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<table>
<thead>
<tr>
<th>CONTENTS</th>
<th>July/August 1989 Supplement</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>AUDIO &amp; HI-FI</strong></td>
<td></td>
</tr>
<tr>
<td>4 Balance indicator (001)</td>
<td>23 Automatic switch (034)</td>
</tr>
<tr>
<td>9 Recording control (009)</td>
<td>24 Head/tail lights for model railway (037)</td>
</tr>
<tr>
<td>11 4-channel mixer (014)</td>
<td>27 Twilight switch (042)</td>
</tr>
<tr>
<td>13 Bucket brigade delay line (019)</td>
<td>29 X-Y plotter interface (044)</td>
</tr>
<tr>
<td>16 Tuneable band-pass filter (022)</td>
<td>29 Timer with audible warning (045)</td>
</tr>
<tr>
<td>20 Sound level attenuator (030)</td>
<td>33 Heating timer (051)</td>
</tr>
<tr>
<td>34 A simple vox (053)</td>
<td>35 High-volume alarm (055)</td>
</tr>
<tr>
<td>34 Low-noise microphone preamplifier (054)</td>
<td>36 Mains-powered timer (056)</td>
</tr>
<tr>
<td><strong>CAR ELECTRONICS</strong></td>
<td></td>
</tr>
<tr>
<td>5 Car lights monitor (003)</td>
<td>37 Single-chip melody generator (057)</td>
</tr>
<tr>
<td>16 Car headlight control (023)</td>
<td>37 Infra-red microphone (058)</td>
</tr>
<tr>
<td>25 Energy control for battery chargers (039)</td>
<td>38 Four-quadrant dimmer (059)</td>
</tr>
<tr>
<td>26 Car alarm (041)</td>
<td>38 Sensor switch and clock (060)</td>
</tr>
<tr>
<td>28 Psychological car lock (043)</td>
<td>39 Mini-drill control (061)</td>
</tr>
<tr>
<td>33 Improved low-fuel indicator (052)</td>
<td>40 Mains failure indicator (063)</td>
</tr>
<tr>
<td><strong>COMPONENTS</strong></td>
<td></td>
</tr>
<tr>
<td>17 nMOS integrated circuits (025)</td>
<td>34 78xx monitor (006)</td>
</tr>
<tr>
<td>21 Fast unity gain opamp (031)</td>
<td>10 9-volt supply (012)</td>
</tr>
<tr>
<td><strong>COMPUTERS</strong></td>
<td>19 Switch-mode voltage regulator (027)</td>
</tr>
<tr>
<td>4 Reset for the PC1640 (002)</td>
<td>23 Low dissipation regulator (035)</td>
</tr>
<tr>
<td>9 Child-proof reset switch (010)</td>
<td>30 High-power zener diode (047)</td>
</tr>
<tr>
<td>12 MSX EPROM (015)</td>
<td>32 Simple variable power supply (050)</td>
</tr>
<tr>
<td>17 Reset protection (024)</td>
<td><strong>RADIO &amp; TV</strong></td>
</tr>
<tr>
<td>22 Printer reset (032)</td>
<td>7 Radio beacon converter (007)</td>
</tr>
<tr>
<td>25 1/o-friendly keyboard (038)</td>
<td>12 Universal squelch (016)</td>
</tr>
<tr>
<td>31 Monitoring temperature with the C64 (049)</td>
<td>14 Fast envelope sampler (020)</td>
</tr>
<tr>
<td>39 Improved low-fuel indicator (052)</td>
<td>19 2-metre transmitter (028)</td>
</tr>
<tr>
<td><strong>ELECTROPHONICS</strong></td>
<td>40 Call tone generator (062)</td>
</tr>
<tr>
<td>5 Vocal eliminator (004)</td>
<td><strong>TELECOMMUNICATIONS</strong></td>
</tr>
<tr>
<td>22 Variable low-pass filter (033)</td>
<td>24 Duplex audio link (036)</td>
</tr>
<tr>
<td>41 Guitar compressor (064)</td>
<td><strong>TEST &amp; MEASUREMENT</strong></td>
</tr>
<tr>
<td><strong>GENERAL INTEREST</strong></td>
<td></td>
</tr>
<tr>
<td>6 Power booster for 7406/7407 (005)</td>
<td>8 Sound level meter (008)</td>
</tr>
<tr>
<td>10 &quot;On&quot; indicator (011)</td>
<td>10 ABS / RMS / LOG converter (013)</td>
</tr>
<tr>
<td>13 Simple temperature indicator (018)</td>
<td>13 48 MHz CMOS oscillator (017)</td>
</tr>
<tr>
<td>18 Programmable switch (026)</td>
<td>15 Noise generator (021)</td>
</tr>
<tr>
<td><strong>POWER SUPPLIES</strong></td>
<td>20 Meter-scale magnifier (029)</td>
</tr>
<tr>
<td>6 78xx monitor (006)</td>
<td>26 TTL supply monitor (040)</td>
</tr>
<tr>
<td>10 9-volt supply (012)</td>
<td>30 nMOS square wave generator (046)</td>
</tr>
<tr>
<td>19 Switch-mode voltage regulator (027)</td>
<td>31 Voltage-controlled oscillator (048)</td>
</tr>
<tr>
<td>23 Low dissipation regulator (035)</td>
<td>42 LC sine wave generator (065)</td>
</tr>
<tr>
<td>30 High-power zener diode (047)</td>
<td>42 Shunt for multimeter (066)</td>
</tr>
</tbody>
</table>
If your amplifier is fitted with two balance controls (as, for instance, the Elektor Electronics Preamplifier for Purists - Ref. 1), it actually offers you a balance control and a level control. A drawback of this is that it is quite difficult to set the balance properly. This may be obviated, however, by replacing the two mono potentiometers by stereo versions, P1 and P2 in the diagram.

One half of the pair, P1a and P2a, assumes the tasks of the removed components. The other half is connected in a bridge circuit. The voltage between the wipers of the potentiometers is then a measure of the balance between the two channels. The lower the potential, the better the balance. If you are interested in knowing the degree of unbalance, connect a centre-zero moving coil meter with a bias resistor between A and B. With this arrangement, zener diodes D1 and D2 may be omitted: they are necessary only with the LED indicator shown in the diagram to prevent the input voltage of the opamp getting too close to the level of the supply voltage.

The circuit around IC1 is a classical differential amplifier. Resistors R5 and R6 provide a virtual earth for the LEDs, which is necessary to ensure that in spite of the asymmetrical supply voltage a positive and a negative output is obtained.

Since the LEDs have been included in the feedback loop of the indicator, the circuit is pretty sensitive. At only 40 mV, that is, just one fourhundredth of the supply voltage, one of the LEDs begins to light already. The maximum current drawn by the LEDs is determined by the values of R5 and R6.


The PC1640 is one of Amstrad’s popular and successful series of compact PCs. It has, unfortunately, a serious deficiency: there is no reset control. Luckily, it is not too difficult to fit this control retrospectively, since every PC, and the 1640 is no exception, has a reset circuit.

In the 1649, the circuit shown below monitors the level of the supply voltage. This voltage is sampled by potential divider R151-R152-R176-R188. If the supply is too low, the Q105, and consequently the Q104, switches off. Transistor Q106 is also provided with too little bias voltage to conduct. The enable input of the CMOS memory, IC134, is linked to the collector of Q105, so that when Q105 is turned off, the memory is disabled. At the same time, the supply to the memory is switched off since Q104 is off.
Only when the supply voltage is at the correct level does Q105 conduct. At the same time, Q104 and Q106 are provided with sufficient base current to switch on. This causes the supply to IC134 to be switched on, which results in the removal of the inhibit on the enable input. The processor receives a signal, PWROK, indicating that the supply level is all right, so that the reset cycle can start. The entire computer is initialized, while all I/O lines are set as required.

To fit a reset control, use is made of the PWROK signal in the following manner. If R152 is short-circuited, PWROK goes low and the processor is reset. Only when PWROK has gone high again will the processor start the reset routine. This is exactly the same cycle that occurs when the computer is restarted. In other words, all the reset control needs to do is to short-circuit R152.

The reset facility is incorporated fairly easily by fitting at some convenient place on the computer a simple push-button switch with a make contact. Connect the two terminals of the switch to the two ends of R152 (clearly marked on the rca) via two short lengths of flexible wire and that's all.

A defect car light is at best a nuisance and at worst a danger. Fortunately, most new cars are equipped with suitable monitors that indicate on the dashboard whether a light is not working. There are, of course, millions of older cars that have no such sophistication and it is for these that the present monitor is intended.

Two special ICs are available from Telefunken that are designed for measuring the current through a light bulb. In practice, detecting whether a current flows through a bulb or not is a most suitable way of determining whether the bulb still works.

If a small resistance is connected in series with the bulb, a small voltage drop will develop across it when the bulb lights (R1 and R2 in the diagram). Each IC can cope with only two bulbs, so that per car three or four ICs are needed. The junction of the bulb and resistor is connected to one of the inputs (pin 4 or pin 6) of the IC. The potential across the resistor is compared in the IC with an internal reference voltage. Depending on which of the two ICs is used, the voltage drop must be about 16 mV (U477B) or 100 mV (U478B). This voltage drop is so small that it will not affect the brightness of the relevant bulb.

The value of the series resistor is determined quite easily. If, for instance, it is in series with the brake light (normally 21 W), the current through the bulb, assuming that the vehicle has a 12 V battery, is 21/12=1.75 A. The resistance must then be 16/1.75=9 mΩ (U477B) or 100/1.75=57 mΩ (U478B).

These resistors may be made from a length of resistance wire (available from most electrical retailers). Failing that, standard circuit wire of 0.7 mm dia. may be used. This has a specific resistance of about 100 mΩ per metre. However, in most cars the existing wiring will have sufficient resistance to serve as series resistor.

LEDs may be connected to the outputs of the IC (pins 3 and 5). These will only light if the relevant car light fails to work properly.

Otherwise properly mixed sounds often suffer from a predominant solo voice (which may, of course, be the intention). If such a voice needs to be suppressed, the present circuit will do the job admirably.

The circuit is based on the fact that solo voices are invariably situated "at the centre" of the stereo recordings that are to be mixed. This means that the voice levels in the left- and right-hand channels are about equal. Arithmetically, therefore, left minus right is zero, that is, a mono signal without voice.

There is, however, a problem: the sound levels of bass instruments, more particularly the double basses, are also just about the same in the two channels. This is because on the one hand low-frequency sounds are virtually non-directional and on the other hand, the recording engineers purposely use these frequencies to give a balance between the two channels.

However, the bass instruments may be recovered by adding those appearing in the left+right signal to the left-right signal. The whole procedure is easily followed in the circuit diagram. The incoming stereo signal is buffered by A1 and A2. The buffered signal is then fed to differential amplifier A3 and subsequently to summing amplifier A5. The latter is followed by a low-pass filter formed by A6. You may choose between a first-order and a second-order filter by respectively omitting or fitting C2. Listen to what sounds better.

The low-frequency signal and the difference signal are applied to summing amplifier A4. The balance between the two is set by P1 and P2 to individual taste.

You may have noticed that the circuit does not contain input
or output capacitors. If you wish, output capacitors may be added without detriment. However, the fitting of input capacitors is not advisable, because the consequent phase shift would adversely affect the operation of the circuit.

(A. Roelen)

005

GENERAL INTEREST

POWER BOOSTER FOR 7406/7407

It often happens that a digital signal is required for controlling a relay, a stepper motor, or other kind of relatively heavy load. This makes it necessary for both the output current and the output voltage of the relevant device to be increased. Some logic devices are provided with constant-voltage open-collector outputs, but these are invariably restricted to 15 V or 30 V.

With a little dexterity, it is possible to provide a 7406 or 7407 with a dedicated open-collector output – see Fig. 1. If you aim for the construction in Fig. 2, the result will take not much more space than a standard 7406. Since the output stage inverts, it is necessary to use the non-inverting 7407 to obtain an inverted output.

The transistor must be chosen in accordance with the wanted output. For most general purposes, the BC546 is perfectly satisfactory (200 mA at 65 V).

(A. Schaffert)

006

POWER SUPPLIES

78XX MONITOR

When a voltage regulator is supplied from a mains adapter, it sometimes happens that its output is too low (because the output from the adapter is too low, or because the voltage has dropped to a low value owing to an overload). It is useful if a warning of that situation is indicated.

The proper operation of the 78xx series of regulators depends on the difference between input and voltage voltage, which must be not less than 3 V (worst case; many regulators are much better).
The voltage drop across the regulator is monitored by IC1. The input and output voltage of the regulator are supplied to IC1 via potential dividers. If the input voltage to the regulator is too low, IC1 goes high, which causes C1 to charge and this turns on T1, so that D2 lights. You may, of course, use a buzzer instead of D2 and R7. The charge on C1 ensures that the LED lights for at least 10 ms. This means that the circuit will react to even very short voltage drops at the input of the regulator. A large ripple that results in a too low input voltage is therefore clearly indicated. The circuit is based on the 7805; the value of R1 must be redimensioned for other members of the 78xx series by the following:

\[ R1 = \left(\frac{2dU}{U_r}\right) + \frac{11}{R2} \]

where \( dU \) is the voltage drop across the regulator and \( U_r \) is the characteristic output voltage of the regulator. It is necessary that \( dU \) is chosen somewhat larger than the actual minimum voltage drop across the regulator to prevent non-operation of the circuit. This is so because the minimum voltage drop across the regulator is constant when the device ceases to function properly until the input voltage returns to normal. In other words, the monitor must be able to react to a voltage drop that is slightly higher than the minimum voltage drop.

### Radio & TV

#### Radio Beacon Converter

The radio beacon band extends from 280 kHz to 516 kHz. Each beacon has its own characteristic AM modulated morse-coded call sign that is transmitted on a specific frequency. To be able to receive distant beacons, the aerial signal is passed through a band-pass filter that effectively suppresses long-wave and medium-wave signals. The filter also converts the aerial impedance, \( Z_{in} \), from about 10 kΩ to the input impedance of mixer IC1, which is about 1 MΩ.

The mixer adds or subtracts the received signal to/from the local oscillator signal, so that the beacon signal can be received on a normal short-wave receiver. The resulting frequencies lie in the range 9.72-9.484 MHz or 10.280-10.516 MHz.

In the construction of the converter some components must be surrounded by a metal shield as indicated by dashed lines on the PCB layout.

The circuit is aligned with the aid of an SSB receiver to which the output of the converter is connected. Tune the receiver to 10 MHz and adjust the oscillator frequency of the converter by means of C8 for zero beat. Next, detune the receiver slightly until a pleasant whistle is heard, which is adjusted for minimum level with the aid of P1. Finally, tune to a beacon transmitting at about 300 kHz and adjust C13 for maximum sound output.

### Parts List

<table>
<thead>
<tr>
<th>Resistors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1, R2 = 40 kΩ</td>
</tr>
<tr>
<td>R3 = 470 kΩ</td>
</tr>
<tr>
<td>P1 = 25 kΩ preset potentiometer</td>
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</tbody>
</table>

<table>
<thead>
<tr>
<th>Semiconductors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1 = zener diode 8.2 V</td>
</tr>
<tr>
<td>IC1 = NE602</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Capacitors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1 = 68 pF</td>
</tr>
<tr>
<td>C2, C6, C15 = 27 pF</td>
</tr>
<tr>
<td>C3, C16 = 1 nF</td>
</tr>
<tr>
<td>C4 = 120 pF</td>
</tr>
<tr>
<td>C5 = 82 pF</td>
</tr>
<tr>
<td>C7 = 220 pF</td>
</tr>
<tr>
<td>C8; C13 = 40 pF trimmer</td>
</tr>
<tr>
<td>C9 = 15 pF</td>
</tr>
<tr>
<td>C10; C11 = 150 pF styrofoam</td>
</tr>
<tr>
<td>C12 = 100 nF</td>
</tr>
<tr>
<td>C14 = 330 pF</td>
</tr>
<tr>
<td>C17 = 1 µF; 16 V</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Inductors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1, L5 = 2.2 mH</td>
</tr>
<tr>
<td>L2, L4; L6 = 4.7 mH</td>
</tr>
<tr>
<td>L3 = 1.5 mH</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Miscellaneous:</th>
</tr>
</thead>
<tbody>
<tr>
<td>X1 = 10 MHz crystal; 30-pF parallel resonance</td>
</tr>
</tbody>
</table>

ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT
Although the NE604 is, strictly speaking, intended primarily for h.f. applications, it may be used for a number of other purposes. One of these is a sound level meter for audio applications as presented here.

Use is made of the IC's signal-strength indicator that is based on an internal logarithmic converter. This enables us to obtain a linear decibel scale, so that the moving-coil meter shown in the diagram may be replaced by a digital instrument.

The signal source is assumed to be an electret microphone that converts ambient noise into an electroical signal. Since this type of microphone normally contains a buffer stage, R7, R8 and C13 have been included to provide the supply voltage for this stage.

The NE604 delivers an output current (at pin 5) of 0-50 μA, which causes a potential difference across R2+R3 of 0-5 V. The range over which the relation between input and output signal is logarithmic is, however, slightly smaller: about 0-4 V. This is equivalent to a sound range of 70 dB.

To compensate the effects of temperature changes, the required resistance of 100 kΩ is formed by two resistors, R2 and R3, and a diode, D1.

Any ripple remaining on the output voltage is removed by R4-C9-C10 before the output is buffered by IC2.

The indicating instrument, here a moving-coil meter, is connected to the output (pin 6) of IC2 via a series resistance, R6+P1. The preset is adjusted to give full-scale deflection (f.s.d.) for an output voltage of 4 V.

Calibrating the meter is a little tricky, unless you have access to an already calibrated instrument. Otherwise, if you know the efficiency of your loudspeaker, that is, how many decibels for 1 W at 1 metre, you can use that as reference. The scale of the meter can then be marked with the (approximate) value. In any case, the meter deflection must at all times be seen as an indication, not as an absolute value: it was not thought worthwhile to add a filter to the circuit to enable absolute measurements to be made.

**Parts list**

Resistors:
- R1: R6 = 2k2
- R2 = 60k4 (E96) or two 120 kΩ resistors in parallel
- R3 = 40k2 or 39 kΩ in series with 1 kΩ
- R4: R5 = 1 M
- R7 = 1 k
- R8 = 10 k
- P1 = 10 k preset potentiometer

Capacitors:
- C1 = 47/63 V radial
- C2 = 15 n
- C3 = 220 μ/16 V radial
- C4-C7 incl. = 10 μ/16 V radial
- C8 = 47 μ/16 V
- C9-C12 = 100 n
- C10 = 4n7
- C11-C12 = 1 n
- C13 = 47 μ/10 V

Semiconductors:
- D1 = 1N4148
- IC1 = NE604
- IC2 = 3130

Miscellaneous:
- S1 = push-button switch with one make contact
- M1 = moving-coil meter, 1 mA
RECORDING CONTROL

The circuit presented here is intended as a recording level control for cassette recorders. It enables the reading on the VU meters to be kept out of the red sector without the necessity of the recorder's level control to be adjusted. This is particularly useful when speech is recorded.

The recording signal is controlled via two voltage dividers, each containing a light-dependent resistor - LDR. Unfortunately, these devices introduce slight distortion, but that is still better than a lot of distortion through overloading of the tape.

The circuit monitors the signal level at the headphone output of the recorder. The signal there is rectified (half wave) and then applied to T1. The speed with which T1 can react is determined by C1. As shown, the attack time is set at 3 ms. The release time of the circuit is determined by R4-C4 and is much longer than the attack time: about 100 ms.

The voltage across C2 determines to what extent the signal to be recorded is attenuated. The charging current through R4 depends on the voltage across the capacitor, so that the light intensity of the LEDs is also dependent on the signal strength.

The light from D1 and D2 falls on to R7 and R8, which form part of the potential dividers that are added to the input circuit of the recorder.

The circuit is easily disabled by opening S1.

CHILD-PROOF RESET SWITCH

The reset switch on a computer is a very important control. If an operating instruction threatens to wreck the internal management of a computer, the reset button is often the only way of avoiding a possible disaster. On the other hand, it may also be the cause of a disaster. After all, one touch on it and hours of work may be negated in an instant, that is, if you do not save your work every fifteen minutes or so. None the less, anyone can have an accident, but it is particularly important that children or pets can not inadvertently operate the control.

The circuit proposed here should put an end to your worries in this respect. Instead of one reset switch, it is necessary to press four switches simultaneously. The chances of this happening by accident or child or pet are so small as to be negligible.

The four switches are placed in positions that make it impossible to operate them all with one hand. Instead, two of them can be operated with the fingers of one hand and the other two with the fingers of the other hand.

As shown, the four switches are connected in series and are intended to replace the existing switch.

(S.G. Dimitriou and F.P. Maggana)
Battery-operated equipment can work on one set of batteries for a long time nowadays. But if it is left on inadvertently, that 'long time' is over very quickly. Moreover, flat batteries are always found at the wrong moment. The circuit proposed here is a sort of aide-memoire. Every two minutes it emits 5–10 pips to indicate that the equipment is still switched on.

Basically, the circuit consists of three rectangular-wave generators and an inverter. The first of the generators is formed by N1 and provides a signal with a period of about two minutes and a pulse duration of around ten seconds. During those ten seconds, the second generator starts operating in a one-second rhythm. Thus, N2 outputs ten pulses every 2 minutes. That output is inverted so that N4, like N2, can only be enabled during the 10-second pulse train from N1. There is a difference, though: during those 10 seconds, N4 is enabled and inhibited ten times and this is what causes the pips.

Do not take the times and number of pulses too literally, because there are wide variances between ICs from different manufacturers. On the other hand, component values are not critical, so that it is fairly easy to adapt the circuit to personal taste or requirements. The buzzer may be a standard Toko type or equivalent. Finally, the current drawn by the circuit is negligibly small.

(R. Kambach)
ing to the root-mean square—RMS—value; and the third corresponding to the logarithmic—LOG—value of the input current, since the IC operates with currents. This makes possible a wide dynamic range: something like 100 dB. Translated into terms of input current, that gives 3 mA (p-p) to 30 nA (p-p). This makes it vital that C1 is a type with a very small leakage current.

The input voltage is converted into current by R1. With a value as shown for this component, the nominal input level is 0 dBV with a reserve up to +20 dBV.

Like the input signal, the output signals are currents that are converted into voltages by opamps A1, A2 and A3, which are connected as current/voltage amplifiers.

To enable the decibel scale to be calibrated accurately, a preset potentiometer, P1 has been included in the A3 stage.

With component values as shown in the diagram, the output voltage from A3 varies at about 33 mV/dB. It should be noted, however, that the dynamic range of the LOG output at 80 dB is rather narrower than the maximum possible 100 dB.

The proposed mixer is designed around four current-driven transconductance amplifiers contained in a Type SSM2024 from Precision Monolithics Inc—PMI. To obtain a low off-set and high control rejection, the four inputs should have an impedance to earth of about 200 Ω. These impedances are obtained from resistors R5–R8 that also form part of a potential divider at each input.

With values as shown in the diagram, the nominal input signal is 1 V (0 dBV). Distortion at that level is about 1% and at lower levels not more than 0.3%.

The amplification of the current-driven amplifiers—CCAs—is determined by the current fed into the control inputs. These inputs form a virtual earth so that calculating the values of the bias resistors (to transform the inputs into voltage-driven inputs) is fairly simple.

With a value for R1–R4 of 10 kΩ, the CCAs are switched off if the potential at the control inputs is lower than 200 mV. Maximum amplification is obtained at a drive current of 500 µA. The voltage at the control inputs is then slightly higher at 0.5 V, so that a maximum control voltage of 5.5 V is needed.

The output currents of the amplifiers are summed by simply linking the output pins (it is that simple with current outputs and completely in agreement with Kirchhoff's rules).

The current-to-voltage converter, IC2, translates the combined output currents into an output voltage. The value of R13 ensures that the amplification of IC2 is unity.
The current drawn by the mixer depends on the setting of the four CCAs and lies between 5 and 9 mA.

The signal-to-noise ratio of the mixer is about 90 dB, while the bandwidth is of the order of 130 kHz. The bandwidth is limited mainly by C1, which is essential to ensure good stability of the current-to-voltage converter.

The SSM2024 operates satisfactorily with a supply voltage between ±9 V and ±18 V, but best results are obtained when it is ±15 V.

The “64 Kbyte RAM extension for MSX computers” (Ref. 1) can house EPROMS, but there is a small problem. To keep the board layout as simple as possible, the data and address lines on the extension board were not connected sequentially. With RAMS that does not matter too much and, in principle, not with EPROMS either, but their programming becomes a bit of a tangle. To prevent that, it is possible to use the auxiliary board shown here, which ensures that the address and data lines are connected to the correct pins of the IC. The board may be adapted to two types of EPROM by means of a wire link. With the link between A and B, a Type 27128 EPROM (16 K) may be used; if a 32 K EPROM is to be used, the link should be laid between A and C.

The two RAMS on the extension board each take two pages (a total of 32 K) in the address memory (pages 0/1 and 2/3 respectively). That means that when a Type 27128 EPROM is used, the data occurs twice on both pages.


This squelch circuit is simple, universally usable and provides a large enough gain to enable it being incorporated in the automatic-gain control (AGC) circuit of a variety of receivers.

The input signal derived from the AGC circuit in a receiver is attenuated by network R1-R2-P1. The signal at the wiper of P1 is taken to the inverting input of opamp A1, which is connected as a comparator. The non-inverting input is provided with a reference voltage of 200 mV by potential divider R9-R10. The output signal of A1 is applied to a Schmitt trigger circuit, A2, via low-pass section R2-C2. This filter ensures that small noise and other interference signals do not affect the correct functioning of the squelch.

Capacitor C3 removes the steep skirt of the output signal of A2, which renders the operation of the AGC rather more pleasant to the ear. The output of A2 is then taken to the base of output transistor T1 via potential divider R7-R8.

The open-collector output of the squelch may be used to suppress the audio frequency output of the receiver.

Since the squelch draws only a small current, less than 10 mA, its incorporation into an existing receiver should not present any problems as far as the power supply is concerned.
TEST & MEASUREMENT

48 MHZ CMOS OSCILLATOR

Crystal oscillators using digital gates generally do not generate frequencies above 30 MHz, because the necessary crystals are not allowed to oscillate on their fundamental frequency.

In the oscillator shown here, the crystal is forced to oscillate on the third harmonic since it is connected in series with a parallel tuned circuit that resonates at the fundamental of 16 MHz. For digital applications, the output of N1 may be enhanced by the use of a second inverter.

The circuit will operate satisfactorily only if non-buffered CMOS devices are used. Gates from the HCU family will enable operation up to 60 MHz.

SIMPLE TEMPERATURE INDICATOR

For the absolute measurement of temperatures a thermometer is indispensable. There are, however, many situations where an absolute value is not needed and a relative indication suffices. The getting hot of an electric drill or vacuum cleaner may be indicated to the user by the lighting, or changing of colour, of a simple indicator light. It would be a further advantage if on such equipment a green light would indicate that all is well as far as temperature is concerned. As the temperature rises, the light should change colour slowly to indicate that the equipment is getting too hot.

The proposed circuit does this and has the additional advantage that it does not need a separate low-voltage supply: it works direct from the mains. The indicator proper is a two-colour LED, D1, while the sensor is a combination of a negative temperature coefficient (NTC) and a positive temperature coefficient (PTC) resistor, R4 and R3 respectively.

At a relatively low temperature, the value of R3 is low and that of R4 is high. During the positive half cycle of the mains voltage, a voltage will exist across R3-D3 that is sufficient high to cause the green section of D1 to light. The value of R3 has been chosen to ensure that during the negative half cycle of the mains voltage the potential across it is too low to cause the red section of D1 to light.

If the temperature rises, the value of R4 diminishes and that of R3 rises. Slowly but surely the green section will light with lesser and lesser brightness, while at the same time the red section lights with greater and greater brightness until ultimately only the red section will light.

Resistor R2 and capacitor C3 ensure that the current drawn by the LEDs does not become too large. This arrangement keeps the dissipation relatively low.

Both R3 and R4 should be of reasonable dimensions, something like 6 mm in diameter - not less. At a temperature of 25 °C, the NTC must have a value of 22-25 kΩ and the PTC one of 25-33 Ω.

The circuit should be treated with great care since it carries the full mains voltage.

AUDIO & HI-FI

BUCKET BRIGADE DELAY LINE

Although bucket-brigade devices (BBBs) are not in the same limelight as many other components, they do exist and are used. One of them, Type MN3004, consists of 512 capacitors and switching transistors. The IC is perhaps best described as an analogue shift register. The capacitors are provided by the drain-gate capacitance of the transistors.

Samples of an analogue signal taken at the input appear 256 clock pulses later at the output. The clock pulses are obtained

ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT
from an associated IC, a type MN3101. This IC also provides a supply voltage, UGG, for the source followers in IC1. Note that the connections of the supply voltages to IC1 and IC2 have not been drawn incorrectly: VDD must be negative with respect to earth (GND pin).

The MN3101 also makes it possible to construct an IC oscillator. Trimmer C9 enables varying the clock frequency and thus the delay time. It is also possible to connect an external oscillator to pin 7 (but note that pins 5 and 6 should be left open).

The clock frequency of the delay line may lie between 10 kHz and 100 kHz. With component values as shown in the diagram, it is about 60 kHz.

Maximum dissipation of the clock generator is about 200 mW, depending on the capacitive load of the memory. If more memories are used, it is possible to reduce the dissipation by connecting resistors in series with pins CP1 and CP2. Lowering the clock frequency will also reduce the dissipation.

The delay time of the circuit is equal to half the number of capacitors divided by the clock frequency: with values as shown, it amounts to 2.56-25.6 ms. The bandwidth of the delay line is roughly 0.33 times the clock frequency. Thus, with a clock frequency of 60 kHz, the bandwidth is 20 kHz and the delay more than 4 ms. In the prototype, these values yielded a signal-to-noise ratio of more than 70 dB and a distortion not exceeding 0.3% (at 1 kHz, 0 dBV). The current drawn was just over 6 mA, but this may increase to 14 mA with increased clock frequencies.

Apart from the delayed input signal, the output signal contains mixing products of the input and the clock. In the prototype (circuit as shown), with a clock of 60 kHz, these products were suppressed in the audio range by 60 dB (R8-C3 and R9-05). It is, however, advisable when a lower clock is used to filter both the input and the output with a filter of at least the fourth order. Try to minimize the distortion by adjusting P1.

The delay line should find application in units for echo, tremolo, vibrator, chorus, reverb, and so on. Another possibility is its use in a compressor to suppress, or at least attenuate, the short periods of overdrive. To that end, delay the input signal prior to the voltage-controlled amplifier (VCA) and drive the circuit that provides the control voltage for the VCA direct. In that application, the delay should be at least equal to the attack time of the compressor.

**Parts List**

- Resistors:
  - C4 = 2 µF
  - C5 = 82 n
  - C6 = 22 µF; 25 V
  - C7; C8 = 47 n; 16 V
  - C9 = 100 p trimmer
  - C10 = 220 n
  - C11 = 100 p
  - C12 = 100 n
  - R1; R2; R13 = 10 k
  - R3; R4; R12 = 100 k
  - R5; R6 = 5 k
  - R7; R8 = 150 k
  - R9 = 100 R
  - R10; R11 = 68 k
  - P1 = 50 k preset
  - PC5; PC6 = PC2

- Semiconductors:
  - D1 = 1N4148
  - IC1 = MN3004
  - IC2 = MN3101
  - IC3 = TL072

**Fast Envelope Sampler**

When a signal is being amplitude-modulated, constant monitoring is necessary to ensure that the maximum modulating frequency does not exceed half the carrier frequency. The fast envelope sampler presented here is basically an advanced AM
demodulator that can be used with modulations where the well-known diode detector with LF filter can not. Phase errors that occur with diode detection are absent in this circuit.

Amplifier IC1 is a buffer with variable (1-11) AC amplification. Opamps A1 and A3 form an active half-wave rectifier that charges capacitor C5 up to the maximum signal level in a half period.

During the zero crossing, detected by A2, the network around N2 generates a short pulse. This pulse ensures that the potential across C5 is applied to C6 via electronic switch ES4. After ES4 has been opened, so that the charge is stored safely in C6, capacitor C5 discharges via parallel-connected switches ES1-ES3 (which are actuated via the network around N3 and across C6. Basically, the circuit is frequency-independent since the clock signal is derived from the carrier. This is the reason that the sampler may be used in such diverse applications as satellite facsimile, radio receivers and speech processors.

The circuit may also be used to good effect in an AGC loop, because the classical problems regarding attack and delay associated with diode detection and LF filter systems do not occur in this type of demodulator.

Since in some applications a direct voltage is needed, for instance, in an AGC loop, the proposed circuit has a DC as well as an AC terminal.

The sampler is intended for operation from a ±5 V supply and draws a current of about 20 mA.

**NOISE GENERATOR**

The output of the sampler is in accordance with the potential.

The noise generator presented here provides constant noise energy over its bandwidth, resulting from the non-linear behaviour of its switching components, more particularly T1. It is very useful for measurements where limited noise bands are required. Varying the ratio R6:R7 and the clock frequency enables the generated noise to be adapted to specific requirements.

Transistors T2 and T3 are current sources. The current through T2 is about ten times the level of that through T3.

Assuming that T4 is on and that the clock input is low, T1 is off and C2 discharges. The capacitor is pulled to about half the supply voltage by the two current sources. When that state is reached, stability ensues because the potential then present at the gate of T4 keeps the FET switched on.

When the clock goes high, T1 is switched on so that C2 is connected between the gate of T4 and earth. Since C2 is only partly charged, the FET is switched off. Transistor T1 is kept switched on by or gate D1-D2 so that the clock pulses are blocked. Capacitor C2 then charges via T3 until the potential across it becomes high enough to switch on T4. Transistor T1 is then switched off and the circuit is ready to receive another clock pulse (or rather a leading edge of one).

Since it is not known when the clock pulse arrives, it is not known to what potential C2 will be discharged by T2 (and countered by T3). It is therefore also not known how long it takes T3 to recharge C2. It follows that it is then not known when the next clock pulse arrives. In other words, the pulse width of the output signal is varying constantly, which is characteristic of a noise signal.

The frequencies contained in the noise signal are limited by the clock signal (higher frequencies than the clock can not occur, although there are harmonics) and the maximum charge and

ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT
TUNEABLE BAND-PASS FILTER

One of the difficulties in the design of higher-order tuneable band-pass filters is achieving correct tracking of the variable resistors in the RC networks. The use of switched capacitor networks can obviate that difficulty as is shown in this filter.

The filter may be divided roughly into two stages: an oscillator that controls the electronic switches and the four phase-shift networks that provide the filtering proper.

The oscillator, based on a 555, generates a pulsating signal whose frequency is adjustable over a wide range: the duty factor varies from 1:10 to 100:1.

Electronic switches ES1-ES4 form the variable resistors whose value is dependent on the frequency of the digital signal. The operation of these switches is fairly simple. When they are closed, their resistance is about 60 Ω; when they are open, it is virtually infinitely high. If a switch is closed for, say, a quarter of the time, its average resistance is therefore 240 Ω. Varying the open:closed ratio of each switch varies the equivalent average resistance. The switching rate of the switches must be much greater than the highest audio frequency to prevent audible interference between the audio and clock signals.

The input signal causes a given direct voltage across C1, so that the opamps may be operated in a quasi-symmetric manner in spite of the single supply voltage. The direct voltage is removed from the output signal by capacitor C10.

The fourth-order filter in the diagram may be used over the entire audio range and has an amplification of about 40, although this depends to some extent on the clock frequency. The bandwidth depends mainly on the set frequency.

The circuit draws a current of not more than 15 mA.

CAR HEADLIGHT CONTROL

It is an annoying fact that you normally only realize that you have left your car headlights on when you want to restart the car only to find that the battery is flat. One of the possible ways of preventing this happening is offered by the present control.

The circuit does not provide a warning but an action: when you switch off the ignition, relay R1 is de-energized and the headlights are switched off, unless you deliberately decide otherwise. That decision is made possible by switch S1, which, when operated, triggers silicon-controlled rectifier Th1 so that R1 is energized. Note that this is possible only when the ignition switch, S2, is off, otherwise the voltage across Th1 is so low, owing to shunt diode D1, that it cannot be triggered. Since, however, the headlights should not normally be switched on when the ignition is off, in most cases S1 will be used only rarely and the switch may then well be omitted altogether.

Relay R1 should be a standard 12 V car type with contacts that can switch up to 25 A.

(H. Huyten)
Most advanced computer programs include preventative measures against (possibly) hasty instructions. Responses such as “are you sure?” and “do you really want to quit?” are familiar to most of us. However, even the cleverest program can not prevent the inadvertent operation of the reset switch and the consequent result of lost data and improperly closed files that cause wasted clusters on the hard disk.

The location of the reset switch on the front panel of many computers asks, of course, for inadvertent operation. Clearly, some means of reset protection is no luxury.

Normally, the reset switch is connected to the mother board of the computer via two wires. One of these is at earth potential and the other is linked to the reset circuit. The protection, whose circuit is shown in the diagram, is inserted between the reset switch and the mother board. The earth connection of the computer must be linked to terminal M of the protection circuit. The protection circuit may draw its power from the computer supply.

When the circuit has been fitted, operation of the reset switch will not result in an immediate restart of the computer. Instead, a buzzer will sound to alert you to the reset operation. The buzzer is actuated for four seconds by monostable IC1a, which is triggered by the reset switch. During those four seconds, the output, pin 5, of IC1a ensures that the reset function, pin 10, of IC1b is disabled. When then the reset switch is operated again, monostable IC1b will be triggered and this will start the reset procedure. Transistor T2 is then switched on for half a second and the buzzer is deactuated via R1 and D4.

The circuit around T1 and N4 ensures that IC1a can accept trigger pulses again ten seconds after the mono time of IC1b has lapsed. This arrangement prevents, say, children operating the reset switch.

**COMPONENTS**

**BICMOS INTEGRATED CIRCUITS**

BICMOS devices combine bipolar and CMOS technologies, providing the best features of both: the fast performance and the 48/64 mA output drive of bipolar devices and the low power consumption of CMOS devices.

In active mode, BICMOS devices operate at about half the supply current of their pure bipolar equivalents. When disabled, they may reduce power consumption by up to 90%. Since at present most BICMOS ICs function as bus interfaces (that are normally disabled), the result may be a system IC-power saving of up to 25%.

Moreover, BICMOS devices use the typical 0.3–3.5 V TTL voltage swings at their output rather than the larger GND-to-VCC swings of CMOS devices. This smaller voltage swing reduces the overall effects of transient voltage noise produced during the simultaneous switching of multiple outputs.

A number of integrated circuits are available in BICMOS technology, including transceivers with registers, pipeline registers, 8/9/10-bit registers, latches and parity bus transceivers. They give the designer additional means of reducing power consumption without compromising advanced performance.

(Source: Texas Instruments)
The programmable switch may be used, for instance, to simulate a data stream or, as shown, to control an analogue multiplexer, IC5. The multiplexer may be used as the basis of a programmable oscillator.

The circuit is based on National Semiconductor keyboard decoder Type MM74C922 (IC1). This device is intended to read a 4x4 matrix keyboard in a simple and fast manner. Apart from a 4-bit output, the IC also has a DATA AVAILABLE output, DA, which is high as long as one of the keys is pressed. The data associated with the last pressed key will remain on the output, even after the key has been released. The speed with which the keyboard is scanned is determined by C1. When this capacitor has a value of 100 nF, as shown in the diagram, the scanning rate is 600 Hz. The anti-bounce period of the keyboard is determined by C2: with a value as shown, the period is about 10 ms (rule of thumb: C2=10C1).

The programming of the RAM, IC2, is fairly simple with the aid of IC1. First, the DA output of the decoder is inverted by N1 to enable it being used as the WRITE pulse. After the key has been released, the RAM is disabled (DA=1, WE=0), and the address counter, IC3, is clocked on by one address. Since the counter is a Type 4040, it reacts to trailing edges and this means that WE has to be inverted again (by N4). It is not possible to use the original DA signal, since this could jeopardize the timing and cause non-defined states. The delay times of the gates ensure that all processes take place in correct sequence. Further delay is provided by the combination of R4 with the capacitance of the clock input.

The programmed data is read with the aid of a separate clock signal generated by an oscillator based on N3. The speed at which the data is read from the RAM is set by P1. Gate N5 is actuated by setting S18 in position A. Contact bounce here is prevented by the combination of R2 and C3, and the hysteresis of N3 and N2. When S18 is in position A, gates N1 and N2 ensure that the RAM is in the read mode (WE=1). At the same time, N2 arranges that the data lines of the RAM are connected as outputs. Gate N4 ensures that either DA or the output of N3 is used as the clock for counter IC3. Furthermore, S18 also arranges the disabling of the data outputs of IC1 to prevent a bus conflict (the RAM now provides the data).

Pin 1 (Q11) of counter IC3 is connected to the clock input, pin 10, via or network R4-D1 to ensure that the counter stops after one cycle. The stopping is indicated by the lighting of D1. Switch S17 enables the counter to be reset; it may also be used to set the counter to zero during programming.

When, during programming, the highest counter position is reached, the clock for the counter is disabled. Continued pressing of the keys will cause overwriting on to the last RAM address.

To obtain a defined counter state on switch-on, network R3-C5 provides a power-on reset of the counter.

Only one half of the RAM is used, since four bits suffice for the proposed circuit. It is, therefore, possible to use a RAM with nibble configuration or use two keyboard encoders.

Finally, the use of 16 separate switches may make the circuit rather expensive: it is far cheaper to use a second-hand 4x4 matrix keyboard.
SWITCH-MODE VOLTAGE REGULATOR

Switch-mode power supplies offer the user the benefit of a much greater efficiency than obtainable with a traditional power supply. The switch-mode regulator presented here has an efficiency of around 85%.

An input voltage of 12–16 V DC is converted into a direct voltage of exactly 5 V. The use of a Type MAX638CPA enables the design and construction of the regulator to be kept fairly simple: only nine additional components are needed to complete the circuit.

Resistors R1 and R2 are used to indicate when the battery voltage becomes low: as soon as the voltage on pin 3 becomes lower than 1.3 V, D1 lights. With values as shown for the potential divider, this corresponds to the supply voltage getting lower than about 6.5 V.

The output of the IC is shunted by a simple IC filter formed by L1, C3 and D2.

The oscillator on board the IC generates a clock frequency of around 65 kHz and drives the output transistor via two NOR gates. The built-in error detector, the 'battery low' indicator or the voltage comparators block the clock frequency, which causes the transistor to switch off.

The IC compares the output voltage of 5 V with a built-in reference (FET). Depending on the load, the FET will be switched on for longer or shorter periods. The maximum current through the FET is 375 mA, corresponding to a maximum output current of 80 mA.

2-METRE TRANSMITTER

The transmitter was designed primarily for use by radio amateurs as a radio beacon and as such it provides a good quality signal free of unwanted harmonics.

Transistor T1, in association with crystal X1, operates as a 36 MHz oscillator. Filter L1-C3 obviates any tendency of the circuit to oscillate at 12 MHz (the fundamental frequency of the crystal).

Circuit L2-C4 is tuned to the fourth harmonic of the oscillator signal (144 MHz). This signal is fed to the aerial via a buffer stage consisting of T2, a double-gated FET. The (amplitude) modulating signal is applied to the second gate of the buffer. The output power of the transmitter has been kept low, about 10–40 mW.

The modulating signal is generated by N1, an oscillator that switches the transmitter on and off via transistor T3. The switching rate lies between 0.1 Hz and 0.5 Hz.

When the output of N1 is low, T3 is switched off, and the transmitter is inoperative because the supply is disabled. When the output of N1 is high, T3 is on and the transmitter operates normally.

The digital pattern at the gate of T2 shapes the modulating signal. Gate N2 generates a square wave at a frequency of 0.1–1 Hz. As long as the output of T3 is high, N4 oscillates at a frequency of about 1 kHz. At the relevant gate of N2 there is, therefore, a periodic burst signal at a frequency of 1 kHz, and this signal is used to modulate the transmitter.

The digital pattern at the relevant gate of T2 may be varied to individual requirements by altering the values of the feedback resistors in the digital chain.

The transmitter is calibrated by setting trimmers C4, C7 and C8 for maximum output power.

Inductors L2 and L3 are wound from 0.8 mm dia enamelled copper wire: L2 = 5 turns with a tap at 1 turn from ground; L3a = 3 turns and L3b = 2 turns. The coupling between L3a and L3b should be arranged for maximum output power.

The circuit draws a current of only 20 mA, enabling the transmitter to be operated from a 9-V PP3 battery for several hours.

ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT
The resolution of moving-coil meters is generally no better than 1%, because the scale normally is given no more than 100 marker lines for the given dimensions: more lines might detract from the legibility. Most digital meters have a resolution of 0.05% or better. The resolution of moving-coil instruments may be improved in two ways: physically enlarging the scale, which is possible only in the factory, or electronically enlarging the scale, which is what this article is all about.

The circuit divides the scale into five sections, each of which is then extended over the full scale. This therefore gives a five-fold improvement in resolution.

The input signal (200 mV = full-scale deflection - f.s.d.) is amplified by IC1 to a value of 2.5 V. The amplified signal is fed to four comparators that divide the input signal into segments of 40 mV. Which segment is indicated on the meter-scale is shown by the LEDs.

The LEDs are shunted by a 10 kΩ resistor, so that the open-collector outputs of the comparators have a well-defined pull-up resistance.

The outputs of the comparators drive two multipliers that, depending on the magnitude of the input signal, provide a direct voltage to buffer IC2. This direct voltage is always a multiple of 0.5 V (since the output of IC1 is 5 x 0.5 = 2.5 V).

There then exists a potential between A and B that is equal to the difference between the input signal and a multiple of 0.5 V. This difference can never exceed 0.5 V over the f.s.d. of the meter. Resistor R4 must therefore have a value that causes a f.s.d. at 0.5 V. The measured value is determined by adding the meter reading to the multiple of 0.5 V indicated by the LEDs.

The circuit is calibrated by first setting the meter to zero reading by P2. Next, apply a voltage equal to one fifth of the f.s.d. (here, 40 mV) to the input. Then, set P3 for minimum resistance, when none of the LEDs should light. Next, set P1 for f.s.d. Finally, adjust P3 until D7 begins to light and the meter reading falls to zero.

(R. Shankar)

**METER-SCALE MAGNIFIER**

**SOUND LEVEL ATTENUATOR**

When the radio or record player is on at a fairly high volume, it is often impossible to hear the telephone or doorbell. A solution to this frequent difficulty is offered by this automatic attenuator. As soon as the doorbell or the telephone rings, it turns down the volume of the audio equipment.

The circuit consists of an optically controlled attenuator and the requisite electronics to connect it to, say, the telephone.

The attenuator is of fairly simple design and is based on a TL074. Its control part, consisting of a current-driven attenuator based on an LT2001 (a combination of an LED and two light-dependent resistors - LDRs - in a common enclosure), is incorporated in the audio equipment.

After the mains has been switched on, C9 causes the resetting of bistable IC1. The high voltage level at pin 6 of the 741 causes T1 and T2 to switch off: this, in turn, results in D2 not conducting and the voltage-controlled current source, T3, delivering maximum current (30 mA) to the LED incorporated in the LT2001. The illuminated LEDs will then have a value of about 1.5 kΩ. The voltage transfer of the attenuators can be preset (once and for all) to exactly 0 dB (at 1 kHz) by P1 and P2.

ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT
Terminals A and B of the circuit are connected direct to the corresponding terminals of the telephone (this may not be allowed in some countries - seek the advice of your local telephone manager), while terminals X and Y are connected across the doorbell terminals. Note that the doorbell must be fed from a 3-24 V transformer. If the telephone or the doorbell rings, the bistable will be set via the relevant opto-isolator (IC4 or IC5).

The low voltage level at the output of IC1 will cause T1 and T2 to switch on. This in turn causes D2 to light and C12 to charge via R23. Owing to the rising potential across the capacitor, the output of the current source will slowly diminish until the minimum value, set by P3, is reached. This has the effect that the TL074 turns down the volume click-free until a reasonable sound level, determined by the setting of P3, is reached.

Pressing switch S1 resets the bistable. This will cause D2 to go out, while the attenuation slowly drops to 0 dB. The attenuator is connected to the control electronics by two wires, P and Q. Thanks to the current drive, this (non-shielded) link may be up to 23 metres (75 ft) long.

The attenuator draws a current of only 10 mA and must be fed from a symmetric ±12 V supply, which may be taken from the audio amplifier. The control circuit needs an asymmetric +12 V supply and draws a current of about 35 mA.

If the LT2001 is difficult to obtain, discrete components may be used: these should, of course, be fitted in a light-tight enclosure.

### COMPONENTS

**FAST UNITY GAIN OPAMP**

A number of operational amplifiers can be used only in circuits that have a certain minimum amplification, because they have been designed with small internal compensation. The advantage of that is that the amplifier is faster.

If we look at two popular opamps, the LF356 and LF357, these characteristics are well illustrated. The LF356 may be used as a unity-gain amplifier with a gain-bandwidth product of 5 MHz and a slew rate of 12 V/µs. The LF357 needs an amplification of not less than 5 and has a gain-bandwidth product of 20 MHz and a slew rate of 50 V/µs.

It is possible to use the LF357 (and similar opamps) with smaller amplifications by using external compensation, yet retaining most of the bandwidth. Normally, this is achieved with a capacitor, but that is not the only, and certainly not the best, method.

An alternative method is illustrated in the diagram. Consider the L165 as a summing amplifier of which one input is connected to earth. It is clear that the second input then forms the input of a unity-gain amplifier whose amplification is determined by R1 and R3 (R3/R1 = 1). The unused input would have provided an amplification of 22 (R3/R2 = 22), which is rather more than the permissible minimum amplification of the L165. The opamp 'believes' that the amplification is higher than required. This has the benefit that the circuit has no tendency whatsoever to oscillate.

The ratio R3:R2 is only of value if R1 is very much larger than R2. Otherwise the amplification is R3/(R1//R2).

Note that the L165 must be fitted on a good heat sink, in spite of its internal thermal overload protection (Iout max. = 3 A).
**PRINTER RESET**

When during a computer print-out something goes wrong with the printer, such as the paper getting snarled up, the only way to stop the operation is normally to switch the machine off. That may be a useful, but certainly not an elegant, method: a reset knob, on the other hand, is.

Nearly all printers with a Centronics interface have a reset input at pin 31 of the Centronics connector (consult the handbook). That input is used in many MS-DOS systems to set the printer to a defined starting state and at the same time to empty the buffer.

The input may, of course, also be used to connect a reset switch to. The diagram in Fig. 1a shows how such a switch may be made quite easily. The 1 kΩ resistor prevents the computer output being short-circuited when the printer is being reset.

Users of the recently published printer buffer (Ref. 1) can fit the switch in the buffer or expand the existing switch so that the printer is reset at the same time as the printer buffer. The circuit is connected to the unused contact of S1 in the printer buffer. The existing reset switch may be expanded as shown in Fig. 1b.

Fig. 1a  Fig. 1b

(1) "Centronics-compatible printer buffer", Elektor Electronics, March 1989, p. 21.

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**ELECTROPHONICS**

**VARIABLE LOW-PASS FILTER**

The Type SSM2045 IC from PMI is an active low-pass filter whose order, Q-factor, cut-off frequency and amplification are set with the aid of control signals.

A possible design for use in electronic music systems is shown in the diagram. To prevent distortion, the level of the input signal is reduced by R1 and a resistor on board the chip to one not exceeding 150 V peak to peak.

Outputs 2 POLE and 4 POLE are each connected to an internal voltage-controlled amplifier (VCA), MIX1 and MIX2 respectively. For optimum off-set and control rejection, these connections are made via resistors R4 and R5. The gain of the VCAs is set by P2, which controls the current that flows through the amplifiers via pins 15 and 16. The maximum current at these control inputs is 250 µA. The balance between the VCAs, and with it the order of the filter, is set by P4 via pin 14. The voltage at this control input can be varied between -250 mV and +250 mV. The input must be driven from an impedance not exceeding 200 Ω. At a drive voltage of 0 mV, the VCAs attenuate the signal by about 6 dB.

The Q-factor depends on the current flowing into pin 17, which is controlled by P1. The input is protected by an internal 18 kΩ resistor. The Q-factor may be set so high as to cause the circuit to start oscillating. This happens when the current is between 120 and 185 µA.

The cut-off frequency can be shifted between 20 Hz and 20 kHz by varying the control voltage at pin 5 between +90 mV and -90 mV by P3. This voltage also determines the frequency of oscillation, enabling a variable oscillator to be created. The resulting sine wave has a distortion of about 1%.

The values of C1−C4 have been chosen to give the
filter a Butterworth characteristic if the current into pin 17 (Q) is zero.

The supply to the IC may be -5 V connected direct to pin 10, or up to -15 V via series resistor R13, through which a current of about 7.1 mA flows. The supply, -Vee, is limited internally by a 6.8 V zener diode.

The output current of the chip is converted into a voltage by IC2. The output of this stage has a small off-set voltage. If this can not be tolerated by the following equipment, the output must be taken via a coupling capacitor.

In the proposed circuit, the values of resistors R2, R6, R8, R9 and R12 have been chosen to allow control of the IC from voltages of 0-5 V or ±5 V.

At an input signal of 0 dBm, the distortion is about 1%, which drops to 0.3% at -6 dBm and 0.03% at -20 dBm. The signal-to-noise ratio is of the order of 80 dB.

AUTOMATIC SWITCH

Many battery-operated instruments, such as digital voltmeters, have a change-over on-off switch. The automatic switch presented here makes full use of that. When the instrument is off, C1 charges fairly quickly via R2 and D1. As soon as the instrument is switched on, C1 discharges slowly via R1. As long as the discharge current exceeds a certain level, T1 is switched on by the voltage across R1 and the supply to the instrument is on. When, after a few minutes, C1 is almost completely discharged, T1 toggles and the supply to the instrument is switched off. The period between switch-on and switch-off may, of course, be varied by different values of C1.

(Ph. Bosma)

LOW DISSIPATION REGULATOR

With the advent of the now well-known three-pin voltage regulators, power supplies are taken for granted. Yet, there are cases where such a regulator is not wholly satisfactory. This type of regulator needs a fairly large potential drop across it (typically not less than 3 V) and draws a relatively high quiescent current (typically 6 mA for a 78xx). The regulator presented here is particularly attractive for battery-operated equipments and offers:

- variable, very stable output voltage;
- low potential drop (some tenths of a volt);
- very small quiescent current (20-30 µA).

In principle, the regulator is a normal series type. The voltage reference is obtained from a common or garden red LED that must draw not less, nor much more, than 5 µA. Even at that low current, an LED has a fairly stable voltage drop. To improve that stability, the current is drawn from the regulated output via R1. Regulation is provided by CMOS opamp Type TLC271. This amplifier operates in the low bias mode, which ensures very low current consumption, by connecting pin 8 to the positive output terminal. The output of the opamp is used as base drive for series regulator T2 via current source T1. This configuration enables good control for only a small voltage swing at the output of the opamp. This is necessary since the slew rate of the opamp in the low-bias mode is pretty poor. The supply for the opamp is also taken from the regulator output. Capacitor C1 therefore serves as a decoupling element for the opamp.

To obtain reliable control, a kind of bootstrap resistor, R5, was found necessary.

The values of R1 and R4 as shown in the diagram provide a variable output voltage of 3-8 V. Higher output voltages, up to a maximum of 16 V, are obtained by increasing R4 by 200 kΩ/V. Resistor R1 should also be increased in value, as long as the current through D1 does not drop below 5 µA.

In this type of circuit, great care should be taken to avoid parasitic capacitances resulting from long connections. These would cause a deterioration of the regulation.

The maximum output current depends mainly on the permissible dissipation in T2 and, therefore, to some extent on the difference between the input and output voltage.
**TELECOMMUNICATIONS**

**DUPLEX AUDIO LINK**

Duplex communication is, of course, not a new technique: it has been used, for instance, in telephone systems for many years. Those systems, however, make use of transformers to achieve duplex - the circuit presented here does it with the aid of electronics. The principle is fairly simple. Two senders impose signals U1 and U2 respectively on to the audio cable. The voltage across the cable is then (U1+U2)/2. The receivers at both sides of the cable deduct their side's sender signal from the cable signal: the result is the signal sent from the other end of the cable. This principle is the basis of the circuit shown. Note that a circuit like that is required at either end of the link.

Opamp A1 is connected as a buffer amplifier and serves as sender. The send signal is imposed on to the cable via R4. Terminating the cable by R4 results in the voltage across the cable being only half the voltage output of A1. This does not detract from the operation of the circuit, however. At the same time, R4 ensures that signals emanating from the other end of the link can not get to the output of A1; if they could, they would be short-circuited by the output.

The receiver is a differential amplifier consisting of opamps A2–A4. The quality of the differential amplifier depends largely on the resistors used in association with the opamps and 1% types are, therefore, essential.

The cable signal, (U1+U2)/2, is applied to one input of the differential amplifier and the (halved) output signal of A1 to the other. Since the differential amplifier has a gain of 6 dB, the received signal applied to K2 has the same level as the original signal.

In practice, the proposed duplex system is not perfect and it is for that reason that, for instance, remnants of the sent signal are detectable in the receiver. Fortunately, these can be removed with the aid of P1. Furthermore, the cable used will load the output slightly capacitively, which causes the compensation voltage at the wiper of P1 to be not wholly in phase with the sent signal.

This effect may be virtually removed with the aid of C1.

The circuit is calibrated by connecting the cable to it and to its twin circuit and injecting a sinusoidal signal at a frequency of 1 kHz and a level of 5 V rms to its input. The input bus of the other circuit must be short-circuited during the calibration. Adjust P2 for minimum signal at K2. Next, increase the frequency of the input signal to 10 kHz and adjust C5 for minimum signal at K2. Repeat the procedure with the other circuit.

The signal suppression at 1 kHz is of the order of 80 dB, while at 20 kHz it is around 60 dB. These are pretty good values for this kind of circuit.

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**HEAD / TAIL LIGHTS FOR MODEL RAILWAY**

The price of a model railway locomotive is directly proportional to its facilities and finish. There are quite a few inexpensive ones on the market of which the finish leaves a lot to be desired. As a rule, manufacturers start their economy with the lights. The circuit presented here enables a DC locomotive to be provided with direction-independent head and tail lights. Since it uses LEDs, a very long life is guaranteed.

The circuit is based on a number of parallel-connected LEDs in a bridge network. The FET at the centre of the bridge ensures a constant current as long as the supply voltage exceeds 4.5 V. The brightness of the lights will, therefore, be independent of the supply voltage.

The LEDs are connected in parallel to keep the minimum operating voltage as low as possible. To ensure good current distribution, the pairs of parallel-connected LEDs should be of the same type and colour.

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**ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT**
I/O-FRIENDLY KEYBOARD

Not all computers have a keyboard, yet it is often essential to have the use of one. Two circuits are presented here that enable a keyboard facility to be produced with the aid of only six or seven i/o lines.

Fig. 1 shows a circuit based on a 74HCT148 and a 74HCT138 that can serve 56 or 64 keys. The circuit in Fig. 2, based on a 74HCT147 and a 74HCT138, can address 72 keys via seven i/o lines. The choice between the two circuits depends on the number of available i/o lines and the wanted number of keys.

In either circuit, the key rows are selected by the bits on the A, B or C input of the HCT138. The combination of these bits determines which of the outputs Y0–Y7 goes low. As long as no key is depressed, the inputs of the HCT148 in Fig. 1 or the HCT147 in Fig. 2 are high. When a key is pressed, the inverted binary information at the output of the ICs show which key it is.

In Fig. 1, the 0 input of the HCT148 is not used, because the code associated with that input is the same as that generated when no key is pressed. The output, pin 14, of this IC is used to detect whether a key has been pressed. It goes low when a key has been pressed.

In Fig. 2, four 1s at the output indicate that no key has been pressed.

ENERGY CONTROL FOR BATTERY CHARGERS

In most automatic battery chargers, the power transformer remains connected to the mains even after the battery (or batteries) has been charged. In many cases, considerable energy savings can be achieved by disconnecting the transformer from the mains when the battery is fully charged. This circuit performs this function for 12 V car battery chargers.

The battery voltage is monitored by an adjustable window comparator around opamps A1 and A2, which are powered by a stabilized supply voltage of 8.2 V (R7-D1). The high and low switching thresholds, UH and UL, are set by presets P1 and P2 respectively. The reference voltage for the opamp is obtained from the divider (D1-R1) and is a function of the battery voltage. With the given values of R1 and R2, a voltage divide factor, D, is obtained

\[ D = \frac{R2}{R1 + R2} = 0.43 \]

Taking into account the series resistors connected to the presets and the use of an 8.2 V supply voltage, the span of P1 is

\[ 7.2 / D = 16.7 \text{ V (max)} \text{ to } 3.8 / D = 8.9 \text{ V (min)} \]
and that of P2 is 

\[ \frac{6.3}{D} = 14.5 \text{ V (max)} \text{ to } \frac{3.3}{D} = 7.7 \text{ V (min)} \]

In practice, it will be desirable to switch the charger off at a battery voltage of 14.0 V and on again when the voltage drops below 12.5 V, corresponding to a window of 1.5 V.

When the battery voltage is lower than \( U_l \), it is, of course, also lower than \( U_h \). This means that both T1 and T2 conduct, so that Relay is energized. Contact K2 switches on the mains to the battery charger, and contact K1 keeps the relay energized even when T2 is switched off and the battery voltage rises to a value between \( U_l \) and \( U_h \). When the battery voltage reaches \( U_h \), both T1 and T2 are turned off, so that the relay is de-energized. After a while, however, it will be switched on again because the battery voltage will have dropped a little owing to the drop across its internal resistance.

Assuming that the required switching levels are \( U_l = 12.5 \text{ V} \) and \( U_h = 14.0 \text{ V} \), the presets are adjusted as follows. Disconnect CI and set P1 for \( U_h \) (max) and P2 for \( U_l \) (min). Power the circuit from a regulated supply set to 12.5 V and adjust P2 until the relay is just energized. Then increase the supply voltage to 14.0 V and adjust P1 until the relay is just de-energized. Finally, connect CI again and connect the circuit to the battery terminals.

The relay must be rated at 12 V DC and 300 Ω. It must have two make or change-over contacts, of which at least one has a voltage rating higher than the mains voltage. One suggested type is the V23037-A2-A101 from Siemens.

The circuit draws 25 mA, which rises to 65 mA when the relay is energized.

(M.S. Dhingra)

### TTL Supply Monitor

The Type LTC1042 IC from Linear Technology is a window comparator that can function with very small currents. This is made possible by the use of sampling techniques that enable the disabling of certain parts of the IC in the non-active phase. This economical behaviour is not so important in the present circuit: the monitor draws a current of rather more than the minimum 100 μA.

The comparator is set with the aid of a bandgap reference diode, D1. The reference voltage of 2.5 V provided by this diode is connected direct to the window centre pin 2.

The width of the window is also fixed with the aid of the reference voltage. Since the circuit is intended to monitor a TTL supply (5 V), the width of the window is arranged at one tenth of this, which is a convenient sub-multiple of the reference voltage. Potential divider R4-R5 gives a voltage of 0.25 V at half-window-width pin 5. This arrangement causes the within window output to go high when the input voltage (Vin at pin 3) exceeds 2.5 V ±10%. The input voltage is held at exactly half the supply voltage by potential divider R1-R2.

Transistors T1 and T2 drive indicator LEDs. When D2 lights, the circuit and the supply voltage are all right. When D3 lights, the supply voltage is too high. If neither of the diodes lights, the supply voltage is too low or even absent.

If you want a LED to indicate that the supply voltage is too low, interchange the function of pins 2 and 3. The above window output then becomes a below window output. Note that the circuit then needs a separate power supply, otherwise the below window LED can not be actuated.

### Car Alarm

The car alarm accepts signals from a variety of sources, including special sensors and the standard switches in a car, such as the door and ignition contacts. The unit has a relay output that enables controlling an acoustic transponder (loudspeaker, buzzer), light, a radio transmitter, and others.

The alarm is remarkably simple to set, because all control functions are performed by a single switch. It is switched on after parking the car by closing S1 when the green LED lights to indicate that the driver has 13 seconds to leave the car. When this delay has lapsed, a yellow LED lights to indicate that the alarm is set. When any of the alarm sensors is actuated (which also happens when the rightful owner of the car opens a door), a red LED lights.

(M.S. Dhingra)
The relay will then be energized after 17 seconds unless switch S1 is opened in the mean time. Since the location of that switch is known only to the car owner, anyone entering the car illegally will set off the alarm.

Two timers Type NE555 are connected as monostable multivibrators. Transistor T1 and associated components form the triggering stage, which is sensitive to both positive and negative triggering signals supplied by the sensors or switches. MOSFETs T2 and T3 serve to provide the 'delayed disable' function and to drive the 'alert' LED respectively.

When the unit is powered, D12 lights, but the alarm remains off for the time defined by time constant R12-C7. Since the capacitor is initially discharged, T3 conducts and pulls the reset pins of both timers to ground until the voltage across R12 has dropped to about 3 V. That level enables both timers, and causes the 'alert' LED to light. When the power is switched off for a second and then on again, C7 is discharged rapidly via D9: this sequence resets the alarm to its initial state.

When enabled, the alarm may be actuated by a low level at input pins 11 or 12 (door switches) or a high level applied to inputs 13 or 14 (ignition system power). The number of inputs may be expanded as required by adding one diode for each sensor or switch. When only one of the inputs is used, it is still recommended to use a diode.

Any trigger pulse or change in the direct voltage at the inputs triggers timer IC1 via C3 and C5. The output pin, 3, of IC1 is driven high so that the 'actuated' LED (D11) lights. Diodes D6 and D8 form an OR function to decouple the outputs of the timers. Capacitor C6 is discharged in preparation of the triggering of IC2. After the delay set by R6-C5, the output of IC1 goes low, thereby triggering IC2 via R9 and C6. The output relay is energized and the transponder is powered by the car battery via the relay contact. The alarm sounds for about 65 seconds (defined by R9-C10). The unit may be reset at any time, however, by operating S1 for at least a second.

(R. Lalic)

**042**

**TWILIGHT SWITCH**

This inexpensive unit switches the light on at dusk and off again at dawn. The circuit has separate time bases for the on and off delays. The dotted line in the diagram divides the circuit into two halves: A, the light-dependent switch around gate N1; and B, the on-off time base around N2 and LED and relay driver T1-T2.

The voltage at junction R2-R9-C1 is inversely proportional to the light intensity measured by light-dependent resistor (LDR) R9. Schmitt trigger gate N1 toggles whenever this voltage reaches one of the input threshold levels. Since the difference between these levels is large compared with the voltage span produced by the potential divider, a variable feedback loop is provided to achieve an effective switching span of 300 to 400 mV. When the output of N1 is high, the voltage at junction R1-P1 is almost equal to the supply voltage. When the output is low, the voltage drops to the level required for the threshold difference at the input of N1.

The output of N1 drives two time base circuits: C2-R4-D1 for the 'on' state and C2-R5-D2 for the 'off' state. These networks switch the output of N2 on and off after the wanted delays. The lamp relay and an indication LED are driven by darlington stage T1-T2, which is controlled by the output of N2.

Capacitor C1 prevents HF signals picked up by the cable between the LDR and the switching unit causing spurious triggering. Because of the high output impedance of the Type 4093, the cable should be a screened type.

The remaining two gates in
the 4093 package may be used for duplicating the time base (part B) to obtain a switching sequence. The values shown for R5-C2 and R4-C2 form a good starting point for dimensioning the delays introduced in this additional time base. The lowest permissible value of R4 and R5 is 47 kΩ, their highest value depends mainly on the leakage current of C2.

During the testing of prototypes of this circuit it was noted that the switching threshold and hysteresis depend on the make of the 4093. Good results were obtained with 50s HCF4093BE; devices from other manufacturers may require the value of R3 to be slightly different. The inputs of the unused gates in the 4093 must be earthed.

The LED is mounted in a suitable waterproof housing, screened from direct light sources.

Preset P1 serves to adjust the light intensity level at which the circuit switches. To prevent too slow a response to changes in the setting, the timer should be aligned with S1 open.

The current drawn by the timer is mainly that drawn by the energized relay.

(R. Lalic)

043

PSYCHOLOGICAL CAR LOCK

The lock circuit is based on elementary psychology rather than on any recent development in electronics.

The lock consists of a 12-key membrane keypad and an associated visual indication circuit. The complete unit is installed in an out-of-the-way position below the dashboard in a car, where it is not too difficult for a potential car thief to spot.

When the keys on the keypad are pressed, the impression is given that the lock will enable the car ignition when all four LEDs light. Because of the special configuration of the lock circuit, this will never happen. Eventually, the would-be thief gets frustrated (we hope) and tries another car, not realizing that the lock circuit is simply not connected to the ignition circuit. Only the rightful owner knows that the car can be started after a special switch, marked, say, 'wiper' is operated. This switch, installed as an accessory on the dashboard, is connected in series with the positive supply line to the ignition coil.

The membrane board must be a type with a common line, not one with a matrix configuration. A suitable keyboard may, of course, also be made from individual keys with numbered caps. The lines marked KO-K9 in the diagram must go to the associated key number on the keypad. Keys * and # are non-connected dummys. The four LEDs are fitted in a row near the keypad, and give an indication that suggests that the combination entered is correct.

As already stated, the circuit makes it impossible for all four LEDs to light simultaneously. This is because pin 3 of N1 is logic high when D1 is on. This means that T3 is off, so that D4 can not light in spite of all other key combinations. It also means that D4 can not light unless D1 is out.

Eight XOR gates in two 4070 packages determine which LEDs are on for a particular code entered via the keyboard. RC networks connected to each keyboard line keep this active for about four seconds after the key has been released. When, for instance, key 1 is pressed briefly, the voltage on line K1 rises to the supply level. Since pin 13 of gate N8 is then the only logic high input of all XOR gates, pin 4 of N2 goes high. This causes T2 to conduct and D2 to light. After about 4 s, the voltage on C1 has dropped to a level that N8 recognizes as logic low, and D2 goes out. Within the 4 s period it will, however, go out the instant key 5 or 7 is pressed. Diode D3 then lights, followed by D2, which remains on for a short period. The two LEDs then go out simultaneously. Note that this functional description applies to only one of many possible combinations.

Apart from its use in cars, the circuit may also find application in games.

The current drawn by the circuit is mainly that drawn by the LEDs that are on during the 4 s interval. In stand-by mode, less than 1 mA is drawn.

(C. Sanjay)
This low-cost circuit can drive two stepper motors and a relay by digital control data supplied from a computer. It can also detect the position of two microswitches and supplies logic levels back to the computer as positional information. This combination makes the interface ideal for use as an X-Y plotter or for building a buggy-style robot.

Circuit IC1 is configured as two set-reset (S-R) latches to provide contact debouncing for the two microswitches, S1 and S2, whose position is detected by computer reading port bits D5 and D6. The relay drive circuit around T1 may be used to switch a solenoid-operated pen on and off under the control of port bit 4.

Motor drive is provided by XOR gates N5-N12, bistables FF1-FF4 and integrated motor coil driver IC3. This combination can drive two unipolar 4-phase stepper motors in half-step mode by setting the direction of rotation with the aid of bits D1 and D3, and a high-to-low logic pulse transition on bit D0 or D2. The control functions are summarized in the table. The motor driver, IC4, is capable of sourcing 500 mA per phase at a maximum motor voltage of 50 V. The ULN2803A has internal diodes that protect it against reverse EMF generated by the motor coils when they are deactuated.

To drive the interface from a computer, set the I/O port for five output bits, DO-D4, and two input bits, D5 and D6, and send the appropriate control signals to the circuit.

The interface requires two supply voltages: 5 V for the logic circuits and the relay driver, and 12 V (+Ve) for the stepper motor coils. Motor 1 is L1-L4 and motor 2 is L5-L8.

<table>
<thead>
<tr>
<th>Step pulse</th>
<th>Motor 1</th>
<th>Motor 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1 0 1 0</td>
<td>L5 L6 L7 L8</td>
</tr>
<tr>
<td>1</td>
<td>1 1 1 0</td>
<td>1 0 1 0</td>
</tr>
<tr>
<td>2</td>
<td>0 1 1 0</td>
<td>1 1 0 1</td>
</tr>
<tr>
<td>3</td>
<td>0 0 1 0</td>
<td>1 0 1 0</td>
</tr>
</tbody>
</table>

Applications of this little circuit include a portable parking meter timer and egg timer. The 14-stage binary ripple counter Type 4060, IC1, has an on-chip oscillator capable of stable operation over a relatively wide frequency range. In the present circuit, the oscillator frequency is determined by an external RC network connected to pins 9, 10 and 11.

When the circuit is switched on with S1, the pulse at junction R4-C2 resets the counter and counting starts. When the count reaches bit 14 (Q13), pin 3 goes high so that the self-oscillating piezo-electric buzzer, a 12 V type, is turned on via driver T1.

The time delay is set with the aid of P1. Time delays of between one minute and two hours are possible by appropriate dimensioning of the timing components:

- 1–30 minutes: C1 = 220 nF; P1 = 500 kΩ
- 1–60 minutes: C1 = 470 nF; P1 = 500 kΩ
- 1–120 minutes: C1 = 470 nF; P1 = 1 MΩ
The timer is powered by a 9 V PP battery. Light-emitting diode D1 does not affect the operation of the circuit and is included merely to show that the timer works. Diode D1 and resistor R3 are, therefore, optional components. A mercury tilt switch may be used for S1 if the unit is to be used as a kitchen timer. The timer is then started by inverting it like a sand-glass.

With the buzzer actuated, the timer draws a current of about 10 mA.

(R.G. Evans)

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**HCMOS SQUARE WAVE GENERATOR**

This pulse generator supplies three rectangular output signals with a duty factor of 0.5. The signals have a fixed phase relation to one another: output 3 is the reference; output 2 has a phase shift of about 180°; and output 1 has a phase shift of about 10°.

The generator is formed by the four bi-directional HCMOS electronic switches contained in a Type 74HC4066 IC. Its operation is based on the fairly accurately defined switching threshold of an HCMOS input. The toggle point for low and high levels lies around Ub/2 to ensure a duty factor of 0.5 and thus a square wave output signal.

When the supply is switched on, C1 is charged via R3 and the on-resistance of ES3. When the voltage of C1 reaches Ub/2, ES4 closes and pulls the control input of ES3 low, causing C1 to discharge via ES2 and R3. When the 'low' switching threshold is reached at the control input of ES4, the generator starts to oscillate.

Considering the oscillator is basically an RC type, its stability is pretty good.

The output frequency of the generator is a function of the control voltage, Uc, as shown by the accompanying characteristic.

---

**HIGH-POWER ZENER DIODE**

Although its regulation characteristics are not as good as those of an integrated voltage regulator, a high-power zener diode has, nevertheless, useful applications, for instance, in shunt regulators.

The circuit shown in the diagram simulates a fairly expensive and hard-to-come-by high-power zener diode. Basically a two-stage current amplifier with a low-power (400 mW) zener reference, it is capable of sinking up to 500 mA at a maximum 25 V. The effective zener voltage of the circuit is about 0.7 V higher than that of the reference device, D1.

Using a preset, P1, instead of a fixed resistor, R1, enables the output voltage defined by D1 to be increased slightly. This substitution is useful when, say, a 6V2 or 9V1 zener is to be simulated and a lower-power type of this rating is not available. The next lower value in the E6 range (here, 5V6 or 8V2) may then be used for D1.

When the voltage across the circuit increases beyond the zener voltage, T1 conducts and provides a base current to T2. This power transistor then passes virtually all the current the 'zener' is capable of sinking. The dimensions of the heat sink for T2 depend on the maximum expected dissipation.

Transistor T1 should be a PNP type with a fairly high current gain, for instance a BC557B or BC559C, while T2 may be almost any medium or high-power NPN type, for example, a BD135 through BD139, BD241, TIP31 or 2N3055.

(C. Sanjay)
The voltage-controlled oscillator (vco) presented here is based on a Type OP80 operational amplifier. This opamp has an exceptionally low bias current of, typically, about 200 fA (femto-ampere = 10^-15 amp) and 2 pA maximum, so that any offset caused by this current is minute. It is, therefore, ideal for use as an integrator, because the operation of that kind of circuit is affected readily by offsets.

The OP80-based integrator in the diagram is used as a vco that is not affected by the polarity of the control voltage. A direct voltage at the input of the circuit will cause C1 to charge. Depending on the polarity of the input voltage, the potential across C1, and thus the output voltage of IC1, will be positive or negative. The speed with which C1 charges depends on the magnitude of the input voltage: this characteristic is used to generate a signal at a voltage-dependent frequency To that end, the output signal of IC1 is applied to a window comparator that has a switching threshold for both the positive and the negative maximum signal. These maxima are set to ±100 mV by P2. In some cases, it may be advantageous for symmetry to split R2 or R3 into a fixed resistor in series with a preset potentiometer.

When one of the comparators toggles, T1 is switched on via N1-N4 so that C1 discharges. This results in a neat sawtooth signal at the output of the circuit, whose frequency depends on the input voltage. Gate N4 ensures that the Fet reacts to both comparators. The other three gates delay the switching signal slightly to ensure that the Fet is switched on long enough to allow C1 to discharge completely.

The Type BSV81 MOSFET is provided with a separate substrate connection that must be linked to the source. Since the substrate is already connected internally to the housing, the device is very sensitive to random radiation, so that the oscillator is best fitted in a small metal enclosure.

If a BSV81 is not obtainable, another MOSFET with very low RSon and very small Cm may be used. If that also is not possible, a junction FET may be tried, but in that case a diode must be connected in series with the gate and a resistor of about 10 kΩ between the gate and the negative supply line. It is important to ensure that the pinch-off voltage level is reached readily. It may well be necessary to experiment with the value of C1.

Correct dimensioning of R1 will enable the relation between input voltage and frequency to be set at, say, 1 Hz/mV. With the input short-circuited, adjust P1 for the lowest possible frequency of the output signal (ideally, f = 0). The maximum input voltage is determined by the peak output current of IC2 (15 mA) and amounts to $15 \times 10^4 \times R1$.

The output signal of the vco is a clean sawtooth signal at a frequency of up to 10 kHz, although higher frequencies are possible. The frequency as a function of the input current is given by:

$$f = \frac{I_{in}}{(U_{top} \times C1)} \quad [\text{Hz}]$$

With values as shown in the diagram,

$$f = \frac{I_{in}}{(3.9 \times 10^{-9})} \quad [\text{Hz}]$$

Finally, note that the supply voltage to the OP80 must under no circumstances exceed ±8 V. The circuit draws a current of typically 4 mA.

---

**MONITORING TEMPERATURE WITH THE C64**

Maplin's module Type FE33L provides an inexpensive and convenient means of monitoring temperature. The module has a built-in A-D converter and an LC display and works from a single 1.5 V battery. Since it is often impractical to take frequent readings manually, the module provides a serial data output that can be used with most microprocessor systems. The combination of hardware and software given in this article enables a C64 computer to use the serial data, within BASIC, via the USR function.

The hardware consists of nothing more than a simple TTL level driver and may be mounted on a small piece of prototyping board. This may be connected to the module by three short wires, while the outputs go to a two-by-twelve 0.156" pitch edge connector for the C64's user port. Pins 5 and 16 of the module should be short-circuited to obtain the maximum sampling rate of one per second. Check all connections before switching the computer on.

The listing provided loads a machine-code program into the small section of RAM above the BASIC ROM at location 49152 ($C000). Note that some lines are very similar to others thus assisting entry. Once this has been RUN (without errors), and SYS 49152 has been entered, the temperature is obtained as follows:

```
TEMP = USR(0) : PRINT TEMP
```

This line can be incorporated into any BASIC program.
The first part of the machine-code program sets the USR vector and all the port B lines to input, while the remaining code is called by the USR function itself. When called, the program waits for the primary clock pulse, which is longer than the others, and then reads in each subsequent bit from the data line. These bits are converted from BCD format into a single floating-point that is returned by the USR function. The software will behave correctly only when the module is in the default °C mode, but this is not a restriction as readings can be converted readily to another scale. If the device is to be used for serious control applications, it must be borne in mind that the software will wait patiently for the primary clock signal to arrive from the module. If the clock signal fails for any reason (for instance, a break in the cable), the control program will be left hanging in an endless loop. It is, therefore, recommended to use a non-maskable interrupt (NMI) generated by the timers on CIA #2 to interrupt the program after a specific duration (for example, greater than the expected sampling time) and return some sort of error condition. For simple applications, this is not necessary and no further programming is required.

(J. Pelan)

### SIMPLE VARIABLE POWER SUPPLY

This low-cost power supply has an output voltage range of 1.5-15 V at a maximum current of 500 mA. Regulation is better than 2% for output currents not exceeding 350 mA. Voltage adjustment is effected by a potentiometer and an acoustic overload indication is provided.

Transistor T4 compares the voltage at the wiper of P1 with the output voltage. When this is 0.65 V higher than the set voltage, T2 is switched on, which removes the base current from darlington power stage T3 - T5. In this manner, the output voltage of the supply is 0.65 V higher than the reference potential at the base of T4, which is derived from a 15 V zener diode, D5.

The voltage at the 18 V, 1 A winding of the external mains transformer is rectified by bridge B1 and smoothed by C1. A simple acoustic overload alarm (B21) is actuated when the output current exceeds around 500 mA. Note that the exact level of actuation depends on the electrical specification of the buzzer, which should be a 24 V, self-oscillating type.

The power supply is, in principle, not short-circuit proof, although the use of a generous heat sink for T5 will enable that transistor to withstand the maximum dissipation of about 20 W for the few seconds that lapse before the supply is switched off.
To obviate radiation, the supply must be fitted in a metal enclosure. Interconnections should be kept as short as possible.

Capacitors C2 and C3 should be tantalum types.

(P. Sicherman)

**HEATING TIMER**

This timer has temperature and time settings. The temperature range is about 150 °C and the time delay is around 25 minutes.

The temperature controller, IC2, is driven by sensor IC1, the familiar Type LM35, which produces an output voltage of 10 mV/°C. This voltage is compared with a reference potential provided by a high-stability, temperature-compensated zener diode, D1. Presets P1 and P2 form the fine and coarse controls respectively for setting the temperature. The comparator switches on T2 whenever the temperature measured by IC1 is below the set value. This causes the relay, Rel, to be energized so that the heater element is powered via the relay contacts.

The timer function is based on oscillator/divider IC3, whose clock frequency is determined by the variable RC network between the P1 and P0 pins (9 and 11). The clock signal, divided by 2 and 21, appears at pins 15 and 3 respectively. Toggle switch S1 selects either of these outputs to give time delays of 6 s to 1.5 min. and 1.5–25 min. These settings are marked A and B respectively.

When the delay set by P3-P4 has lapsed, the oscillator in IC3 is disabled by the high level at the pole of the time selector switch. At the same time, T1 is switched on, T2 is switched off, and the (active) buzzer sounds to indicate that the set time has lapsed. The relay is de-energized via T2 and the heater is disconnected from its supply. The timer may be reset while the heater is on by pressing S2.

Some accurate calibration is required in the temperature controller. Connect a digital voltmeter between earth and junction R3-P1 and adjust P2 to obtain a voltmeter reading of 100 mV (=10 °C). Set P1 as appropriate by actually measuring the temperatures at which the relay is energized. Next, set P3 to minimum resistance and S1 to position A and adjust P4 to obtain a time delay of 5-6 seconds after S2 has been pressed. The time delays are set by P3 with the aid of an accurate watch. This procedure is not required for position B, since in this delays 16 times as long as in position A are provided automatically. If a simple thermostat only is required, the timer circuit, T1, T3 and T4 may be omitted.

The circuit is powered by a regulated 5 V supply and draws a current of about 30 mA with the relay inoperative.

The coil resistance of the relay must be not less than 400 Ω.

The temperature sensor must be fitted, of course, at some distance from the heating element.

(C. Sanjay)

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**IMPROVED LOW FUEL INDICATOR**

The indicator described here obviates the flickering of the 'low fuel' warning light on the dashboard caused by vehicle movement. The indicator ensures that the light remains extinguished until the duty factor of the signal supplied by the fuel sensor is smaller than 0.5. When that happens, the light comes on and remains on until the fuel tank has been filled to a sufficient level. The present circuit tests the light by causing it to flash for about 5 seconds every time the ignition key is turned to start the engine.

The signal processor is switched on together with the ignition. Initially, C1 is discharged, enabling oscillator N2 via inverter N1. One input of N3, pin 12, is connected to R3-C3, the time constant of which is equal to that of R1-C1. If the output of the fuel sensor is high, pin 12 of N3 is held high via R2-C3. The 1.5 Hz signal from oscillator N2 is inverted by N3 and passed to darlington
lamp driver T1-T2. After the delay introduced by R1-C1 has lapsed, gate N1 disables oscillator N2 so that the warning light goes out.

When the fuel sensor inside the fuel tank supplies pulses owing to vehicle movement, C3 is charged via R3 and discharged via R2-R3. When the duty factor (on/off ratio) of the level sensor signal drops below 0.5, the potential across C3 becomes high enough to enable the lamp driver, so that the 'fuel low' warning light is on permanently without any flicker.

(R. Lalic)

A VOX is a voice-operated switch that is often used as a substitute for the press-to-talk switch on a microphone. The one described here can be connected to almost any audio equipment that has a socket for an external loudspeaker. The actuation threshold is set by the volume control on the AF amplifier that drives the VOX.

The (loudspeaker) signal across R2 is capacitively fed to the base of T1. Resistor R3 limits the base current of this transistor when the input voltage exceeds 600 mV. Diode D1 blocks the positive excursions of the input signal, so that $U_{EB}$ cannot become more negative than about 0.6 V.

The output relay is driven by darlington T2. Resistor R4 keeps the relay disabled when T1 is off. The value of bipolar capacitor C2 allows it to serve as a ripple filter in conjunction with T2. Resistor R5 limits the base current of T2 to a safe level.

The switching threshold of the VOX is about 600 mV across R2. The maximum input voltage is determined by the maximum permissible dissipation of R2 and R3. As a general rule, the input voltage should not exceed 40 V (p-p).

The current drawn by the VOX is mainly the sum of the currents through the relay coil and through R5. The resistor may carry up to 100 mA when the VOX is overdriven.

(S.G. Dimitriou and F.P. Maggana)

The microphone preamplifier described is based on the SSM2015 from Precision Monolithics Inc. (PMI), which offers a very high gain and very low noise (1.3 nV/√f). It is intended for use with balanced input signals and is capable of providing an amplification of 10-2,000, depending on the value of R4. With $R_5 = R_6 = 10 \, k\Omega$, the amplification, $A$, is calculated from

$$A = \frac{20,000}{R_4} \times 3.5$$

With values as shown in the diagram, the amplification is, therefore, about 1,000.

Resistor R3 sets the operating point of the differential input amplifier and thus determines the bandwidth and the slew rate. A value of 33 kΩ gives near-optimum values of these characteristics, but results in a relatively high input bias current of 4.5 µA (with $R_3 = 150 \, k\Omega$, the current is only 1 µA). Moreover, the input noise
level, particularly the current-noise contribution, has increased a little. Nevertheless, the preamplifier has a signal-to-noise ratio of 95 dB, measured with short-circuited + and - inputs and an output level of 0 dBV. Resistor R3 enables the source impedance to be matched to the input of the differential amplifier: if \( Z = 600 \Omega \), R3 has an optimum value of 33 k\( \Omega \). With a 600 \( \Omega \) resistor across the input terminals, the signal-to-noise ratio is of the order of 86 dB.

The tables and the curve in Fig. 1 give a number of values for the bias resistor and compensation capacitors C2 and C3.

The differential inputs of the SSM2015 are of the floating type, so that external resistors, R1 and R2, are required to give a suitable DC setting. In single-ended (unbalanced) applications, care must be taken to prevent off-sets arising from biasing differences between the inputs, owing to different impedances (ground for one input, and the source for the other). Resistors R1 and R2 cause common-mode noise and must not be given a higher value than shown in the diagram.

Off-set compensation with the aid of preset P1 is required for the given value of R3, which results in the available amplification of 1,000. The value of R3 depends on the gain setting – see Table 2. Capacitor C4 compensates the on-chip input current regulator, and C1 suppresses HF signals.

Distortion of the preamplifier at 1 kHz and 0 dBV was measured at less than 0.006% and less than 0.01% at a test frequency of 10 kHz. The half-power bandwidth is 180 kHz at 3 V across 1 k\( \Omega \). Common-mode rejection at 50 Hz is greater than 100 dB.

Pam states that the output of the SSM2015 is not intended to drive long lines: capacitive loads greater than 150 pF should be decoupled by a 100 \( \Omega \) resistor in series with the output (but note that R5 must remain connected to pin 3).

(Precision Monolithics Inc.)

Table 1

<table>
<thead>
<tr>
<th>R3</th>
<th>C3</th>
<th>C2</th>
</tr>
</thead>
<tbody>
<tr>
<td>27-47 k( \Omega )</td>
<td>15 pF</td>
<td>15 pF</td>
</tr>
<tr>
<td>47-68 k( \Omega )</td>
<td>30 pF</td>
<td>5 pF</td>
</tr>
<tr>
<td>68-150 k( \Omega )</td>
<td>100 pF</td>
<td>250 k( \Omega )</td>
</tr>
</tbody>
</table>

Table 2

<table>
<thead>
<tr>
<th>R3</th>
<th>27-47 k( \Omega )</th>
<th>47-68 k( \Omega )</th>
<th>68-150 k( \Omega )</th>
</tr>
</thead>
<tbody>
<tr>
<td>A = 10</td>
<td>P1 = 500 k( \Omega )</td>
<td>250 k( \Omega )</td>
<td>250 k( \Omega )</td>
</tr>
<tr>
<td>A = 100</td>
<td>P1 = 500 k( \Omega )</td>
<td>100 k( \Omega )</td>
<td>100 k( \Omega )</td>
</tr>
<tr>
<td>A = 1000</td>
<td>P1 = 250 k( \Omega )</td>
<td>100 k( \Omega )</td>
<td>50 k( \Omega )</td>
</tr>
</tbody>
</table>

055

**HIGH-VOLUME ALARM**

When this alarm is actuated by a low-level signal at input EN, the (HF) loudspeaker produces a number of 4-tone sequences separated by quiet intervals. Each sequence sounds louder than the previous one to give the alarm a very distinctive character. The peak output is reached after about 28 seconds.

As long as the EN input is logic high, counters IC1a and IC1b remain reset and interval oscillator N2 and tone generator N3 are disabled. The alarm is then off.

When EN is actuated, the oscillator and the two counters are enabled. Counter IC1a is clocked with pulses (PRF = 8 Hz) from N2. Gates N1 and N4 at counter outputs Q0 and Q3 cause T1 to be turned off during eight consecutive clock cycles from IC1a.

---

**Table 2**

<table>
<thead>
<tr>
<th>R3</th>
<th>27-47 kΩ</th>
<th>47-68 kΩ</th>
<th>68-150 kΩ</th>
</tr>
</thead>
<tbody>
<tr>
<td>A = 10</td>
<td>P1 = 500 kΩ</td>
<td>250 kΩ</td>
<td>250 kΩ</td>
</tr>
<tr>
<td>A = 100</td>
<td>P1 = 500 kΩ</td>
<td>100 kΩ</td>
<td>100 kΩ</td>
</tr>
<tr>
<td>A = 1000</td>
<td>P1 = 250 kΩ</td>
<td>100 kΩ</td>
<td>50 kΩ</td>
</tr>
</tbody>
</table>
During the next eight cycles, the transistor is alternately switched on and off as illustrated in the timing diagram. The loudspeaker sounds only when T1 conducts.

Since output Q3 of counter IC1a is connected to the CLK input of counter IC1b, the latter is incremented by the 16th negative pulse transition. In practice, this means that counter IC1b is clocked after each tone sequence. The most significant outputs of counter IC1b drive the 3-bit selection inputs of analogue multiplexer IC3. Since Q0 is not used, IC1b requires two clock pulses to enable the multiplexer to connect the next input, Xn, to the output, X. The seven resistors at the multiplexer inputs cause the base voltage of T3 to increase after each alternate tone sequence. As a result, the voltage across the loudspeaker increases, so that the alarm sounds louder.

Transistor T1 switches the successive beeps on and off. Resistor R16, capacitor C5 and transistor T2 prevent abrupt switching of the voltage across the loudspeaker when T1 is turned off, and ensure that the sound level reverts slowly to the level set by the resistor at the relevant multiplexer input.

After the alarm is actuated, the output volume rises sevenfold. Diodes D1, D2 and D3 cause counter IC1b to stop at state 1110, so that the multiplexer passes the full positive supply voltage at input X7 to volume control T3. The alarm then sounds continuously at maximum volume, requiring the power supply to provide a peak current of up to 1.25 A through the loudspeaker. The resultant sound is ear-piercing.

Presets P1 and P2 serve to set the interval repetition rate and the sound frequency respectively.

(C. Sanjay)

MAINS POWERED TIMER

This timer may be inserted in a power line to provide a controllable delay before a load is energized. It was developed to work in conjunction with a passive infrared movement detector as part of an intruder alarm.

The mains voltage is reduced by C3 and rectified to give about 30 V across C1. This potential charges C2 slowly via R4-P1. When UC2 reaches about 14 V, electronic switch T1-T2 actuates a solid-state relay (a Type S202DS from Sharp). When the mains voltage is removed, C2 discharges rapidly via D6 and R10. The delay extends from 15 s (P1 set to minimum resistance) to 5 min (P1 set to maximum resistance).

The solid-state relay needs cooling in accordance with the current drawn by the load: at up to 1 A no heat sink is re-
quired; at 1–3 A (max), a 5×5 cm heat sink is advisable.

During the building of the circuit, due consideration must be
given to safety since many parts will be at mains potential. For
instance, fitting the unit in an ABS or other man-made fibre enclo-
sure is a must. If a potentiometer is used for P1, its spindle should
be an insulated type. If a preset is used, it must not be accessible
through a hole in the enclosure.

Switch S1 is a DPST type that disconnects the circuit from the
mains. Nevertheless, the only way of working on the circuit in
safety is by taking the plug out of the mains socket and allowing
C3 sufficient time to discharge.

(D. Dwyer)

---

**SINGLE-CHIP MELODY GENERATOR**

This melody generator, based on a Type 4093 CMOS
Schmitt trigger, may be used in alarms, doorbells and
cars (audible reverse gear or lights on indicator).

Three of the four NAND gates in the 4093 are con-
nected in series by RC networks. Oscillation is effected
by feedback of the output signal of N4 to the input of
N2. The logic high levels produced by the cascaded
gates in the oscillator circuit are used for biasing one
of associated diodes D1, D2 and D3. The relevant
diode connects one of frequency-determining capaci-
tors C1–C3 to tone oscillator N1. The audio signal
available when S1 is pressed is applied to comple-
mentary transistor pair T1–T2 that drives the loud-
speaker.

The frequency of the emitted tone may be adjusted
to individual taste by preset P1.

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**INFRA-RED MICROPHONE**

The circuit of this microphone was originally designed to mon-
tor a 7-segment display used on the flight deck of a Boeing 737.
That display used filaments, so that an IR detector was an obvious
choice. In the Boeing, it was connected to a small portable
recorder: the sensor therefore acted as a microphone that reacted
to IR light instead of to sound. This idea of an IR microphone is
made more tangible by housing the device in the shell of the plug
as shown in the photograph.

The IR microphone consists of a Type BP104 photodiode con-
nected to the inputs of a DC-coupled operational amplifier, whose
gain is determined by R1.

The device may be used 'to listen' to the visual world around
us. It is particularly effective where sources of noise, such as
incandescent light bulbs, are switched off. A gas flame, such as
that of a cigarette lighter, is manifested as a soft breeze. A cozy fire
burning in the grate comes out as a real hurricane. This means
that the microphone may be used as an acoustic fire alarm, but
that is about the only application we can think of. However, the
circuit is intended more to give us an opportunity to see our envi-
ronment from a different angle. If the BP104 is replaced by a
BPW34, the sensitivity of the device is shifted from the infra-red
to the visible spectrum.

The current drawn by the circuit depends to some extent on
the supply voltage and should be 2–5 mA.
FOUR-QUADRANT DIMMER

This very special mains-operated dimmer for domestic or industrial lights is not available in proprietary form: it enables brightness control of two groups of lights in one operation. The possible combinations of brightness are shown in the table. It will be clear that it is not possible to obtain continuous control of brightness in the two groups.

The rotary switch selects the resistor in a given network and thus the brightness of the relevant group of lights. No resistor means that the group is off; a short-circuit gives maximum brightness, and resistors of 10 kΩ and 18 kΩ mean intermediate brightness. The diodes prevent the groups affecting one another.

The 64 µH choke, L1, and the 150 nF capacitor across the bridge rectifier prevent the dimmer causing interference in other equipment connected to the mains.

If the triacs are fitted on a heat sink rated at 12 K/W, up to 500 W per group may be controlled. It is, of course, essential that the enclosure in which the dimmer is fitted provides ample cooling: a fair number of slots or holes in it are, therefore, essential: these should not permit the circuit elements to be touched.

The switch should have a non-metallic spindle: this is not only safer than a metallic one, but it also enables the easy removal of the end-notch so that the switch may be rotated continuously instead of having to be returned to the first stop every time it is operated.

It is recommended that the mains on-off switch S2 is fitted with a built-in 'on' indicator bulb: this shows at a glance whether the circuit is on even though S1 may be in the off position.

Finally, do bear in mind that this circuit carries mains voltage in many places: good workmanship and insulation are, therefore, of the utmost importance.

(C.G. Mangold)

SENSOR SWITCH AND CLOCK

One Type TL084 IC and an old quartz watch enable the construction of a de luxe on-off switch. Two of the four opamps contained in the TL084 (A1 and A2) are used to amplify the input signals from the sensors hundredfold (with the component values as shown in the diagram). Just touching the sensors with a finger causes a good 50 Hz input signal (hum). Note that the amplification drops rapidly with rising frequency.

Diodes D5 and D6 rectify (single-phase) the 50 Hz signal. Since the diodes are connected in anti-phase, touching the 'off' sensor results in a positive potential across C10, whereas touching the 'on' sensor gives a negative potential across C10.

Opamp A4 is connected as an inverting bistable, so that a nega
The circuit described here is intended as a revolution control for small DC motors as fitted, for instance, in small electric drills (such as used in precision engineering and for drilling printed-circuit boards, among others). The behaviour of these motors, which are normally permanent magnet types, is comparable to that of independently powered motors. In theory, the RPM of these motors depends solely on the applied voltage. The motor adjusts its RPM until the counter EMF generated in its coils is equal to the applied voltage. There is, unfortunately, a drop across the internal resistance of the motor and this causes the RPM to drop in relation to the load. In other words, the larger the load, the larger the drop across the internal resistance and the lower the RPM.

The present circuit provides a kind of com-
compensation for the internal resistance of the motor: when the current drawn by the motor rises, the supply voltage is increased automatically to counter the fall in RPM.

The circuit is based on an enhanced voltage regulator consisting of IC1 and T1, which provides a reasonably large output current (even small drills draw 2-5 A). The 'onset' supply voltage, and thus the RPM, is set by P2. Because of emitter resistor R1, the currents through IC1 and T1 will be related to one another in the ratio determined by R1 and R2. Owing to this arrangement, the internal short-circuit protection of IC1 will also, indirectly, provide some protection to T1.

As soon as the current drawn exceeds a certain value, T2 will be switched on. This results in a base current for T3 so that R5 is in parallel (well, more or less) with R6. This arrangement automatically raises the output voltage to counter a threatened drop in RPM. The moment at which this action occurs is set by P1, so that the present circuit can be adapted pretty precisely to the motor used.

If only very small motors are likely to be used, the power supply (transformer and bridge rectifier) may be rated rather more conservatively. As a guide, the current in the transformer secondary should be about one and a half times the maximum DC output current.

(G.J. Lammertink)

**062**

**CALL TONE GENERATOR**

Amateur VHF relay stations are normally actuated by a 1750 Hz call tone. This may give problems when the relevant sending equipment has no internal call tone generator, or it has one whose frequency is not sufficiently accurate, or whose tone duration is not long enough to securely energize the relevant relay.

These problems can be overcome by the stand-alone generator described here. Simply placed in front of the microphone, it makes absolutely certain that the relay station is actuated.

The generator consists of a quartz oscillator, a frequency counter and a buffer-amplifier, all contained in just two CMOS ICs. It is powered by a 9 V PP3 battery, from which it draws a current of around 5 mA.

Gates N1 and N2 form an oscillator that is controlled by a 3.2768 MHz crystal and provides clock pulses to IC2 which is connected as a programmable scaler. Diodes D1–D5 determine the divide factor of 1872. Counter output Q1 thus provides the wanted 1750 Hz signal, which is buffered by N3–N6 before being applied to a piezo electric buzzer. Capacitor C3 suppresses any harmonics, while R4 determines the volume of the output signal.

(N. Körber)

**063**

**MAINS FAILURE INDICATOR**

When the mains voltage is present at the input terminals, the transistor in the optocoupler is on, T1 is off and silicon-controlled rectifier TH1 is in the conducting state. Since both terminals of the piezo electric buzzer are then at the same potential, the buzzer is inactive. If the mains voltage drops out, transistor T1 conducts and causes one of the terminals of the buzzer to be connected to earth; the thyristor remains in the conducting state. In this situation there is a large enough potential difference across both the buzzer and D5 to cause these elements indicating the mains failure both audibly and visibly.

When the mains is restored, the circuit returns to its original state. A touch on the reset button then interrupts the current through the SCR so that the thyristor goes into the blocking state, and the other terminal of the buzzer is connected to ground.

(ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT)
The unit is powered by a 9 V PP3 battery and draws a quiescent current of 1.7–2.5 mA. It is important that the enclosure is a well-insulated type.

Finally two points to note. If by accident the circuit to the optocoupler and R2 is broken, electrolytic capacitor C2 may be damaged since it will be charged well above its 25 V rating. Secondly, where a plug is used for the mains connection, it is advisable to solder a 1MΩ resistor across C1 so that this capacitor does not retain its charge after the plug is removed from the mains socket.

**Parts list**

<table>
<thead>
<tr>
<th>Resistors:</th>
<th>Semiconductors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1 = 15 kΩ; 2 W</td>
<td>D1–D4 = 1N4004</td>
</tr>
<tr>
<td>R2, R8 = 1 kΩ</td>
<td>D3 = LED</td>
</tr>
<tr>
<td>R3 = 4 kΩ</td>
<td>T1 = BC547B</td>
</tr>
<tr>
<td>R4, R7 = 10 kΩ</td>
<td>Th1 = BRX46</td>
</tr>
<tr>
<td>R5 = 5 kΩ</td>
<td>IC1 = OPY1264B</td>
</tr>
<tr>
<td>R6 = 100 kΩ</td>
<td>Miscellaneous:</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Capacitors:</th>
<th>Switch with 1 make contact</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1 = 220 nF; 630 V</td>
<td>BZ1 = piezo buzzer 9 V</td>
</tr>
<tr>
<td>C2 = 4µF; 25 V</td>
<td>9-V PP3 battery</td>
</tr>
</tbody>
</table>

---

**064**

**ELEKTOR ELECTRONICS JULY 1989 SUPPLEMENT**

**GUITAR COMPRESSOR**

The control of this compressor is based on the dependence of the dynamic resistance of a diode on the current flowing through it. The heart of the present circuit is the diode bridge D1–D4, which behaves as a variable resistance controlled by the current flowing in T1.

The input signal is applied to preamplifier stage A1 via low-pass filter R1–C1 that removes any HF noise from the input. Switch S1 in the feedback loop of A1 sets the amplification to 1 (position
A), 6 (C) or 11 (B). The amplified signal is applied to the diode bridge direct via R12 and C5, and inverted via inverter A2, capacitor C6 and resistor R13. The two signals are summed by the bridge, amplified (in A3) and then split again into two, one of which is inverted by A4. The positive half cycles of the two signals are used to switch on 12 and T3 respectively. Capacitor C11 is then charged via R12. When the potential across this capacitor reaches a certain level, T1 is also switched on, after which a control current flows through the bridge via R21. This current lowers the resistance of the bridge so that the signal is attenuated (compressed). At the same time, the LED lights to indicate that the signal is being compressed. Capacitor C12 prevents any DC voltage from reaching the output.

The output signal is taken from the wiper of P1. Low-pass section R20-C13 limits its bandwidth to 12 kHz.

Switch S2 enables the selection of various decay times of C11. The values shown in the diagram have in practice proved to be the most useful. Nevertheless, these values are subjective and may be altered to personal taste and requirements.

(W. Teder)

---

**LC SINE WAVE GENERATOR**

This compact LC oscillator offers a frequency range of about 1 kHz to almost 9 MHz and a low-distortion sine wave output.

The heart of the circuit is series-resonant circuit L1-C2-C3 in the feedback loop of amplifiers T1-T2. Transistor T2, which is connected as an emitter follower, serves as impedance converter, whereas T1, connected in a common base circuit, is a voltage amplifier whose amplification is determined by the impedance of L1 in its collector circuit and the emitter current. The feedback loop runs from the collector of T1 via the junction of capacitive divider C1-C2, source follower BS170 and the input impedance formed by R1 and C4. The whole is strongly reminiscent of a Colpitts circuit. The signal is also taken to the output terminal via C5.

Of particular interest is the amplitude control by the current source. The signal is rectified by two Schottky diodes, smoothed by C9 and then used to control the current through T3. The gain of amplifier T1 is therefore higher at low input levels than at higher ones. This arrangement ensures very low distortion, since the amplifier can not be overdriven.

The resonant frequency may be calculated from

\[ f = \frac{1}{2\pi\sqrt{LC}} \]

With values as shown, it extends from 863 Hz (L1 = 10 nH) to 8.630 MHz (L1 = 100 nH).

The unit may be used to measure the Q of inductors. To that end, a potentiometer is connected in parallel with L1 and adjusted so that the current through the amplifier is doubled. The Q is then calculated from

\[ Q = \frac{R_p}{2\pi f L} \]

---

**SHUNT FOR MULTIMETER**

The current range in multimeters, particularly the more inexpensive ones, is restricted by the load limits of the internal shunts to 1-2 A. The photo shows how easily a precision heavy-duty resistor from Dale or RCL (0.1 Ω; 20 W; 1%) may be used as an external shunt. These resistors were not designed for this purpose, but they are much cheaper than custom made shunt resistors. The 20 W rating applies only, by the way, if a heat sink is used: without that its rating is only 8 W.

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