilities have become familiar. When the game is played a lot, it is best to supply the power via a small mains supply or a nicad. The power consumption is 90 mA when all the LEDs are on. When the game is played less frequently it will be sufficient to use two 4.5 V dry cells.

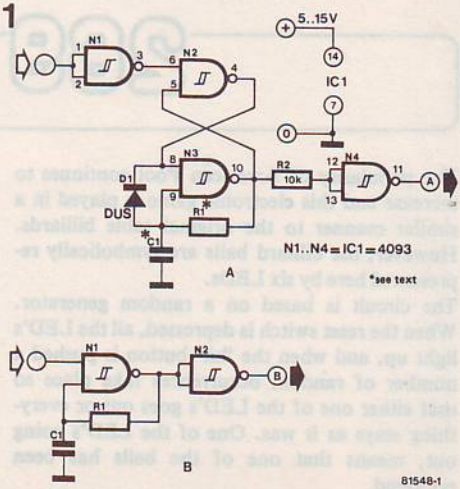
H.J. Walter

270 Constant pulse width oscillator

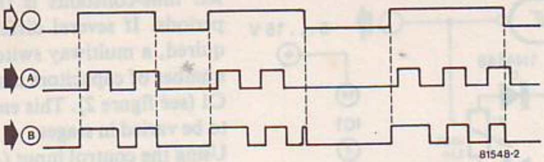
Switching oscillators on and off can sometimes give rise to problems due to the fact that the first or last (or both) pulses can vary in width from maximum down to almost non-existent. In most cases, it is probably true to say that the very narrow pulse or 'spike' will be the one to cause the problems. It all stems from the fact that the oscillator switch off time is invariably not synchronised with oscillator output.

Figure 1B shows the usual simple gate oscillator that is often found in digital circuits, probably chosen for its simplicity and economy as much as anything. However, it does suffer from the problems mentioned above as can be seen from the waveform illustration in figure 2 Ⓑ. The variation in pulse width is readily apparent. The last pulse in the second set may prove to be too small for some gates to 'see' while others in the system can. The result could be an extensive 'red

1

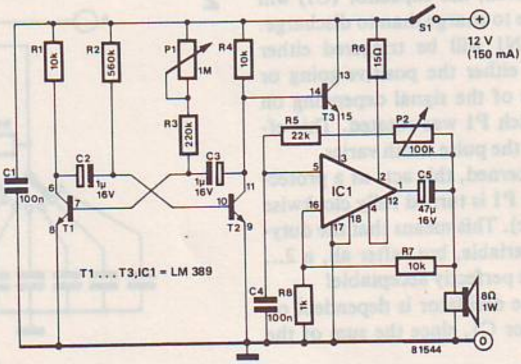


2



herring' chase.
An effective solution to the problem is shown in the circuit in figure 1A where the simple gate oscillator is coupled to an RS flipflop. Diode D1 prevents capacitor C1 from charging during the time that the oscillator is switched off. This ensures that the first pulse at the output has the same width as those following. This is illustrated in figure 2 (A). The sacrifice of two extra gates may well be worth the benifits that this constant pulse width oscillator can provide.

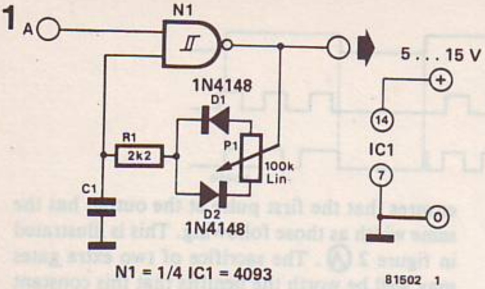
271 Single IC siren



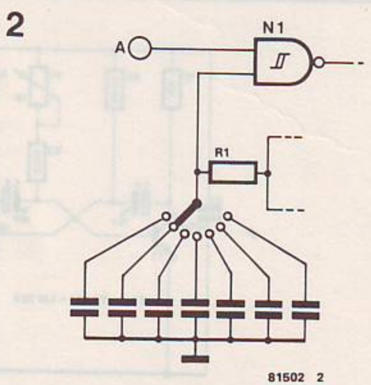
Circuits that produce some sort of noise appear to be highly popular with many readers. A possible reason for this is that correct circuit operation is verified audibly and without the need for test equipment, in other words, the circuit does something in a physical sense.
This particular siren is very simple and easy to construct since it is built around a single IC, the LM 389 from National Semiconductors. This IC contains an audio power amplifier, similar to the LM 386, together with three uncommitted NPN transistors.
The two transistors T1 and T2 form the basis of an astable multivibrator with a frequency variable between 1 and 7 Hz. The preset P1 is used to adjust this. The amplifier is also configured as a squarewave oscillator and its output drives the loudspeaker at frequencies variable between 250 Hz and 1500 Hz. The amplifier, however, is switched on and off by the multivibrator via transistor T3. The result is a pulsed siren like sound. The frequency of the audible tone is adjusted by the preset P2.

272 Pulse generator with variable duty-cycle

A single 4093 CMOS IC is eminently suitable for constructing a simple pulse generator. The IC contains four Schmitt-triggers. By adding a resistor, two diodes, a capacitor and a potentio-

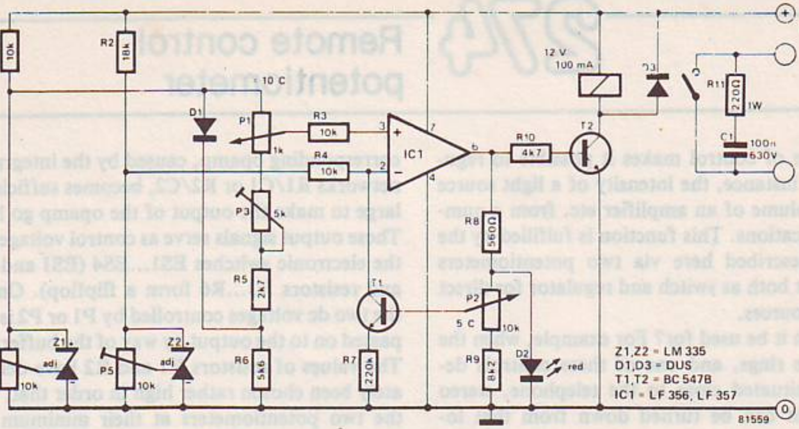


meter, one of the four gates can be used to produce an oscillator with a set frequency and a variable duty-cycle. The pulse duration is determined by the RC time-constant of the network consisting of capacitor C1 and resistors R1 + P1. When the wiper of the potentiometer is in the mid position a perfectly symmetrical squarewave signal is obtained at the output. If, however, the setting of P1 is altered, the capacitor (C1) will take a different time to charge than to discharge. As a result, gate N1 will be triggered either sooner or later on either the positive-going or negative-going edge of the signal depending on the direction in which P1 was rotated. This effectively means that the pulse width varies.
As far as R1 is concerned, this acts as a protective measure in case P1 is turned fully clockwise (minimum resistance). This means that the duty-cycle is not 100% variable, but, after all, a 2... 98% range should be perfectly acceptable!
The frequency of the oscillator is dependent on the value of capacitor C1, since the sum of the RC time-constants is the same for both half-periods. If several different frequencies are required, a multiway switch with a corresponding number of capacitors may be included to replace C1 (see figure 2). This enables the pulse duration to be varied in stages.
Using the control input (A) the entire unit can be incorporated in a logic circuit. If the input voltage is logic zero, the output will be logic one; if, however, the input level is logic one, the oscillator will start to operate. If the control input is not required, this may be omitted by either linking the input to the junction of R1/C1 or to the positive rail of the power supply (logic one).
Although the edge of the output pulse is fairly steep already, it can be further improved by connecting one of the other gates in the same IC to the output. The second gate will then act as an inverter.



273 Differential switch

There is virtually no doubt that interest in this particular circuit will increase as the price of electricity – and of energy in general – continues to rise. The differential switch is able to measure the difference in temperature between two points and, depending on the temperature difference, it will switch a relay on or off. The relay can then be used, for example, to activate a circulation pump. There are numerous applications for the circuit. It can be used in combination with solar heating panels or solar collectors and it can also be used to control the pump in central heating systems. In the latter case, one sensor is placed in the return pipe while the other is situated in the hot water outlet pipe close to the boiler. As soon as the boiler switches on, a temperature difference is created and the pump also switches on.
The attractive feature of this design is the fact that both the temperature difference and the hysteresis of the unit can both be set independently, so that they do not affect each other. Moreover, the adjustments are virtually linear, therefore the potentiometer settings can be relied on to give consistent results. A LED has been included in the circuit to give an indication of when the relay is actually on.
The temperature sensors are two LM 335s



(National Semiconductor). This IC can be looked upon as being a zener diode whose voltage increases by 10 mV per °C. Therefore, at room temperature the zener voltage is equal to: (273 + 20) x 10 mV = 2.93 V.

The temperature transducers incorporate calibration connections, which make it possible to set the output voltage (at 20 °C) to the value mentioned above. In the same way, undesirable differences between the sensors can be corrected. It is also possible to disregard the adjustment input of one of the sensors (by not connecting it) and to adjust the other sensor to give the same characteristics as the first. This can make construction and setting up considerably simpler.

The principle of operation is as follows: The voltages from the two sensors are directly compared by IC2. When the temperature – and thus the voltage – of Z1 becomes greater than that of Z2, the output of IC2 goes high lighting LED D2 and activating the relay via transistor T2. If potentiometer P1 has not been turned fully up, a higher input voltage is required to operate the comparator and the relay will therefore be activated at a higher temperature difference. There is a potential drop of about 0.6 V across diode D1. Approximately 100 mV of this remains across P1 (the actual voltage drop across P1 can be adjusted by means of P3). The 100 mV corresponds to about 10°C, so in effect P1 can be adjusted over a range of 10°C. Sensor Z1 must therefore be 10°C warmer than Z2 with P1 at the lowest setting in order to activate the relay.

Once the pump has been switched on by the relay, the temperature of the sensor close to the boiler will drop due to the circulation of the water. This could result in the circuit switching itself off almost immediately. Obviously, this

situation is undesirable and for this reason potentiometer P2 has been included to adjust the amount of hysteresis by a maximum factor of 5°C. With P2 set in the centre position the circuit has a hysteresis of 2.5°C. This means that if P1 has been set to, say, 5°C, the relay will be activated when the temperature difference reaches 5°C, but will not turn off until the difference in temperature is 5°C – 2.5°C = 2.5°C.

LED D2 should be a red one with an operating voltage of about 1.3 V. The supply voltage for the circuit is not critical and can deviate by a few volts. The circuit diagram shows a supply voltage of 12 V because relays operating at this voltage are readily available. Transistor T2 is only allowed to dissipate a maximum of 100 mA and for this reason the current rating of the relay should not exceed this value.

The actual temperature at which the circuit operates can be calculated from the voltage across Z1 and Z2, if a thermometer is not available.

274 Remote control potentiometer

This type of control makes it possible to regulate, for instance, the intensity of a light source or the volume of an amplifier etc. from a number of locations. This function is fulfilled by the circuit described here via two potentiometers which act both as switch and regulator for direct current sources.

What can it be used for? For example, when the telephone rings, and one of these controls devices is situated close to the telephone, stereo equipment can be turned down from that location, provided the stereo system incorporates dc-controlled ICs such as the TCA 730 or TCA 740. As already mentioned, the circuit can also be connected, via opto-couplers, to a light source and therefore act as a dimmer control. No doubt inventive readers can think of many more applications for the unit.

When the potentiometer setting is altered, an electronic switch will automatically close allowing the dc voltage level on the wiper of the potentiometer to be passed through to the output.

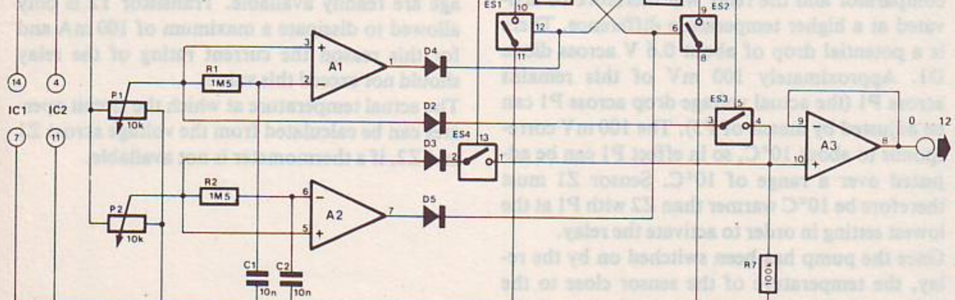
How does it work? The 'hot end' of the two potentiometers (P1 and P2) is kept at about 12 V by means of zener diode D1. As the input range of opamps A1 and A2 is 0 V to 13.5 V, this gives sufficient protection against input overload. When the setting of one of the potentiometers is altered, the potential difference between the inverting input and the non-inverting input of the corresponding opamp, caused by the integration networks R1/C1 or R2/C2, becomes sufficiently large to make the output of the opamp go high. These output signals serve as control voltages for the electronic switches ES1...ES4 (ES1 and ES2 and resistors R3...R6 form a flipflop). One of the two dc voltages controlled by P1 or P2 is also passed on to the output by way of the buffer A3.

The values of resistors R1 and R2 have deliberately been chosen rather high in order that, with the two potentiometers at their minimum setting, the output of the opamps will be low. Diodes D2 and D3 are included so that the flipflop does not return to its original state during the actual transition.

One of the not so good aspects of the circuit should also be mentioned at this stage. This is the fact that when it is necessary to set the previously unadjusted potentiometer to a low dc output, it must first be quickly turned up and then down. This may appear slightly awkward at first, but it will not take long to get the hang of it.

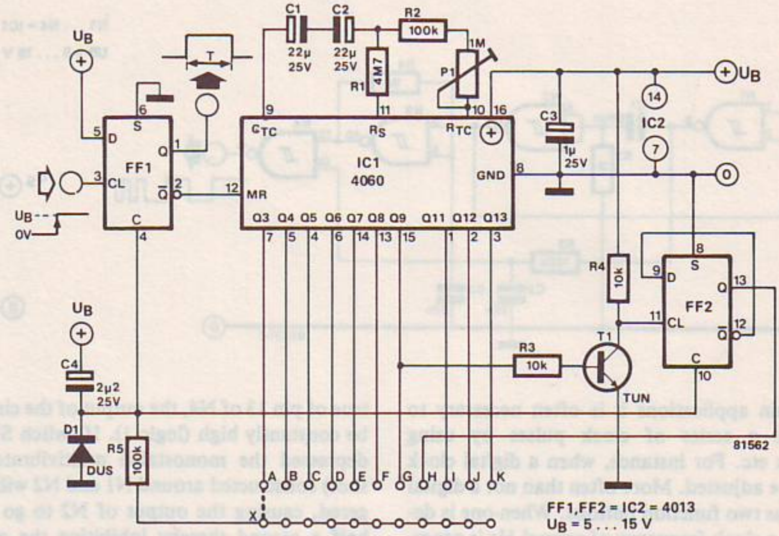
If the connecting leads to the circuit are rather on the long side, it will be necessary to include a 10µ /16 V capacitor between the 'hot' ends of the potentiometers and ground.

R. Behrens



ES1 . . . ES4 = IC1 = 4066
A1 . . . A3 = ¾IC2 = LM324, CA324
D2 . . D5 = 1N4148

275 Long-period timer



This monostable multivibrator can be seen as a long-period alternative for timer circuits which incorporate the well-known 555. It allows the user to obtain on-off periods varying from 20 seconds to about 60 hours.

The design is quite simple. It consists of a start/reset part, a 'slow' oscillator and a series of flip-flops. Most of this can be found, ready for use, in IC1. For the internal oscillator of IC1, only a further two capacitors (C1, C2), two resistors (R1, R2) and potentiometer P1 are needed. The output signals of the counter in the IC can be reached via the Q outputs. Rather curiously, Q10 is not included and a 'Q10' is constructed by the addition of T1 and FF2. The timer is set in motion by the leading edge (positive-going) of the clockpulse to pin 3 of FF1. The $\overline{Q}$ output (pin 2) then becomes '0', and the oscillator in IC1 starts. The Q outputs of IC1 will then become 'high' in turn to the timing of the oscillator frequency which is adjustable between 2.5 and 25 seconds by rotating P1. Depending on which one of points A, B, C...K is connected to point X, a logic '1' will be sent to the clear input (pin 4) of FF1 via R5 after a short or very long period. This way, the flipflop is cleared, the $\overline{Q}$ output becomes '1', and the oscillator stops. The 'timer' will start up again only after a new start pulse reaches pin 3 of FF1.

Due to the very large number of possibilities, there is a formidable choice of periods to chose from. When points A and X are connected, the time set with P1 can be varied to range from 20 sec. to 3.5 min., with connection B-X this range becomes 40 sec. to 7 min., and so on.

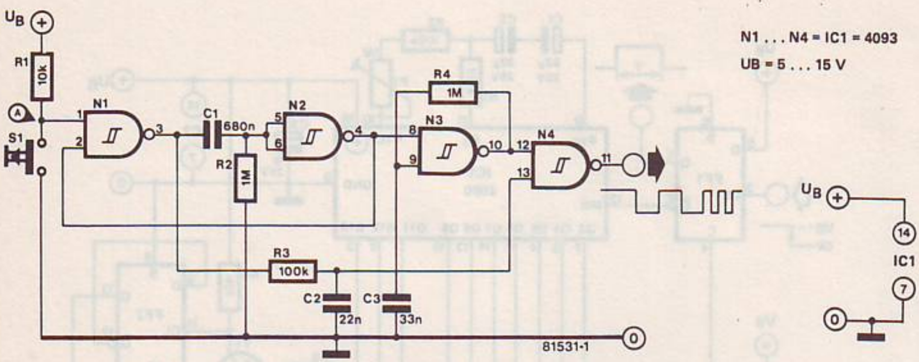
The period can be calculated exactly with the help of a simple formula:

$$T = (M - 0.5) \times 25 \cdot 10^{-6} \times (R2 + P1),$$

in which T is the time and M the selected dividing factor. This factor is 2^3 for connection A-X, 2^4 for B-X, 2^5 for C-X, and so on. For connection K-X, the dividing factor is 2^{13} and, substituting this value in the formula, the respectable period of about 60 hours is obtained.

276 Novel clock control

1



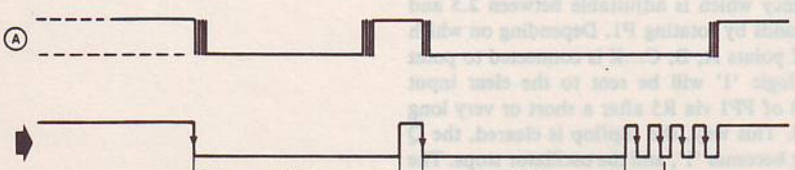
In certain applications it is often necessary to generate a series of clock pulses by using switches etc. For instance, when a digital clock has to be adjusted. More often than not a digital clock has two function buttons. When one is depressed a clock frequency of several Hz is generated enabling the clock to be set to roughly the correct time very quickly. When, on the other hand, the second button is depressed, only one clock pulse is generated so that the clock can be set accurately. Why then two buttons if they both appear to perform what amounts to the same function?

As a matter of fact, there is no real need to have two buttons. The circuit described here fulfils the same task by only using one. Everything now depends on how long the button is depressed for. If the switch is held down for less than half a second, only one clock pulse will be generated. If, however, it is depressed for longer, a clock frequency of 30 Hz will appear at the output of the circuit.

The circuit works as follows. When switch S1 is open the clock generator constructed around N1 will oscillate at a frequency of 30 Hz. However, since the output of N1 is logic zero, as is also true of pin 13 of N4, the output of the circuit will be constantly high (logic 1). If switch S1 is now depressed the monostable multivibrator (one-shot) constructed around N1 and N2 will be triggered, causing the output of N2 to go low for half a second thereby inhibiting the oscillator N3. The output of N1 will now be high, so that the two inputs of N4 are also high. This means that the output of the circuit (N4) will be low (the first clock pulse). If S1 is still depressed after the time delay of the monostable has elapsed, the output of N1 will remain high, the output of N2 will be high, therefore N3 will now oscillate and the pulse train thus produced will be fed to the output of the circuit. If, on the other hand, S1 had been released before the end of the half second period, Pin 13 of N4 would have become logic zero the moment that N3 started to oscillate again. As a result, the output would go high once more.

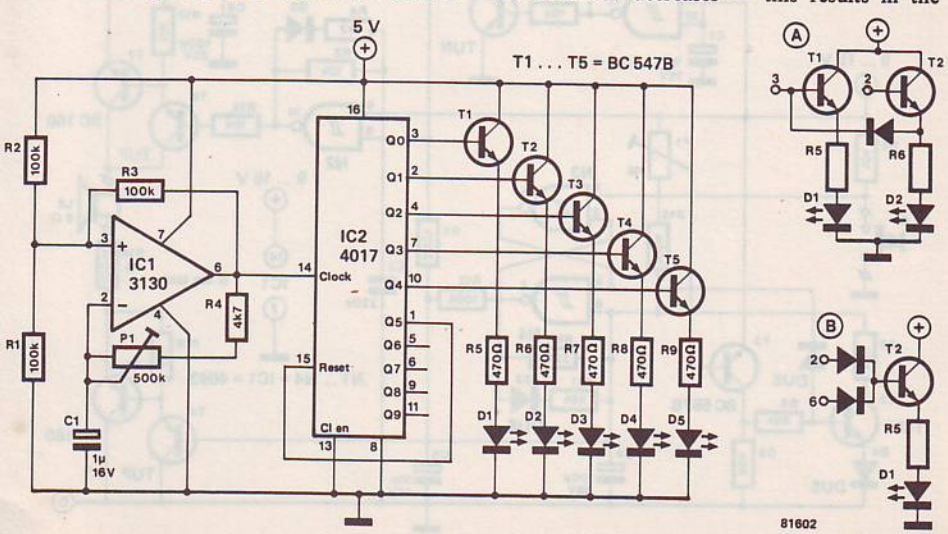
The waveforms involved are shown in figure 2. The vertical lines in signal A represent the contact bounce caused by switch S1. This contact bounce is suppressed by the RC network R3 and C2.

2



277 Novel flashing light

As you may have already guessed, this circuit represents no more or less than a method of illuminating a series of LEDs. The diagram shows five LEDs, but this can be extended to a maximum of ten by connecting the reset line (shown connected to output 5) to the next output, or, if ten LEDs are used, by omitting the reset line altogether. Of course, for each extra LED a corresponding output stage consisting of a transistor and a resistor will also have to be added.
What can the circuit do?
In the simplest configuration, all the LEDs light up in turn. The rate at which this happens is determined by the setting of potentiometer P1. Other sequences can be found by incorporating some clever little 'tricks'. However, the basic configuration can certainly be useful, for example, in the case of a model of a road obstruction where they use those yellow warning flashers which light up one after the other. The illustration at Ⓐ shows how the LEDs can be made to light and remain lit in sequence by simply including a diode between each output stage. The cathode of the diode is connected to the base of T1 and the anode is connected to the emitter of T2. A second diode is connected between the base of T2 and the emitter of T3 and so on. This simple arrangement is quite adequate to produce the desired effect.
LEDs which light up in a 'to-and-fro' fashion can be obtained by connecting the bases of the transistors to the outputs of IC2 in the following manner: The cathode of the first diode is connected to the base of T1 and the anode is connected to pin 3 of the 4017. Two diodes are connected to the base of T2, one goes to pin 2 and the other to pin 6. Similarly, T3 is connected to pins 4 and 5, T4 is connected to pins 7 and 1, and finally T5 is connected to pin 10.
By altering the pin numbering a totally 'random' display sequence can be obtained. Remember that if more outputs are to be used the reset connection must be moved to the next, unused output.
The effects can also be combined: all sorts of 'weird and wonderful' sequences can be obtained by placing diodes between the outputs of IC2 and the transistor bases and between the base of one and the emitter of the next.
As it is quite likely that this circuit will be constructed by model builders who may have relatively little electronic experience, a brief explanation of how the circuit works may be advantageous.
Opamp IC1 generates squarewave pulses by continuously charging and discharging capacitor C1. If the resistance of potentiometer P1 is increased the charge and discharge time for this capacitor increases and therefore the frequency of the oscillator decreases – this results in the

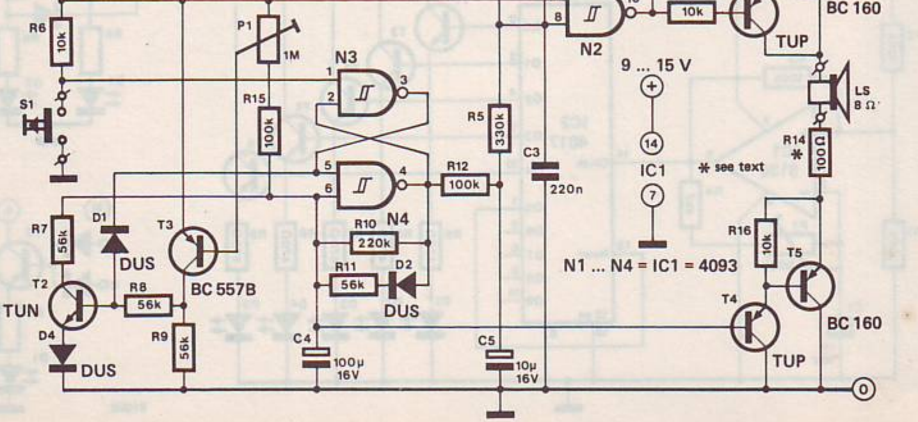


LEDs being lit for longer periods.
The squarewave pulses are fed to a divide-by-ten counter, IC2. The outputs of this counter each go high in turn after every clock pulse. The previous output, which was high, now goes low and the LED connected to it, via the resistor and transistor, will go out, while the next LED in the sequence will come on. Finally, when output five goes high the counter is reset causing a new sequence to be started (LED D1 lights up).

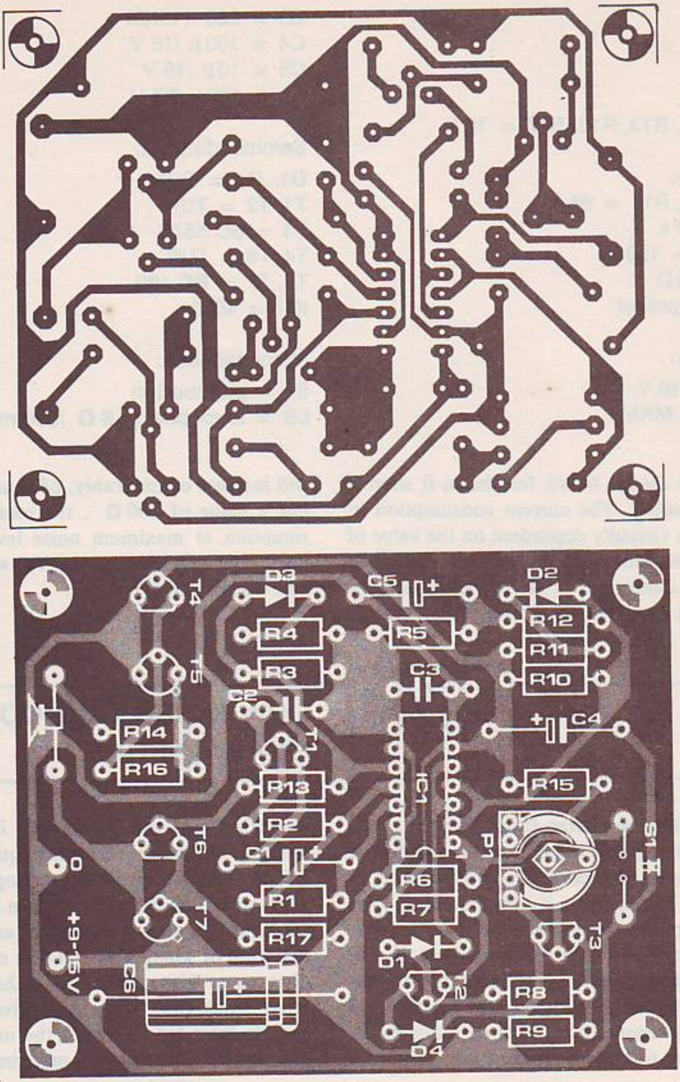
278 'Hi-fi' siren

The title 'siren' above a circuit diagram usually implies a circuit for a two-tone or multi-tone 'horn' which sounds continuously and can be used for an alarm of one sort or another. However, the siren described here is not particularly suitable for that application, but no doubt inventive readers will find a use for it! The hi-fi siren simulates, as accurately as possible, a passing police patrol car with its siren going full blast.
What happens when a police car approaches? At first the siren is only heard very faintly. The noise increases gradually until at the moment it passes the observer the sound reaches maximum intensity. After this the noise decreases immediately while at the same time the pitch gets lower due to the 'Doppler effect'.
By examining the circuit diagram, we can see how this is accomplished electronically. The oscillators constructed around gates N1 and N2 constitute the actual two-tone siren. Initially, the base voltage of transistor T4, and thus the emitter voltage of T5, is approximately equal to the supply voltage. Therefore, no current passes through the loudspeaker. When the start button, S1, is depressed, the flipflop constructed around N3 and N4 changes state and the potential across capacitor C4 decreases slowly. This causes the emitter voltage of T5 to fall also, so that the current through the speaker will start to build up. This current is in fact switched on and off by T6 and T7 to the timing of the double oscillator transistors T6 and T7 in the rythm of the double (N1/N2), which in effect produces the required sound.
The current, and therefore the noise, continues

1



2



to increase gradually due to the falling voltage across C4. At a certain moment this voltage reaches the negative trigger threshold of N4, so that the flipflop is reset and the output of N4 goes high. At this particular moment the voltage on the emitter of T5 is at a minimum and the noise level is at a maximum. As the output of N4 has now gone low, the frequency of the oscillator constructed around N2 is lowered somewhat, thereby simulating the Doppler effect. Capacitor C4 will then slowly recharge up to the supply voltage level, leading to a gradual lowering of the noise level.

The circuitry around transistors T2 and T3 ensures that the noise level increases very slowly at first, but it will then speed up. This makes the sound more realistic. The resultant effect can be adjusted by means of potentiometer P1 and should be done 'by ear'. To do this, the potentiometer is turned fully clockwise, so that the wiper is at full supply voltage level. The start button is then depressed and the noise produced is listened to. The potentiometer is then turned back a little way and the pushbutton depressed once more. This procedure should be repeated until the siren sounds as realistic as possible.

Parts list

Resistors:

R1 = 1 M
R2, R3, R6, R13, R16, R17 = 10 k
R4 = 4k7
R5 = 330 k
R7, R8, R9, R11 = 56 k
R10 = 220 k
R12, R15 = 100 k
R13 = 100 Ω
P1 = 1 M preset

Capacitors:

C1 = 1µ /16 V
C2 = 68 n MKM
C3 = 220 n MKM
C4 = 100µ /16 V
C5 = 10µ /16 V
C6 = 100µ /25 V

Semiconductors:

D1...D4 = DUS
T1, T2 = TUN
T3 = BC 557B
T4, T6 = TUP
T5, T7 = BC 160
IC1 = 4093

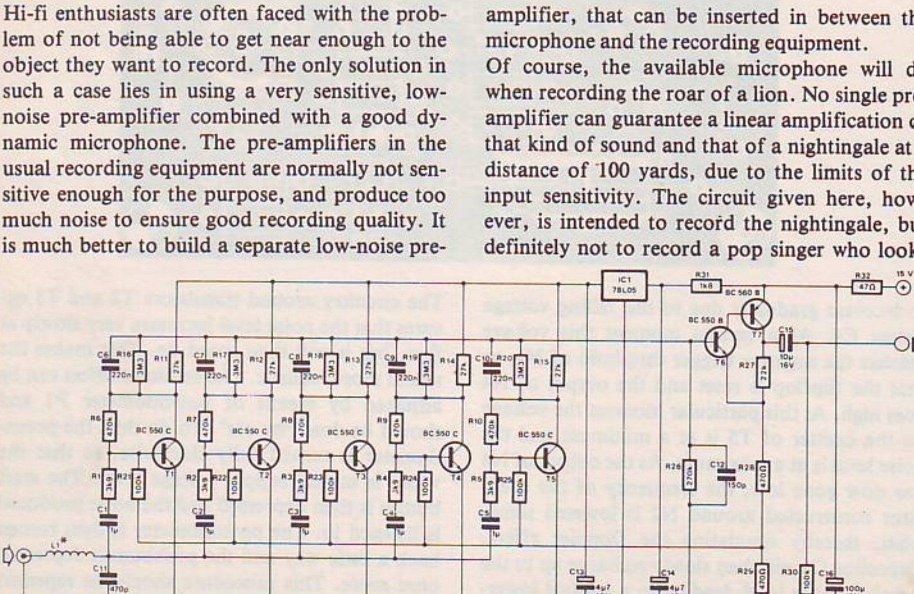
Miscellaneous:

S1 = pushbutton
LS = loudspeaker 8 Ω /500 mW

The printed circuit board for the hi-fi siren is given in figure 2. The current consumption of the circuit is virtually dependent on the value of resistor R14. This resistor can be reduced in value to a minimum of 27 Ω if a louder siren is required. In that case the current consumption will increase considerably, of course. When R14 has a value of 100 Ω, the total current consumption at maximum noise level is approximately 60 mA (U_B = 15 V), while at rest it amounts to only a few mA.

279 Low-noise microphone pre-amplifier

Hi-fi enthusiasts are often faced with the problem of not being able to get near enough to the object they want to record. The only solution in such a case lies in using a very sensitive, low-noise pre-amplifier combined with a good dynamic microphone. The pre-amplifiers in the usual recording equipment are normally not sensitive enough for the purpose, and produce too much noise to ensure good recording quality. It is much better to build a separate low-noise pre-amplifier, that can be inserted in between the microphone and the recording equipment.
Of course, the available microphone will do when recording the roar of a lion. No single pre-amplifier can guarantee a linear amplification of that kind of sound and that of a nightingale at a distance of 100 yards, due to the limits of the input sensitivity. The circuit given here, however, is intended to record the nightingale, but definitely not to record a pop singer who looks



as if he is about to swallow his microphone.

Any transistor produces noise, but to combat this as effectively as possible, noise-free types from a good manufacturer should be selected. On top of that, the circuit should make optimum use of the transistor noise characteristic. It will be noticed immediately that the circuit described here contains 5 parallel stages. The noise generated by each stage can be added in the manner of a vector power, that is, as the square root of the sum of the squares of each component. As the noise of each stage is slightly out of phase, the resultant figure will be slightly lower than that of each stage. Mathematically, this reduction can be expressed as $\sqrt{n}$, so that in this case it is equal to $\sqrt{5}$, in other words, slightly less than 2.3 times. In terms of decibels it can be said that the reduction in noise amounts to 7 dB when compared with a single stage.

A further method of reducing noise is to limit the current through the transistors to a minimum. This is in fact done here, as can be seen from the 1.5 mA (stereo) value. This is even less than the 2 mA required by the voltage regulator IC1, which is included to reduce the supply voltage for the amplifier stages to 5 V. The lower power consumption produces a higher signal-to-noise ratio at the cost of considerable harmonic distortion. Local (R6 to R10) and overall (R21 to R25) feedback ensures the removal of this distortion. The circuit gives excellent results in spite of the 1% distortion which is unavoidable during recording.

What kind of results can therefore be expected from this circuit? The input sensitivity, to give an output of 60 mV, is about 0.13 mV, which is adequate for most applications. The gain of the circuit amounts to around 475 times. Audible distortion occurs with an input signal level of about 8 mV (or greater!) so that fairly strong sound levels can be processed. The frequency response of the preamplifier, between the 3 dB points, is 20 Hz to 45 kHz, the upper limit being determined by the value of capacitor C12. The effect of L1/C11 is negligible as they are included only to prevent interference from local radio stations. Coil L1 consists of a couple of turns of copper wire through a ferrite bead, and can be omitted when there is no likelyhood of radio interference.

This design gives an improvement in signal-to-noise ratio of about 12 dB when compared with conventional designs, making it possible to record that nightingale's song from a distance of 100 m.

With the aid of this preamplifier, the enthusiast can make music recordings or outside recordings of really amazing quality. The microphone and recording equipment will naturally contribute greatly to the actual quality achieved.

As the gain of the preamplifier is dependent on the value of resistor R27, the gain can be altered by selecting a different value for this resistor. However, strictly speaking, the collector voltage of transistor T7 needs to be kept at a level of 7.5 V. Effectively, the value of R26 will also have to be altered. A smaller value for R27 means a larger value for R26. For a gain of 200, R27 should have a value of 10 kΩ : and the value of R26 should be 680 k Ω .

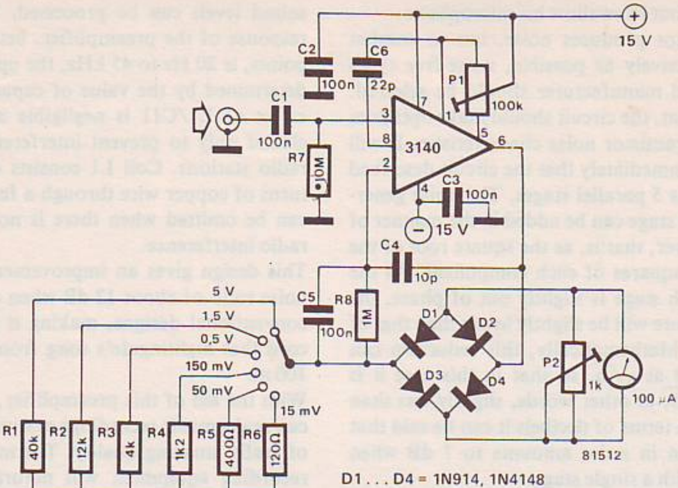
By incorporating a resistor in series with L1, the input impedance can be increased, but the same effect can be obtained by changing the values of R1...R5. It will be obvious that all resistors should be low-noise metal film types after having gone through all this trouble to eliminate the noise generated by the transistors.

P. de Bra

280 Extended range milli-voltmeter

A multimeter, as the name implies, is a multi-purpose measuring tool, although it has its limits. For example, its range for measuring AC in the audio band is usually inadequate, and the sensitivity, internal resistance and frequency response of the cheaper moving-coil multi-purpose instrument normally leaves quite a lot to be desired. The widerange millivoltmeter described here closes that gap in a very simple and elegant way. The instrument can be used to measure alternating current of frequencies between 100 Hz and 500 kHz. When using MOS-FET input opamps the input impedance at all measurement ranges will amount to 10 M Ω .

At the lowest measuring voltage of 15 mV the sensitivity is such that there is a full-scale deflec-



tion on the 100 μ A meter.

The opamp serves both as measurement amplifier and active rectifier. The level of amplification is determined by the switched resistors R1 to R6. With the instrument set at a particular sensitivity range, the value of a resistor can be determined simply by dividing the input voltage for a full deflection by 100 μ A. When, for example, at the measuring range of 150 mV, a 200 mV range is expected to be needed, resistor R4 should be changed to a value of 2 k.

Because the bridge rectifier diodes D1 to D4 are located in the feedback loop of the amplifier, there is compensation for the threshold voltage of the diodes, for which reason the mV scale responds in a linear fashion.

The meter is zeroed with the aid of P1 and the input short-circuited, while the measuring range is determined by P2. The latter requires a calibration voltage that can be obtained from a small mains transformer with a secondary voltage of slightly less than 5 V. At this level, the voltage can be measured quite accurately with the aid of the multimeter. The calibration voltage should then be connected to the extended-range milli-voltmeter set at 5 V, and the reading of the 100 μ A instrument is then adjusted, using P2, to the value of the calibration voltage. The other measuring ranges are then set simultaneously corresponding to the tolerance of the resistors R1 to R6.

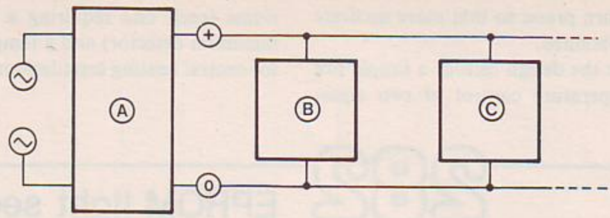
When the circuit is used to extend or supplement an existing multimeter, the moving-coil part of the multimeter should be used in the 100μ A range. The best power supply to use in that case is about 9 V, obtained from two small 9 V dry cells which will last quite a long time as the power consumption is very low.

281 Temperature alarm

This design allows a simultaneous check of a maximum of four temperatures. The individual detectors are connected to the main control unit by means of a pair of wires. When one of the detectors registers an abnormally high or low temperature, an alarm sounds. The positioning of resistor R10 determines whether the alarm is activated when the registered temperature rises above the preset temperature setting or whether it sounds when it falls below the required temperature.

When R10 is incorporated between the positive supply line and the output of IC3, the alarm will be triggered when the measured temperature exceeds that of the setting of P1. In this situation the output of IC3 falls and the detector starts to

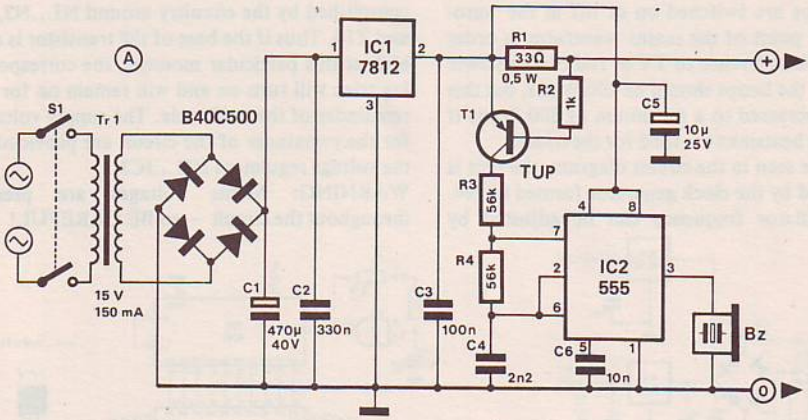
1



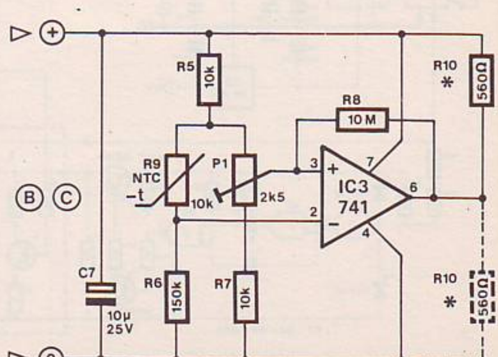
A = principal
B = secondary 1
C = secondary 2

draw a current of around 20 mA. This means that the voltage across R1 becomes greater than 0.6 V causing transistor T1 to conduct and the alarm to sound. This occurs because as the temperature rises the resistance of the NTC resistor (R9) decreases taking the inverting input of IC3 higher than the non-inverting input (voltage-wise). This in turn causes the output of the op-amp to go low.
If resistor R10 is placed between the output of IC2 and ground, the reverse happens. In this instance, as the temperature drops the resistance of R9 increases taking the inverting input of IC3

2



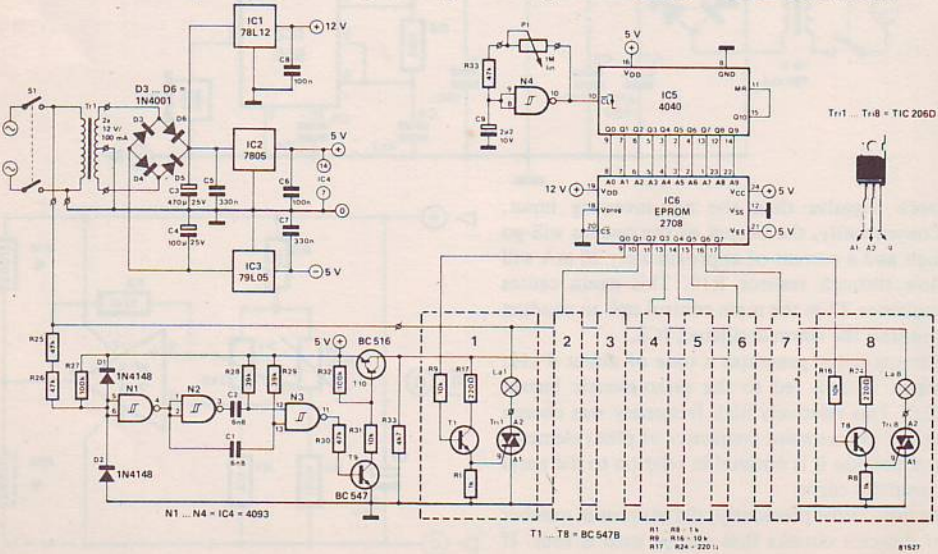
more negative than the non-inverting input. Consequently, the output of the opamp will go high and a current of approximately 20 mA will flow through resistor R10. This again causes transistor T1 in the main control unit to conduct and start the alarm oscillator, IC2.
The oscillator generates a tone of about 4 kHz which is then fed to the piezo-electric buzzer (Bz). This relatively high frequency was chosen to suit the resonant frequency of piezo elements and because it is optimal in relation to the aural sensitivity curve.
As mentioned previously, the maximum number of detector circuits that can be used is four. If any more were included the quiescent current would exceed that of the 'alarm current' and the design would not function correctly. In retrospect, the alarm current can not be increased as in that case the output current of the opamp would be exceeded – with detrimental results!
If another form of alarm is required, IC2 and its associated components can be omitted and the transistor can be used to control a relay or other similar device. The sensitivity of the circuit, in other words the temperature at which the device operates, can be adjusted by means of the preset potentiometer P1. It may well be advantageous



to use a multi-turn preset so that more accurate settings can be obtained.
Applications for the design include a simple fire alarm, the temperature control of two aquariums (each one requiring a maximum and a minimum detector) and a temperature regulator for central heating installations.

282 EPROM light sequencer

The circuit described here is an eight channel light sequencer. The information for each channel (on or off) is contained in a 2708 type EPROM (these are coming down in price every day!). There are therefore 1024 steps before the 'program' repeats itself. It is possible to program the EPROM yourself if you have access to an EPROM programming device. Otherwise you can use an EPROM which has already been programmed for a computer. This will cause the eight lamps to turn on and off in the strangest of sequences. A test with the monitor program of the Junior Computer showed this to work very well.
The lamps are switched on or off at the 'zero-crossing' point of the mains waveform in order to avoid interference to TV or radio. The power rating of the lamps should be 200 Watts, but this can be increased to a maximum of 800 Watts if adequate heatsinks are used for the triacs.
As can be seen in the circuit diagram, the unit is controlled by the clock generator formed by N4. The oscillator frequency can be adjusted by means of potentiometer P1. The output of the clock generator drives a binary counter, IC5, which counts up from zero to 1023 in binary. The outputs of IC5 are fed to the address inputs of the EPROM, IC6. The contents of each address location are therefore read out sequentially. A logic '1' on a data output of the EPROM will turn the corresponding lamp on via the associated driver transistor and triac. When the output is logic zero, the lamp will be extinguished.
The supply voltage for transistors T1...T8 is only present for about 300 μ s at the zero-crossing of the mains voltage. This is accomplished by the circuitry around N1...N3, T9 and T10. Thus if the base of the transistor is also high at this particular moment, the corresponding triac will turn on and will remain on for the remainder of the half-cycle. The supply voltages for the remainder of the circuit are provided by the voltage regulators IC1...IC3.
WARNING: Mains voltages are present throughout the circuit – so BE CAREFUL!

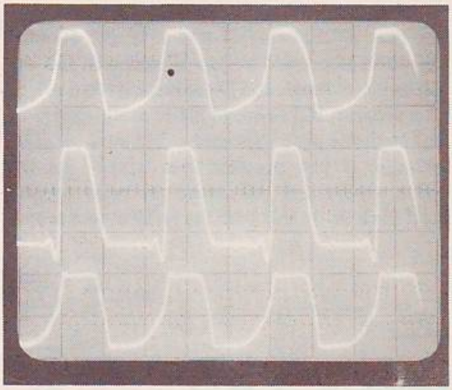


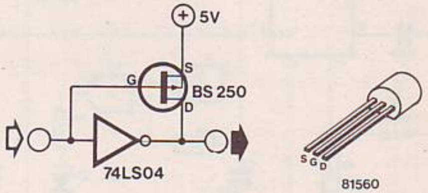
283 Fast TTL interface

It is surprising how often TTL ICs are asked to perform a difficult (if not impossible) task. It is, of course, gratifying to know that experimenters have so much confidence in the capabilities of the TTL logic family. But, on the other hand, why drive the devices to extremes when they are not likely to survive? More often than not the output of this type of IC is overloaded, or they are used, mistakenly, to drive CMOS ICs directly. In the latter case, the guaranteed active pull-up TTL output voltage (2.4 V) is lower than the minimum CMOS input voltage required to guarantee switching (3.5 V). Also, the input impedance of CMOS ICs is essentially capacitive, which means that the slew rate of the TTL output signal will suffer at high frequencies.

The upper trace in the photograph shows the effect when a TTL output is capacitively loaded (220 pF in this instance). The negative-going edges of the signal are still quite acceptable, as TTL outputs can 'sink' more current than they can 'source'. However, the output current will be reduced at the same rate at which the voltage level increases causing the positive-going edge to level off. As would be expected, CMOS ICs do not react favourably to this type of signal. This is because CMOS ICs will not switch until the input level has reached approximately half the supply level, unless the manufacturer has taken the appropriate measures to avoid this. As a result, a delay of about 40 ns will be incurred and the poor signal edge at the input is bound to lead to a far optimum edge at the output.

In the case of the lower signal shown, an effort has been made to improve the positive-going edge by including a pull-up resistor between the TTL output and the positive supply voltage. This resistor (220 Ω) will certainly improve the positive-going edge of the signal, but at the expense of the negative-going edge. The ideal solution, therefore, is to incorporate a VFET which will only operate during the positive-going section of the waveform. The middle signal shows that the amplitude will now be the same as the full supply voltage and the positive-going edge reaches its peak very rapidly despite the capacitive load.





Note: the oscilloscope in the photograph was set up as follows:

vertical deflection – 2 V/div.
horizontal deflection – 100 ns/div (in other words, the frequency of the signal on display is 4 MHz!!).

284 Automatic soldering iron switch

We all know that it is quite easy to forget to turn off the soldering iron. If you then go out of the house with other things on your mind, it is just possible to return to find a heap of smoking rubble. This would be extremely upsetting, especially when this sort of calamity can easily be avoided. The results are not usually so dramatic, but the least you can expect is a bigger electricity bill. The circuit here will iron the problem out and will repay the effort of making it in a very short time.

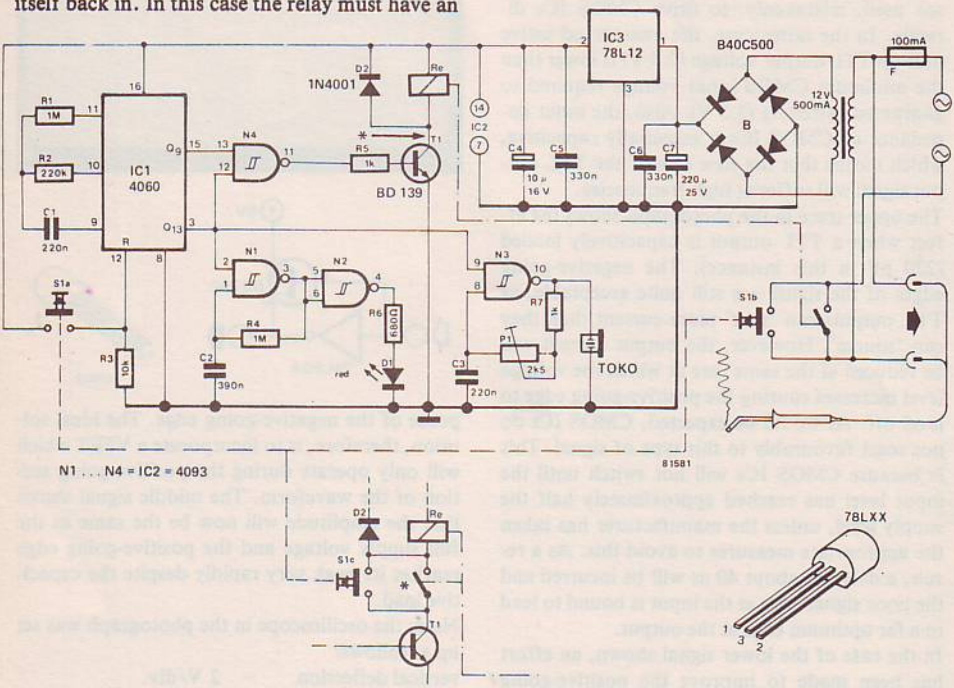
The circuit operates as follows. IC1 is an oscillator divided by 2^{13} which generates a time interval of about a quarter of an hour. At the end of this time a LED flashes and a buzzer sounds. Unless S1 is pressed within 50 seconds, the circuit switches itself, and the soldering iron, off. If S1 is pressed, IC1 re-starts the 15 minute period.

Although the prototype circuit behaved satisfactorily, it is just possible that 'spikes' on the mains supply line will cause the relay to pull itself back in. In this case the relay must have an extra contact, to positively switch it off. This then requires a third pole of S1, to bridge this contact at switch on.

The switching can be done using two separate push switches which must be pressed simultaneously.

If a relay is available with an operating voltage other than 12 volts, then the circuit can be operated at the relay voltage, by changing the power supply and regulator. The supply voltage must, however, be kept between 3 and 18 volts.

M.A. Prins



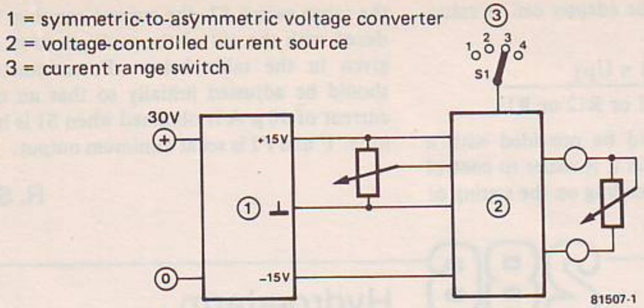
285 Constant current adapter

It is quite often the case that the electronics enthusiast requires a constant current source. When such a need does arise, for example for test purposes, it is a piece of equipment that is not usually available. However, it is not necessary to construct an entire constant current source for each application. It is sufficient to

Table

S1	i	P1 x
1	10 µ A...100µ A	10 µ A
2	100µ A...1 mA	100 µ A
3	1 mA...10 mA	1 mA
4	10 mA...100 mA	10 mA

1



have an adapter that can be connected to an existing power supply whenever a constant current source is required.

The proposed circuit has another useful application: the asymmetric-to-symmetric power supply converter at the front end of the constant current adapter can be used separately to power an amplifier or similar circuit.

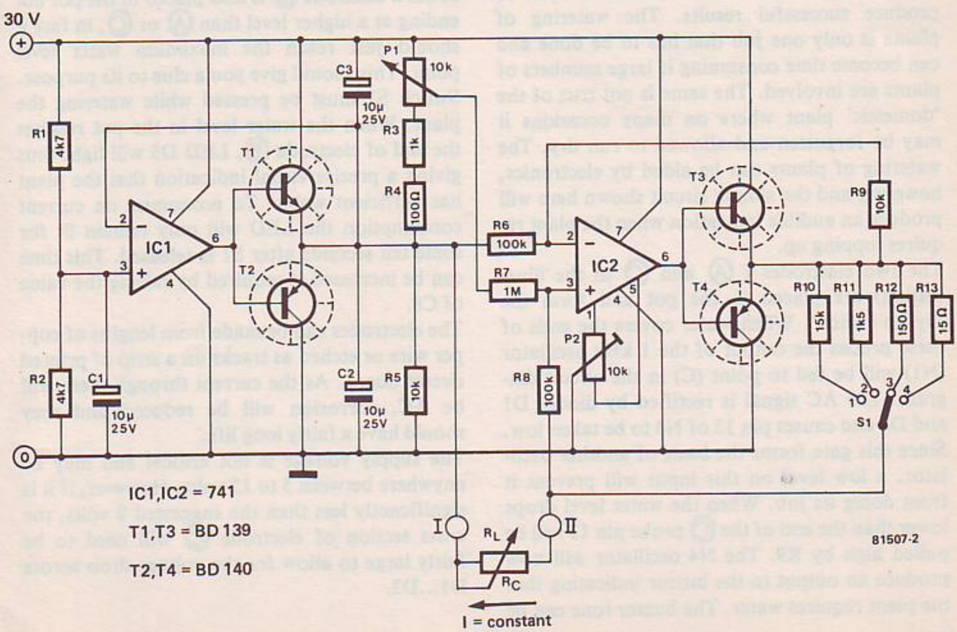
Most amateur constructors own a mains power supply with a variable output voltage of up to 30 V and a current output of around 200 mA (or greater). This supply can then be used to power the constant current adapter.

The asymmetric-to-symmetric converter consisting of IC1 and transistors T1 and T2 provides an effective output voltage of ± 15 V across the two capacitors C2 and C3. This symmetrical supply can be used separately provided the required output current is no greater than about 50 mA.

Let us now examine the constant current adapter itself in greater detail. The asymmetric-to-symmetric voltage converter is required to power the operational amplifier IC2. This opamp is used as the current source and is controlled by the potential divider consisting of potentiometer P1 and resistors R3 and R4. Potentiometer P1 can be adjusted to give an output voltage of between 1.5...15 V.

A constant current will flow through load resistor R_L, which is dependent on the voltage setting of P1 and on the range selected by switch S1. The circuit is such that, regardless of the actual range, the current through R_L is determined by the setting of P1. Transistors T3 and T4 simply form a buffer stage.

2



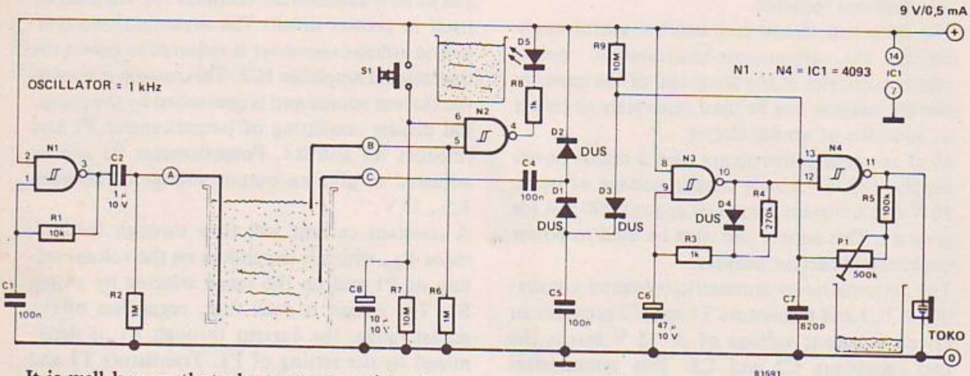
The output current of the adapter can be calculated from the formula.

$$I = \frac{0.1 \times U_{P1}}{R10 \text{ or } R11 \text{ or } R12 \text{ or } R13}$$

Potentiometer P1 should be provided with a scale from 1 to 10 so that it is easier to control the desired current. Depending on the setting of the range switch S1, the current can then be deduced with the aid of the multiplication factor given in the table below. Potentiometer P2 should be adjusted initially so that an output current of 10 μ A is obtained when S1 is in position '1' and P1 is set at minimum output.

R. Storn

286 Hydro-alarm



It is well known that plant growers with a large 'collection' have to work very hard in order to produce successful results. The watering of plants is only one job that has to be done and can become time consuming if large numbers of plants are involved. The same is not true of the 'domestic' plant where on many occasions it may be forgotten and allowed to run dry. The watering of plants can be aided by electronics, however, and the simple circuit shown here will produce an audible indication when the plant requires topping up.

The two electrodes (Ⓐ and Ⓒ in the illustration) are placed in the pot and form the 'hydro switch'. When water covers the ends of these probes the output of the 1 kHz oscillator (N1) will be fed to point (C) in the circuit diagram. The AC signal is rectified by diodes D1 and D2 and causes pin 13 of N4 to be taken low. Since this gate forms the basis of another oscillator, a low level on this input will prevent it from doing its job. When the water level drops lower than the end of the Ⓒ probe pin 13 will be pulled high by R9. The N4 oscillator will now produce an output to the buzzer indicating that the plant requires water. The buzzer tone can be varied by adjustment of the 500 k preset.

A third electrode Ⓑ is also placed in the pot but ending at a higher level than Ⓐ or Ⓒ, in fact it should just reach the maximum water level point. This should give you a clue to its purpose. Switch S1 must be pressed while watering the plant. When the water level in the pot reaches the end of electrode Ⓑ, LED D5 will light thus giving a precise visual indication that the plant has sufficient water. To economise on current consumption the LED will only remain lit for some ten seconds after S1 is released. This time can be increased if required by raising the value of C8.

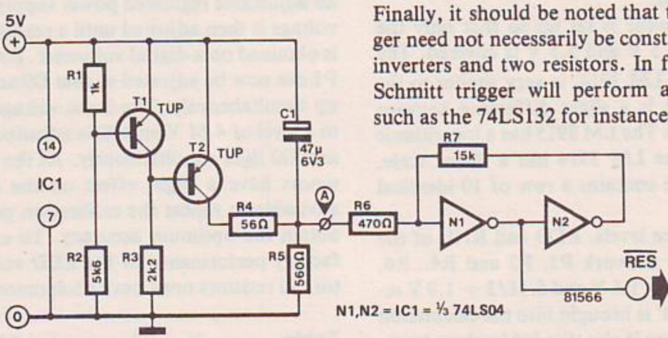
The electrodes can be made from lengths of copper wire or etched as tracks on a strip of printed circuit board. As the current through them will be AC, corrosion will be reduced and they should have a fairly long life.

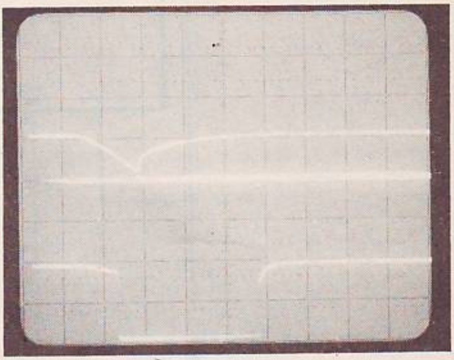
The supply voltage is not critical and may be anywhere between 5 to 15 volts. However, if it is significantly less than the suggested 9 volts, the cross section of electrode Ⓒ will need to be fairly large to allow for the voltage drop across D1...D3.

287 Automatic reset

This circuit is designed to generate a reset pulse whenever the supply voltage is switched on and, in addition, whenever there is any substantial interference 'spikes' on the power supply lines. Most digital circuits (and microprocessor systems in particular) have to be reset for a certain period after the supply voltage is initially switched on. In this instance, an active low reset pulse is generated which will remain logic zero for 30 ms after the supply is switched on. At the end of this duration the supply will be fully on.

The 'automatic reset' works as follows. The section of circuit around transistors T1 and T2 ensures that the voltage across capacitor C1 remains 0 V until the supply voltage reaches a level of 4.5 V. As soon as the supply voltage reaches this value, the transistors will stop conducting and capacitor C1 will gradually charge up via resistor R5. As a result, the voltage at point A will slowly drop from around 4.5 V to 0 V, causing the Schmitt trigger formed by N1 and N2 to be triggered and the output $\overline{RES}$ will go high. In other words, the $\overline{RES}$ output will most definitely be low for about 30 ms after switch-on until the supply voltage level has attained its correct value (+4.75 V for TTL).





As a side-line, the circuit also allows for a reset pulse to be generated if ever the supply voltage should drop below 4.5 V for any reason. In certain computer applications this could well be vital, as such an interference spike, however short, could well erase a large section of memory! For this reason a noise generated reset pulse could well be useful, as the computer operator then knows that the program will always be started from 'square one'.

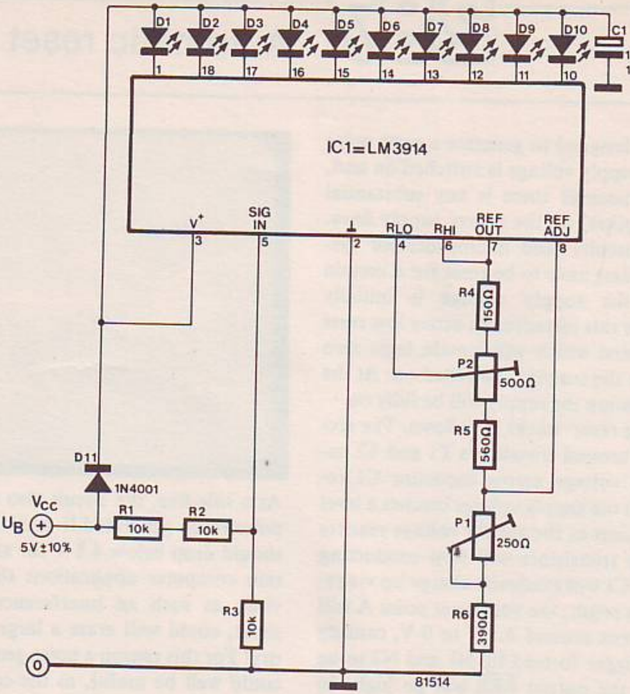
The photograph shows a characteristic form of interference spike and the reset pulse that is generated from it.

Finally, it should be noted that the Schmitt trigger need not necessarily be constructed from two inverters and two resistors. In fact, any type of Schmitt trigger will perform adequately here, such as the 74LS132 for instance.

288 LED voltage monitor

A stable power supply is absolutely essential for the correct operation of computers and TTL circuits. A voltage fluctuation of 10% is certainly not tolerable, therefore it is prudent to keep a regular check on the supply voltage level.

Because of their lack of resolution and accuracy



it is inadvisable to use analogue panel meters to monitor the power supply voltage. Besides this, a fluttering pointer is hardly the best choice for a warning device. The LED voltage monitor solves all these problems.

The voltage monitor is set up so that only the range between 4.5 V and 5.5 V is covered. The device used, the LM 3914, is very similar to the LM 3915. There is a slight difference between the two however: The LM 3915 has a logarithmic scale whereas the LM 3914 has a linear scale. The latter device contains a row of 10 identical 1 k Ω resistors.

The two reference levels, RLO and RHI, of the potential divider network P1, P2 and R4...R6, are set to 4.51/3 = 1.5 V and 5.41/3 = 1.8 V respectively. The '3' is brought into the calculation as the input voltage is also divided by three by resistors R1...R3. The table shows which LEDs will light for the corresponding input voltage once the circuit has been set up correctly.

For a clear warning indication it is best to use red LEDs for D1 and D10 and green ones for the rest. It may also be useful to use a different colour (orange) for D5 and D6 as an indication of the nominal voltage level.

The power supply for the circuit can be taken from the voltage to be monitored as the current requirement is only 20 mA. Diode D11 is included to protect the circuit against reverse input polarity.

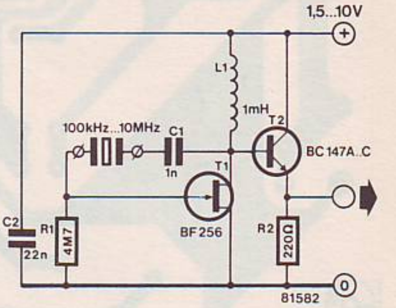
To calibrate the circuit it must be connected to an adjustable regulated power supply. The input voltage is then adjusted until a reading of 5,41 V is obtained on a digital voltmeter. Potentiometer P1 can now be adjusted so that D9 and D10 light up simultaneously. The input voltage is then set to a level of 4.61 V and P2 is adjusted so that D1 and D2 light simultaneously. As the internal resistors have a slight effect on the circuit it is advisable to repeat the calibration procedure to obtain the optimum accuracy. To ensure satisfactory performance of the LED voltage monitor, all resistors must have a tolerance of 5%.

Table	V_{CC} (V)	LED
	4.51...4.60	D1
	4.61...4.70	D2
	4.71...4.80	D3
	4.81...4.90	D4
	4.91...5.00	D5
	5.01...5.10	D6
	5.11...5.20	D7
	5.21...5.30	D8
	5.31...5.40	D9
	5.41...5.50	D10

289 Crystal oscillator... ...for low voltage supplies

It is very easy to construct a crystal oscillator using a field effect transistor. This particular circuit operates at relatively low supply voltages, from 1.5 volts upwards, and was tested with common-or-garden quartz crystals with frequencies ranging from 100 kHz to 10 MHz.

The crystal is connected between the drain and gate of the field effect transistor, T1, and operates in the parallel resonance mode. Coil L1 is included to improve the frequency range. Furthermore, it helps as an additional parallel coil for those crystals which are not particularly suited for this application and which do not feel like oscillating. Capacitor C1 is the series 'padding' capacitor for the crystal. The necessary feedback and the 180° phase shift is provided by the internal input and output capacitances of the FET. The output signal is buffered by transistor T2.



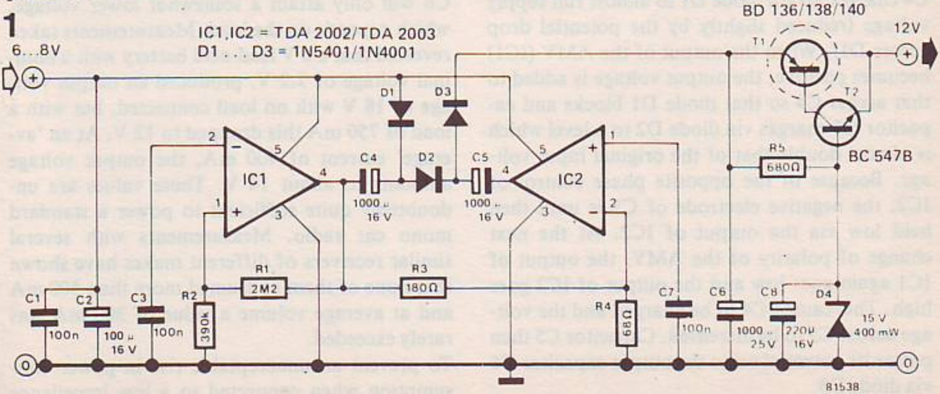
This circuit was tested with the following range of crystals: 100 kHz, 1 MHz, 4 MHz, 6 MHz, 8 MHz and 10 MHz. The circuit can be used in a variety of applications due to its low supply voltage requirements (1.5 V minimum).

290 6 to 12 volt converter

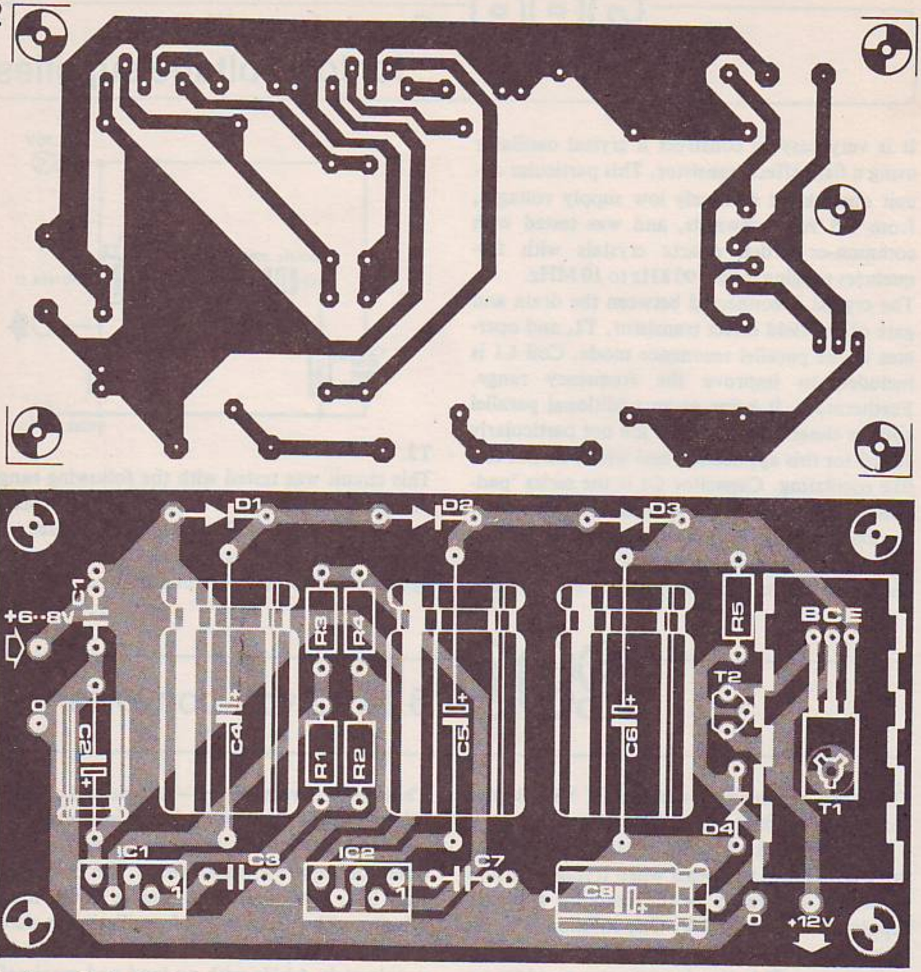
There are an amazing number of VW Beetles and Fords on the road which still operate on a 6 V battery. In such vehicles (and motorcycles) there are always problems when trying to install a modern car radio as they require a power supply of at least 10.7 V. One solution is to incorporate a 6 to 12 V converter of the type described here. This simple converter provides an output of around 700 mA and is relatively inexpensive to construct.

These two characteristics – simple and cheap – arise from the concept of the circuit which contains two integrated audio power amplifiers and does not require a transformer. The first amplifier, IC1, functions as an astable power multivibrator. The frequency of oscillation is determined by the value of capacitor C3 and is approximately 4 kHz with no load and maximally 6 kHz when a load is applied. The output signal of a second amplifier, IC2, is identical to that of

1



2



the first, albeit 180 degrees out of phase.

When the output voltage of IC1 is low, capacitor C4 charges up via diode D1 to almost full supply voltage (reduced slightly by the potential drop across D1). When the output of the AMV (IC1) becomes positive, the output voltage is added to that across C4 so that diode D1 blocks and capacitor C5 charges via diode D2 to a level which is almost double that of the original input voltage. Because of the opposite phase control of IC2, the negative electrode of C5 is until then held low via the output of IC2. At the next change of polarity of the AMV, the output of IC1 again goes low and the output of IC2 goes high. This causes C4 to be charged and the voltage across C5 to be increased. Capacitor C5 then passes its potential on to the output capacitor C6 via diode D3.

In theory, therefore, the final effect of the circuit is to treble the input voltage, but in practice C6 will only attain a somewhat lower voltage, which depends on the load. Measurements taken revealed that a 6 V lead-acid battery with a nominal voltage of 7.2 V, produced an output voltage of 18 V with no load connected, but with a load of 750 mA this dropped to 12 V. At an 'average' current of 400 mA, the output voltage amounts to about 14 V. These values are undoubtedly quite sufficient to power a standard mono car radio. Measurements with several similar receivers of different makes have shown that none of them consumed more than 500 mA and at average volume a value of 300 mA was rarely exceeded.

To prevent an unacceptable rise in power consumption when connected to a low impedance

Parts list

Resistors:

R1 = 2M2
R2 = 390 k
R3 = 180 Ω
R4 = 68 Ω
R5 = 680 Ω

Capacitors:

C1, C3, C7 = 100 n
C2 = 100 μ /16 V
C4, C5 = 1000 μ /16 V
C6 = 1000 μ /25 V
C8 = 220 μ /16 V

Semiconductors:

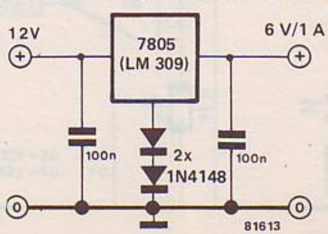
T1 = BD 136/138/140
T2 = BC 547B
D1, D2, D3 = 1N5401/1N4001
D4 = 15 V/400 mW zener diode
IC1, IC2 = TDA 2002/TDA 2003

load, the converter is provided with an additional limiter stage consisting of a 15 V zener diode and a complementary Darlington circuit (transistors T1 and T2). This arrangement limites the maximum voltage to about 14,2 V. At the same time, capacitor C8 connected to the two transistors reduced the ripple of the output voltage to less than 50 mV under full load conditions. During practical trials no effect of the oscillating frequency of the converter on the quality of radio reception was noticed

The printed circuit board for the converter is shown in figure 2. Due to its small size, construction of the circuit should not pose any problems. Both IC amplifiers and transistor T1 can be kept sufficiently cool if these components are mounted (with mica washers) on a common heatsink along the longest side of the board. The heatsink should be as large as the board itself and should be mounted at 90 degrees to the board in order to guarantee an optimum heat transfer. Both IC amplifiers contain integrated protection circuitry against short circuits and thermal overload, so that the worst need not be feared if the unit is subjected to overload or overheating.
Either the TDA 2002 or the TSA 2003 can be used for the amplifiers. The TDA 2003 has the edge on the 2002 due to a few improved characteristics. The same holds true for the diodes; the 3 A diodes (1N5401) are best suited because less voltage is dropped across them. When 1N4001 types are used, a loss in output voltage of 0.5 V to 1 V should be expected.
If the values of capacitors C4, C5 and C6 are increased to 200 μ F, the maximum output current is raised by about 100 mA. For even higher output currents, two converters can be connected in parallel. In that case, the limiting stage (R5, C8, D4, T1 and T2) is omitted from the second board and a connection made between the two positive electrodes of the two C6s. Transistor T1 can then be one of the following types; BD 236, BD 238, BD 204, BD 288 or BD 438. The maximum current that can be obtained by connecting two converters in parallel is nearly doubled to about 1.3 A, therefore stereo or cassette radios can be installed in 6 V cars quite easily.

291 12 V to 6 V converter

After having described how a modern 12 V car radio can be installed in a vehicle with a 6 V system with the aid of a 6 V to 12 V converter, it may be a good idea to look at the other side of the coin – where a lower voltage is required from a 12 volt system. The most common application for this type of converter is when portable cassette recorders are to be used in the car. Many of these require a supply voltage of between 5...8 volts.
The simplest and most obvious solution is to use an integrated voltage regulator. Apart from utilising 6 volt types such as the 7806, it is also possible to use a 5 volt version (7805 of LM 309) and boost its output voltage by including two diodes



in the common lead as shown in the circuit diagram. Depending on the type of diodes used, this will produce an output voltage between 6 V and 6.5 V. The maximum output current of the types mentioned is 1 A. It is important to ensure that the regulator is sufficiently cooled by means of a suitable heatsink.

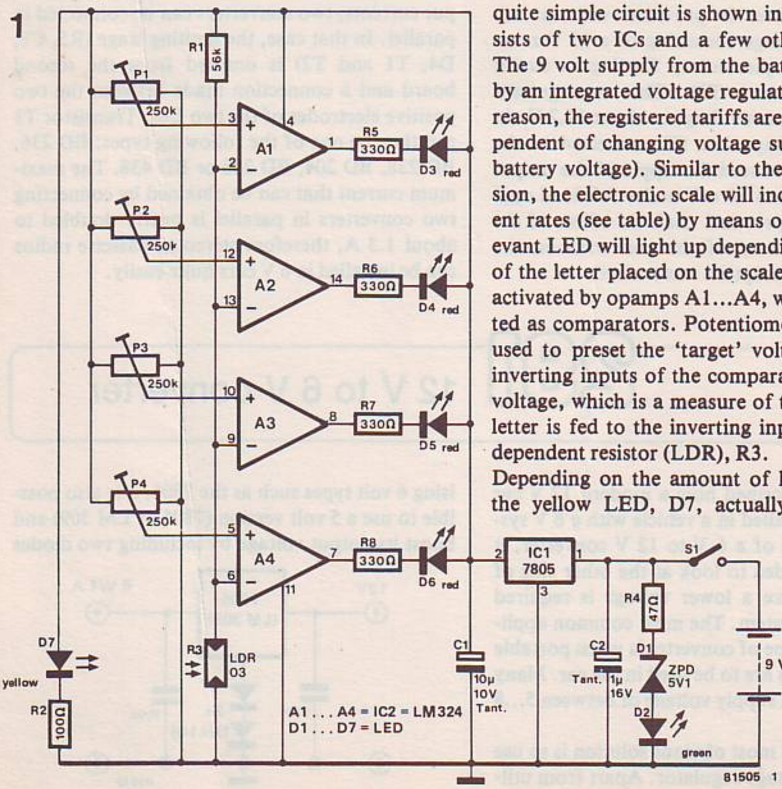
Combined radio/cassette players very often require a slightly higher voltage of 7.5 V. In this instance, either the 7808 can be used, or the 7805 with four diodes in series with the common lead.

292 Post office letter scales

No doubt, everyone is familiar with the kind of scales to be found on Post Office counters. A letter or small parcel is placed on it and the weight and the required postage can be read off it. There is no reason why this can not be accomplished electronically. Of course, some mechanical ingenuity will be required as the equipment will include some form of scale or balance mechanism. Moreover, the electronic scales should look similar to the mechanical counterpart. If the latter were not the case, nobody would think of putting a letter on to weight it!

Table	LED	weight in g.
	–	less than 60
	D3	60...100
	D3, D4	100...150
	D3...D5	150...200
	D3...D6	over 60

1



Let us first consider the electronic details. The quite simple circuit is shown in figure 1. It consists of two ICs and a few other components. The 9 volt supply from the battery is stabilised by an integrated voltage regulator, IC1. For this reason, the registered tariffs are reasonably independent of changing voltage supplies (a falling battery voltage). Similar to the mechanical version, the electronic scale will indicate five different rates (see table) by means of LEDs. The relevant LED will light up depending on the weight of the letter placed on the scales. The LEDs are activated by opamps A1...A4, which are connected as comparators. Potentiometers P1...P4 are used to preset the 'target' voltage at the non-inverting inputs of the comparators. The actual voltage, which is a measure of the weight of the letter is fed to the inverting inputs via the light dependent resistor (LDR), R3.

Depending on the amount of light radiated by the yellow LED, D7, actually falling on the

Parts list

Resistors:
R1 = 56 k
R2 = 100 Ω
R3 = LDR 03
R4 = 47 Ω
R5...R8 = 330 Ω
P1...P4 = 250 k preset

Capacitors:
C1 = 10µ /10 V tantalum
C2 = 1µ /16 V tantalum

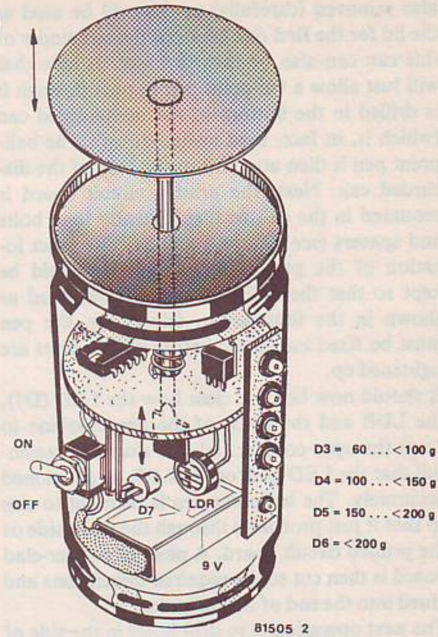
Semiconductors:
D1 = 5V1/400 mW zener diode
D2 = green LED
D3...D6 = red LED
D7 = yellow LED
IC1 = 7805
IC2 = LM 324, CA 324

Figure 1. The circuit diagram of the scales. It consists of two integrated circuits and a few extra components. LEDs D3...D6 indicate the various weights and postal charges. The opto-coupler consists of LED D7 and the LDR R3.

Figure 2. An exploded view of the scales. The tip of the ballpoint pen controls the amount of light that falls on the LDR.

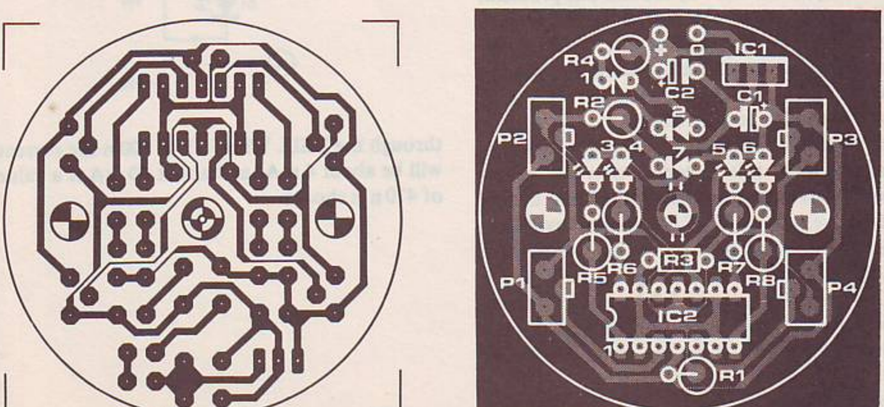
Figure 3. The printed circuit board and component layout for the scales. LED D7, the LDR and the wiring are soldered to the underside of the board.

2



LDR, a certain voltage will appear at all four inverting inputs of the comparators. When this voltage is identical to, or greater than, that preset at the non-inverting input, the relative opamp output will go low. As the anodes of the LEDs are connected to the positive aide of the power supply, the corresponding LED will light up.
The mechanical construction of the scales (see figure 2) is reminiscent of the good old 'canned circuits' issue (December 1980). In fact, the best thing to use for this purpose is an old beer or soft drinks can. Actually, two are required. The lid of the first one (with the ring-pull) is removed and discarded. The bottom of the second one is

3



also removed (carefully) as this will be used as the lid for the first one later on, the remainder of this can can also be thrown away. A hole that will just allow a ballpoint pen to pass through it is drilled in the bottom of the undiscarded can (which is, in fact, used upside-down). The ballpoint pen is then attached to the base of the discarded can. Next, the printed circuit board is mounted in the can with sufficiently long bolts and spacers (see figures 2 and 3). The exact location of the printed circuit board should be kept so that the ballpoint pen is positioned as shown in the illustration. Obviously, the pen must be fixed into place before all the bolts are tightened up.

It should now become clear how the LED (D7), the LDR and the point of the pen combine to form the opto-coupler. At this stage it is essential that the LED and/or the LDR be positioned accurately. The ballpoint pen is then cut to size so that it just protrudes through the underside of the printed circuit board. A piece of copper-clad board is then cut to the required dimensions and glued into the end of the ballpoint.

The next operation is to drill holes in the side of the can to accommodate the LEDs, the switch (S1) and the holes whereby the preset potentiometers can be adjusted. It is probably easiest to mount the LEDs together on a small piece of Veroboard. Connections to the main board can then be made via a length of ribbon cable, which, once soldered, can be glued into place. The 9 V battery can be stuck to the side of the can with a piece of double-sided sticky tape. Obviously, the positioning of the potentiometer adjustment holes must be determined very accurately. Once everything has been packed in and all the soldering completed, the scales are ready for use.

The table shown here should be attached to the outside of the can next to the LEDs, so that the amount of postage required for the particular letter can be read off immediately. The scales are calibrated by adjusting the four trimmer potentiometers with known weights on the scales. LED D3 should light up over 60 g, D4 over 100 g, D5 over 150 g and D6 over 200 g.

K. Hense

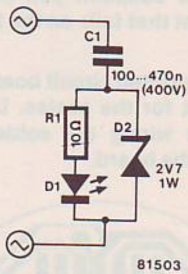
293 Mains LED

The very long life span of LEDs make them eminently suitable for on/off indicators. However, their use on mains voltages has been restricted since a low operating voltage prevents their direct connection with the mains supply in the manner of a neon.

Fortunately, there is a way around this problem, the AC resistance of a capacitor can be used to limit the current. No power is lost in the capacitor at all, since the current passing through the capacitor and the voltage across it are 90° out of phase with respect to each other.

Zener diode D2, acting as an ordinary forward-biased diode in this instance, prevents excess voltage levels appearing across the LED during the negative half-cycle of the mains waveform. If the circuit is switched on during the positive half, D2 will prevent the voltage across LED D1 and R1 from rising above 2.7 V. If an ordinary diode were used here, as in the earlier circuit, the LED is likely to go to the big scrap box in the sky.

The value of C1 determines the current passing



through the LED. When C1 = 100 n the current will be about 4 mA, and about 20 mA if a value of 470 n is chosen.

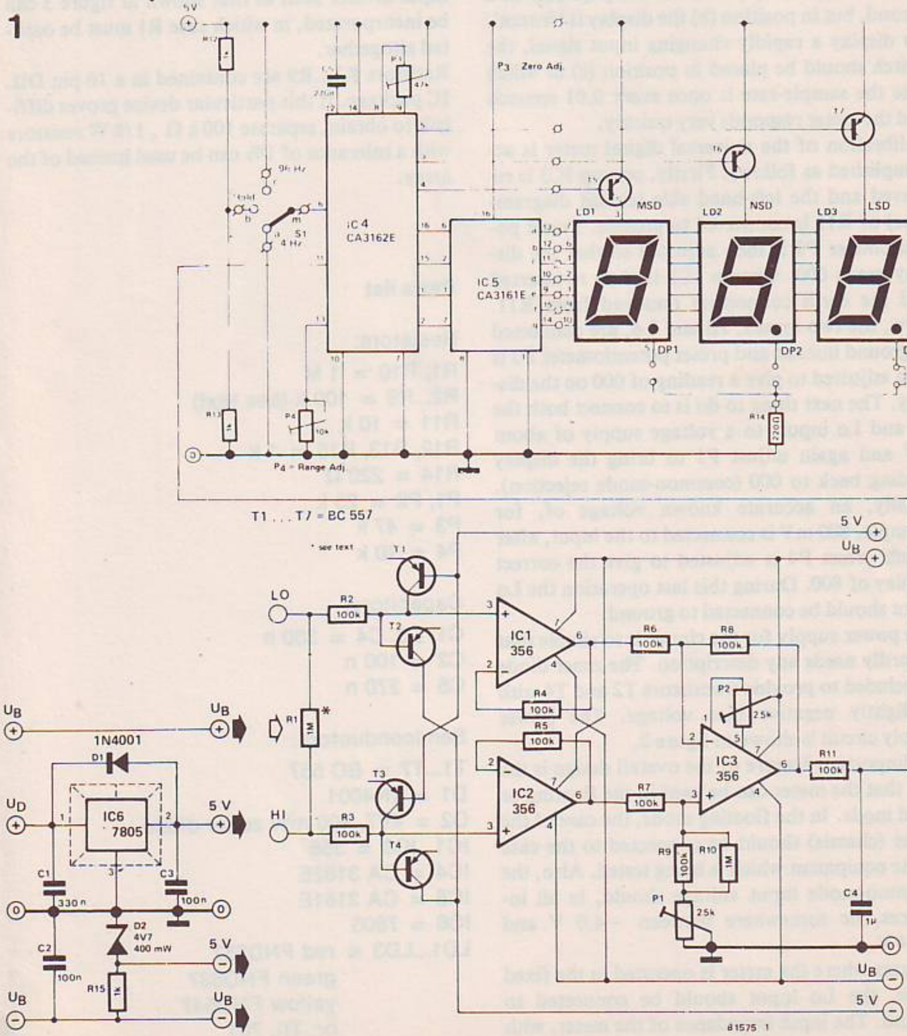
294 Universal digital meter

This digital meter is a great improvement on previous designs through the inclusion of an input stage containing J-FET opamps. This avoids various problems such as an unstable zero-reading. The J-FET inputs provide a very high input impedance and instead of the normal protection diodes, the circuit contains transistors connected as diodes. The transistors used have a low leakage current (1 nA).

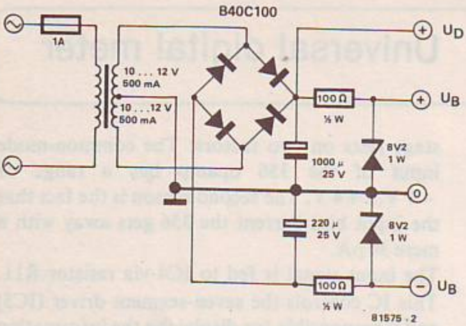
The reason for selecting such an extensive input stage rests on two factors: The common-mode input of the 356 opamp has a range of −4 V... +4 V. The second reason is the fact that the input bias current the 356 gets away with a mere 30 pA.

The input signal is fed to IC4 via resistor R11. This IC controls the seven-segment driver (IC5) and is responsible for displaying the information once it arrives. The three position switch (S1), situated to the left of IC4 on the circuit diagram,

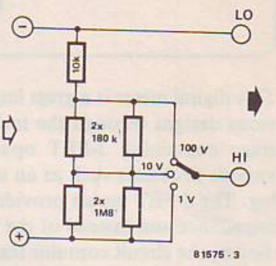
1



2



3



varies the sample-rate of the incoming data. In position (a) a sample is taken every quarter of a second, but in position (b) the display is 'frozen'. To display a rapidly changing input signal, the switch should be placed in position (c) in which case the sample-rate is once every 0.01 seconds and the meter responds very quickly.

Calibration of the universal digital meter is accomplished as follows: Firstly, opamp IC3 is removed and the left-hand side (circuit diagram-wise) of R11 is connected to ground. Preset potentiometer P3 is then adjusted so that the display reads 000. Opamp IC3 is then re-inserted and the earth connection removed from R11. Now, the two inputs, Hi and Lo, are connected to ground instead and preset potentiometer P2 is then adjusted to give a reading of 000 on the display. The next thing to do is to connect both the Hi and Lo inputs to a voltage supply of about 3 V and again adjust P1 to bring the display reading back to 000 (common-mode rejection). Finally, an accurate known voltage of, for example, 800 mV is connected to the input, after which preset P4 is adjusted to give the correct display of 800. During this last operation the Lo input should be connected to ground.

The power supply for the circuit is so simple that it hardly needs any description. The zener diode is included to provide transistors T2 and T4 with a slightly negative bias voltage. The power supply circuit is shown in figure 2.

An important feature of the overall design is the fact that the meter can be used in the floating or fixed mode. In the floating mode, the case of the meter (chassis) should be connected to the case of the equipment which is being tested. Also, the common-mode input voltage should, in all instances, lie somewhere between −4.0 V and +0.4 V.

In cases where the meter is operated in the fixed mode, the Lo input should be connected to ground. The input impedance of the meter, with resistor R1 in place, is 1 MΩ. If desired, an input divider such as that shown in figure 3 can be incorporated, in which case R1 must be omitted altogether.

Resistors R2...R9 are contained in a 16 pin DIL IC package. If this particular device proves difficult to obtain, separate 100 kΩ, 1/8 W resistors with a tolerance of 1% can be used instead of the array.

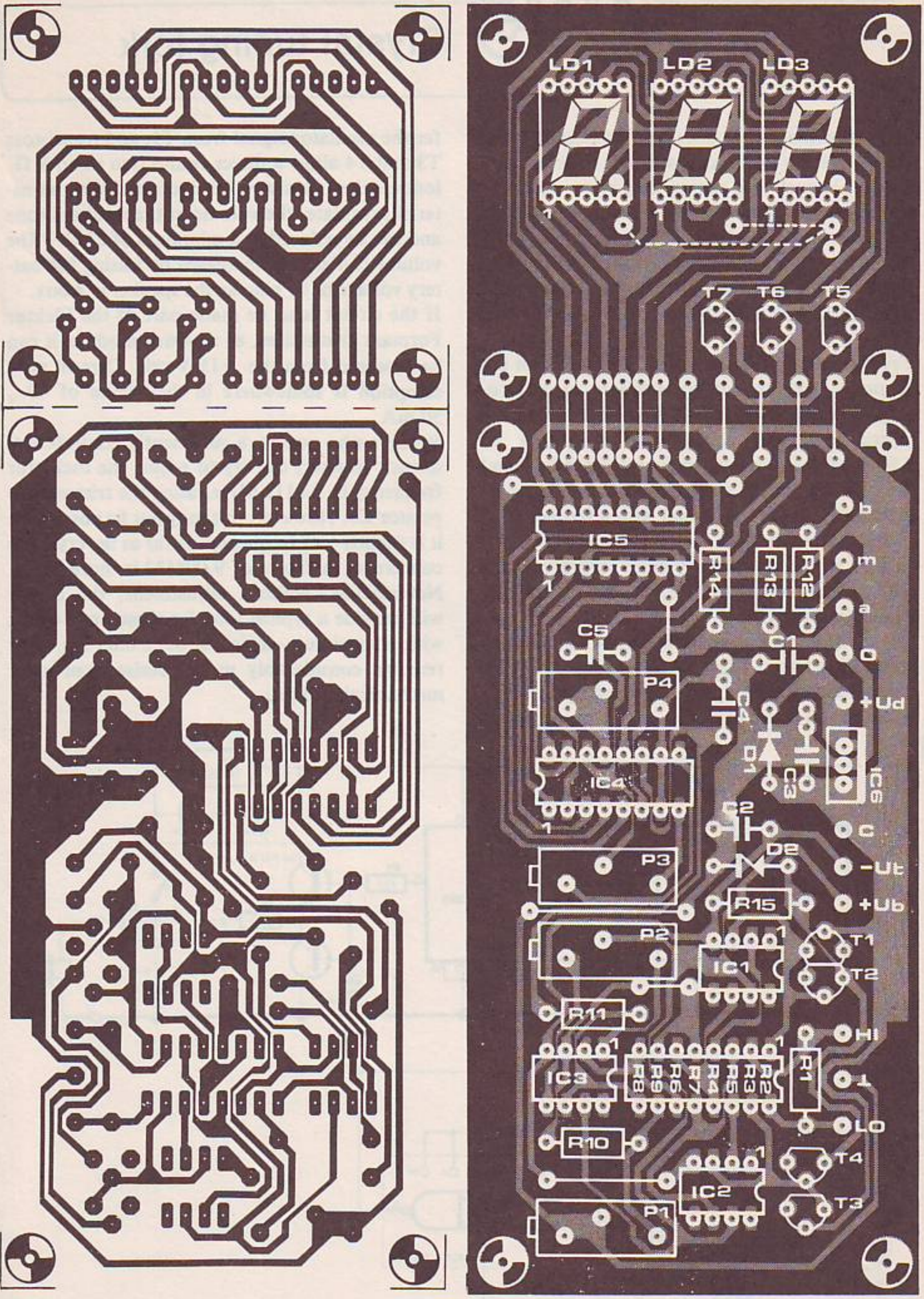
Parts list

Resistors:
R1, R10 = 1 M
R2...R9 = 100 k (see text)
R11 = 10 k
R12, R13, R15 = 1 k
R14 = 220 Ω
P1, P2 = 25 k
P3 = 47 k
P4 = 10 k

Capacitors:
C1, C3, C4 = 330 n
C2 = 100 n
C5 = 270 n

Semiconductors:
T1...T7 = BC 557
D1 = 1N4001
D2 = 4V7, 400 mW zener diode
IC1...IC3 = 356
IC4 = CA 3162E
IC5 = CA 3161E
IC6 = 7805
LD1...LD3 = red FND557
green FND537
yellow FND547
or: TIL 701

4



LD1
LD2
LD3
T7
T6
T5
IC5
R14
R13
R12
b
m
a
0
+Ud
C5
C1
P4
C4
D1
C3
IC6
IC4
C2
c
D2
-Ut
P3
R15
+Ub
P2
IC1
T1
T2
R11
Hi
IC3
R8 R9 R6 R7 R4 R5 R3 R2
R1
Lo
R10
IC2
T4
T3
P1

295 Crystal tuning fork

The musician, who does not possess 'perfect pitch', will no doubt occasionally have to use a tuning aid such as a tuning fork or an electronic 440 Hz sound generator. Low frequency 440 Hz oscillators are not ideally suited for use as 'electronic tuning forks' because of their inherent instability. For this reason, the crystal controlled oscillator described here may well offer the best solution.

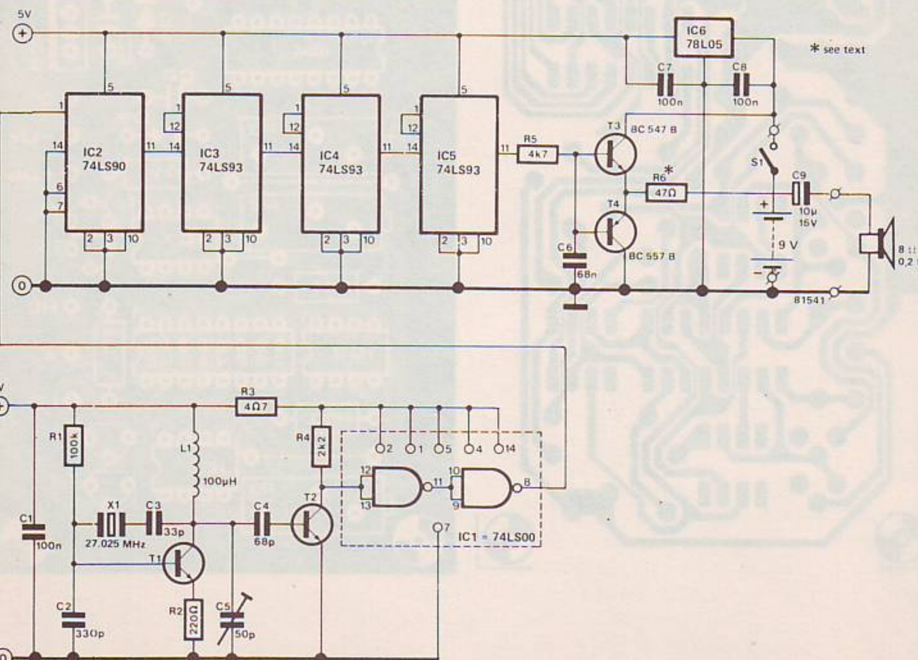
Crystals with a frequency of 27.025 MHz for use in model radio control transmitters, are reasonably inexpensive and readily available. Such a crystal oscillates at a basic frequency of 9.0083 MHz, which when first divided by 5 and then by 12^{12} yields a tone with a frequency of 439.86 Hz. Ideally, a crystal with a frequency of 27.035 MHz, channel 7 of American CB transceivers, would give slightly more accurate results (440.02 Hz), but these are, as yet, not readily available in the UK!

The division by 2^{12} (= 4096) can be obtained by connecting twelve flipflops in series (IC3...IC5), while the oscillator signal is divided by 5 in IC2. Transistor T2 and the gates in IC1 serve to buffer the oscillator signal from T1, and transistors T3 and T4 allow a direct connection to an 8 Ω loudspeaker. Resistor R6, indicated with an asterisk, regulates the volume of the resultant tone and can be reduced to a minimum of 22 Ω . The volume can also be increased by raising the battery voltage or by placing the speaker in a box.

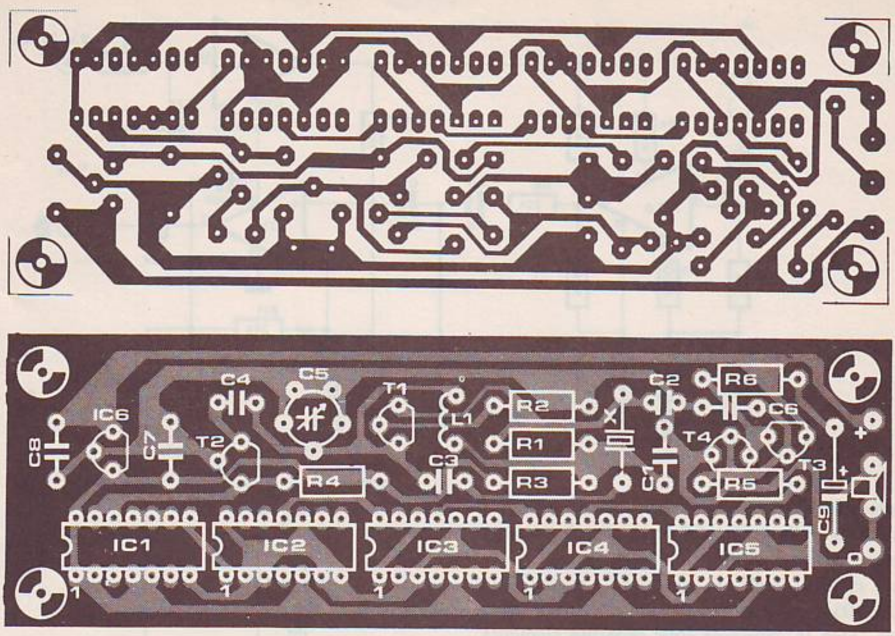
If the circuit is to be made part of the Elektor Formant synthesiser, as an extra module, it can be powered from the +15 V rail. Current consumption is somewhere in the region of 40... 50 mA.

Readers who possess a sufficiently accurate frequency counter, can try to adjust the oscillator frequency to 9.0112 MHz, using the trimmer capacitor C5. However, this may not be possible as it is slightly 'off frequency' as far as the crystal is concerned. A figure of 9.008332 is more likely. Nevertheless, without adjustment, the circuit will provide a typical tone frequency of 440 Hz with a maximum deviation of ±0.05 Hz. This remains considerably more precise than most mechanical devices.

1



2



Parts list

Resistors:
R1 = 100 k
R2 = 220 Ω
R3 = 4.7 Ω
R4 = 2k2
R5 = 4k7
R6 = 47 Ω *

Capacitors:
C1, C7, C8 = 100 n
C2 = 330 p
C3 = 33 p
C4 = 68 p
C5 = 50 p trimmer
C6 = 68 n
C9 = 10 µ /16 V

Semiconductors:
T1, T2 = BF 198, BF 199, BF 494
T3 = BC 547B
T4 = BC 557B
IC1 = 74LS00
IC2 = 74LS90
IC3...IC5 = 74LS93
IC6 = 78L05

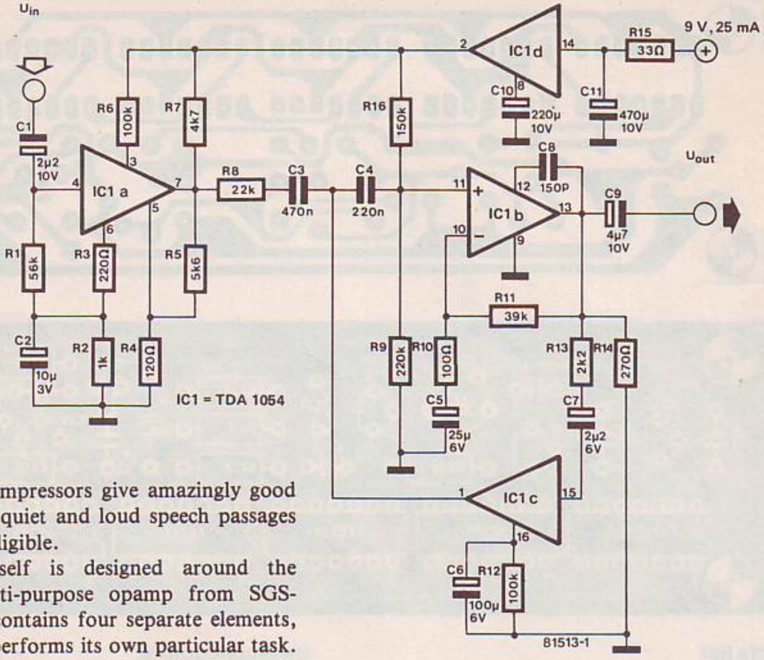
Miscellaneous:
L1 = 100 µ H
X1 = 27.025 MHz crystal (with holder)
S1 = single-pole switch
LS = 8 Ω /0.2 W loudspeaker

* see text

296 Microcompressor

Dynamic range compressors can be used in any device that requires a constant audio output level. The first example that comes to mind is the automatic record level control in cassette recorders. The compressor can also be employed, however, in such items as amateur radio equipment, discotheques, babyphones and intercom systems to ensure optimum intelligibility and to prevent any damage to amplifiers and loudspeakers. When used in combination with a

1



microphone, compressors give amazingly good results as both quiet and loud speech passages are equally intelligible.

The circuit itself is designed around the TDA 1054 multi-purpose opamp from SGS-Ates. This IC contains four separate elements, each of which performs its own particular task. IC1a is a preamplifier which is used to boost the input signal to about 50 x (1 + R5/R4). Opamp IC1b is also used as a preamplifier, but this has a gain of 400 x (1 + R11/R10). The function of IC1d is to remove any ripple from the supply voltage, while IC1c takes care of the actual automatic level control.

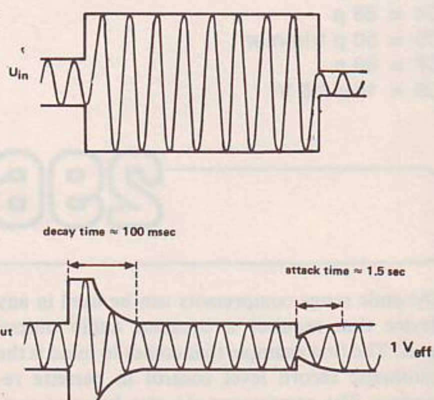
A good compressor should compress the entire signal in a linear fashion, in other words, not by simply 'clipping' the top of the waveform. This can be accomplished by making the amount of level reduction dependent on the largest amplitude appearing in the input signal. To do this, the amplitude of the output signal is monitored and when this rises above a certain level, attenuation is introduced.

The attenuator, IC1c, is driven via resistor R13 and capacitor C7 and acts as a variable resistance between the junction of C3/C4 and ground. This ensures that the input to the output amplifier, IC1b, is sufficiently attenuated when the output level rises above about 1 V rms. Capacitor C7 gives the system delay time which is necessary to ensure that control is adequately fast to follow the envelope of the signal waveform, but not fast enough to respond to the waveform itself. This capacitor therefore determines the attack time of the circuit. The decay time depends on the values of capacitor C6 and resistor R12. This delay time must be much longer in order to maintain a reasonably consistent sound level.

A graphic illustration of how the circuit works is shown in figure 2. The times given are valid for the component values in figure 1. The attack and decay times can be modified according to personal taste by changing the values of C7 and C6 respectively.

The input of the microcompressor is suitable for low signal levels, for instance from microphones. The input impedance is approximately 50 k . Higher input signal levels can be con-

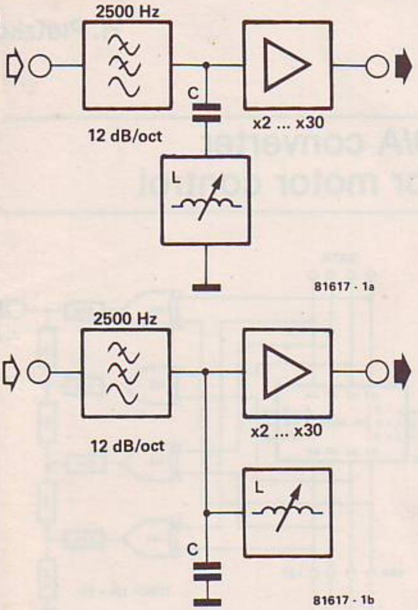
2



nected directly to resistor R8, in which case the entire circuit before R8 can be omitted.
The supply voltage can be increased to 12 V if desired, but this will mean that the rating of the electrolytic capacitors will have to be increased also.

297 Active notch or CW filter

1

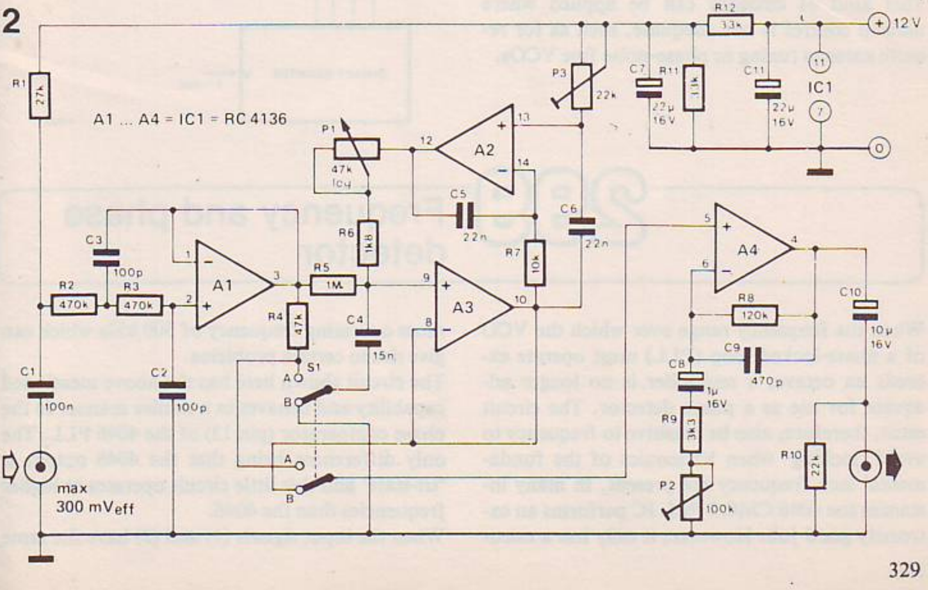


In general, the majority of low-cost shortwave receivers have poor selectivity. Usually, reception is accompanied by a number of interfering signals, or more than one transmission is heard at the same time (especially with CW transmissions, which have a very narrow bandwidth). An add-on extra for the existing inexpensive set could therefore prove very useful, particularly if a (much) more expensive one is not available (or the funds thereof).

The filter described here is an active equivalent for an LC tuned circuit and can be operated in the parallel mode (peaking function) or the series mode (notch function) as can be seen from figure 1. The filter is connected to the audio output of the receiver or, if present, the tape recorder output. The output of the filter is sufficient to drive a pair of headphones directly, provided their impedance is 600 Ω or greater.

The simulated inductor consists of the circuitry around opamps A2 and A3 (see figure 2). A 12 dB per octave filter for the input signal is provided by A1, while A4 acts as an amplifier with an adjustable gain of 2...30 times.

2



The resonant circuit can be switched between the parallel and series operating modes by means of S1. The filter can be tuned over the range 300... 400 Hz by means of potentiometer P1. With switch S1 in position B, preset potentiometer P3 should be rotated to the position just prior to where the circuit starts oscillating. This calls for a precise alignment and therefore P3 should be a multi-turn type. The gain of the circuit can be adjusted by means of potentiometer P2. This can be a preset type if a fairly constant level of background noise is to be expected, otherwise a normal (volume control) potentiometer is preferred.

When the circuit is actually used, the following points must be remembered:

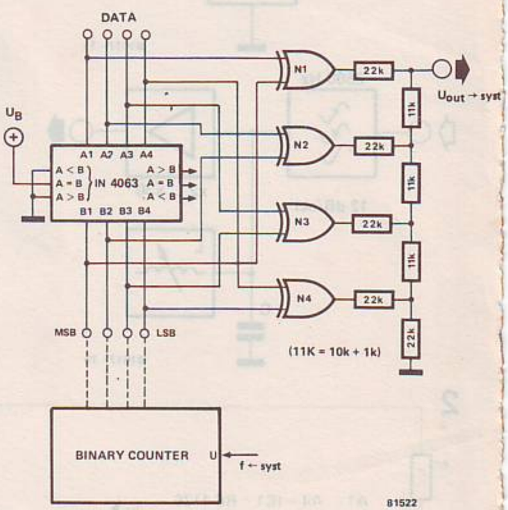
1) the effects of frequency drift will be pronounced
2) as the AGC circuit responds to a much wider bandwidth than that of the filter, the recovered signal could show 'alien' pumping action.

H. Pietzko

298 D/A converter for motor control

In many instances, it is quite sufficient to control motors with an on/off type action. However, if a more 'linear' control is required, the circuit given here may prove useful. The four bit magnitude comparator (4063) inputs are EX-ORed and fed to a resistive divider chain to provide an output voltage which is proportional to the difference of the comparator input values (with an accuracy of four bits). These values can be derived, for instance, from a binary counter.

The outputs of the comparator (A > B, A = B, A < B) can be used to (indirectly) turn the motor on or off and/or to reverse the polarity of the supply voltage in the case of DC motors. This kind of circuitry can be applied where varicap control is not adequate, such as for remote antenna tuning or phase-noise free VCOs.



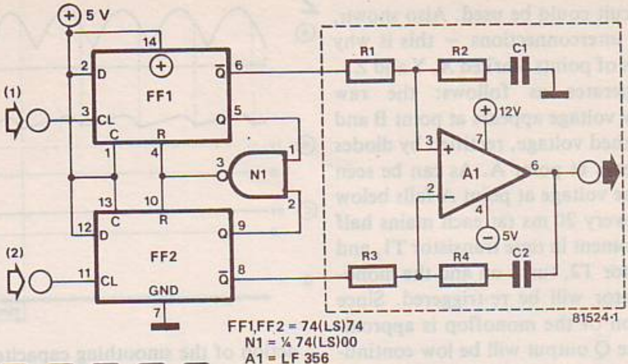
299 Frequency and phase detector

When the frequency range over which the VCO of a phase-locked loop (PLL) must operate exceeds an octave, a multiplier is no longer adequate for use as a phase detector. The circuit must, therefore, also be sensitive to frequency to avoid 'locking' when harmonics of the fundamental lock frequency are present. In many instances the 4046 CMOS PLL IC performs an extremely good job. However, it only has a maximum operating frequency of 500 kHz which can give rise to certain problems.

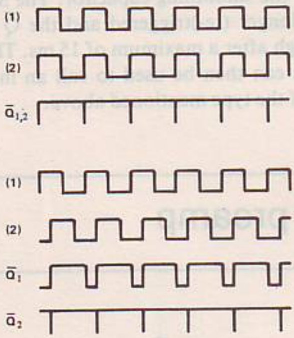
The circuit shown here has the above mentioned capability and behaves in a similar manner to the phase comparator (pin 13) of the 4046 PLL. The only differences being that the 4046 output is 'tri-state' and this little circuit operates at higher frequencies than the 4046.

When the input signals (1) and (2) have the same

1



2

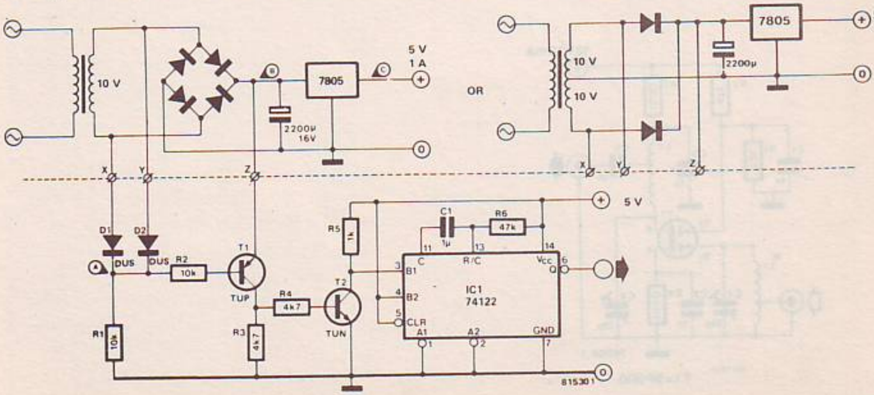


frequency and phase relationship, the two flipflops are both reset simultaneously. Should the phase shift between the two input signals alter, the reset timing will also change. In this instance, the average voltage at the $\overline{Q}$ output of one of the flipflops will be greater than that at the other. This is clearly illustrated in the timing diagram.
The dc level at the output of the differential amplifier, A1, is used to control the VCO. The actual values of components R1...R4, C1 and C2 depend on the frequency of operation.

300 Power failure forecaster

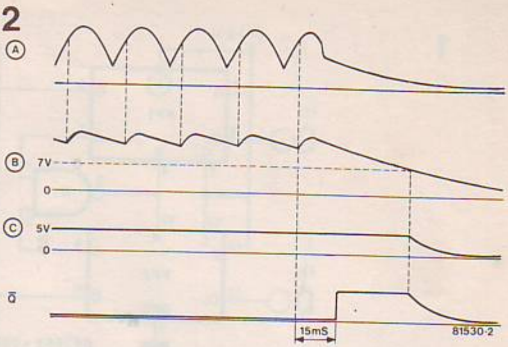
This circuit can be extremely useful in, amongst other things, microprocessor systems. Should the main power voltage fail, the circuit will provide a logic high level at its output a short time before the supply disappears completely. This time delay, although short, can be sufficient to take emergency measures. For instance, storing the data contained in certain of the internal processor registers into a low-power standby random access memory (battery powered).
In the circuit diagram, above the dotted line, there are two examples of 5 volt power supplies

1



in which the circuit could be used. Also shown, are the relevant interconnections – this is why there are two sets of points marked X, Y and Z.
The circuit operates as follows: the raw smoothed supply voltage appears at point B and the raw unsmoothed voltage, rectified by diodes D1 and D2 appears at point A. As can be seen from figure 2, the voltage at point A falls below that at point B every 20 ms (at each mains half cycle). At this moment in time transistor T1, and therefore transistor T2, turns on and the monostable multivibrator will be re-triggered. Since the pulse duration of the monoflop is approximately 15 ms, the $\overline{Q}$ output will be low continuously as long as the main supply voltage is applied. However, as soon as the main voltage fails, the voltage at point A will become lower than that at point B immediately due to the action of the smoothing capacitor. The MMV is then no longer (re-)triggered and the $\overline{Q}$ output will go high after a maximum of 15 ms. This output pulse can then be used to call an interrupt routine of the type mentioned above.

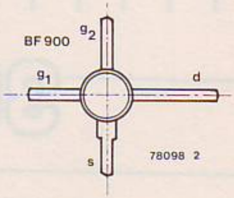
2



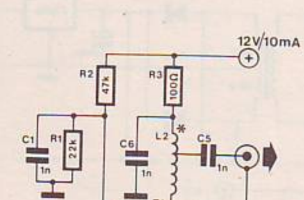
301 VHF preamp

Designing a preamp for the VHF waveband (around 100 MHz) is not always an easy matter. This circuit, however, is both relatively simple to use and is inexpensive. It has the advantage of a fairly large bandwidth (2 MHz) and good noise figure (2.5 dB). The preamp has a large dynamic range and a gain of 20 dB at a frequency of 144 MHz.
L1 and L2 are air-cored coils with an internal diameter of 6 mm and consist of 4 turns of 1 mm silverplated copper wire. L1 is tapped one turn from the earthy end, whilst L2 has a tap one turn from the end nearest R3. Ceramic types are recommended for the four 1 n capacitors.

2

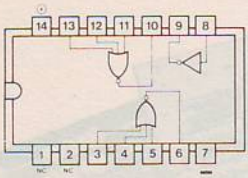


1

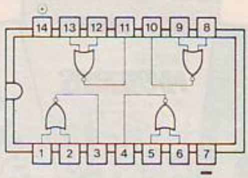


BLANK

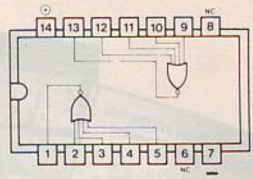
DUAL 3 INPUT NOR GATE PLUS INVERTER
4000



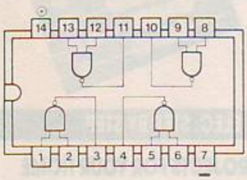
QUADRUPLE 2 INPUT NOR GATE
4001



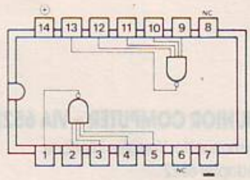
DUAL 4 INPUT NOR GATE
4002



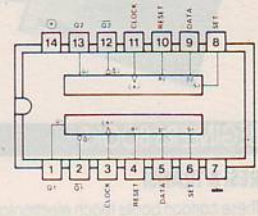
QUADRUPLE 2 INPUT NAND GATE
4011



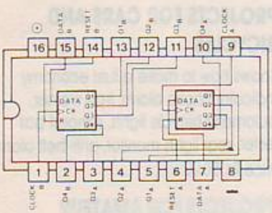
DUAL 4 INPUT NAND GATE
4012



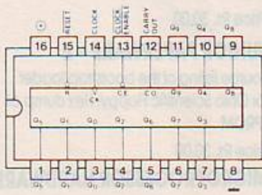
DUAL D FLIP FLOP
4013



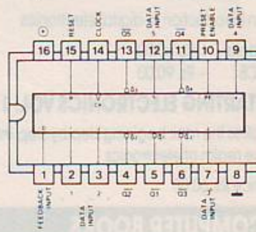
DUAL 4 BIT STATIC SHIFT REGISTER
4015



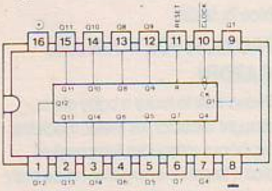
DIVIDE BY 10 SYNCHRONOUS COUNTER
4017



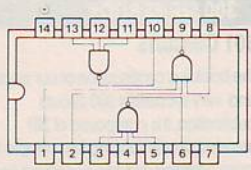
SYNCHRONOUS PRESETTABLE DIVIDE BY 'N' COUNTER
4018



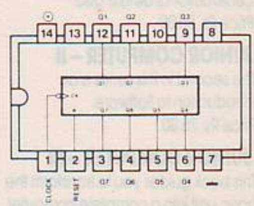
14 BIT BINARY RIPPLE COUNTER
4020



TRIPLE 3 INPUT NAND GATE
4023

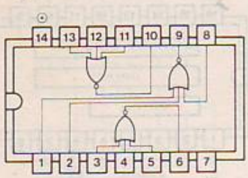


7 STAGE BINARY RIPPLE COUNTER
4024



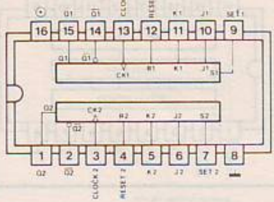
TRIPLE 3 INPUT NOR GATE

4025



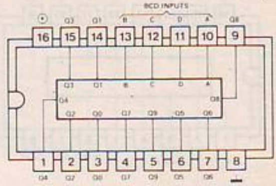
DUAL JK FLIP FLOP

4027



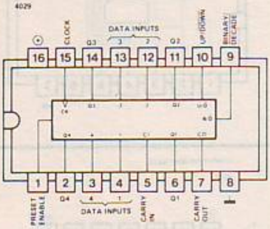
BCD TO DECIMAL DECODER

4028



SYNCHRONOUS PRESET TABLE BINARY/DECADE UP/DOWN COUNTER

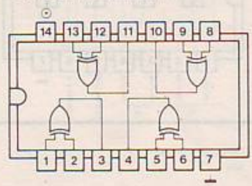
4029



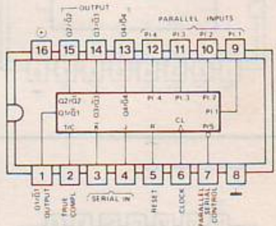
QUADRUPLE 2 INPUT EXCLUSIVE OR GATES

4030

4070 low power TTL compatible (fan out = 2)

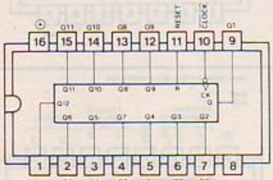


4035
4 BIT
PARALLEL IN/PARALLEL OUT
SHIFT REGISTER



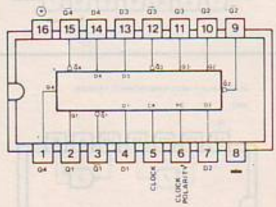
12 BIT BINARY RIPPLE COUNTER

4040



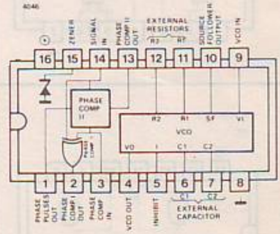
QUAD CLOCKED "D" LATCH

4042



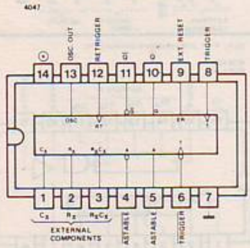
MICROPOWER PLL

4046



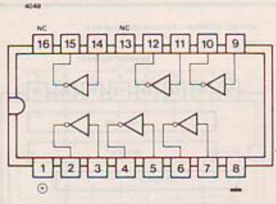
MONOSTABLE/ASTABLE MULTIVIBRATOR

4047



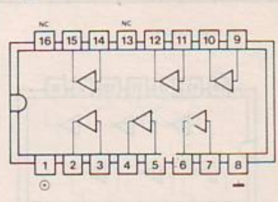
HEX INVERTING BUFFER

4049



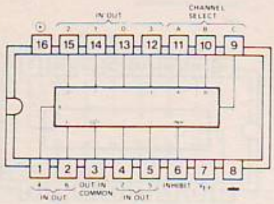
HEX BUFFER

4050



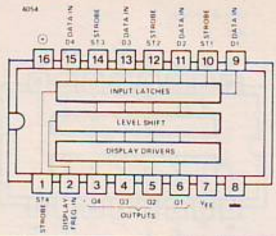
8 CHANNEL ANALOGUE MULTIPLEXER DEMULTIPLEXER

4051



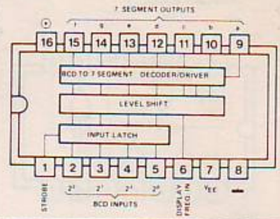
LCD DRIVER

4054



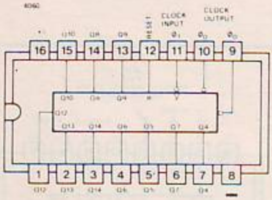
BCD TO 7 SEGMENT DECODER/DRIVER

4056



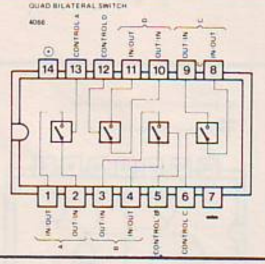
14 BIT BINARY RIPPLE COUNTER AND OSCILLATOR

4060



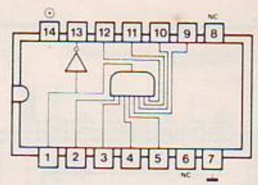
QUAD BILATERAL SWITCH

4066



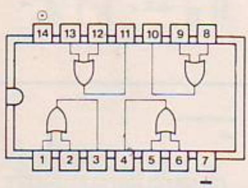
8 INPUT AND/NAND GATE

4068



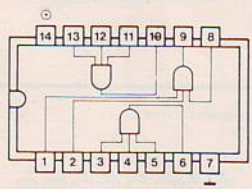
QUADRUPLE 2 INPUT OR GATE

4071



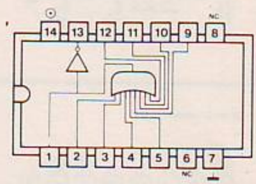
TRIPLE 3 INPUT AND GATE

4073



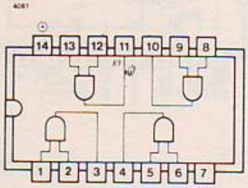
8 INPUT OR/NOR GATE

4078



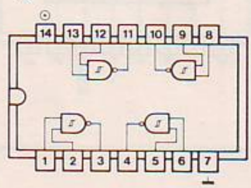
QUADRUPLE 2 INPUT AND GATE

4081



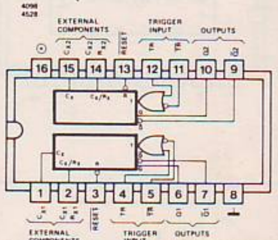
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER

4093



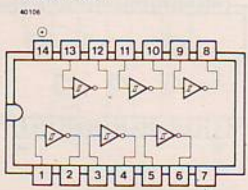
DUAL MONOSTABLE MULTIVIBRATOR

4098
4528



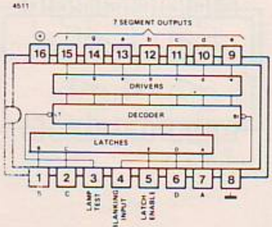
HEX SCHMITT TRIGGER

40106



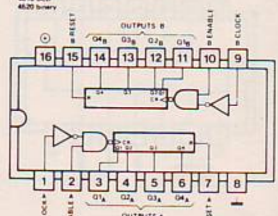
BCD TO 7 SEGMENT LATCH/DECODER/DRIVER

4511



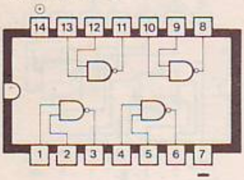
DUAL 4 BIT SYNCHRONOUS UP COUNTERS

4518 BCD
4520 binary



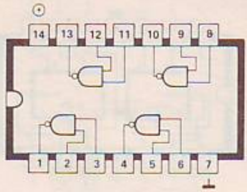
QUADRUPLE 2 INPUT NAND GATES

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



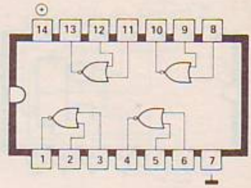
QUADRUPLE 2 INPUT NAND GATE WITH OPEN COLLECTOR OUTPUT

7401



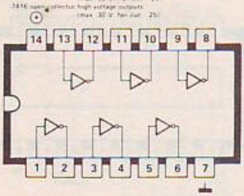
QUADRUPLE 2 INPUT NOR GATES

7402
7428 power driver (fan out : 30)



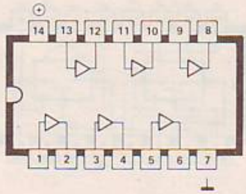
HEX INVERTERS

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)



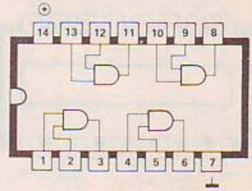
HEX BUFFER/DRIVER WITH OPEN COLLECTOR HIGH VOLTAGE OUTPUTS (max. 30 V; fan out : 25)

7407



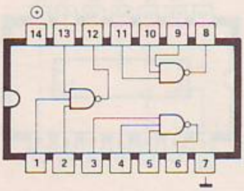
QUADRUPLE 2 INPUT AND GATES

7408
7409 open collector outputs



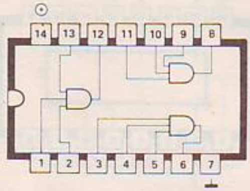
TRIPLE 3 INPUT NAND GATES

7410
7412 open collector outputs



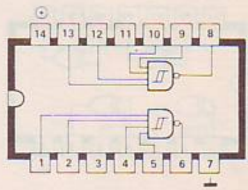
TRIPLE 3 INPUT AND GATE

7411



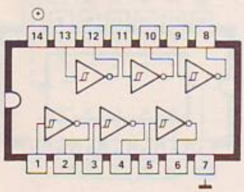
DUAL 4 INPUT NAND SCHMITT TRIGGER

7413



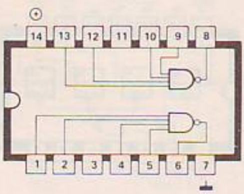
HEX SCHMITT TRIGGER INVERTER

7414



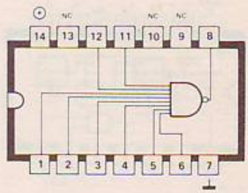
DUAL 4 INPUT NAND GATES

7420
7440 power driver (fan out : 30)



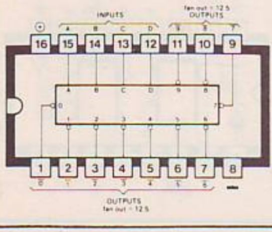
8 INPUT NAND GATE

7430



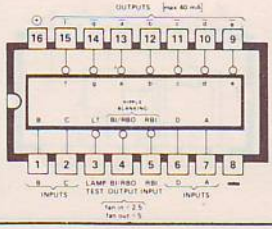
BCD TO DECIMAL DECODER/DRIVER WITH OPEN COLLECTOR OUTPUTS (max. 30 V)

7445



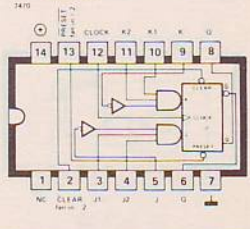
BCD TO 7 SEGMENT DECODER/DRIVER

7447



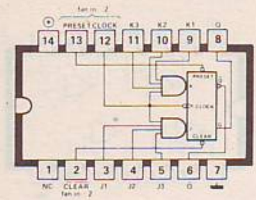
AND GATED J-K POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7470



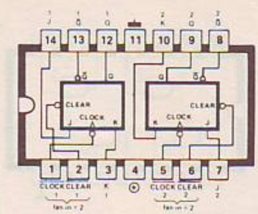
AND GATED J K FLIP FLOP WITH PRESET AND CLEAR

7472



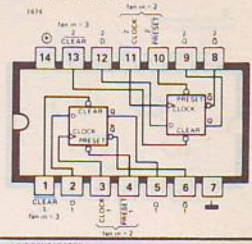
DUAL J K FLIP FLOP WITH CLEAR

7473



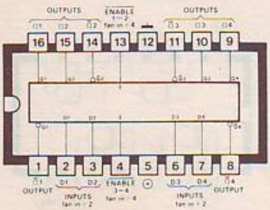
DUAL D TYPE POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7474



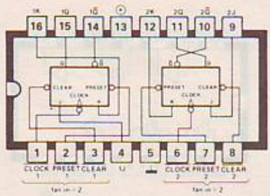
4 BIT BISTABLE LATCH

7475



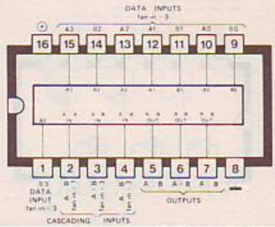
DUAL J K MASTER SLAVE FLIP FLOP WITH PRESET AND CLEAR

7476



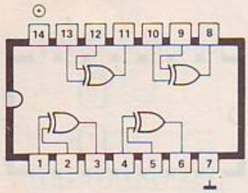
4 BIT COMPARATOR

7485



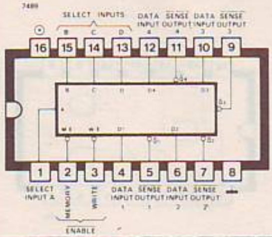
QUADRUPLE 2 INPUT EXCLUSIVE OR GATE

7486



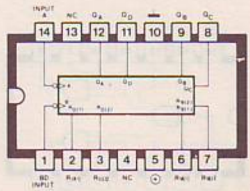
64 BIT READ WRITE MEMORY (fan out = 7.5)

7489



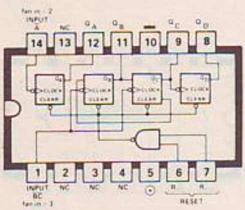
DECADE COUNTER

7490



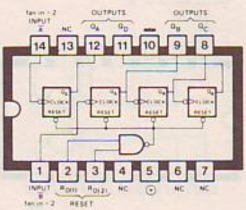
DIVIDE BY TWELVE COUNTER (÷ 2 and ÷ 6)

7492



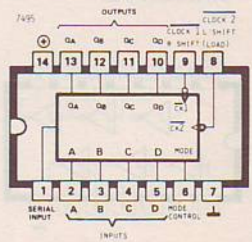
4 BIT BINARY COUNTER

7493



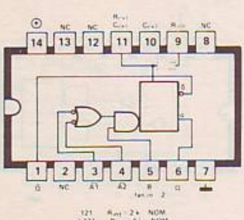
4-BIT PARALLEL-IN PARALLEL-OUT SHIFT REGISTER

7495



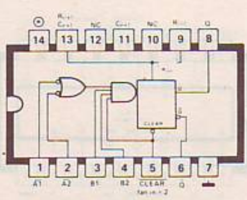
MONOSTABLE MULTIVIBRATOR

74121



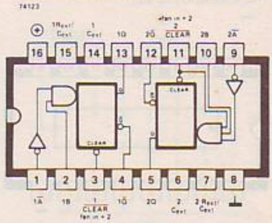
RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74122



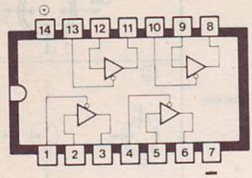
DUAL RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74123



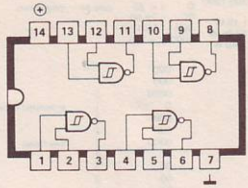
QUAD BUFFER (3 STATE)

74125



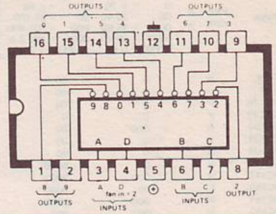
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER

74132



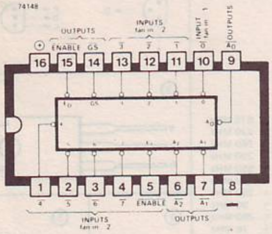
BCD TO DECIMAL DECODER DRIVER
(outputs max 60 V max 7 mA)

74141



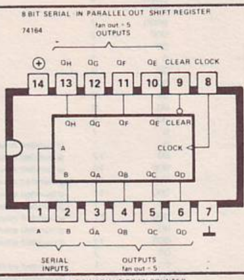
PRIORITY ENCODER

74148



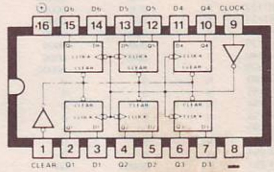
8 BIT SERIAL IN PARALLEL OUT SHIFT REGISTER

74164



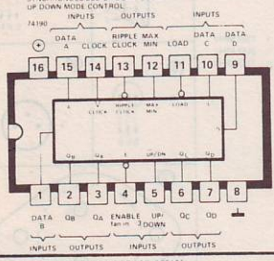
HEX D FLIP FLOP WITH CLEAR

74174



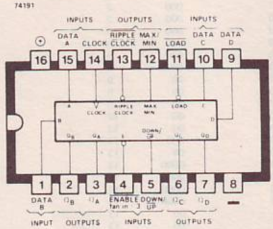
SYNCHRONOUS BCD UP DOWN COUNTER WITH UP DOWN MODE CONTROL

74190



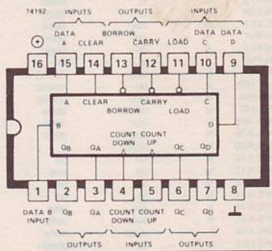
SYNCHRONOUS 4 BIT BINARY UP/DOWN COUNTER

74191



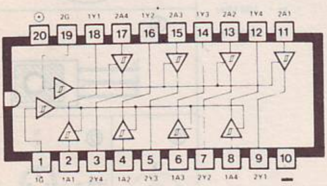
SYNCHRONOUS UP DOWN DECADE COUNTER

74192



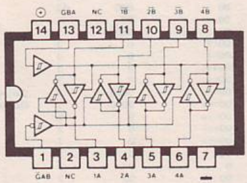
OCTAL BUFFER AND LINE DRIVER (3 STATE)

74LS241



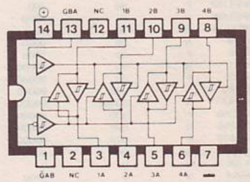
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS242



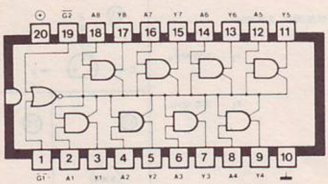
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS243



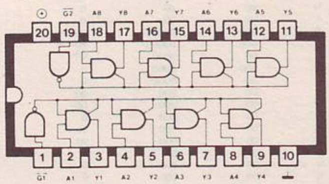
OCTAL BUFFER (3 STATE)

81LS95

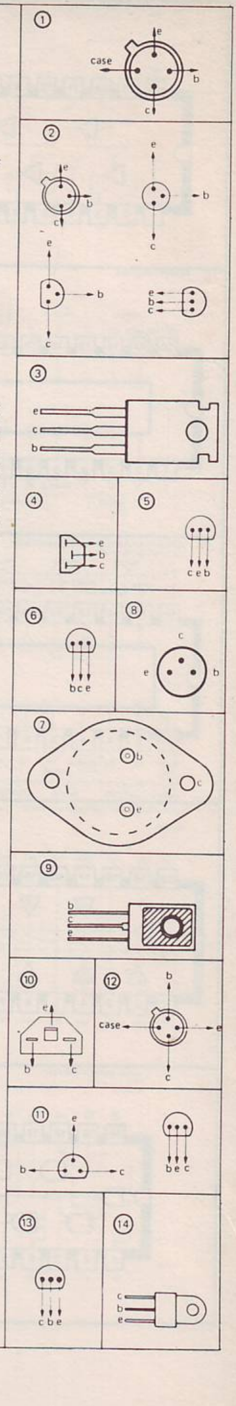


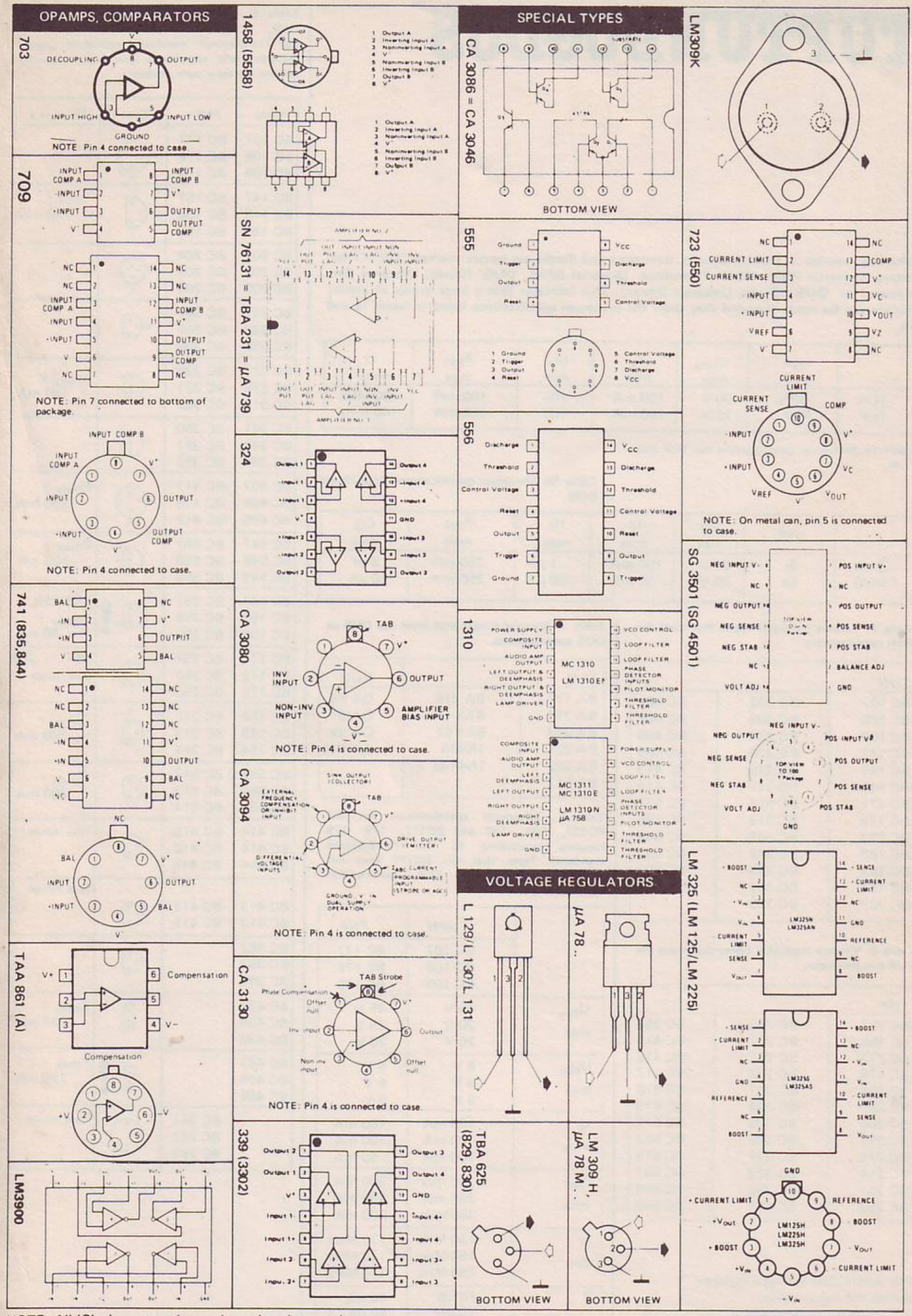
OCTAL BUFFER (3 STATE)

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