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THE MAGAZINE FOR THE RADIO & ELECTRONICS EXPERIMENTALIST

July/August 1990

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Elektor Electronics

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Battery tester
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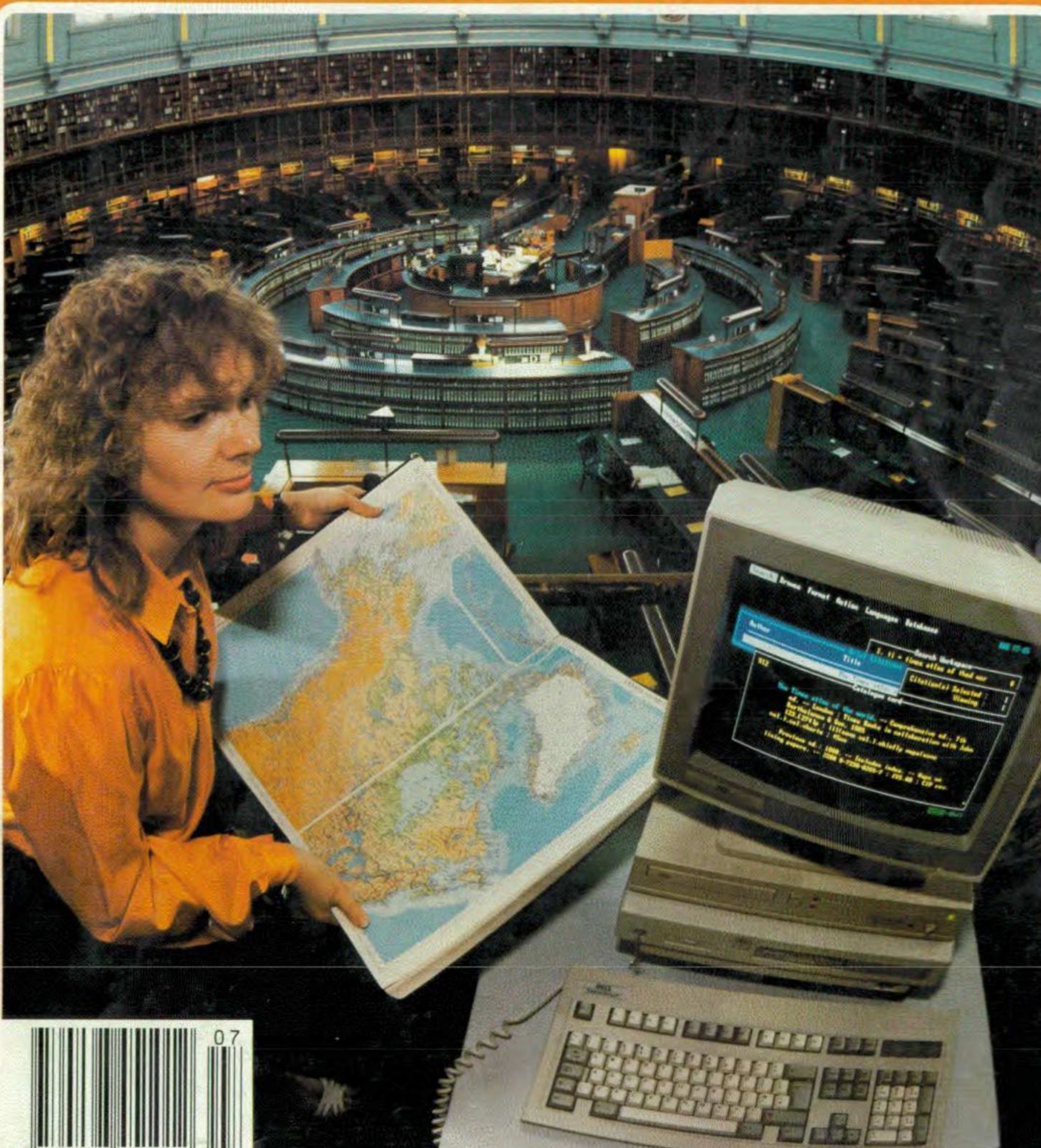
Square-wave generator
TTL-level 100 MHz quartz oscillator
SCART-plug FM mini sender
Versatile NiCd battery charger



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Elektor Electronics

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- High-current H_{fe} tester
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- Isolation amplifier
- Inter-IC communications
- LF-HF test probe
- AC motor control

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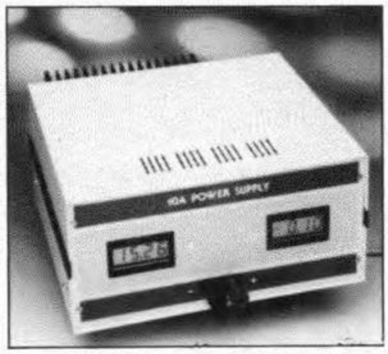

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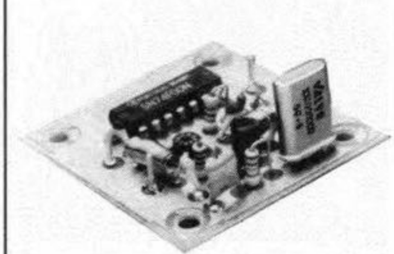

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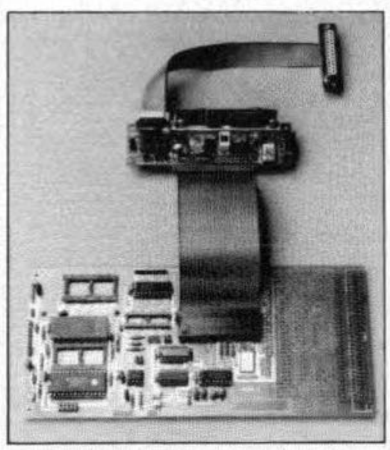

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001 ELECTRONIC DOORMAN

The electronic doorman enables a door to be opened automatically after a predefined delay from the moment a bell has been rung. It is intended for use in, for instance, waiting rooms and offices. The idea behind it is that the person who normally operates the door-opening control on his/her desk need not interrupt his/her work to open the door to visitors, patients or clients.

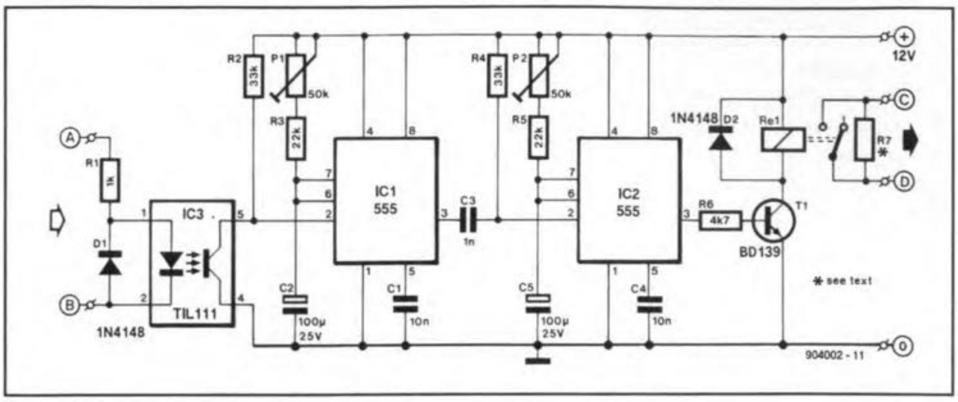
The voltage that serves to trigger the circuit is obtained by connecting inputs 'A' and 'B' in parallel with the bell, electric chime or buzzer. When a signal is detected, pin 2 of timer IC1 is pulled low by the phototransistor in optocoupler IC3. The time delay introduced by the timer may be set roughly between 3 s and 6 s with P1. After this delay, pin 3 of IC1 reverts to low. The trailing edge of this signal is converted to a short trigger pulse by C3-R4.

A second timer, IC2, is triggered and introduces a second delay of between 2 s and 6 s, set with P2. During this delay, the high output level at pin 3 causes the door opener to be actuated via driver T1 and relay Re1. The relay contacts, C and D, are connected in parallel with the existing door-opening switch.

Resistor R7 eliminates inductive voltage peaks when the doorman is switched off to prevent erroneous triggering of the timers. Its value must be determined empirically. Clearly, it should not be so low as to cause the door to be opened at the moment the relay contacts are connected to the switch on your desk.

The presets on the board allow the wait time and the 'door open' time to be set to individual requirements. In practice, a 'door open' time of about 4 s gives the best effect. ■

(*R. Dischler*)



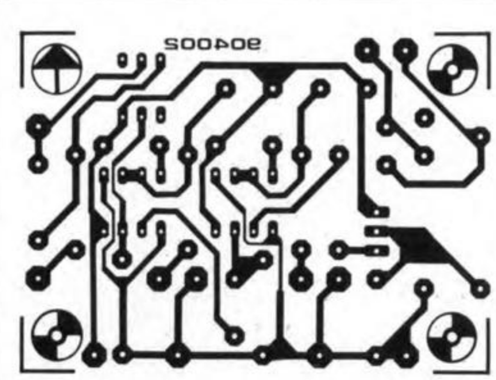
PARTS LIST

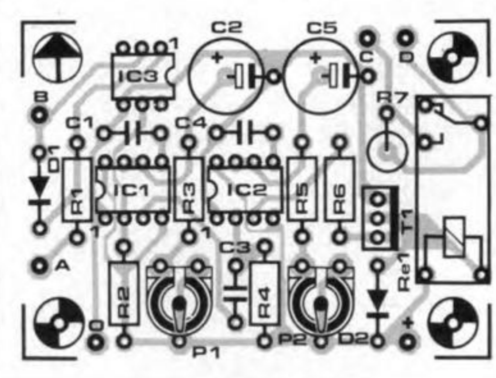
Resistors:
R1 = 1k0
R2, R4 = 33 k
R3, R5 = 22 k
R6 = 4k7
P1, P2 = 50 k preset

Capacitors:
C1, C4 = 10 n
C2, C5 = 100 µF, 25 V, radial
C3 = 1 n

Semiconductors:
IC1, IC2 = 555
IC3 = TIL11
D1, D2 = 1N4148
T1 = BD139

Miscellaneous:
Re1 = 12 V relay for PCB mounting (e.g. Siemens V23127-B00--A101)
PCB Type 904002





002 NUMERICAL PULSE-WIDTH CONTROL

The control enables the pulse-width of a clock signal to be set with thumb-wheel switches. The pulse widths are set in five ranges: 1–999 µs; 0.01–9.99 ms; 0.1–99.9 ms; 1–999 ms; and 0.01–9.99 s. These ranges overlap to some extent and this is done on purpose to make available settings like 5.46 ms or 45.8 ms. The circuit has an error detector that gives a visual indication if the set pulse-width exceeds the period of the input clock signal. The permissible timing error is ±0.1 µs in all ranges.

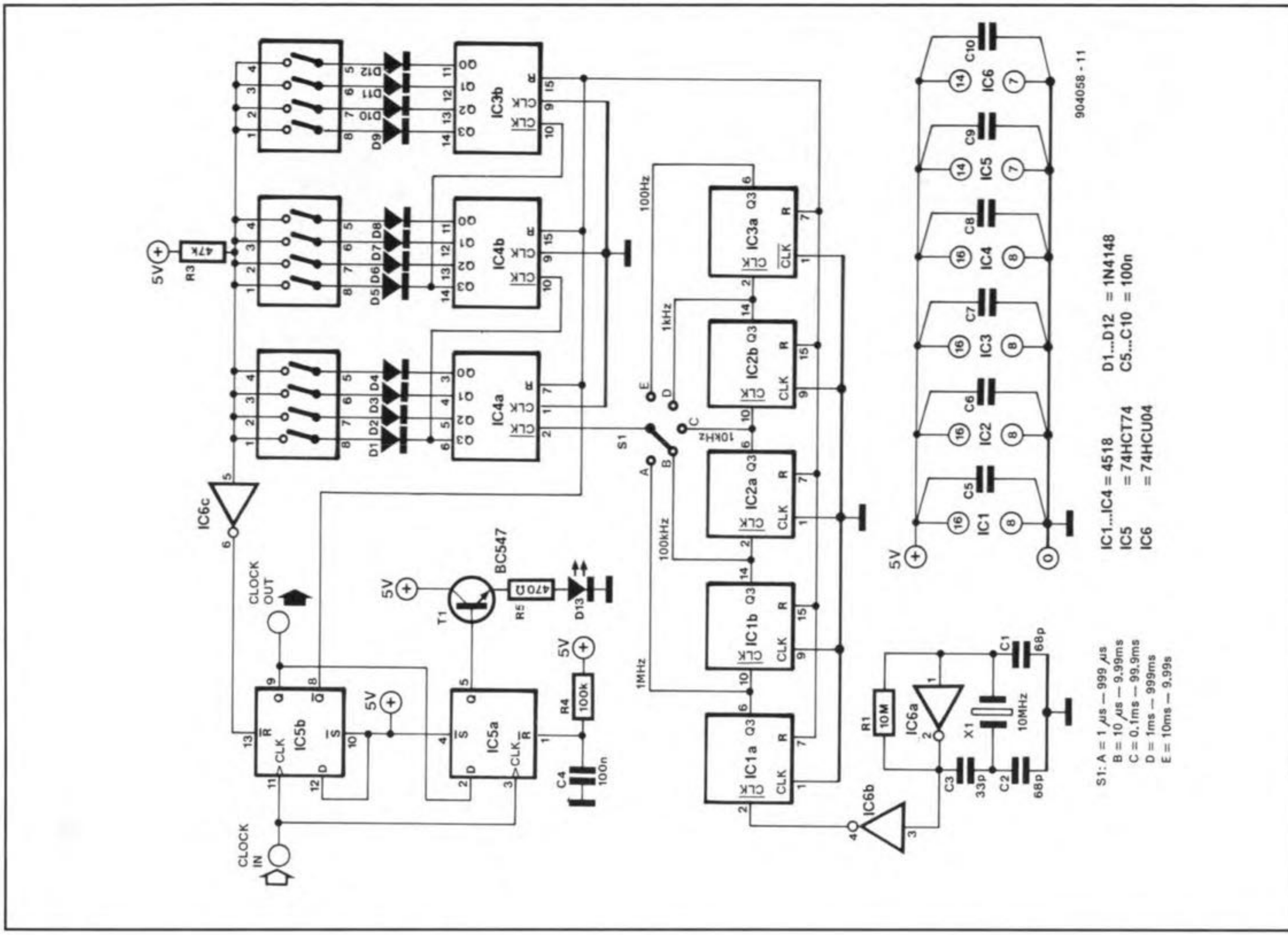
HCMOS inverter IC6a and quartz crystal X1 form a 10 MHz oscillator whose output signal is divided by IC1, IC2 and IC3a to 1 MHz, 0.1 MHz, ...100 Hz. The required range is selected by S1. The signal is used to clock IC4 which, together with IC3b, forms the pulse-width counter.

The outputs of the pulse-width counter are applied to diodes and thumb-wheel switches. The AND function so created causes the output of IC6c to go low only when the count state in IC4 and IC3b is equal to the number set with the thumb-wheel switches.

The circuit operates at the leading edge of the input clock signal applied to IC5a. When a leading edge occurs, the $\overline{Q}$ output of IC5a goes low, thereby enabling counters IC1–IC4. The Q output, which forms the output of the circuit, goes high.

When the set time is reached, the output of IC6c goes low. Bistable IC5a is immediately reset whereupon its $\overline{Q}$ output goes high and the counters are reset. Consequently, the output of IC6c reverts to high so that IC5a is enabled again and ready to be clocked by the next leading edge at its CLK input. Meanwhile, the Q output of IC5a goes low and this marks the end of the output pulse.

S 5



If a leading edge occurs while the Q output of IC5a is high, a logic 1 is clocked into IC5b whereupon D13, the ERROR indicator, lights. It goes out again as soon as the error condition is ended by a change of range or set value.

IC6 is an unbuffered type that must not be replaced by an HC or HCT equivalent, otherwise the reliable operation of the oscillator is not guaranteed.

It should be noted that pulse-width settings between 0.1 µs and 99.9 µs with an input signal of 10 MHz may not work in all cases because the AND function formed by diodes D1–D12 and the associated pull-up resistor, R3, may not be fast enough.

The output pulses from IC5b may have to be cleaned or reshaped to eliminate overshoot.

The circuit draws a current of not more than 10 mA from a 5-V supply. ■

(*C. Sanjay*)

PARTS LIST

Resistors:

R1 = 1k0
R2, R4 = 33 k
R3, R5 = 22 k
R6 = 4k7
P1, P2 = 50 k preset

Capacitors:

C1, C4 = 10 n
C2, C5 = 100 µF, 25 V, radial
C3 = 1 n

Semiconductors:

IC1, IC2 = 555
IC3 = TIL11
D1, D2 = 1N4148
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Miscellaneous:

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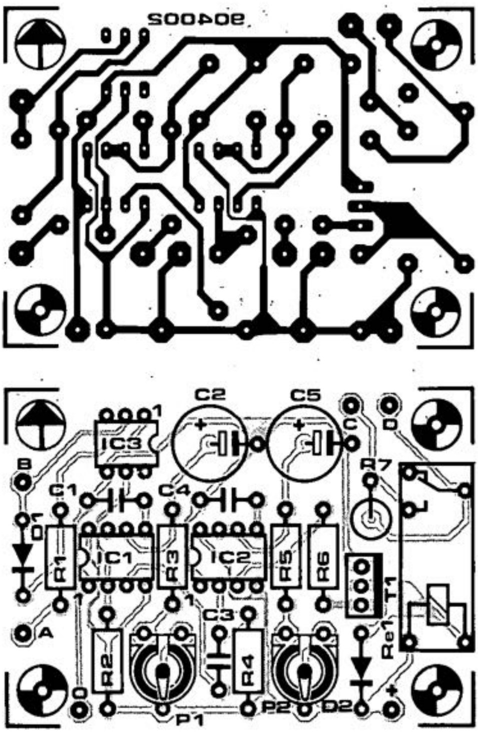


A standard voltage regulator Type 7805 is inexpensive and easily available, but its maximum current of 1 A can at times prove a handicap. However, this current may be increased by adding a power transistor (T3 in the circuit diagram below) on a heat sink. When the current drain is small, the 7805 continues to function as before. When the current rises above 15 mA, however, the potential drop across R4 is large enough to switch on T3. This transistor is protected against short circuits by T2. When the current through the MJ2955 rises above 3 A, the voltage drop across R3 is large enough to switch on T2. This limits the base-emitter voltage of T3, so that the output current can not increase much more.

In parallel with T2 is a transistor, T1,

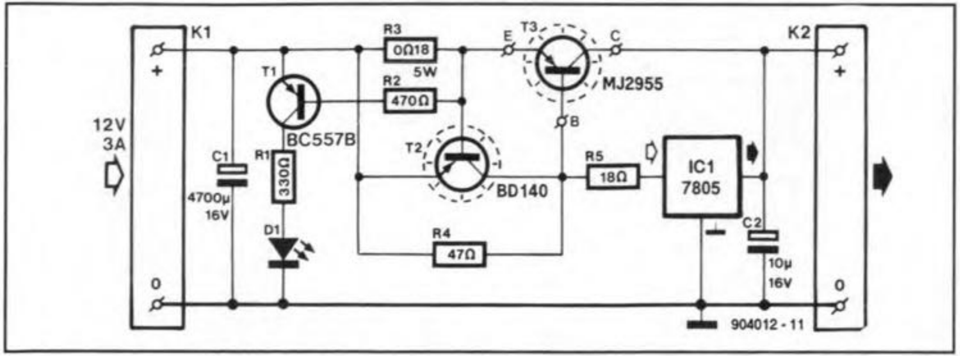


Fig. 1. Circuit diagram of the power supply.

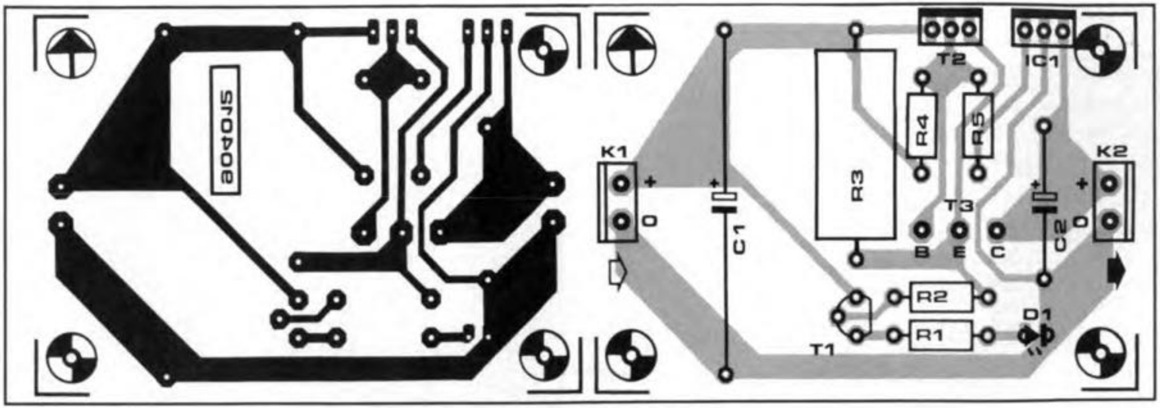


Fig. 2. The printed circuit board for the power supply.

that switches on an LED as soon as current limiting occurs. Resistor R5 has been added to limit the current through the regulator as soon as the current limiting circuit operates, since R4 is then short-circuited by T2: in the absence of R5, the full current would flow through the 7805.

Of course, there is a price to be paid for the higher output current: the input voltage must be 10 V for an output current of 3 A, instead of 8.5 V for currents up to 1 A.

The current limiting comes in relatively gradually: when the output is short-circuited, a current of up to 6 A may flow for a short period. Obviously, that situtation should not be allowed to last for long.

In the construction, take care that T2 and T3 are insulated from the heat sink. The 7805 does not really need a heat sink, but it does no harm to fit it also to the heat sink. If you follow the component layout in Figure 2 above, you should not experience any difficulties with the remainder of the construction. ■

(*K. Walters*)

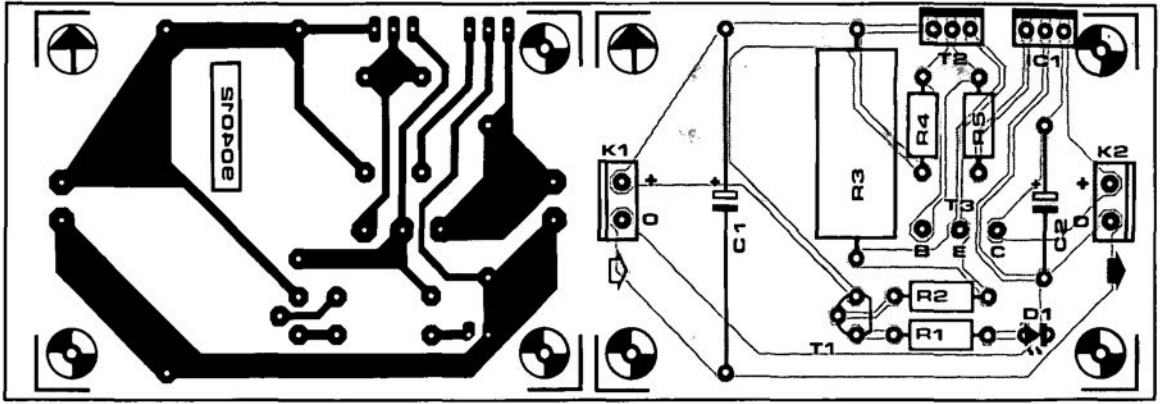
PARTS LIST

Resistors:
R1 = 330 Ω
R2 = 470 Ω
R3 = 0Ω18; 5 W
R4 = 47 Ω
R5 = 18 Ω

Capacitors:
C1 = 4700 μ; 16 V
C2 = 10 μ; 16 V

Semiconductors:
D1 = LED, red
T1 = BC557B
T2 = BD140
T3 = MJ2955
IC1 = 7805

Miscellaneous:
K1, K2 = 2-way PCB connector
heat sink 2–3 K/W



890405
K1
+
0
C1
R3
R4
R5
T2
IC1
T3
B
E
C
C2
K2
+
0
R2
R1
T1
D1

004 AF ISOLATING AMPLIFIER

An isolating amplifier, also called buffer, is used to match two dissimilar impedance points and isolate one stage from a succeeding one in a cascaded system, and thus prevent undesirable interaction between them.

The present isolating amplifier has a bandwidth of 40 Hz to 40 kHz and a distortion of not greater than 1% at a 1 kHz signal of 70 mV r.m.s. The current drawn by each section is not greater than 10 mA.

The amplifier is based on an opto-isolator that provides the separation between the two sections. The LED in the isolator, a Type CNY21 or IL10, is driven by opamp IC1, a Type LF356.

Because the feedback resistor, R2, follows the LED, a large portion of the distortion produced by the LED is suppressed by the opamp. The bias current for the LED is adjusted with P1. In the present circuit, the level of this current is chosen at 1 mA, since that gives a reasonable compromise between the overall power consumption and the non-linear distortion.

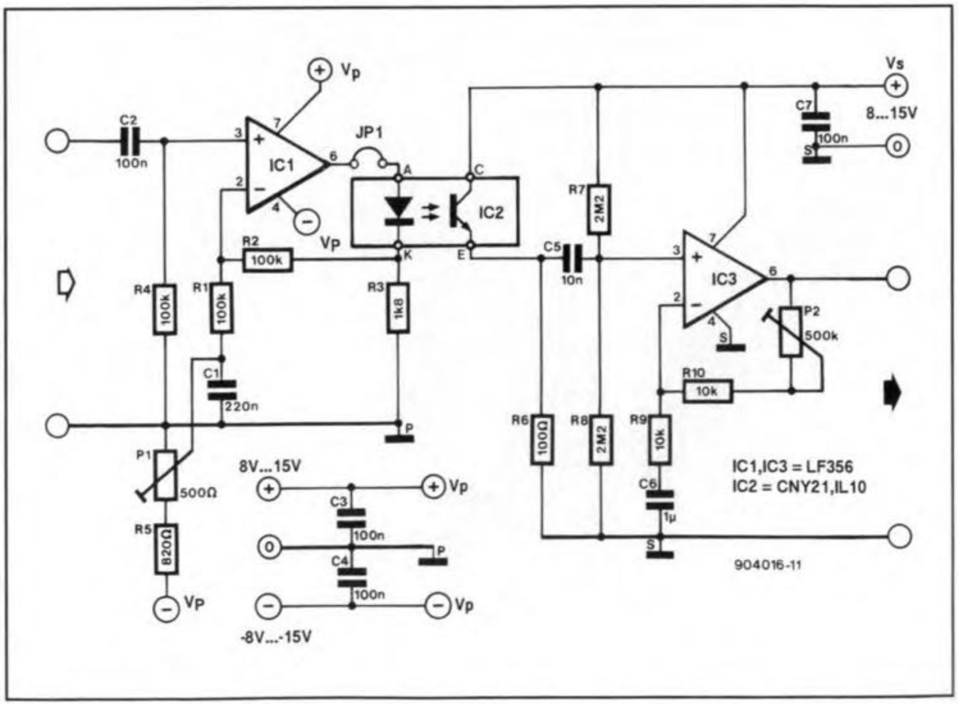
The bias-current setting is not the only factor that determines the total distortion: the alternating current through the LED also plays a part. This is the reason that the primary section of the amplifier has been designed to cause an a.c. through the LED whose level is about 10% of that of the bias current at a signal level of 70 mV r.m.s. (100 mV p-p). When this input level is exceeded, the distortion increases significantly, and it is, therefore, necessary, to limit the input level to the value stated.

The direct and alternating currents through the LED, I_L and i_L respectively, are calculated from:

$$I_L = U_{P1}\,(R2+R3)/R1R3$$

$$i_L = u_i\,(R1+R2+R3)/R1R3$$

where U_{P1} is the voltage at the wiper of P1 and u_i is the input voltage.



Adjustment of the primary section is effected by setting P1 to obtain a reading of 1 mA through a milliammeter connected in place of JP1.

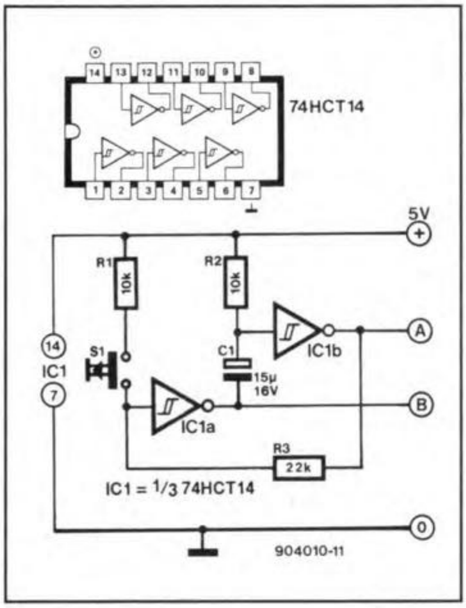
The signal received by the photo-transistor in the opto-isolator is amplified by a second LF356 whose gain is controlled by P2. After the LED current has been set, this preset may be adjusted to ensure unity gain of the entire amplifier.

The circuit needs two completely separate power supplies and this means two transformers or one transformer with two isolated secondary windings. The primary section needs a symmetrical supply, whereas a single 8–15 V supply will suffice for the secondary section. ■

(J. Ruffell)

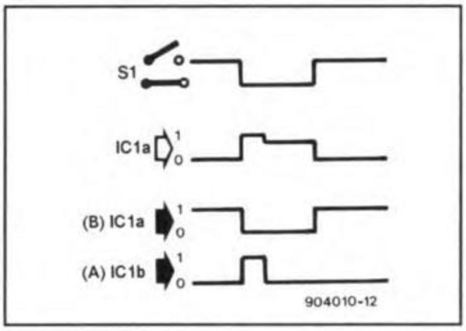
DEBOUNCING CIRCUIT WITH TWO OUTPUTS 005

Any switch or key in a digital circuit may cause problems because mechanical contacts bounce up and down a few times before they close. Normally, this weakness is negated by an RS bistable, but this article shows that it also may be achieved by a monostable.



The two gates in the circuit diagram form a monostable with a mono time of 100 ms (the bounce time of a key is typically 20 ms).

In quiescent operation, the input of inverter IC1b is at the level of the supply voltage, so that its output is low. This low level is connected to the input of IC1a via R3. The output of IC1a is thus high and C1 is not being charged.



When the switch, S1, is closed, the input of IC1a goes high because R1 has a smaller value than R3. The output of IC1a then becomes zero, which is immediately connected to the input of IC1b via C1. This low level remains at the input during a time determined by R2-C2. Any bounce of the switch during this time has no effect whatsoever, because the output of IC1b, and thus the input of IC1a, is high.

When a switch or key is released, it will be noticeable at output B but not at output A, because C1 needs time to discharge. Only after it has discharged, can the monostable be triggered again.

The gates should be CMOS types, preferably of the HC/HCT series. The circuit works best with Schmitt trigger inverters, although most run-of-the-mill inverters work perfectly well.

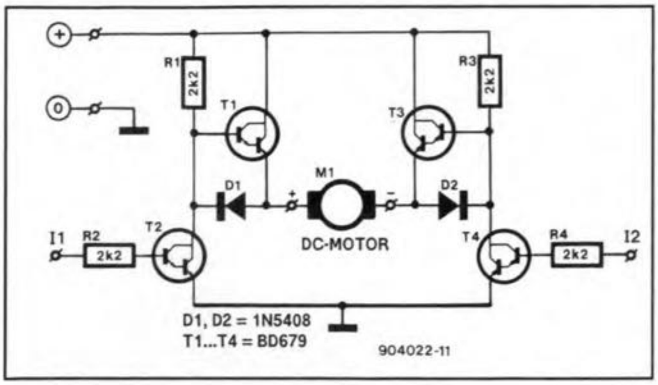
The current drawn from the supply is negligible. ■

(*From an idea by H. Smits*)

CLOCKWISE AND ANTI-CLOCKWISE DC MOTOR CONTROL 006

This straightforward circuit, based on four darlingtons, enables a d.c. motor to rotate clockwise or anti-clockwise under the control of two digital signals provided by, say, a computer.

As may be seen from the diagram, the circuit consists of two identical sections. Concentrating on the left-hand section, when a high logic level (+5 V) is applied to input I1, T2 is switched on and a current can flow to earth via D1. T1 is cut off, because its base is negative with respect to its emitter owing to the voltage drop across the diode (–0.6 V). When a low logic level (0 V) is applied to I1, T2 is cut off and T1 obtains base current via R1. The motor can then draw current via T1.



The right-hand section operates in an identical manner.

By applying different logic levels to the inputs, that is, logic high to I1 and logic low to I2, or vice versa, the motor may be made to rotate clockwise or anti-clockwise, as the case may be. When the levels at the inputs are identical, the motor is at a standstill.

With component values as shown, motors needing up to 45 V at 2 A may be controlled. However, when the current exceeds 0.5 A, the transistors need heat sinks.

The circuit may be used to control the motor speed by pulse-width modulation. This requires a constant level at one input (depending on the direction of rotation), while the pulses are applied to the other input. ■

(*R. Mennis*)

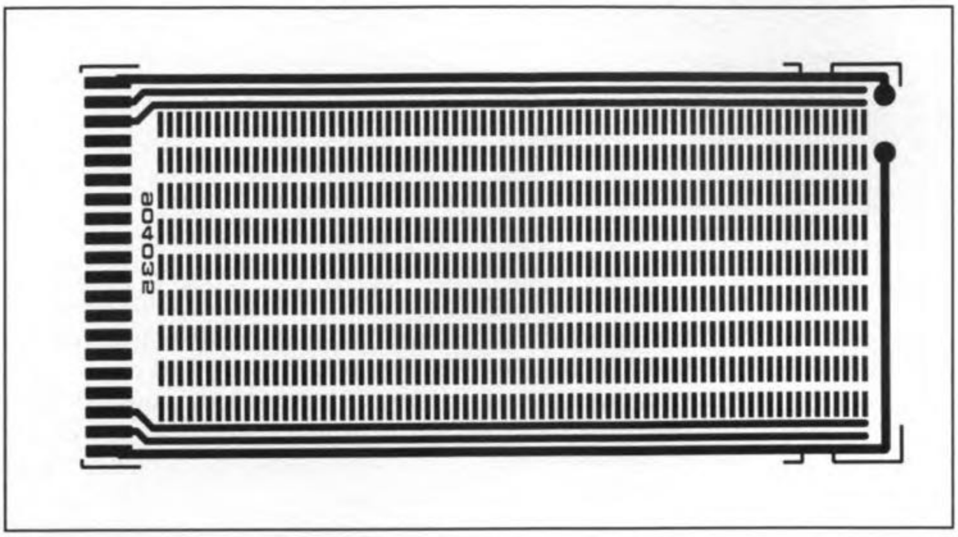
007 SMT PROTOTYPING BOARD

904035

As more and more electronics enthusiasts appear to have overcome their initial doubts, misgivings and fears of working with surface-mount technology (SMT) components, there is a growing demand for a universal board that allows prototype SMT circuits to be assembled quickly and reliably.

Since SMT component have no wire terminals, they can be fitted only by being soldered direct to the copper pads. The board presented here provides pads that are arranged in a pattern that enables virtually all types of SMT component to be accommodated. ■

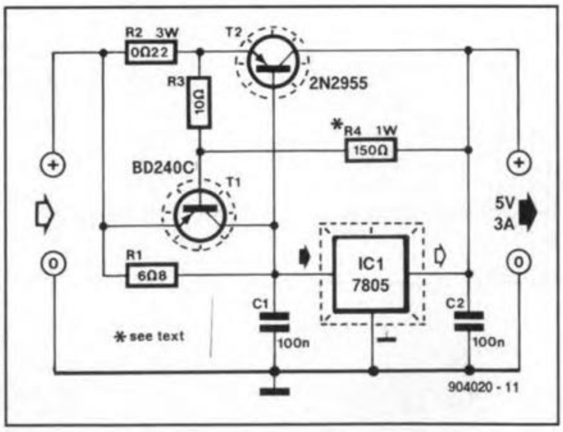
(M. Fabisch)



008 FOLDBACK VOLTAGE REGULATOR

The usual series 7805 and 7812 three-pin voltage regulators are excellent for normal applications. If currents up to 3 A are required, an additional transistor, such as T2 in the diagram, is used. That solution works well, but the overall dissipation in case of a short-circuit can get fairly high. That creates a difficulty, particularly when a Type 7812, 7815 or 7824 is used. This difficulty may be overcome by so-called foldback regulation. This ensures electronically that the maximum current is reduced when the output voltage decreases. In the prototype, the maximum current with the output short-circuited was only 0.5 V so that overheating did not occur.

Only a few additional components are needed for foldback regulation. In the diagram, T1 provides current limiting. As soon as the voltage drop across R2+R3 becomes greater than 0.6–0.7 V, the transistor is switched on, which reduces the base current of T2 to virtually zero. The voltage regulator is then more or less on its own, but it has very good thermal protection and limits its output current well before any harm is done. The voltage at which the protection circuits come into operation is the sum of the potentials across R2 and R3. Resistors R3 and R4 form a voltage divider for the potential across T2. The dissipation in T2 is directly proportional to the collector-emitter voltage, which is thus used here to control the current. In this way, the regulation characteristic is a function of the level of the input voltage.



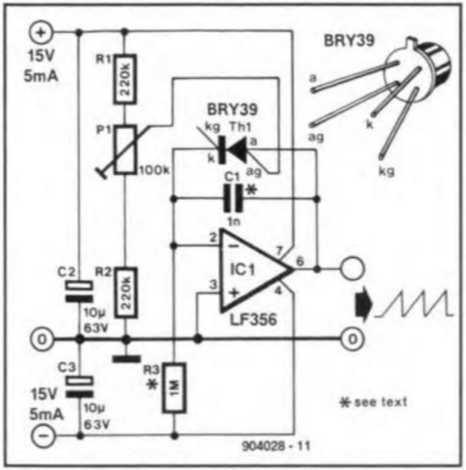
It is educational to experiment with the values of R2 and R3. When a short-circuit occurs, the drop across R3 should be large enough to drive T1 into virtual saturation. There is then practically no output current.

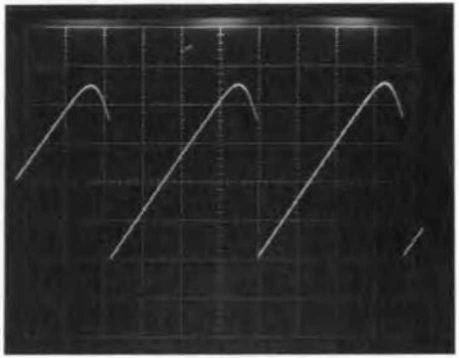
During testing, it will be noticed that 78xx regulators can withstand currents that are considerably larger than specified by the manufacturer (1–1.5 A), that is, until they get hot, when the maximum current level decreases ■

(Fairchild application)

009 LOW-FREQUENCY SAWTOOTH GENERATOR

The most noteworthy element in this circuit is Th1, which is known in data books by no fewer than three different names: thyristor tetrode, programmable uni-junction transistor (PUT) and silicon-controlled switch. In fact, the BRY39 is a four-layer (p-n-p-n) component. One of the characteristics of this type of component is that the junction of the two outer layers, that is, anode and cathode, begins to conduct when the potential across it exceeds a certain value. It ceases to conduct when





X: 0.2 ms/div; Y: 2 V/div; zero line: 4 V.

the current flowing from anode to cathode drops below a given level.

The sawtooth generator, which makes use of this property, is little more than an integrator of which the input is connected permanently to the negative supply rail. If it were not for Th1, the output of IC1 would rise slowly from 0 V to +15 V after switch-on and stay there. This voltage would also remain across C1 and Th1. This does not happen, however, because before the output of IC1 reaches +15 V, Th1 is switched on, which causes C1 to discharge rapidly so that the output of IC1 drops to 0 V. It does not reach that level, however, because before then the current through Th1 is too low to keep the device conducting. As soon as Th1 is switched off, the output of IC1 will rise slowly to +15 V, and so the action continues.

The voltage at which Th1 switches on may be preset within certain limits with P1. If it is set at 8.3 V (the positive peak of the sawtooth), the period would be of the order of 0.5R1C1, but this could, of course, be set to the required value with P1.

Note, however, that because of other properties of Th1 the value of R3 must remain between 500 kΩ and 2.2 MΩ. The value of C1 may lie between 1 nF and 200 µF. If large values are used, it is advisable to connect a 15 Ω resistor in series with the capacitor to limit the peak current during the discharge. ■

(*R. Sanjay*)

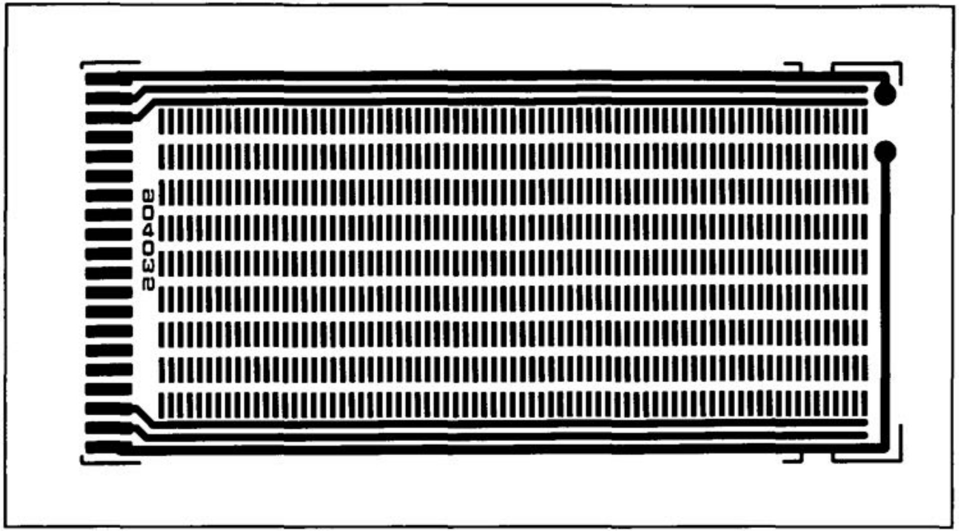
007 SMT PROTOTYPING BOARD

904035

As more and more electronics enthusiasts appear to have overcome their initial doubts, misgivings and fears of working with surface-mount technology (SMT) components, there is a growing demand for a universal board that allows prototype SMT circuits to be assembled quickly and reliably.

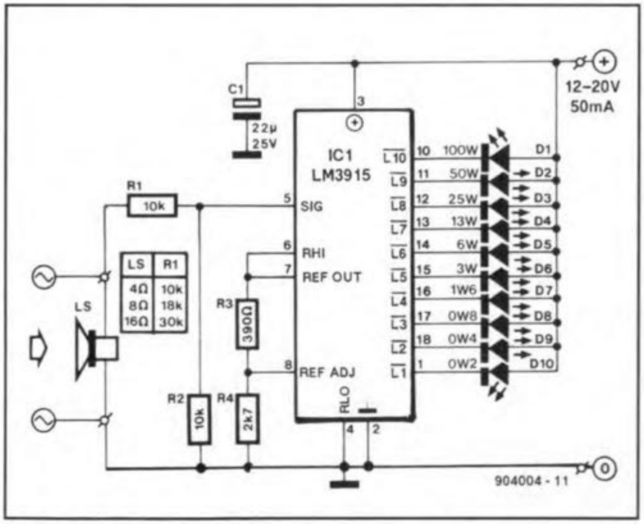
Since SMT component have no wire terminals, they can be fitted only by being soldered direct to the copper pads. The board presented here provides pads that are arranged in a pattern that enables virtually all types of SMT component to be accommodated. ■

(M. Fabisch)



AUDIO POWER INDICATOR 010

The LM3915 from National Semiconductor contains virtually everything for making a simple, yet reliable, audio power meter with a bar-type read-out. The circuit has one drawback in that it requires a separate power supply. This is compensated, however, by the fact that it is pretty sensitive (0.2 W min.) and does not degrade the sound quality in any way, since it does not present an additional load to the amplifier (in contrast, many inexpensive AF power indicators derive their display current from the amplifier).



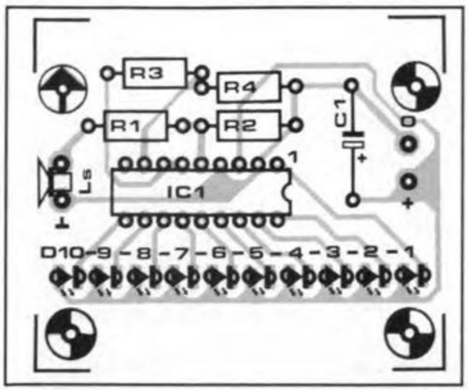
PARTS LIST

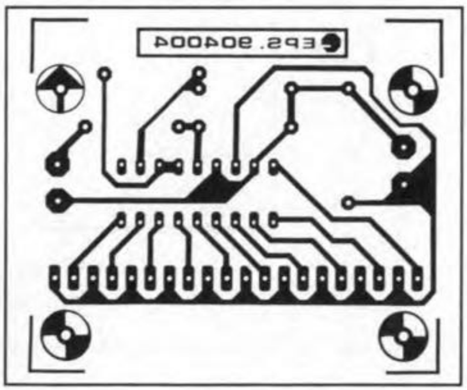
Resistors:
R1 = 10 k (see text)
R2 = 10 k
R3 = 390 Ω
R4 = 2k7

Capacitors:
C1 = 22 µF, 25 V, axial

Semiconductors:
D1–D10 = LED, rectangu lar
IC1 = LM3915

Miscellaneous:
Enclosure: Pan-Tec Type PIN1064-01

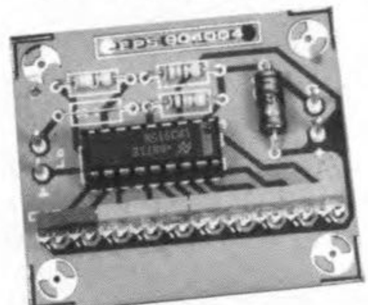




The value of resistor R1 depends on the loudspeaker impedance as shown in the table inset in the circuit diagram. The resistor may be replaced by a wire link in the relevant position on the PCB if it can be fitted inside the plug that connects the indicator to the loudspeaker. This makes it convenient to use the indicator with loudspeakers of different impedance: use a dedicated cable for each impedance.

For use with stereo systems, the circuit is either built in duplicate or the signals across the loudspeakers are applied to two series-connected resistors R1, whose common junction is connected to pin 5 of IC1. The latter method may raise some eyebrows, but it works fine in practice.

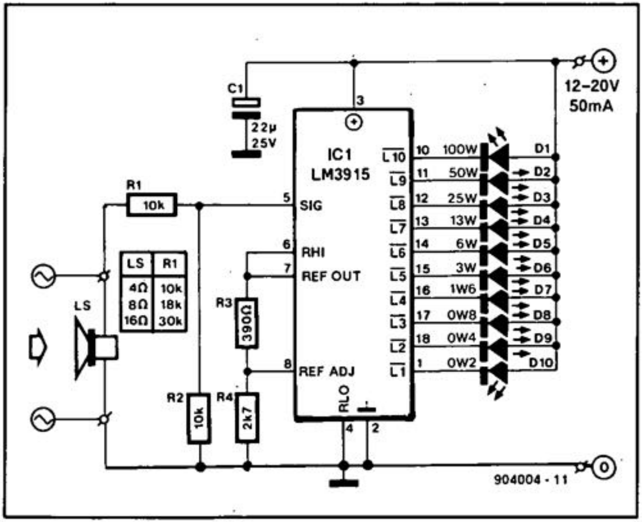
The power supply for the indicator is



derived from a simple a.c. mains adaptor that provides a d.c. output of 12–20 V.

Finally, it should be noted that the actual power measurement is an approximation only since the LM3915 reacts to the positive half-cycles of the signal only. This causes the top LED in the bar to light at a slightly reduced intensity. ■

(*K. Walters*)



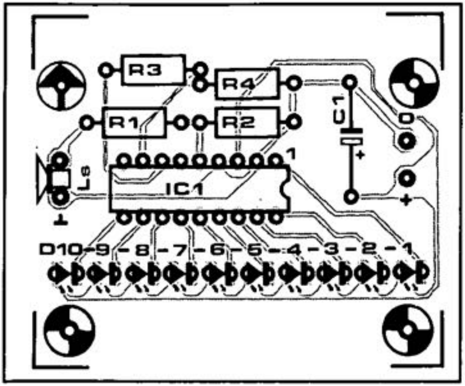
PARTS LIST

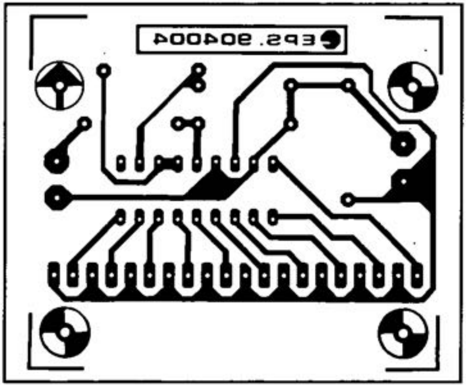
Resistors:
R1 = 10 k (see text)
R2 = 10 k
R3 = 390 Ω
R4 = 2k7

Capacitors:
C1 = 22 µF, 25 V, axial

Semiconductors:
D1–D10 = LED, rectangu lar
IC1 = LM3915

Miscellaneous:
Enclosure: Pan-Tec Type PIN1064-01





011 LINEARLY VARIABLE POWER SUPPLY

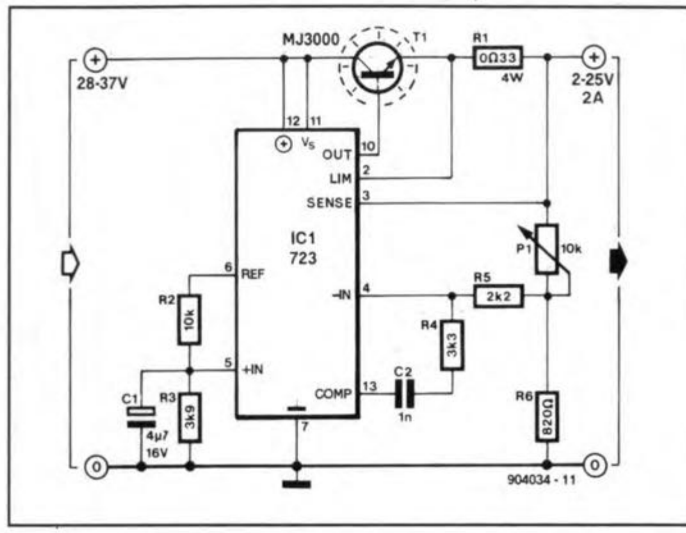
In most variable power supplies, the correlation between the setting of the wiper of the relevant potentiometer and the output voltage is non-linear. This results in the output voltage either having to be measured or indicated on an integral meter. If the relation were linear, a simple linear scale on the potentiometer would suffice.

The only difference between the present and a traditional power supply is that the wiper and the 'earthy' terminal of the relevant potentiometer are inter-linked. This simple fact allows a number of equivalent circuits to be improved. It is, of course, necessary to use a linear potentiometer.

Some parameters of the circuit: input-voltage: 28–37 V; output voltage: adjustable from 2 V to 25 V; maximum output current: 2 A.

It should be noted that the dissipation of the MJ3000 can rise to 50 W: it is, therefore, necessary that this component be fitted on a 1.5 K/W heat sink ■

(N. Körber)



012 POWER SUPPLY DOWN TO ZERO VOLTS

Most power supplies, particularly switch-mode types, are not designed to provide output voltages smaller than a volt or two. It is often desirable, however, especially in experimental work, to raise the output voltage slowly from 0 V.

The circuit shown here may be used with almost any power supply. The use of an auxiliary voltage, provided by T1-R3-D6, makes the supply act as if its output is equal to the internal reference voltage of 5.1 V, whereas in fact it is lower.

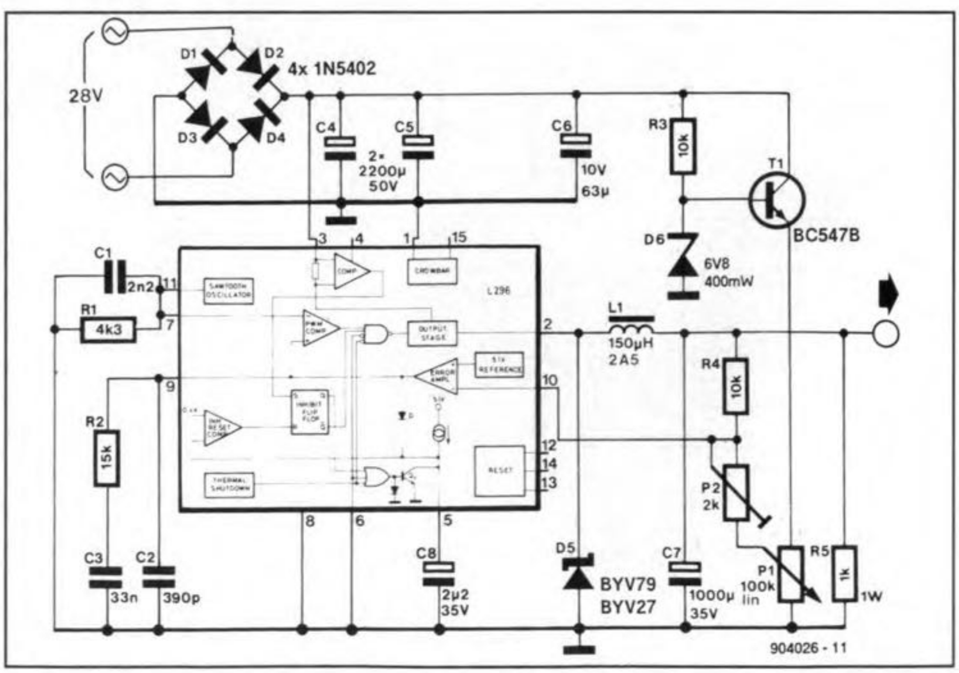
When the wiper of P1 is at earth potential, the circuit is an ordinary supply whose output may be varied with P2 between 5.1 V and 30 V. For the present purposes, P2 is permanently adjusted to give an output of 30 V: from now on the output will be varied by P1.

When the wiper of P1 is at earth potential, the output voltage is 30 V. Opening the potentiometer causes a larger and larger voltage from wiper to ground, which is derived from the auxiliary source based on T1. From the regulator, it therefore seems as if the output voltage increases, which, of course, is not so. The output voltage thus drops further and further the more P1 is opened.

In the prototype, an auxiliary voltage of 6 V was rather too low to reduce the output voltage to zero. A zener of 8.2 V was found more suitable.

The Type L296 regulator can provide a current of up to 2 A. A Type L4960 may also be used, b ut the maximum current is then somewhat lower. ■

(SGS application)



PULSER 013

A pulser is a special type of pulse generator that produces short-duration (fast) pulses. Such pulses are very useful for test and measurement purposes. For instance, in conjunction with an oscilloscope, they enable short-circuits and breaks in PCB tracks to be located quickly.

The pulser also makes it possible for the logic level to be ascertained at any point in a circuit where the probe is held. When spring-loaded switch S1 is pressed, a 1 μs pulse is generated at the opposite logic level from that at the relevant point in the circuit under test. Because of C5, a current of up to 500 mA may be drawn for the duration of the pulse.

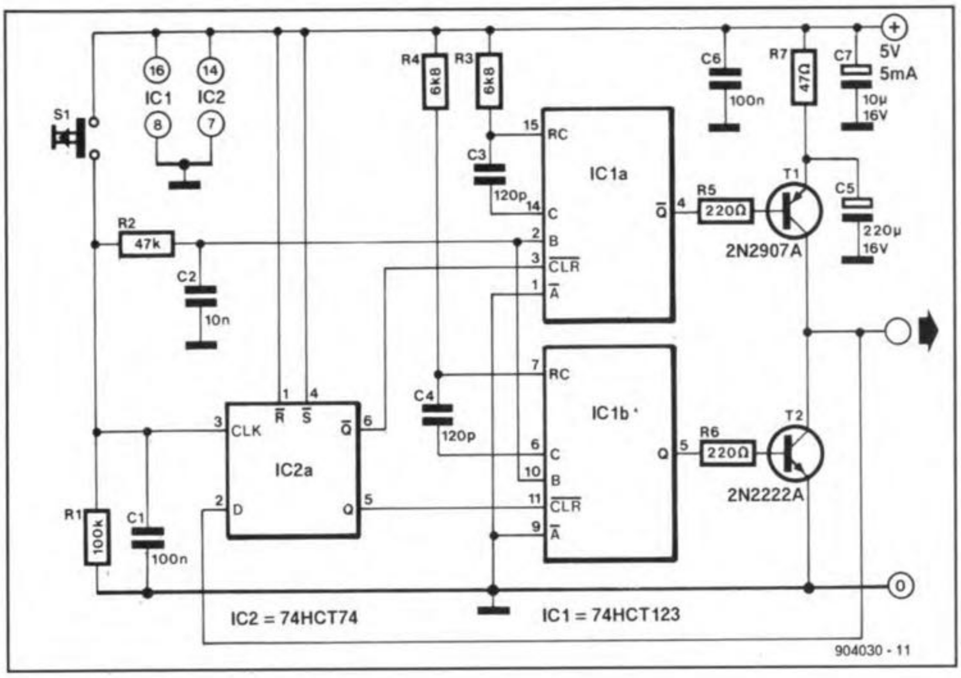
The logic level at the test point is ascertained by IC2a, a D-type bistable to whose D input the measured signal is applied. At the instant S1 is pressed, the bistable receives a clock pulse upon which the signal at the D input is transferred instantly to the Q output and—inverted—to the $\overline{Q}$ output.

Two monostables, IC1a and IC1b, drive the two output transistors that generate the pulse.

The signal from S1 is applied to the B inputs of the monostables via delay line R2-C2. The monostables can be triggered by a leading edge at their B input only when their $\overline{CLR}$ input is logic high. Therefore, depending on the logic level at the D input of IC2a, one of the monostables will be disabled. The other is triggered and switches on the relevant transistor at its output for 1 μs. If, when S1 was pressed, the logic level at the probe was '1', transistor T2 is switched on by IC1b. If the level had been '0', T1 would have been switched on by IC1a.

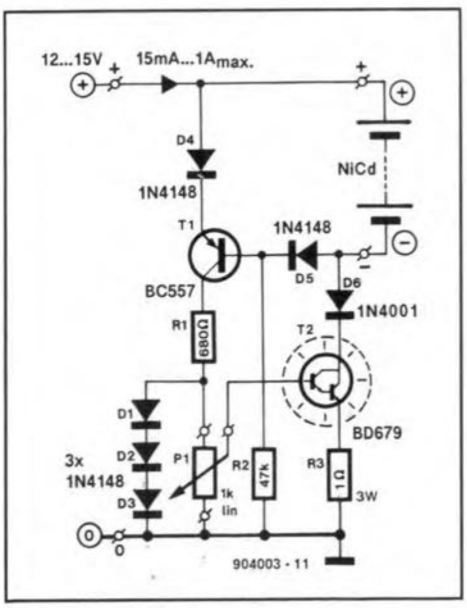
Capacitor C5 ensures that sufficient current is provided at the test point. When no pulses are generated (pressing S1 generates only one pulse), C5 recharges via R7.

During operation, the pulser draws on average not more than 5 mA from a 9 V battery. ■



(C. Sanjay)

DE LUXE NICD BATTERY CHARGER 014



With circuit values as shown, this charger allows up to seven series-connected NiCd batteries to be charged. This number may be increased by raising the input voltage by about 1.65 V per additional battery. Provided T2 is fitted on an appropriate heat sink, the input voltage may be increased up to 25 V.

In contrast to many conventional NiCd battery chargers on the market, the present one provides protection against polarity reversal. That is, when the battery terminals are connected the wrong way to the charger, no charging current is produced.

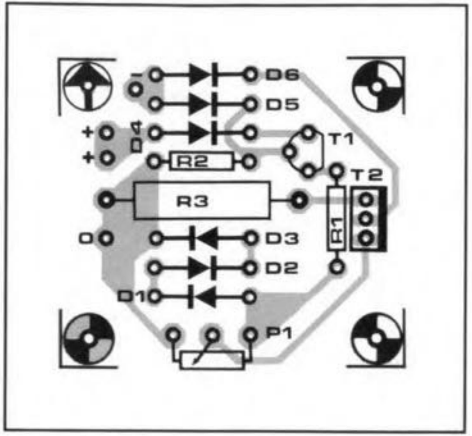
Another useful property of the charger is that it does not load the battery when it is switched off.

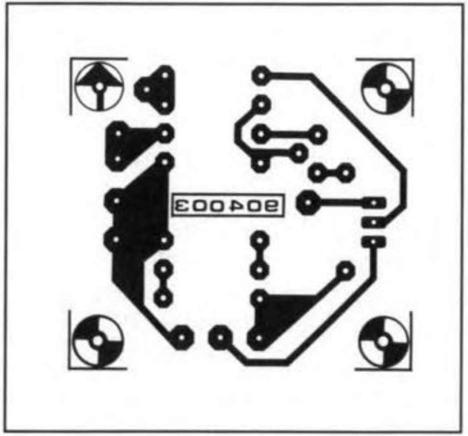
NiCd batteries are normally charged over a period of 14 hours at a current equal to about 1/10th of the battery capacity. In other words, a 500 mAh battery is usually charged at 50 mA for 14 hours. Normally, a slightly longer period of charge will not harm the battery. If, however, the current is higher than indicated, the charging time must be reduced proportionally to prevent damage to the cells.

The level of charging current is controlled between 0 mA and 1 A with preset P1. The current may be measured with a voltmeter across R3, which, to avoid calculations, has a convenient value of 1 Ω.

Operation of the circuit is straightforward. Transistor T1 is on when a battery is connected with correct polarity and also if the output terminals are open-circuit. The collector current of this transistor results in a reference voltage of about 2.1 V across

S 12





PARTS LIST

Resistors:
R1 = 680 Ω
R2 = 47 k
R3 = 1 Ω, 3 W
P1 = 1 k linear potentiometer

Semiconductors:
D1–D5 = 1N4148
D6 = 1N4001
T1 = BC557B
T2 = BD679

Miscellaneous:
Heat sink for T2

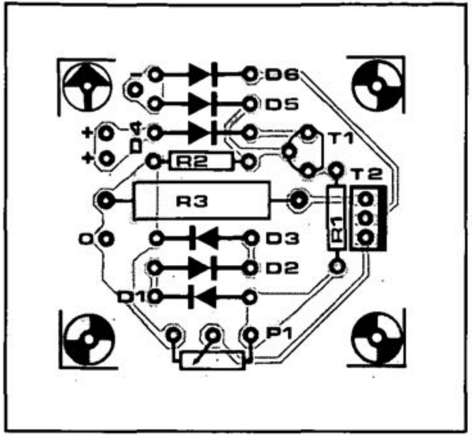
diodes D1–D3. Part of this voltage is applied to darlington transistor T2 via P1. The emitter resistor of T2, R3, provides the constant-current function. Note that T2 must be fitted on a heat sink.

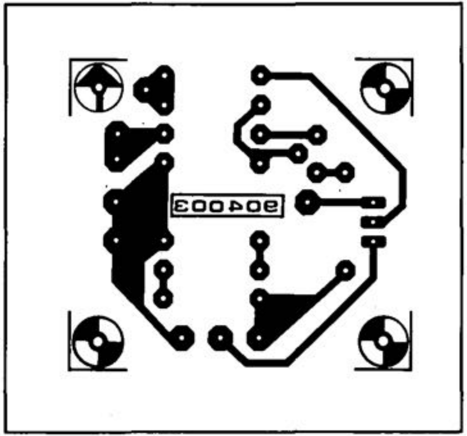
In case the BD679 is difficult to obtain, it may be replaced by any n-p-n medium power darlington with a collector voltage/current specification of 30 V at 2 A.

The maximum output current may be increased above 1 A by lowering the value of R3.

The charger draws a quiescent current of 15 mA at 12 V. ■

(*Ever Ready application*)





PARTS LIST

Resistors:
R1 =680 Ω
R2 = 47 k
R3 = 1 Ω, 3 W
P1 = 1 k linear potentiometer

Semiconductors:
D1–D5 = 1N4148
D6 = 1N4001
T1 = BC557B
T2 = BD679

Miscellaneous:
Heat sink for T2

015 LOW-LOSS VOLTAGE REGULATOR

The voltage regulator in portable, battery-operated equipment must have minimum dissipation, good temperature stability, and be able to deliver an adequate current. These requirements are met by the IRF9530 MOSFET series regulator from Precision Monolithics Inc. (PMI). One of its main attributes is the low drive current required: the two Type OP90 opamps that drive it and at the same time generate the reference voltage draw only 20 μA.

Circuit IC2 compares a portion of the output voltage existing across divider R6-R7 with the reference voltage provided by IC1. Its output adjusts the gate potential of the IRF9530 in a manner that ensures a constant drain voltage of 5 V. Resistor R7 is a pull-up resistor. At output currents between 0 A and 1 A, the gate voltage lies between 3.75 V and 1.9 V respectively.

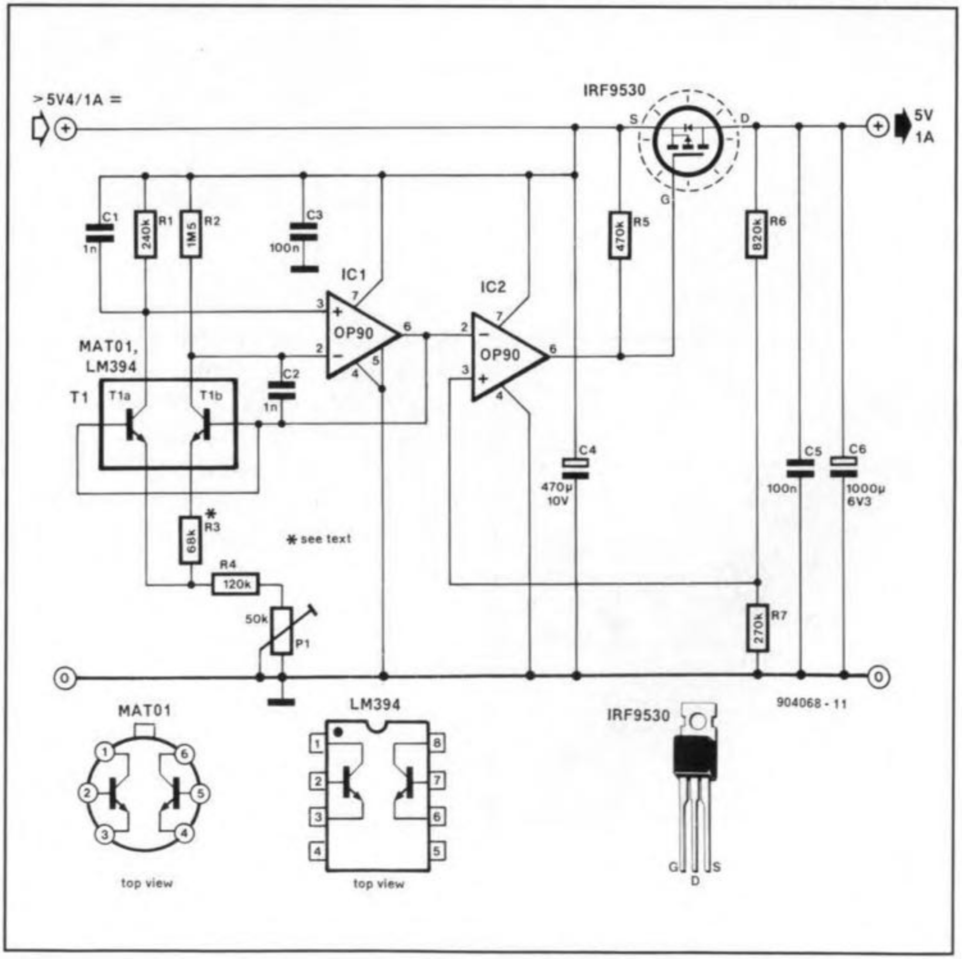
Transistor T1 and circuit IC1 form a band-gap voltage reference. Because of its high stability and low current drain (about 5 μA), this type of reference is eminently suitable for this application

Owing to the presence of R3 in the emitter circuit of T1b, this transistor draws less current than T1a. Because of this, the base-emitter voltage of T1b will be somewhat smaller than that of T1a. None the less, the collector voltages of the two transistors are practically equal since the value of R2 is much higher than that of R1. These collector voltages are applied to the two inputs of IC1 and their (magnified) difference is fed back to the bases.

The ratio R1:R2:R3 has been chosen to ensure good compensation of the temperature coefficient of the transistors.

The optimum operating point of the circuit coincides with a reference voltage of 1.23 V.

The circuit was originally designed with a Type MAT01 in the T1 position, but it will work satisfactorily with most types of dual transistor. It should be noted, however, that some other types were found very sensitive to tolerances. In some cases, this prevented the circuit from stabilizing on a fixed operating point, which resulted in the output remaining high or low. To en-



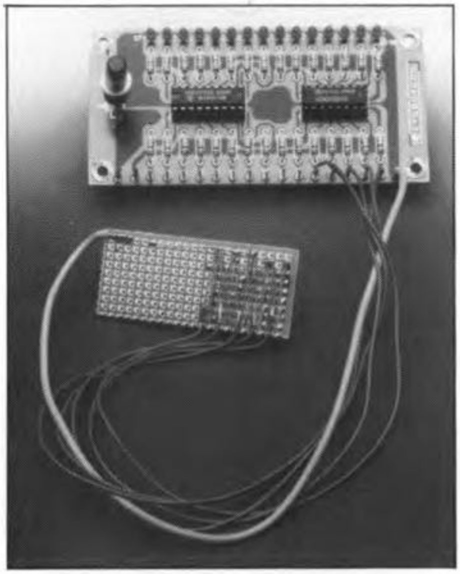
sure a well-defined output voltage at power-up, the offset adjustment (pin 5) of IC1 is connected to earth. This results in the output of IC1 always being logic high on power-up.

When a dual transistor other than the MAT01 is used, it is advisable to adapt the values of R3 and R4 by disconnecting pin 6 of IC1 from the bases of the transistors. Next, apply a voltage of 1.23 V (via a voltage divider) to the bases and vary the values of R3 and R4 (if necessary) until IC1 just does not toggle. Finally, test the temperature stability with a small blow lamp.

The IRF9530 only needs a heat sink if the input voltage exceeds 6.25 V.

The circuit is intended for use with input voltages from NiCd or low-capacity lead-acid batteries, which provide an inherently stable voltage. Therefore, the circuit does not contain any ripple suppression components. Furthermore, large variations in current are compensated only slowly. These small drawbacks may be avoided by the use of a standard reference and a faster opamp in the IC2 position. That will, however, result in an increase in current, which in the circuit as shown is only 45 μA. ■

(*PMI application*)



The fluid level indicator may be used, for instance, in the fresh water tank on board yachts, or in mobile homes or caravans.

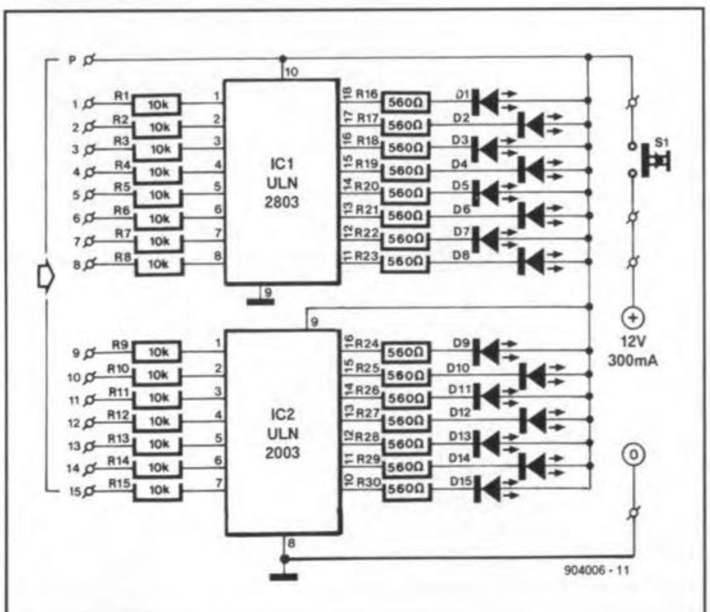
The drivers in IC1 and IC2 are darlington transistors (of which the ULN2803 has eight and the ULN2003, seven). The base of each of these is connected to a sensor, formed by a carbon rod, or a strip of aluminium or copper. These sensors are fitted at the appropriate level in the fluid vessel.

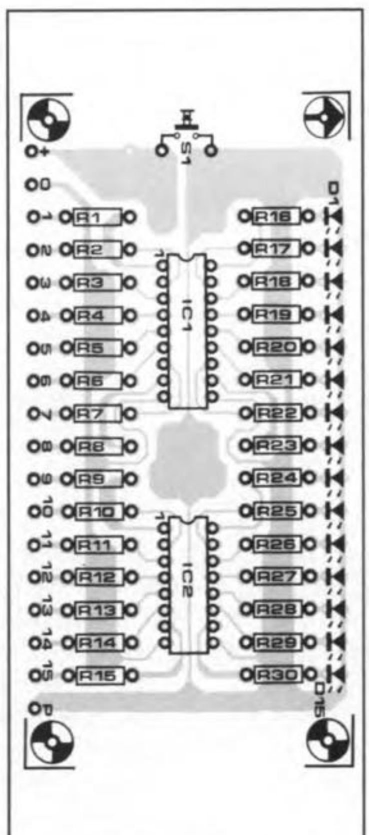
The 15 LEDs ate the outputs of the drivers are arranged to display a bar that provides an easy-to-read fluid level indicator when S1 is pressed. Because of the relatively high current drain (about 300 mA) and the limited capacity of the battery on board the yacht or vehicle, it is strongly advisable not to use the indicator continuously (this would also cause rapid erosion of the sensors).

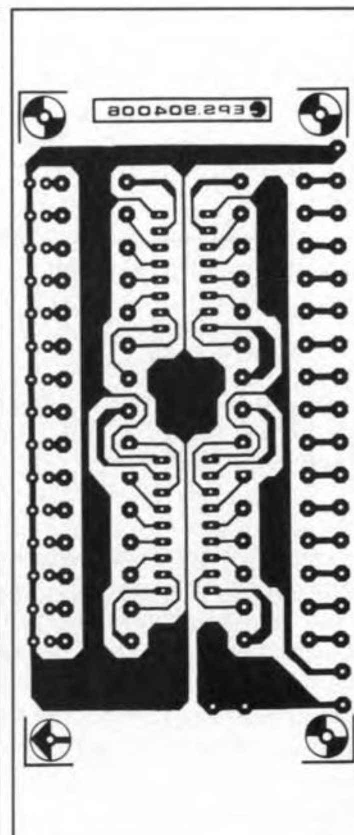
Since the ULN drivers are capable of supplying a peak current of up to 500 mA, one or more of the LEDs may be replaced by a relay, small buzzer, or other means of providing an audible warning for too high or too low a fluid level.

The ULN 2003 and ULN2803 may be replaced by similar devices from the same family of power drivers, for instance, the ULN2005 and ULN 2805, or the ULN2001 and ULN 2801. A word of warning if the 2001 and 2801 are used: DO NOT CONNECT any of the inputs of these devices direct to +12 V, since this is likely to destroy the relevant IC. ■

(D. Lorenz)





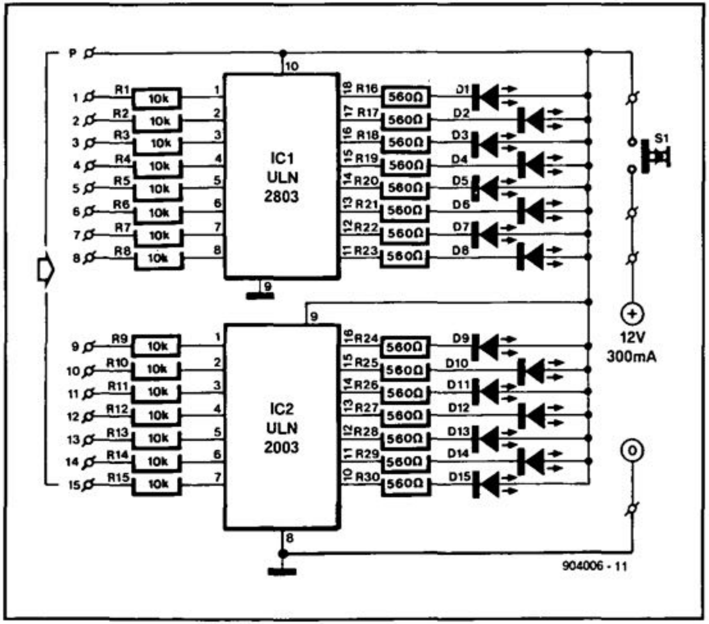


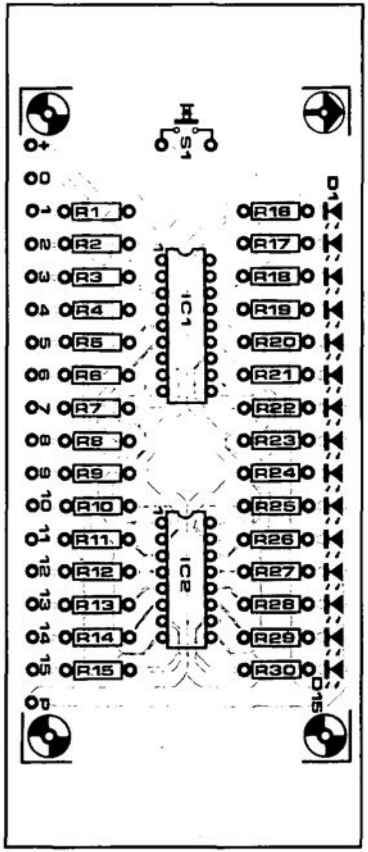
PARTS LIST

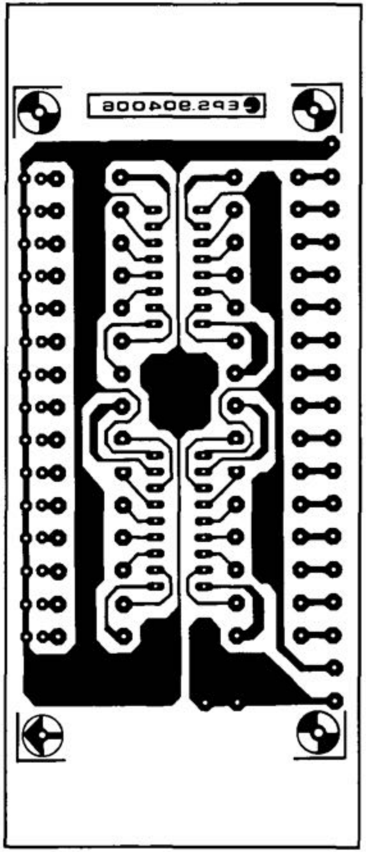
Resistors:
R1–R15 = 10 k
R16–R30 = 560 Ω

Semiconductors:
D1–D15 = LED, 3 mm dia., red
IC1 = ULN2803
IC2 = ULN2003

Miscellaneous:
S1 = push-to-make button switch







PARTS LIST

Resistors:

R1–R15 = 10 k

R16–R30 = 560 Ω

Semiconductors:

D1–D15 = LED, 3 mm dia., red

IC1 = ULN2803

IC2 = ULN2003

Miscellaneous:

S1 = push-to-make button switch

017 AF OUTPUT LEVEL INDICATOR

Described here is a maximum-level indicator that does not, or hardly, react to fast peak powers. It is intended to be driven by the output stages of an AF amplifier.

The section based on T1 is a traditional voltage regulator for the incoming supply. The values of R1 and R2 depend on the level of the supply voltage as shown in the table. The transistor does not need a heat sink.

The remainder of the circuit consists of two identical sections for the left- and right-hand channels respectively. Only the left-hand channel section is discussed.

The incoming signal is rectified by D4 and D5 and then charges C4. This capacitor discharges via R5 and R6.

Circuit IC1a acts as a comparator and is connected as a Schmitt trigger. It compares the rectified input signal with a reference voltage that is set with P1. Diode D3 gives this voltage an offset of 0.7 V, which is necessary in view of the hysteresis of IC1a. The hysteresis is determined by R6 and R7, but is reduced by the d.c. feedback network R8-R9.

The LED is driven by an emitter follower, T2.

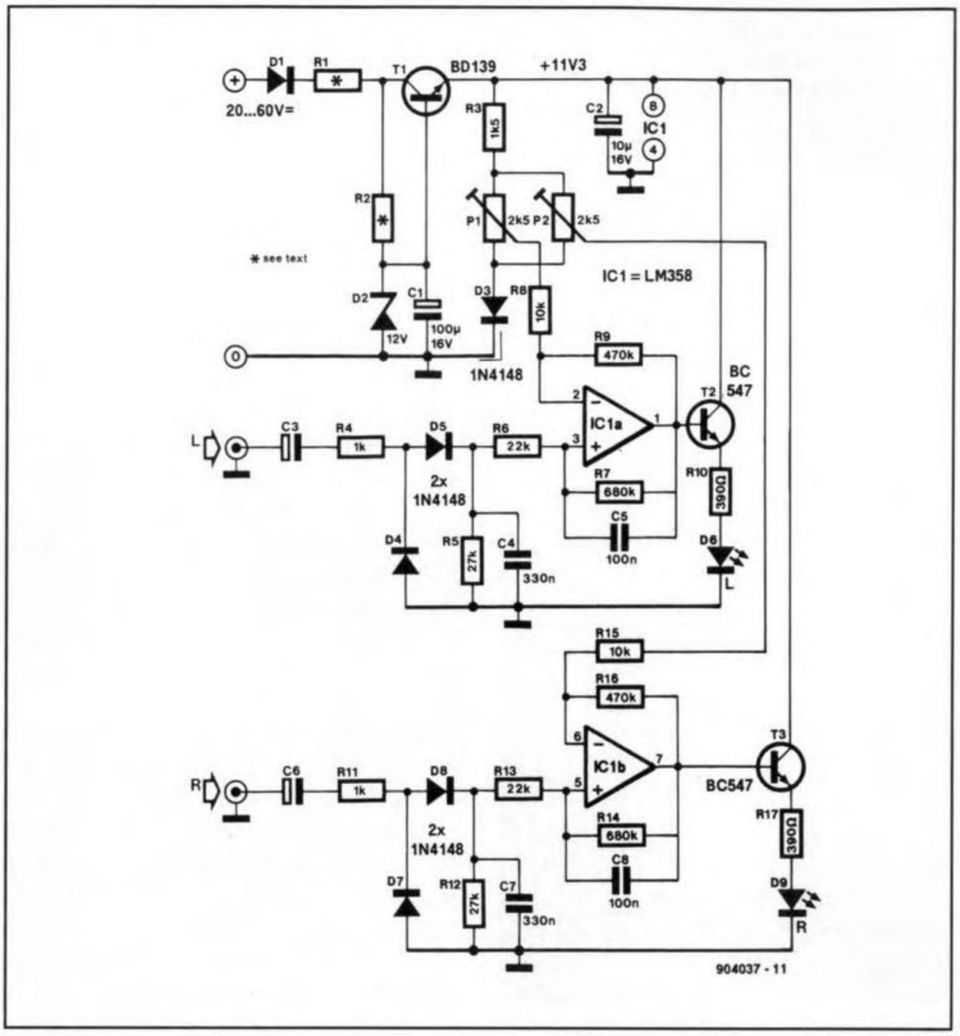
The opamp is a Type LM358, which has the advantage of performing well even if operating with an asymmetrical supply when the inputs are at earth potential.

The signal input may lie between 2 Vpp and 10 Vpp. The toggle level of the circuit is arranged with preset potentiometers. The hysteresis is 400 mV over the whole range. The circuit is triggered at the same level over the frequency range 100 Hz to 15 kHz.

The circuit reacts to short bursts (such as those from the kettle drum) only if their amplitude is at least 50 per cent higher than the trigger voltage.

The single supply voltage may have a value of between +20 V and +60 V. Current drain varies between 10 mA (quiescent) and about 50 mA when both LEDs light. ■

(*A. Ferndown*)



Supply voltage	R1	R2
20 V	330 Ω, 1 W	3k3
30 V	470 Ω, 1 W	3k3
40 V	560 Ω, 2 W	3k3
50 V	680 Ω, 5 W	5k6

018 INFRA-RED DETECTOR I

Almost all households have at least one infra-red remote control unit, be it for the viode or audio equipment. Unfortunately, it sometimes happens that the control does not fucntion properly and it is then difficult to ascertain whether the receiver or sender is at fault. The detector described here can help by telling us whether the sender works or not.

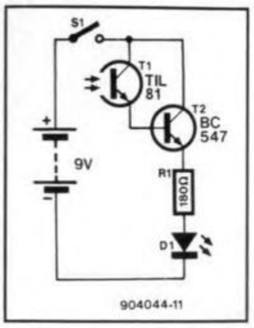
The IR light from the sender is detected by T1, an IR photo transistor. When IR light falls on to T1, it switches on T2 and this results in the LED lighting at the rhythm of the IR signals.

The brightness of the LED depends on the strength of the IR light falling on to T1 so that the remaining capacity of the batteries may be estimated.

Although a Type TIL81 is used here in the T1 position, virtually any IR photo transistor is suitable.

Since the current through the LED is fairly small, it is advisable to use a high-efficiency type. ■

(*R. Systermans*)



ELECTRONIC 4-POSITION SELECTOR

019

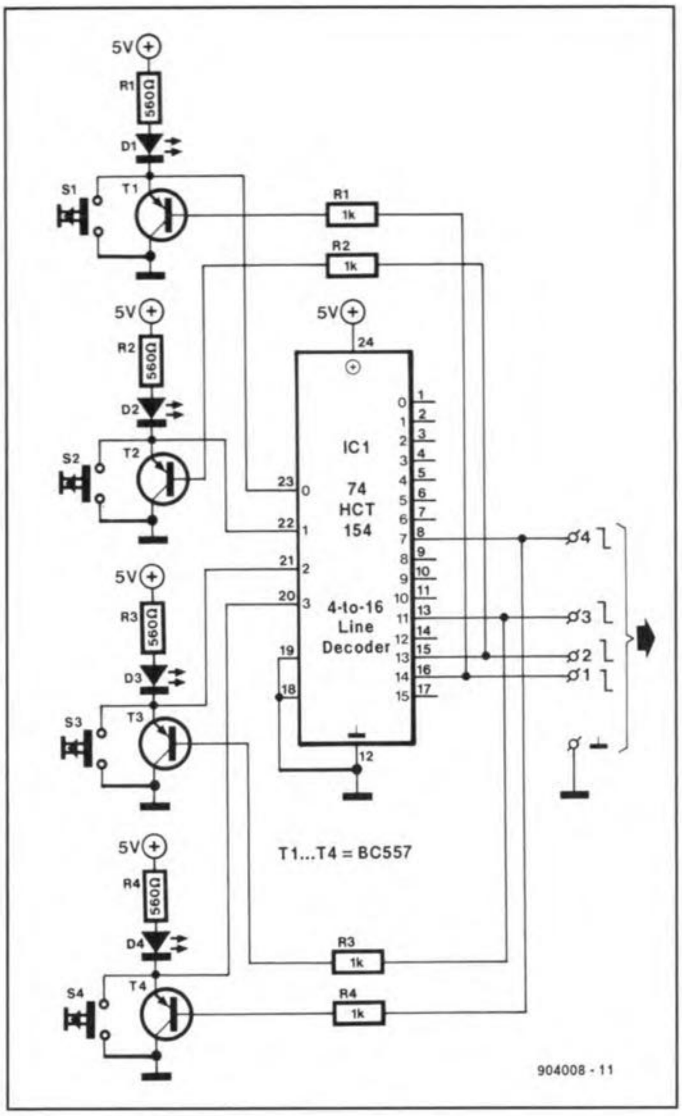
A 4-to-16 decoder is eminently suitable for constructing a simple four-position selector with integral debouncing of the switches.

The circuit is a cunning design that uses few components. The switches are connected to the inputs of the decoder via a resistor, an LED and a transistor. The transistor is in parallel with the switch, so that pressing the button has effect only when the transistor is off.

The transistors are driven by four outputs of the decoder via 1 k resistors.

When the supply is switched on, none of the transistors conducts, so that the inputs of IC1 are '1111' and pin 17 of the decoder is logic high. If in this situation one of the switches is pressed, the associated input bit to IC1 becomes logic low. This changes the input code and the output associated with the switch goes high, that is, pin 16 for S1, pin 15 for S2, pin 13 for S3 and pin 8 for S4.

The signal at the relevant output pin is used to switch on the transistor associated with the closed switch. From then on, the transistor assumes the task of the switch, so that this may be released. This condition is indicated by the relevant LED.



If one of the other switches is then pressed, the input code to IC1 contains two zeros. This actuates a decoder output that is not fed back to the transistors. The conducting transistor is then switched off, which results in the decoder input code changing once again, but this time to the correct code associated with the currently pressed switch. The decoder output associated with that switch then goes high and switches on the relevant transistor, which assumes the task of the switch.

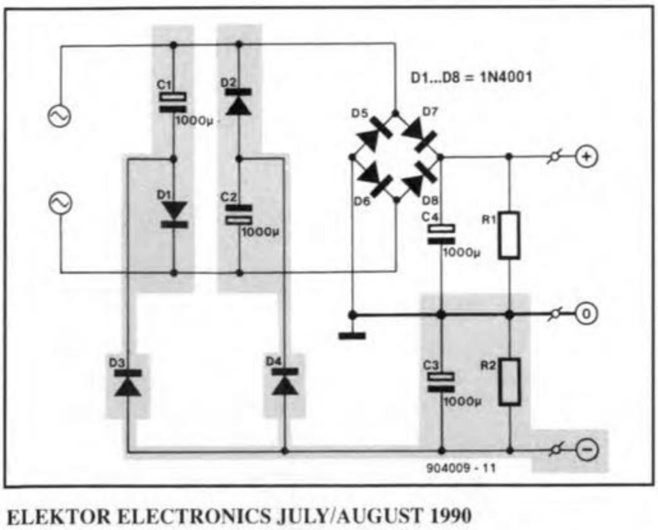
The outputs of the circuit may be used, perhaps via an additional amplifier, to actuate four relays.

The current drawn by the circuit is determined primarily by the LEDs: as shown, the circuit draws about 10 mA. ■

(*H. Smits*)

POWER SUPPLY EXTENSION

020



It happens frequently that a circuit powered by a single supply voltage is extended and then needs a second supply voltage of opposite polarity to that already available. That requirement can be met by the circuit described here.

Normally, an auxiliary negative voltage would be provided by half-wave rectification by C1, D1, D3, and C3, but it is better to use full-wave rectification, because of the higher current and smaller ripple, and that is effected by the addition of C2, D2 and D4.

In the diagram, bridge rectifier D5–D8 and smoothing capacitor C4 provide the existing positive supply voltage. The added components provide a negative voltage at practically the same level as the positive supply, but that depends on the value of C1 and C2 and the current required. With values as shown, the negative supply can provide up to about 200 mA.

It should be noted that in this type of

circuit the current drawn from the positive supply must always be larger than that from the negative supply. If the positive supply is not loaded, the negative supply can not provide any current! Note also that in the diagram R1 and R2 are the respective loads, not actual resistors.

If it is required that the current from the negative supply is greater than that from the positive rail, the circuit should be inverted. The bridge rectifier should then provide the negative voltage and the additional components, the positive voltage. All diodes and capacitors should also be inverted. ■

(*K. Walters*)

021 INEXPENSIVE TRANSISTOR TESTER

Resistance checks on suspect transistors normally fail to provide a conclusive answer as to whether the transistor is all right or defect. Moreover, such checks mean that the transistor must be removed from its circuit and connected in six different ways, that is, b-e test, b-c test and c-e test, each two times with reversed meter polarity.

The tester presented here allows the transistor under test (TUT) to remain in circuit, provided the circuit has relatively high resistance values.

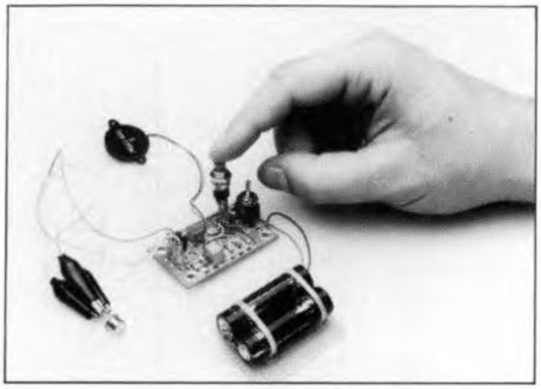
The tester is suitable for p-n-p as well as n-p-n transistors and also accepts darlingtons. The selection between p-n-p and n-p-n is made by reversing the supply voltage with S2.

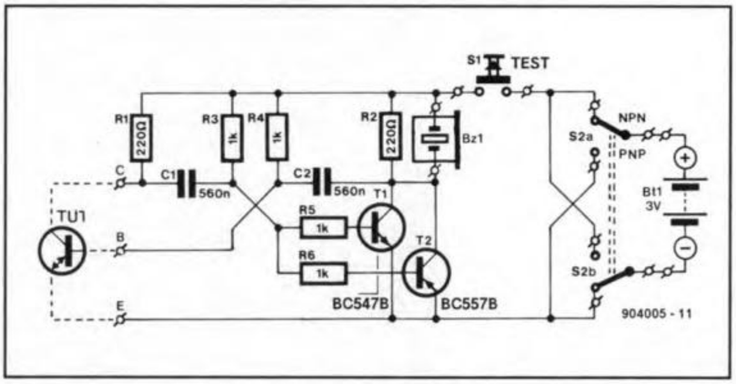
When the necessary connections are made, the transistor under test forms part of a collector-coupled astable. The transistor is almost certainly all right if it enables the astable to oscillate at about 2 kHz, which is indicated by buzzer Bz1. It should, however, be noted that transistors with relatively low current gain may pass this test and still be defect.

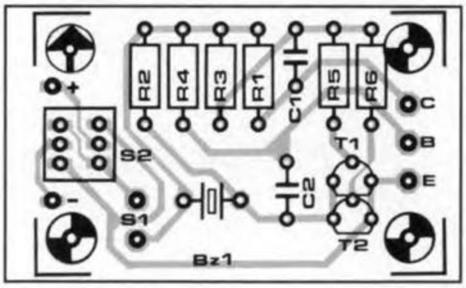
The small size of the printed-circuit board, and the use of two series-connected 1.5 V batteries allow the tester to be housed in a compact case. Connections to the TUT are by means of flexible test leads terminated in small crocodile clips.

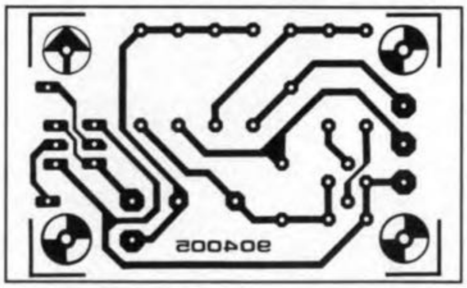
In operation, the tester draws a current of about 20 mA. ■

(J. Ruffell)









PARTS LIST

Resistors:
R1, R2 = 220 Ω
R3–R6 = 1 k

Capacitors:
C1, C2 = 560 n

Semiconductors:
T1 = BC547B
T2 = BC557B

Miscellaneous:
S1 = miniature push-button
S2 = miniature DPDT switch
Bz1 = passive piezo-ceramic buzzer Type PKM11-4A0
2 batteries, 1.5 V

022 POWER SUPPLY INPUT ADAPTOR

This circuit enables the input voltage of a power supply to be adjusted in accordance with the required output voltage to eliminates undue power dissipation in series regulators. This is achieved by monitoring the output voltage andaltering the unregulated input voltage as necessary with the aid of two relays.

The sensing input, derived from the positive supply terminal, is applied to potential divider P1-R1-R2-R3-R4. Comparators IC1a, IC1b and IC1c detect respectively when the voltage at the relevant junction of the divider is 1/4, 1/2 or 3/4 of the maximum sensing input set with P1.

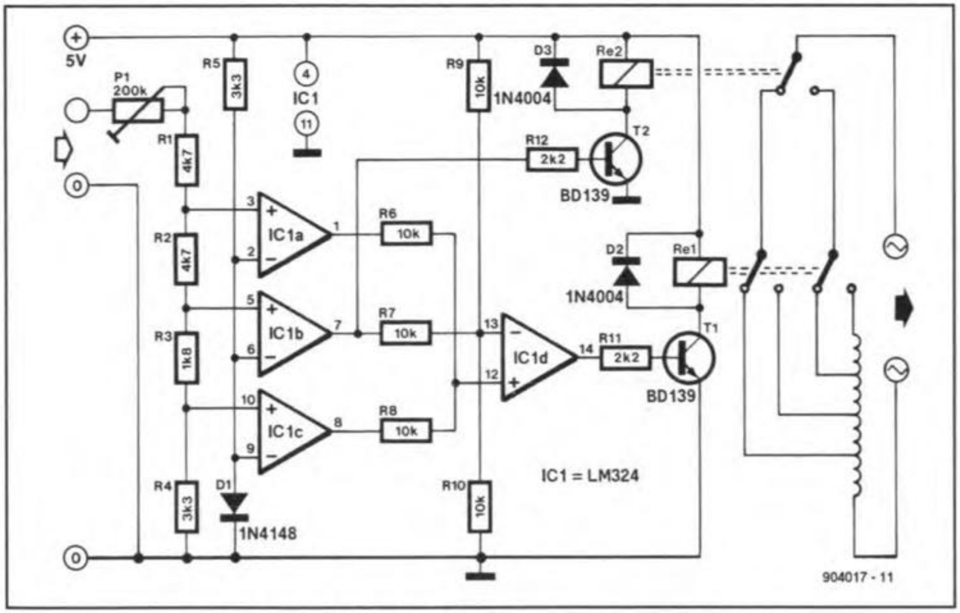


Fig. 1. Circuit diagram of the power supply input adaptor.

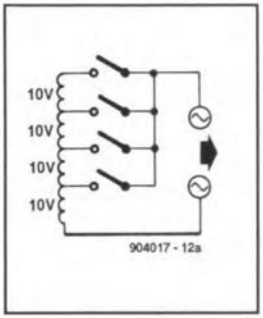


Fig. 2a.

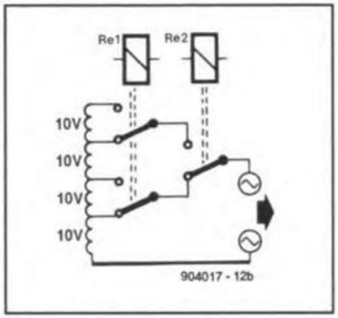


Fig. 2b.

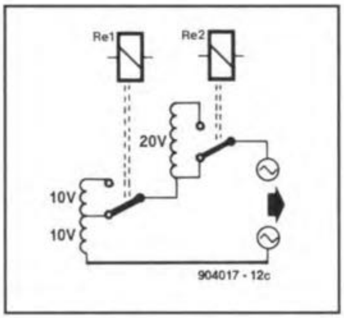


Fig. 2c.

The voltage drop across diode D1 serves as a 0.7 V reference.

Double-pole relay Re1 is energized (a) when the output of IC1a is high and that of IC1b is low, and (b) when the output of IC1c is high.

Condition (a) pertains when the output voltage of the supply is between 1/4 and 1/2 of its maximum value, while (b) occurs when the output voltage is greater than 3/4 of its maximum value.

When IC1a is the only opamp whose output is high, the +ve input of IC1d is held at half the supply voltage. Since in that condition the –ve input is at 1/3 of the supply voltage, IC1d toggles and this causes Re1 to be energized.

If the output of IC1b also goes high, the –ve input of IC1d becomes 2/3 of the supply voltage, the comparator returns to its original state and Re1 is de-energized.

When the output of IC1c also goes high, IC1d toggles again and the relay is re-energized.

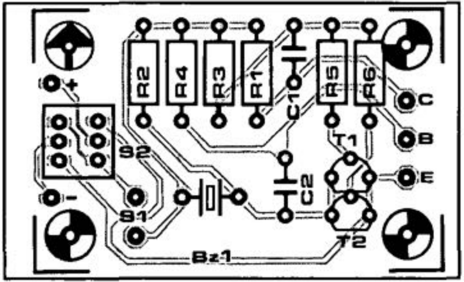
To prevent unwanted voltage dips when the circuit switches between the taps on the secondary winding of the supply transformer, it is better to use two relays as shown in Fig. 2b. This arrangement has advantages over that in Fig. 2a in which one relay controls four contacts. An even better arrangement is shown in Fig. 2c, since that obviates the use of a DPDT relay. Note, however, that a mains transformer with two separate separate secondary windings is required.

Preset P1 is adjusted to give the required supply output at which the relay is actuated. Switching levels of 10 V, 20 V and 30 V, for instance, are obtained with P1 set to a value of 125 kΩ.

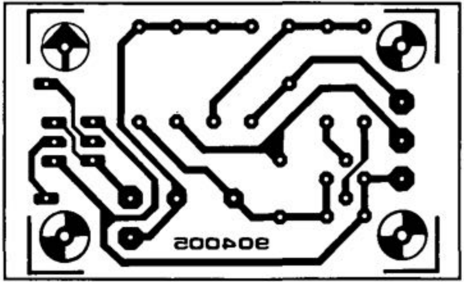
The adaptor is particularly suitable for use with fairly large power supplies, for example, 40 V, 5 A types. The fourfold reduction in the unregulated input voltage results in a decrease in dissipation from 200 W to 50 W.

The adaptor draws a current of 5 mA, excluding that required by the relay(s). ■

(K. Walters)



R2
R4
R3
R1
C1
R5
R6
+
C
B
E
T1
T2
C2
S2
S1
-
Bz1

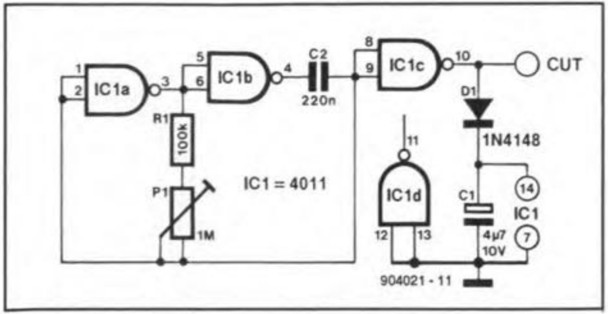


904005

REPEATING "FIRE" BUTTON

023

A joystick is an indispensable aid with most computer games. Apart from the stick that enables movement to the left, the right, upward and downward, the unit also contains one or two "fire" buttons. In some games, there is a lot of firing to be done, and this means continuous pressing and releasing of the fire button.



To prevent your getting a "tennis thumb" or "tennis finger", it is advisable to use the circuit presented here, which provides automatic actuation of the "fire" button.

Two gates of a Type 4011 chip, IC1a and IC1b, form an astable multivibrator, whose frequency is varied with P1. The generated square wave pulses are applied to the "fire" button via gate IC1c, which is connected as a buffer.

Power for the circuit is derived from the computer via D1 and C1.

The circuit is small enough to be housed in the base of the joystick. ■

(*K. Nischalke*)

024 MIDI SIGNAL REDISTRIBUTION REVISITED

The "MIDI signal redistribution" unit we published in May 1987 (p. 22) has even more possibilities than we thought at the time.

If the redistribution unit is used to the full, there is a veritable mass of interconnecting cables. Any changes, and a number of cables have to be rerouted. There is, however, a way of avoiding a lot of this work, as will be described here.

Requirements are: one MIDI redistribution unit and four two-pole, four-position switches. Connected as shown in the diagram, these make it possible to interconnect four MIDI instruments in a variety of ways without a vast number of cables. As a bonus, far fewer DIN connectors are needed than for the original set-up.

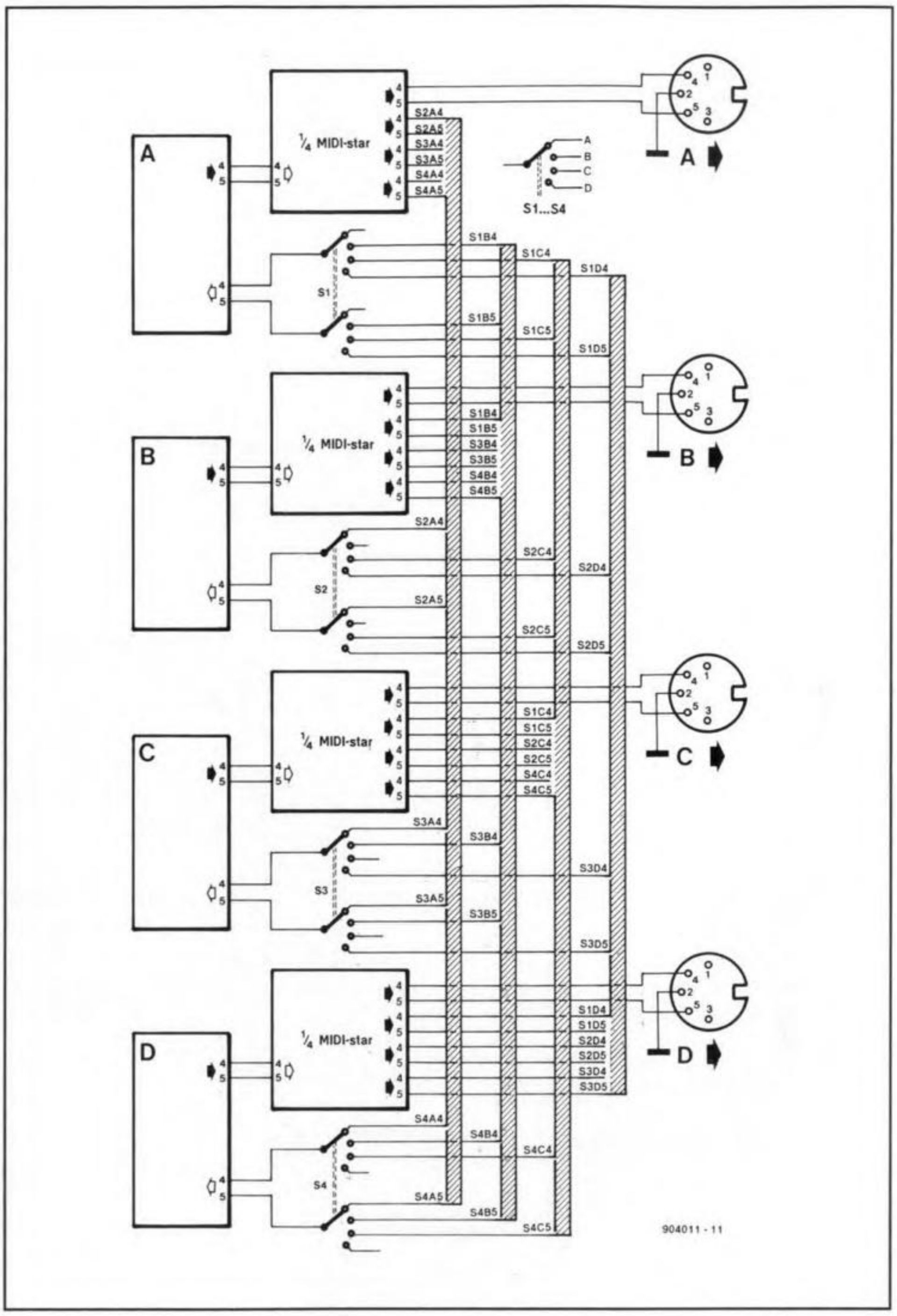
In the preset circuit, the redistribution unit is used as a four-way throughput. The original switches, S1 and S2, are not required: they are replaced by a simple wire link between contacts M and 1. A wire link is also required between '2' and '3'. The remainder of the wiring is as shown in the present diagram.

From each quarter of the unit, cable connections run to the switches in the other quarters. Since the outputs of the quarters carry the same signals, it does not matter which switch is connected to a particular output. This is the reason that the connections are shown as a 'bus', in spite of their being called '4' or '5' (that is, the pin numbers of the DIN connectors). The result is a much clearer circuit diagram.

A summary of the properties of the circuit is therefore:

- each MIDI output may be connected to the inputs of the other instruments;
- each input may be switched off;
- on each instrument, a MIDI output remains externally available. ■

(J. Blankaert)

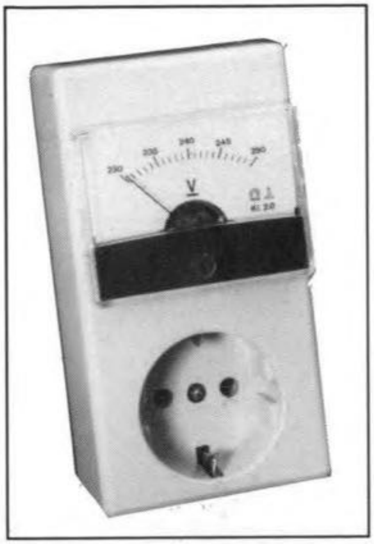


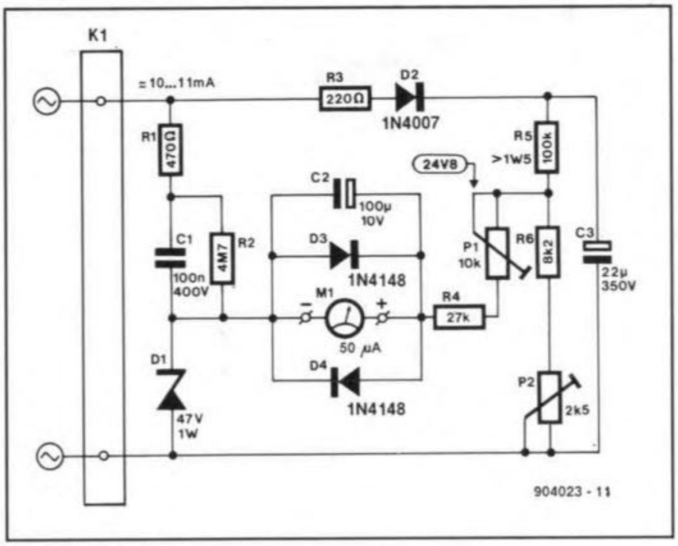
025 MAINS VOLTAGE METER

The design of this meter is in direct response to the intention of most electricity generating boards in Europe of standardizing on a mains voltage of 230 V, 50 Hz. The change from the current 220 V or 240 V (mainly UK and Eire) will be made gradually during the present decade.

The instrument provides an accurate indication of the mains voltage on an analogue meter, the scale of which ranges from 210 V to 230 V or from 230 V to 250 V, depending on the current mains voltage. This means that the current mains voltage is read at the centre of the scale. Once the mains voltage in your locality has been changed to 230 V, it will be a simple matter of adjusting the scale appropriately.

A 50 μA moving-coil meter is connected between two voltage sources. One, the reference, is divider R1-C1-R2-D1. The other, divider R5-R6-P1-P2, is variable and





WARNING

Since the circuit carries dangerous voltages at many points, great care must be taken in providing the necessary insulation. Never work on the circuit while it is connected to the mains. Observe all precautions in regard of electrical safety and make sure that no part of the circuit can be touched while it is being aligned or operated. The entire circuit, including the meter, must be housed in a suitable ABS enclosure.

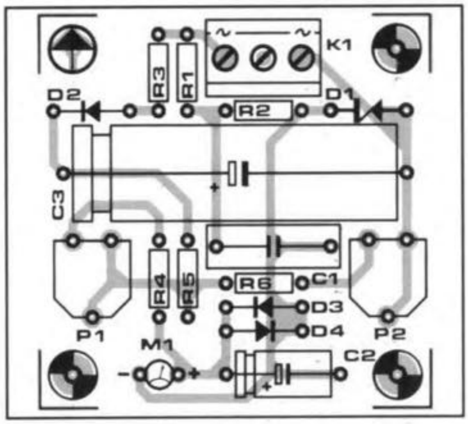
PARTS LIST

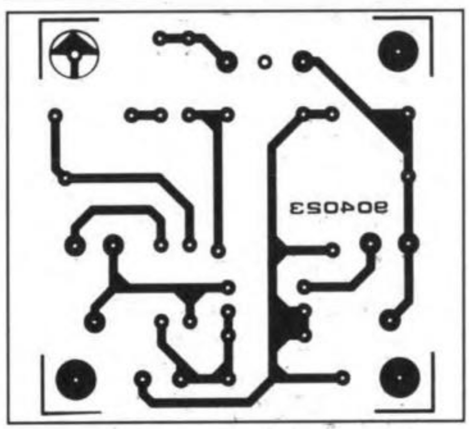
Resistors:
R1 = 470 Ω
R2 = 4M7
R3 = 220 Ω
R4 = 27 k
R5 = 100 k, 1.5 W
R6 = 8k2
P1 = 10 k preset
P2 = 2k5 preset

Capacitors:
C1 = 100 nF, 400 V
C2 = 100 µF, 10 V
C3 = 22 µF, 350 V

Semiconductors:
D1 = zener diode, 47 V, 1 W
D2 = 1N4007
D3, D4 = 1N4148

Miscellaneous:
M1 = 50 µA moving-coil meter
K1 = 3-way PCB terminal block
Enclosure, ABS, e.g., Bopla SE432DE





is used for calibration.

Resistor R1 protects zener diode D1 by limiting the charging current of C1 to a safe value when the circuit is first connected to the mains.

Both voltage sources provide half-wave rectification of the mains voltage.

The value of capacitor C2 determines the meter response to relatively fast variations of the mains voltage. Its value may lie between 20 µF and 220 µF. Since the reference voltage is not a direct voltage, a change in the value of C2 may require the circuit to be readjusted.

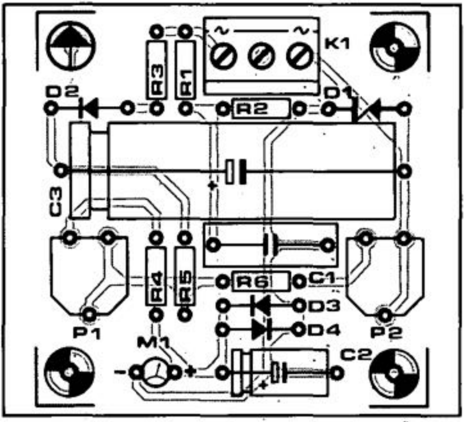
The circuit may also have to be readjusted after a time to compensate for the drift caused by the heat generated by D1 and R5. Since R5 is located quite close to R4 on the PCB, it may in some cases be necessary to use a 27k4 metal film typresistor in the R4 position.

Connect a variable transformer to the input of the circuit and set its output to the minimum expected mains voltage, that is, 210 V or 230 V (UK and Eire). Adjust P2 until the meter reads 0 µA. Next, increase the transformer output to the maximum expected mains voltage, that is, 230 V or 250 V (UK and Eire). Adjust P1 until the meter reads 50 µA.

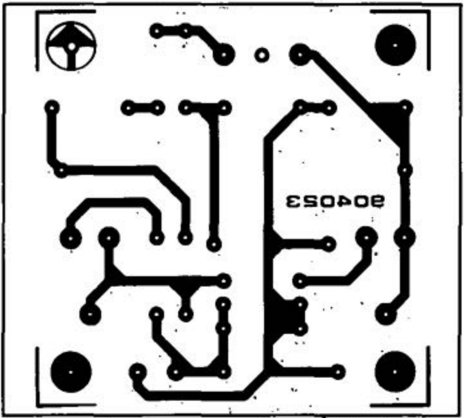
The circuit may be aligned without the use of a variable mains transformer by connecting the primary of a mains transformer with an open-circuit secondary voltage of about 10 V to the mains, and the secondary winding in series with the mains. When the voltages across the secondary and primary windings are in phase, a total voltage of 230 V or 250 V (UK and Eire) is obtained. When the connections of the secondary winding are reversed, 210 V or 230 V is obtained. In both cases, either P1 or P2 should be adjusted as described when a variable mains transformer is used.

It is recommended to provide the meter with indication marks at, say, 10 V increments. ■

(T. Giffard)



K1
R3
R1
D2
R2
D1
C3
R4
R5
R6
C1
D3
D4
P1
P2
M1
C2



904023

VARIABLE DIVIDER

026

The divider is based on a dual hex counter Type 74HCT393 and two 4-bit comparators Type 74HCT85. The division factor is set between 2 and 256 with two hex thumb-wheel switches.

The signal to be divided is aapplied to the clock input of the first counter, IC2a via IC1a. The outputs of the counter are compared by IC4 with the setting of the first thumb-wheel switch. When the data at the Q inputs and P inputs are identical, the = output of IC4 becomes logic high, provided that the = input (pin 3) is high, which here is always the case.

The = output (pin 6) of IC4 is linked to the = input of IC5, so that the comparator output of this IC can be high only when the data at the inputs of IC4 are identical.

The input signal is combined via the OR gates in IC3 with the outputs of IC2a in a way that arranges for the second counter, IC2b, to receive a clock pulse when IC2a has counted to 16 (and has thus returned to zero).

Circuit IC5 compares the content of IC2b with the position of the second thumb-wheel switch. Only when both comparators have identical inputs, that is, when the position of the counters corresponds to the setting of the two thumb-wheel switches, does a '1' appear at pin 6 of IC5.

The counters are then reset via IC1b and IC1c, after which the counting and comparing can start afresh.

When pin 6 of IC5 is high, the output of the divider is a positive pulse whose width is equal to half the period of the input signal, irrespective of the set divide factor. Therefore, the output signal is not symmetrical. If symmetry is required, a D-type bistable should be added at the output. The divider output then serves as the clock for the bistable ($\overline{Q}$ output linked to the D input). Remember, however, that the bistable divides the signal by 2.

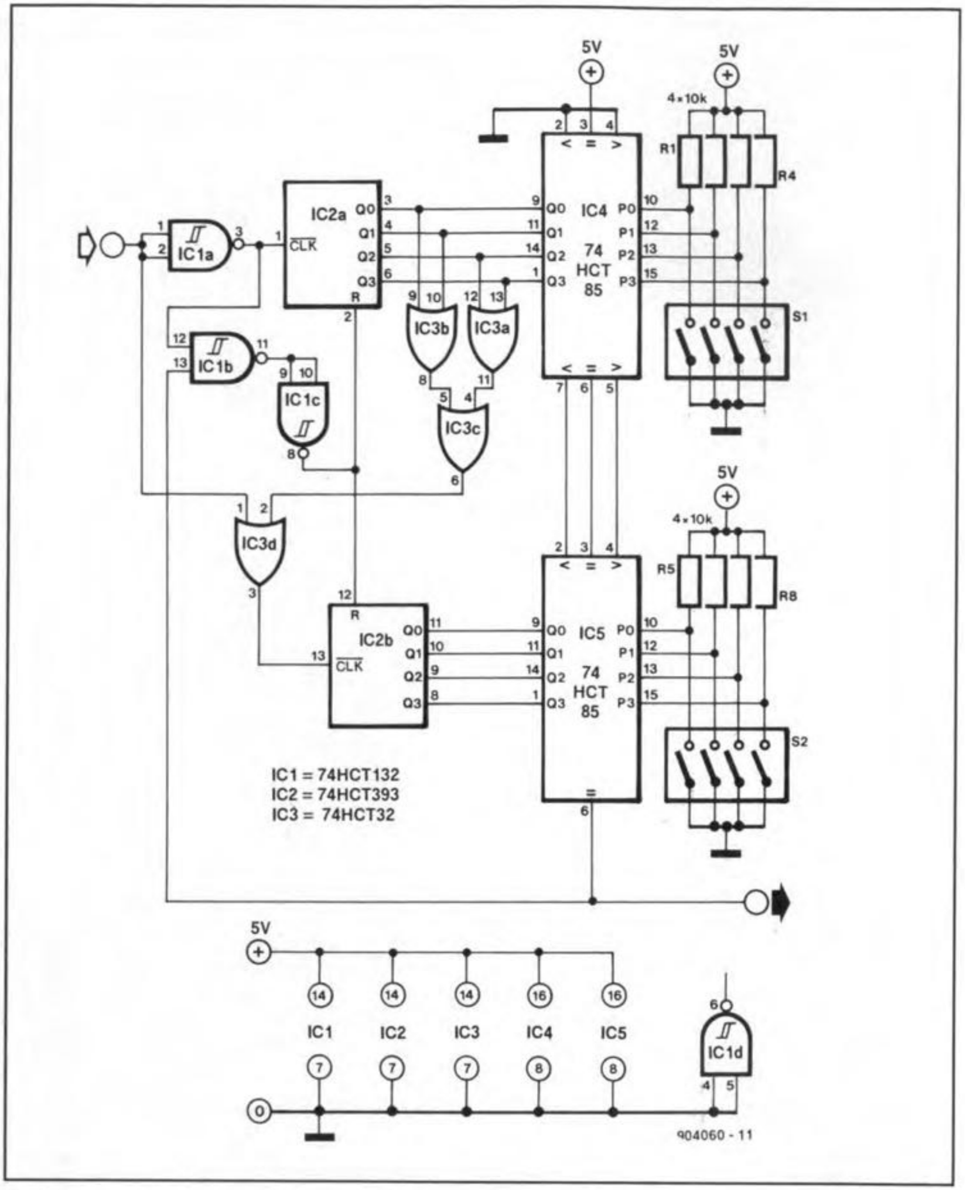
When, after the counters have been reset, the input signal becomes low, it would be expected that the second counter is immediately set to '1'. This is prevented, however, by the still active reset section .

The highest counter position is 256 and that is reached when both thumb-wheel switches are set to 0. Both counters then go their entire range.

It is possible to extend the divider by adding one or more stages consisting of a counter, comparator and thumb-wheel switch, and connecting this (them) to the preceding counter via the four OR gates. The reset signal must, of course, always emanate from the last stage.

The thumb-wheel switches may be replaced by standard DIL switches. ■

(H. Smits)



027 BEDSIDE LIGHT WITH AUTO OFF

This simple bedside light is suitable for telling the time at night, findings one's way to the door, and so on.

It is based on the well-known CMOS Type 7555 timer, connected as a monostable, whose time constant, τ, with values as shown gives a delay of about 5 minutes before the light goes out. Different delays may be calculated from

$\tau = 0.69\ R1C1$ [seconds]

where R1 is in ohms and C1 in farads.

Resistors R2 and R3 serve as pull-ups on the trigger and reset pins respectively.

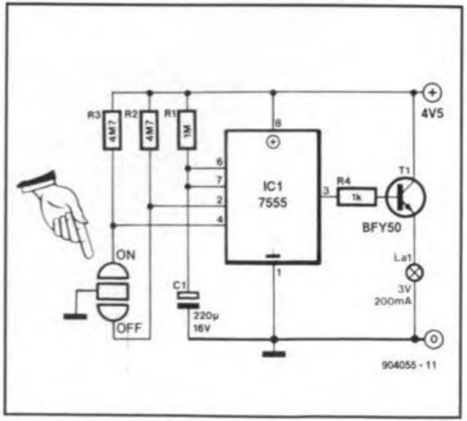
Touch pads are used to trigger the monostable or to reset it before the time delay has lapsed.

The output of the timer drives T1 via R4. The transistor can switch up to 250 mA without needing a heat sink.

Depending on the application, higher or lower wattage bulbs may be used, but bear in mind the cost of the batteries!

Standby current of the prototype was 35 µA at a supply voltage of 4.5 V. ■

(*R.G. Evans*)



THRIFTY FLASHING LED

028

The adjustable flashing rate and low off-state current drain make this circuit eminently suitable for use as an on-off indicator where battery power is at a premium, or as a fake car alarm.

Transistors T1 and T2 form a relaxation oscillator whose frequency may be set between 1 Hz and 10 Hz by P1.

The LED has an on-time of about 5 ms, so that it may draw a relatively high current to produce intense flashes.

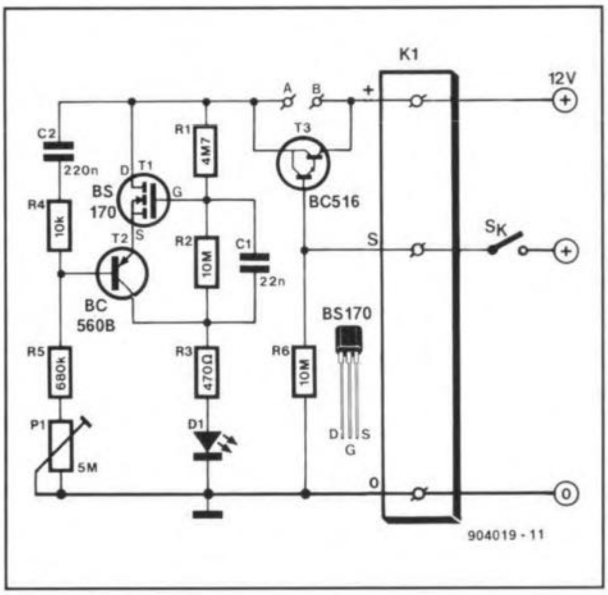
Because of the low duty factor, 0.005 at 1 Hz and 0.05 at 10 Hz, the average current drawn by the circuit is between 0.1 mA and 1 mA at a supply voltage of 12 V. This compares favourably with special high efficiency flashing LEDs that typically require an average current of 2 mA, depending on their series resistor.

The circuit may be operated from supply voltages between 6 V and 25 V.

Resistor R6 and transistor T3 should be omitted, and a wire link fitted between 'A' and 'B', if the circuit is used as an 'on' indicator.

In the fake car alarm application, T3 is switched off when the contact key, Sk, is closed. The oscillator then draws a current of not more than 2 μA and the LED does not flash. When Sk is opened, T3 draws base current via R6, starts to conduct and actuates the oscillator. ■

(J. Ruffell)



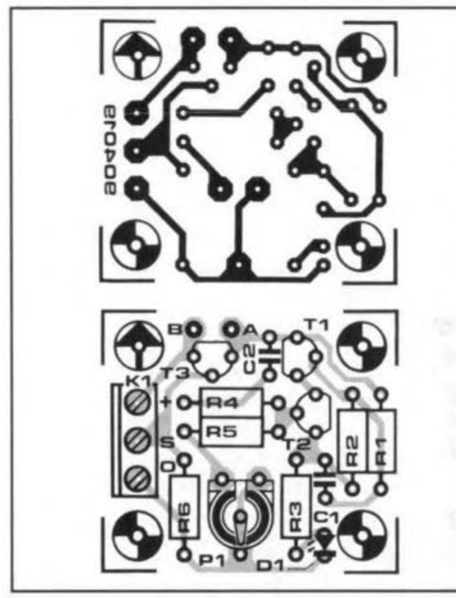
PARTS LIST

Resistors:
R1 = 4M7
R2 = 10 M
R3 = 470 Ω
R4 = 10 k
R5 = 680 k
P1 = 5 M, preset pot meter

Capacitors:
C1 = 22 n
C2 = 220 n

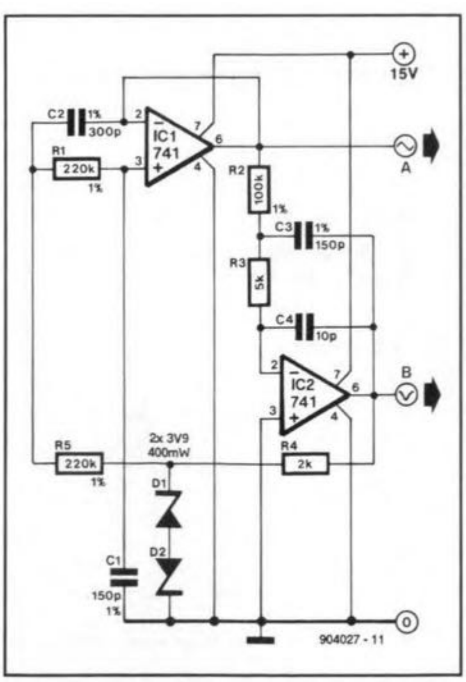
Semiconductors
D1 = LED, 5 mm, red
T1 = BS170
T2 = BC560B
T3 = BC516

Miscellaneous:K1 = 3-way PCB terminal block



STABLE SINE WAVE OSCILLATOR

029



Sine wave oscillators normally generate near-perfect sine waves, but the stability of their output signal is often not very good. The circuit presented here is aimed at improving the stability.

The improvement is brought about by limiting the fed-back output signal by two series-connected zener diodes, D1 and D2.

The oscillator proper consists of two sections: IC1, R1, R5, C1 and C2 form a second-order low-pass filter, while IC2 is connected as an integrator.

The sinusoidal output signal of IC2 becomes trapezoidal after the limiting action and is then applied to the low-pass section. This means that the amplitude of the fedback signal is constant, so that the peak value of the sine wave output of the oscillator is also constant.

The circuit provides two sine wave outputs at A and B respectively that are mutually 90° out of phase. The waveform at output B is slightly purer than that at A.

The third harmonic is about 40 dB down on the fundamental.

With values as shown, the circuit generates a signal at a frequency of 3.3 kHz with a peak-to-peak value of 11 V. The frequency may be altered by changing the values of C1, C2 and C3 proportionally.

The circuit draws a current of 3 mA at a supply voltage of 15 V. ■

(National Semiconductor application)

PARTS LIST

Resistors:

R1 = 4M7
R2 = 10 M
R3 = 470 Ω
R4 = 10 k
R5 = 680 k
P1 = 5 M, preset pot meter

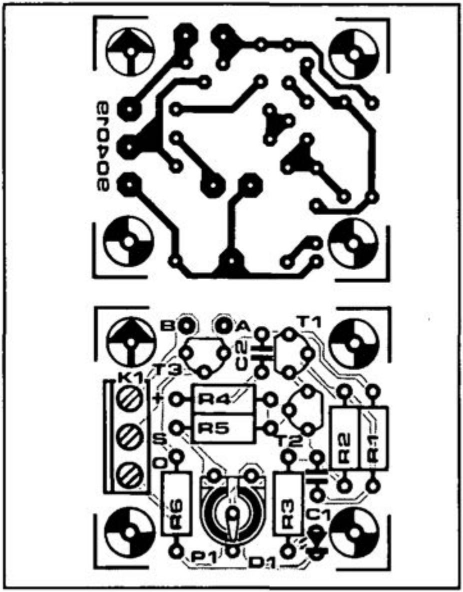
Capacitors:

C1 = 22 n
C2 = 220 n

Semiconductors

D1 = LED, 5 mm, red
T1 = BS170
T2 = BC560B
T3 = BC516

Miscellaneous:K1 = 3-way PCB terminal block

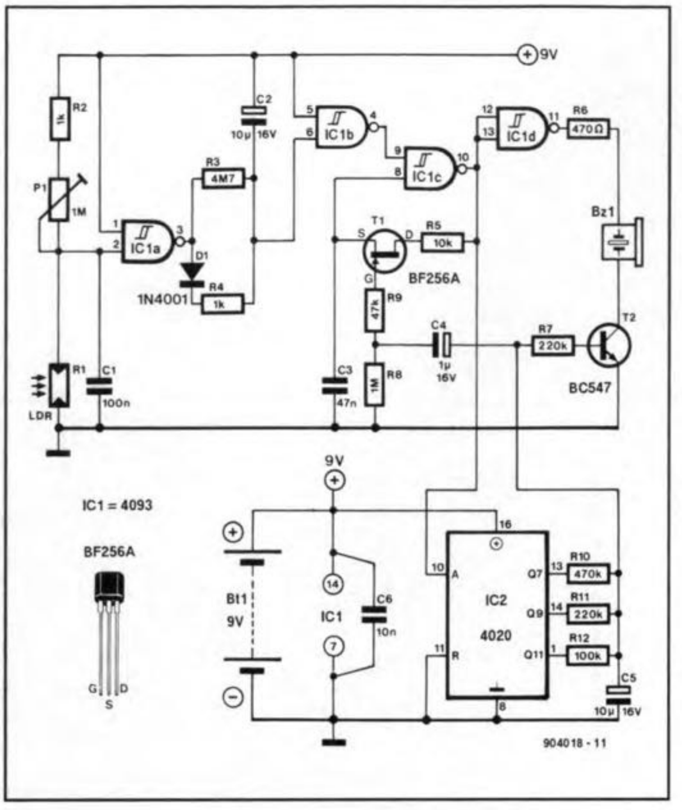


030 ELECTRONIC MOSQUITO

This circuit behaves like a real mosquito: as soon as it gets dark, it begins to buzz irritatingly. As soon as it gets light, it becomes mute once more, so that, again like a real mosquito, it is very difficult to pinpoint.

The potential at input 2 of IC1a determines whether the mosquito hums or not. The level of this voltage depends on the resistance of photovaristor R1 and the setting of P1. If the level is higher than the trigger threshold, pin 3 goes low, which causes C3 to charge via R3.

It takes about 90 seconds for C3 to become fully charged and during this time pin 6 of IC1b is high. As soon as the voltage at this pin drops below the lower threshold of IC1b, the output (pin 4) of the gate goes high. This results in IC1c commencing to oscillate. A square-wave signal is then produced at the output of the oscillator, which is buffered by IC1d and then applied to the buzzer. The buzzer operates as long as T2 is switched on.



The output of the oscillator is used also as a clock for 14-bit binary counter IC2. The consequent signals at outputs Q7, Q9 and Q11 of this circuit are added together and used to drive T2. The whining tone is therefore not continuous but totally random. This random signal is used also by T1 to influence the output of IC1c to a small extent.

As soon as it gets light, the potential at pin 2 of IC1a drops, which results in C2 discharging rapidly via D1 and R4, so that the oscillator ceases at once.

The circuit draws a current of only 2–5 mA, so that a 9-V battery will last for quite some time.

(J. Beckers)

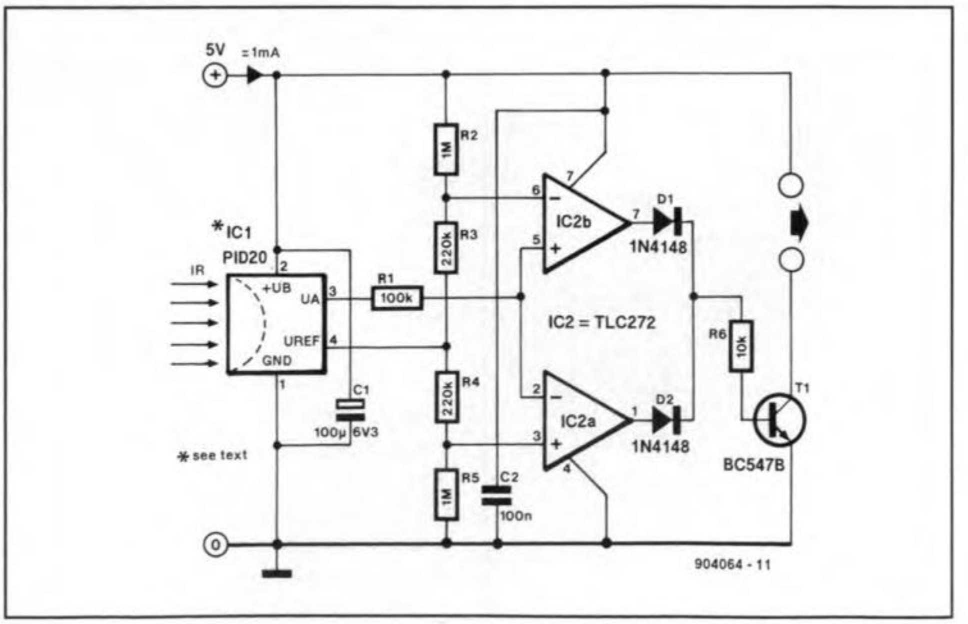
031 INFRA-RED DETECTOR II

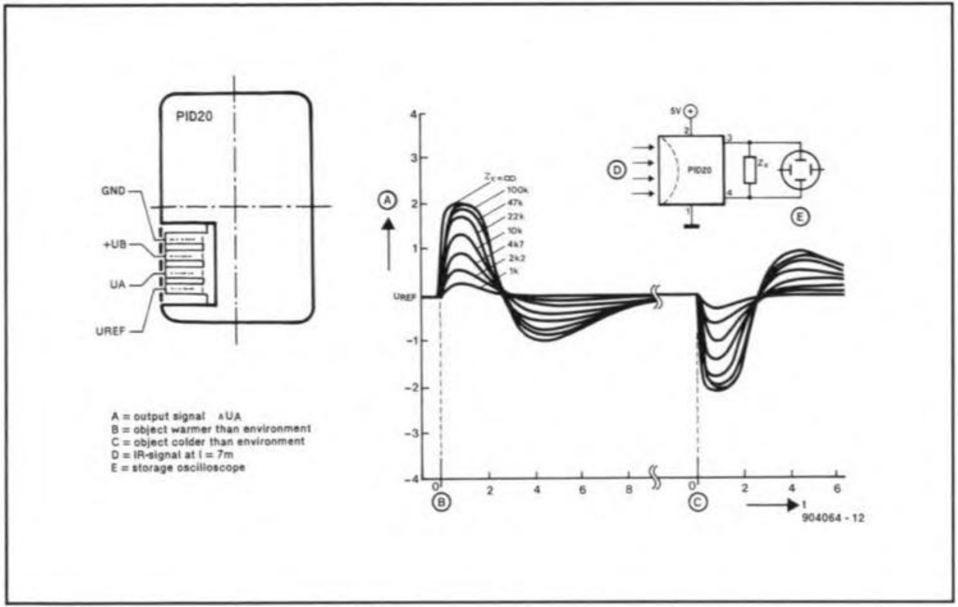
Siemens' Type PID20 is a passive infra-red detector that transforms heat radiation into electrical pulses. Adding two opamps and some miscellaneous components to the device is sufficient for the construction of a efficient infra-red (IR) detector.

The magnitude of the output signal of the PID20 is determined by the load at its pins 3 and 4 as illustrated in Fig. 2 that shows the output at loads from 1 kΩ to ∞. From the curves, it is clear that a reasonably high load is needed for a good output signal. In the circuit, both output pins are loaded by 100 kΩ (R1 at pin 3 and R3 shunted by R4 at pin 4).

The output signal at pin 3 is compared with a reference voltage equal to half the supply voltage. This reference voltage is derived from potential divider R2-R3-R4-R5. The output increases at the approach of an object that is warmer than the surroundings or at the removal of an object colder than the surroundings. Note the ringing in Fig. 2 after a change in surroundings has taken place. The sudden appearance of a warm object causes the output voltage first to rise and, after a few seconds, to drop, sometimes even below the reference voltage level. This behaviour needs to be taken careful account of in practice.

The change in voltage at the output of





the sensor is compared by IC2a and IC2b with voltages respectively 0.5 V below and 0.5 V above the reference voltage. Depending on the output level, one of the comparators toggles and switches on T1. This transistor may be used, for instance, to drive a relay (up to 100 mA). It is also possible to connect a transistor (and diode) to each opamp, so that it is immediately seen whether a source has caused an increase or decrease in heat.

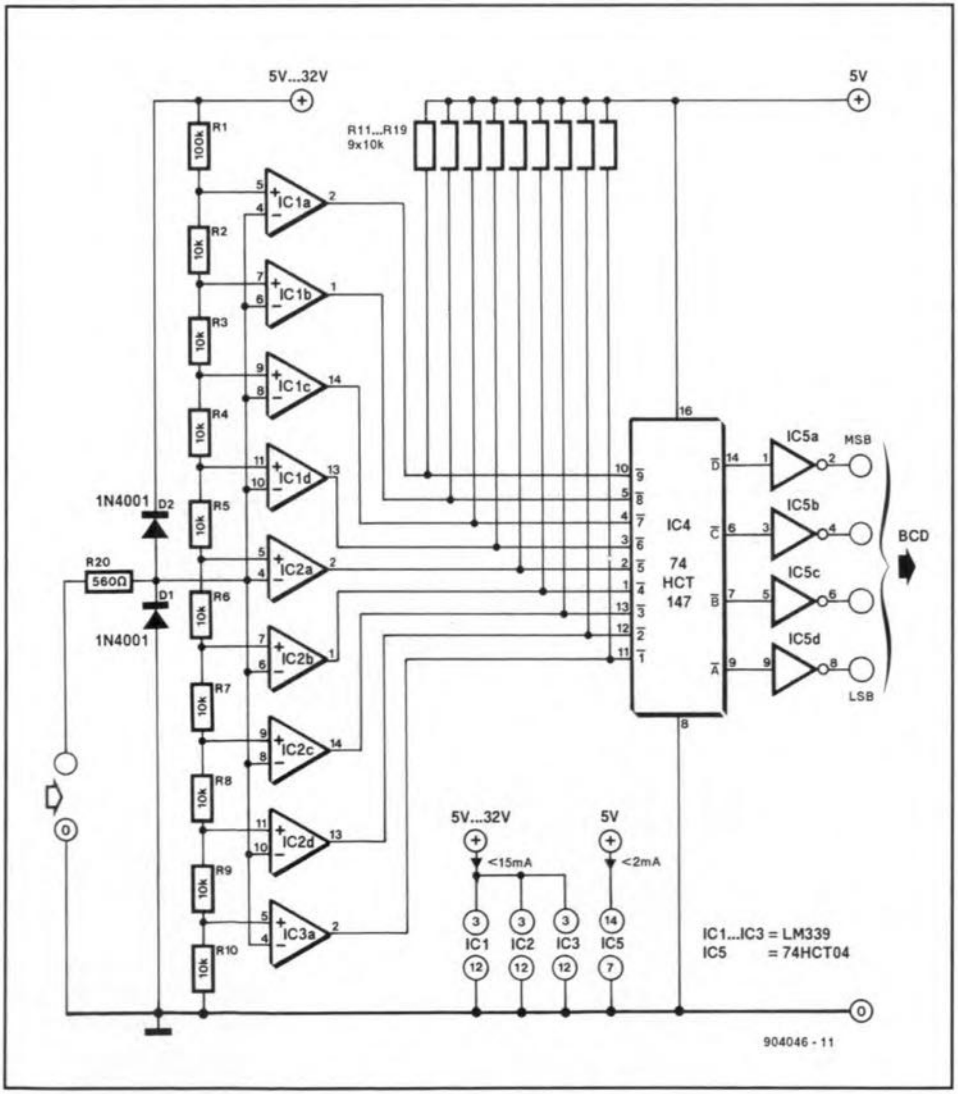
As shown, the circuit draws a current of about 1 mA (of which roughly 0.2 mA is on account of the sensor). The current drain may be reduced by replacing the TLC272 by a TLC27L2.

The sensor should be fitted by means of a special connector (ask your dealer). ■

(*Siemens application*)

4-BIT BCD FLASH CONVERTER 032

There are many ways of designing an analogue-to-digital converter: the present design is aimed particularly at a high speed. The speed is determined primarily by the reaction time of comparators (here about 1 μs). The design has a drawback: the number of components is direcly proportional to the number of levels that the converter can cope with. In the present design, this is limited to ten (if none of the comparators detects the exceeding of a level, the input is below the lowest level, and this is recognized as zero [10th level]).



The open-collector outputs of the comparators are connected to a priority detector, IC4, which is a kind of decimal-to-BCD decoder. If, however, several inputs are low at the same time, priority is given to that BCD code that corresponds to the input with the highest number. Four NOT gates invert the BCD code generated by IC4.

Because the comparators have open-collector outputs, they can work from a higher supply voltage than IC4 and IC5. Pull-up resistors R11–R19 must, therefore, be connected to the +5 V rail and NOT to the comparator supply rail.

The voltage levels with which IC1-IC3 compare the input level are provided by potential divider R2–R10.

The input range of the converter may be modified by changing the value of R1. It should be noted, however, that the upper threshold, that is, the level at pin 5 of IC1, must always be at least 2 V lower than the supply voltage. With values as shown, and a supply of 12 V, the upper threshold is 5.68 V and each step is 632 mV. ■

(*P. Coster*)

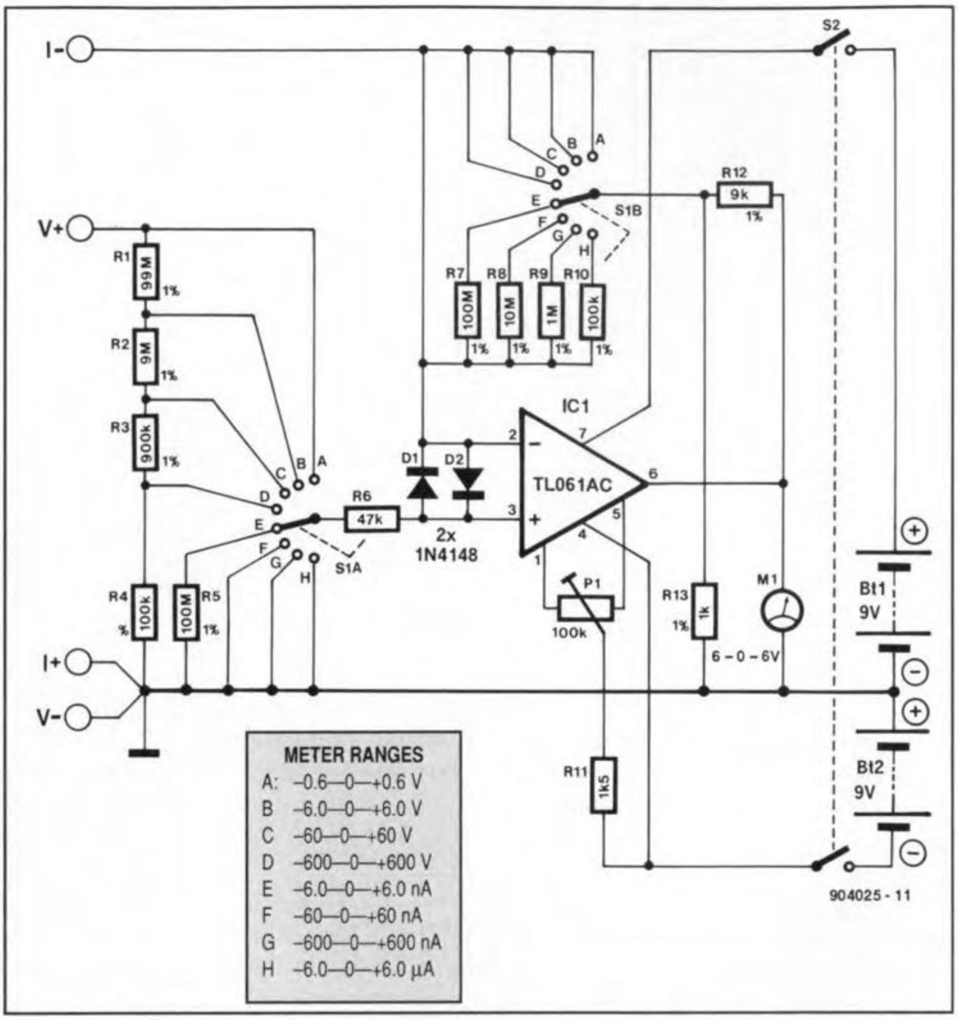
033 HIGH-RESISTANCE MULTIMETER

Although the multimeter whose circuit is shown here is fairly simple to make, you may experience difficulties in obtaining some of the components, since most retailers do not stock 99 MΩ or 100 MΩ resistors. It then becomes matter of making up these resistors or omit the associated (current) ranges.

The use of these high-resistance components is made possible by the use of a Type TL061 opamp, which has very high-resistance inputs.

For the voltage ranges, the opamp is connected as a non-inverting amplifier with an ampliciation of 1+R12/R13=10. If the meter has a full-scale deflection (f.s.d.) of 6 V, the input sensitivity is 600 mV. Input attenuator R1–R4 provides additional ranges up to 600 V.

When the meter is used to measure current, the unknown current will flow through one of the resistors R7–R10, depending on the selected range. Since the value of these resistors is much higher than that of R13, the output voltage is equal to 10× the potential drop caused by the unknown current across R7–R10. With the values as shown, this gives current ranges of 6 nA, 60 nA, 600 nA and 6 µA. These ranges are fairly small, but that is because the opamp can not deliver more current and also, in order to keep the circuit simple, R7–R10 must be much larger than R13. Note that these resistors have nothing to do with the input resistance of the instrument. The I+ terminal is at earth potential and the I terminal is a virtual-earth point: that results in a very low-resistance input (as should be the case).



Apart from the unknown current, the input offset current of the opamp also flows through R7–R10. To eliminate the effect of this current (some nA) in the most sensitive range (R7), resistor R5 has been added.

Since a centre-zero meter and symmetrical supply were used, the polarity of the input signal does not matter. The meter is set to zero with P1.

Resistor R6 and diodes D1 and D2 serve to protect the opamp against too high input voltages.

The supply is formed by two 9-V batteries that, since the current drain is less than 1 mA, will last a long time. ■

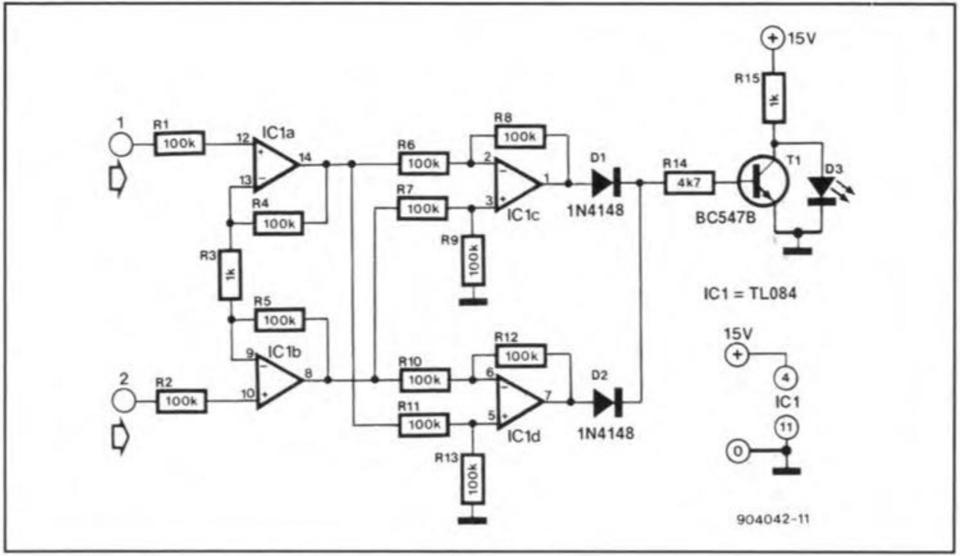
(*Texas Instruments application*)

034 SUPPLY VOLTAGE MONITOR

In much equipment, such as computers, amplifiers, and the like, where high tolerance components are used, it is important to know at all times whether the supply voltage is within specification. The circuit presented here was designed to show just that. It does so by comparing two d.c. voltages. As soon as there is a difference of more than 10 mV between these, the indicator LED goes out to signal that something is amiss. It is, therefore, necessary to use a very accurate reference voltage, corresponding to the required voltage level.

Circuits IC1a and IC1b form a difference amplifier: the voltage to be monitored is applied to input 1 and the reference voltage to input 2. Differences between these two potentials as small as 10 mV are sufficient to cause one of the two comparators, IC1c or IC1d, to toggle.

The output of IC1d goes high when the voltage level at input 1 is 10 mV or more greater than that at input 2. Comparator IC1c toggles when the voltage level at input 1 is 10 mV or more lower than that at input 2. The indicator LED is switched off via diode D1 or D2, as the case may be,



and transistor T1.

The level of the supply voltage to the circuit is not important: it may be ±10 V or, as here, +15 V. The important thing is that the input voltages are always at least 1.5 V lower than the supply voltage. Thus, if a ±10 V supply is used, the input voltages must lie between –8.5 V and +8.5 V. This is to prevent the common-mode suppression becoming too small for the opamps to handle the signals.

The 100 kΩ resistors must have a tolerance of not more than 1%, again to prevent the common-mode suppression of the opamps from being degraded..

The current drain of the circuit is 25 mA for a supply voltage of 15 V. ■

(*K. Walters*)

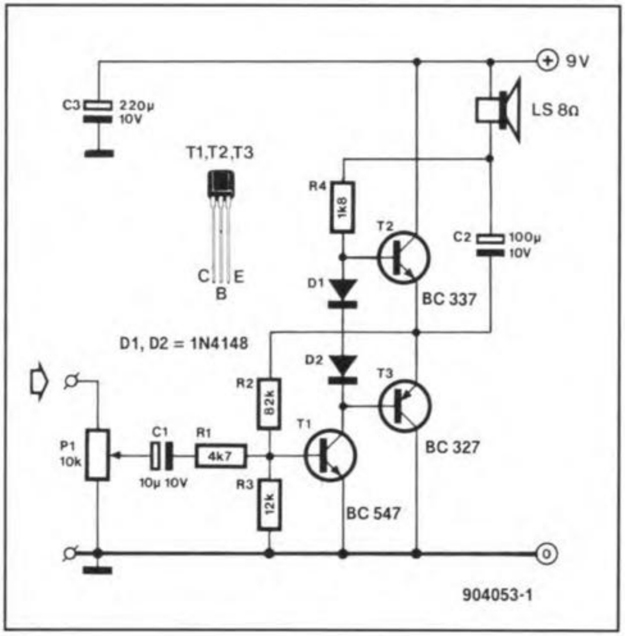
MINIATURE AF AMPLIFIER

035

The tiny a.f. power amplifier presented here provides an output of up to 250 mW and may be used in a number of applications, for instance, in a stereo version as a booster for personal radios.

The design is straightforward: a BC547 transistor drives a balanced power amplifier consisting of a BC337 and a BC327.

The quiescent current is arranged by diodes D1 and D2. Owing to the simplicity of the circuit, the quiescent current varies with temperature. This drawback is particularly noticeable when the output transistors get much hotter than the diodes. In that case, either the output power has to be reduced or the output transistors must be fitted on heat sinks. Another solution is inserting 0.47 Ω resistors in the emitter circuits of the output transistors.



The amplification is determined by the values of R1 and R3 and, of course, P1. With values as shown, and dependent on the setting of P1, the amplification is about 15. This may be varied by changing the value of R1. It is not recommended to alter the values of R2 and R3, since these resistors determine the d.c. operating point of the amplifier.

The input sensitivity for a power output of 250 mW into 8 Ω and an amplification of 15 is about 95 mV. The circuit then draws a current of around 180 mA. ■

(*P. Coster*)

50/75-OHM DRIVER

036

The 50/75 Ω driver is based on a Type OP64, which was designed specifically for use in pulse and video applications. Because of internal compensation, the IC is stable at amplification factors of 5 or greater. Since the output can deliver a current of up to 80 mA, the IC can be loaded by 150 Ω (i.e., a 75 Ω system) with a supply voltage of ±15 V without limiting taking place.

The circuit in Fig. 1 is intended as a simple output amplifier in a 75 Ω system or, if the value of R1 is changed, as an impedance adaptor. With the values of R2 and R3 as shown, the amplification is 5. The bandwidth depends to a large extent on the board layout. According to the manufacturers' data, the roll-over frequency , at this level of amplification, may be as high as 20–30 MHz.

The circuit in Fig. 2 is intended to couple a 75 Ω system to a 50 Ω system. The gain in this set-up is 0.5 dB. Potential divider R4-R5 ensures an output impedance of 50 Ω and attenuates the output of IC1 (=5× input signal) to make the overall amplification unity. At the same time, R4 ensures that the load on the OP64 does not become too large in a 50 Ω system.

The OP64 is very fast: with an amplifi-

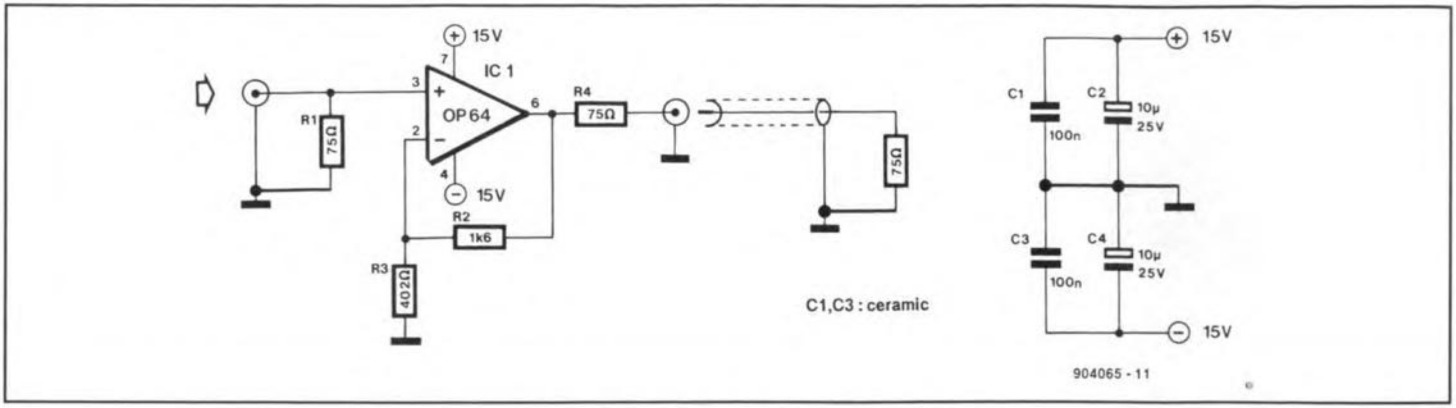


Fig. 1. Simple output amplifier or impedance adaptor.

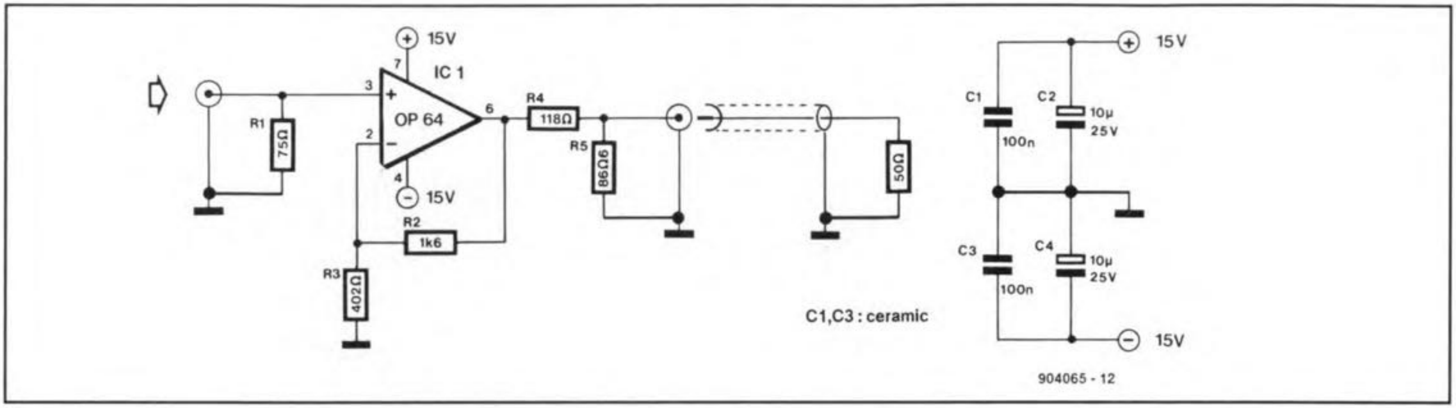


Fig. 2. Converter for coupling a 75 Ω system to a 50 Ω system.

cation of 5, the slew rate at a leading edge is about 135 V/µs and at a trailing edge, around 120 V/µs. There is then some overshoot, but that disappears when the amplification is increased to 10.

At higher frequencies, the output impedance of the OP64 increases to some extent: to about 20 Ω at an amplification of ×10. The impedance may be increased to to about 2 kΩ via the disable pin 6 (which is active low). This is a handy way of limiting the current drawn by the IC. The normal current drain is 6–6.5 mA, but in the disable condition this reduces to about 0.5 mA for the positive half and around 0.12 mA for the negative half.

According to the data sheet, the IC is protected against short circuits for about 10 seconds, so care must be taken that no lasting short circuits or overloads occur.

In a board layout of the circuit, the current through the load must not return via the input earth, but via a common earth, to which the decoupling capacitors should also be connected. ■

(*T. Giffard*)

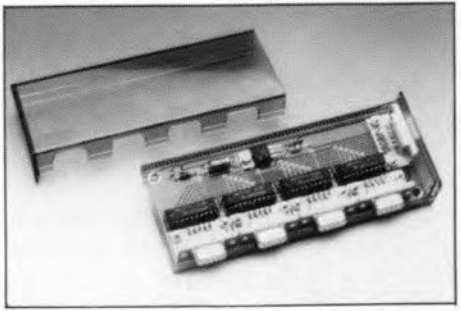
037 FOUR-MONITOR DRIVER FOR PCS

The driver described here allows up to four monitors for IBM PCs or compatibles to be driven by a single display adaptor card. This card and associated monitors must be types that operate with TTL (digital) signal levels, such as:

- CGA—colour graphics adaptor;
- Hercules card—monochrome adaptor;
- EGA—enhanced graphics adaptor.

Applications of the circuit may be found in teaching computer science, PC slide shows, presentations and demonstrations.

The circuit consists of four octal three-state buffers Type 74HC541: IC1–IC4. The TTL levels from the computer are applied to the inputs of the (permanently enabled) buffers via connector K5, which is a 9-way sub-D male type.



The signals carried by this connector depend on the display adaptor card used in the PC as shown in the accompanying

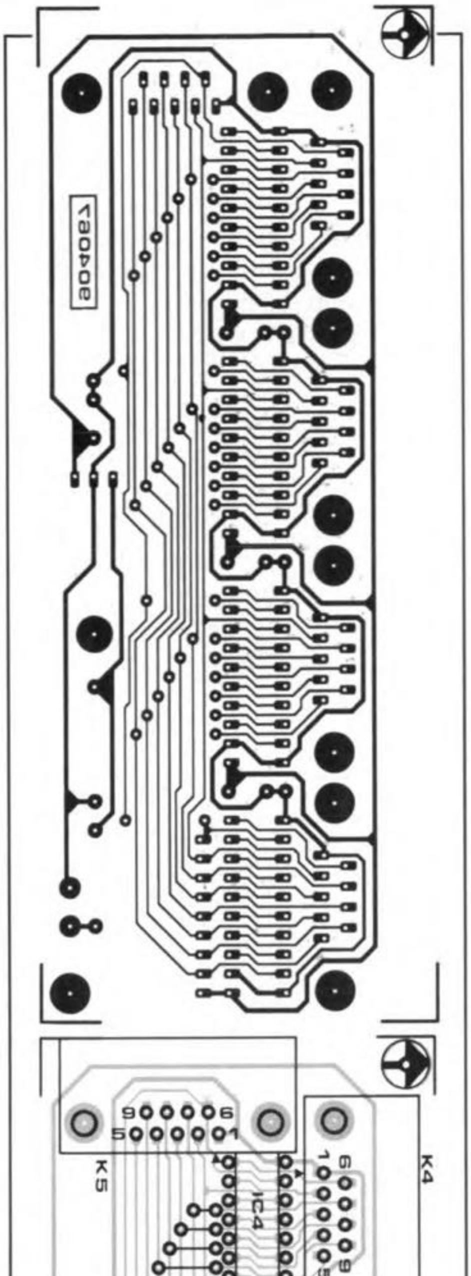
PARTS LIST

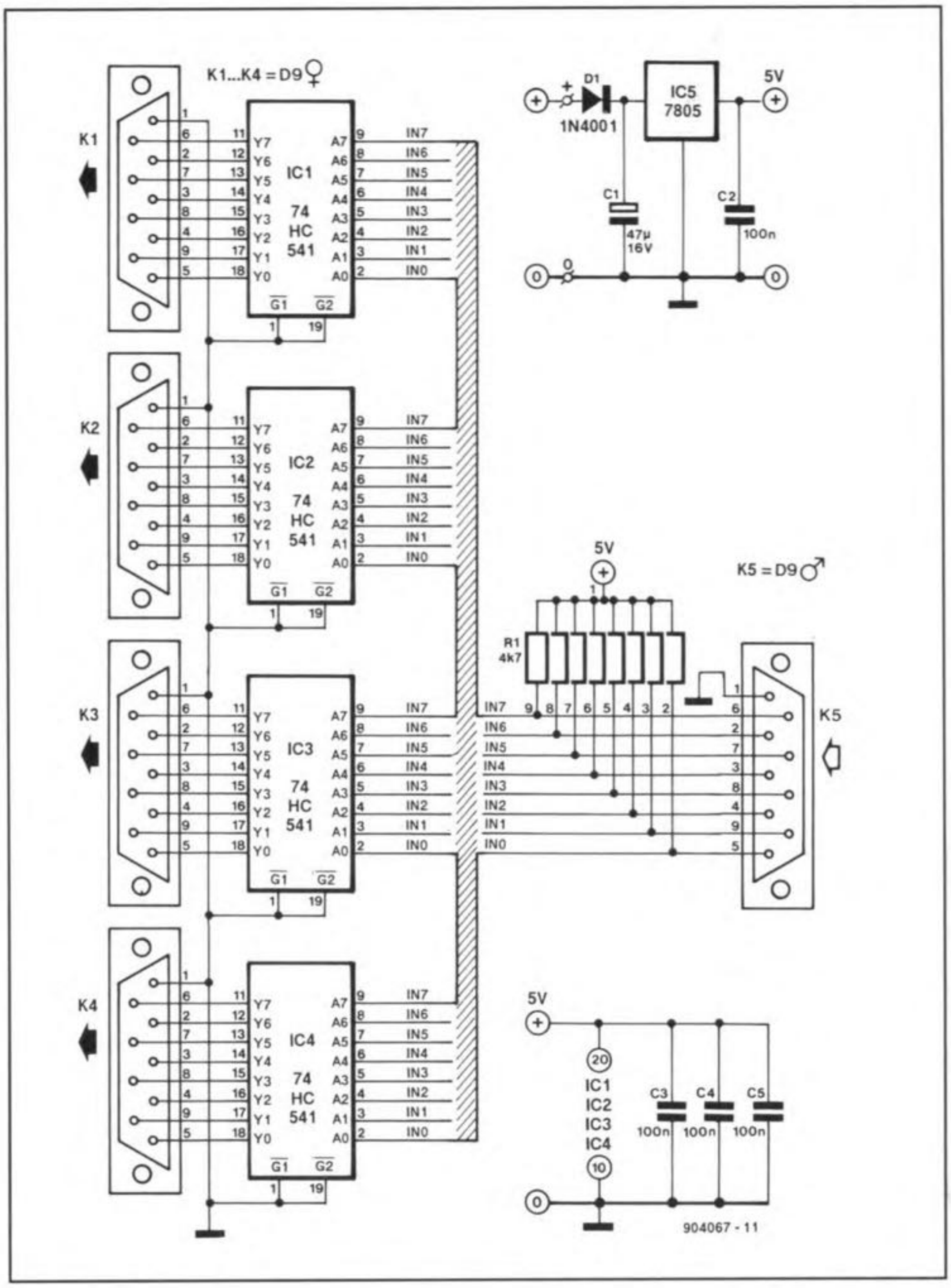
Resistors:
R1 = 8-way SIL resistor array, 4k7

Capacitors;
C1 = 47 µF, 16 V
C2–C5 = 100 n

Semiconductors:
D1 = 1N4001
IC1–IC4 = 74HC541
IC5 = 7805

Miscellaneous:
K1–K4 = 9-way female sub-D connector for PCB
K5 = 9-way male sub-D connector for PCB
Enclosure Heddic Type 222





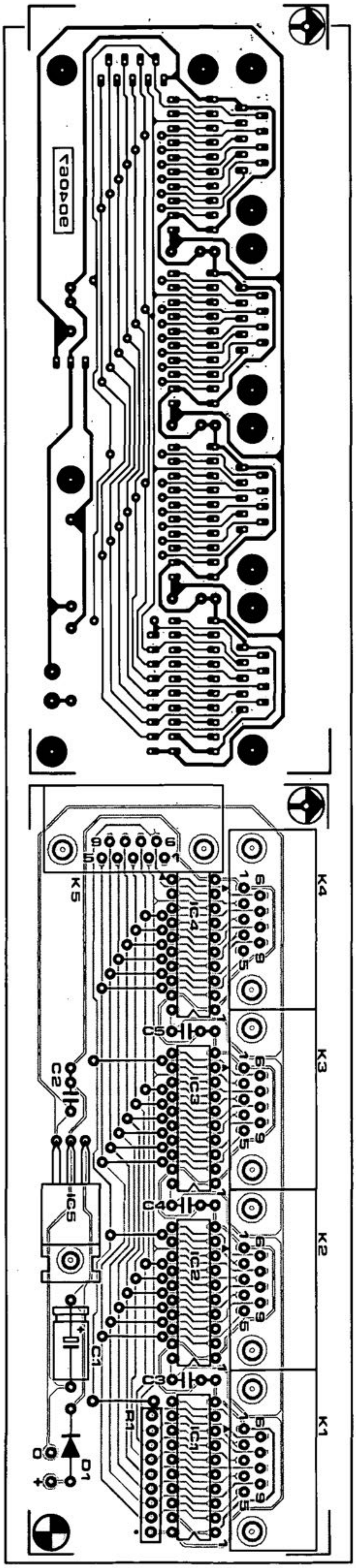
TABLE

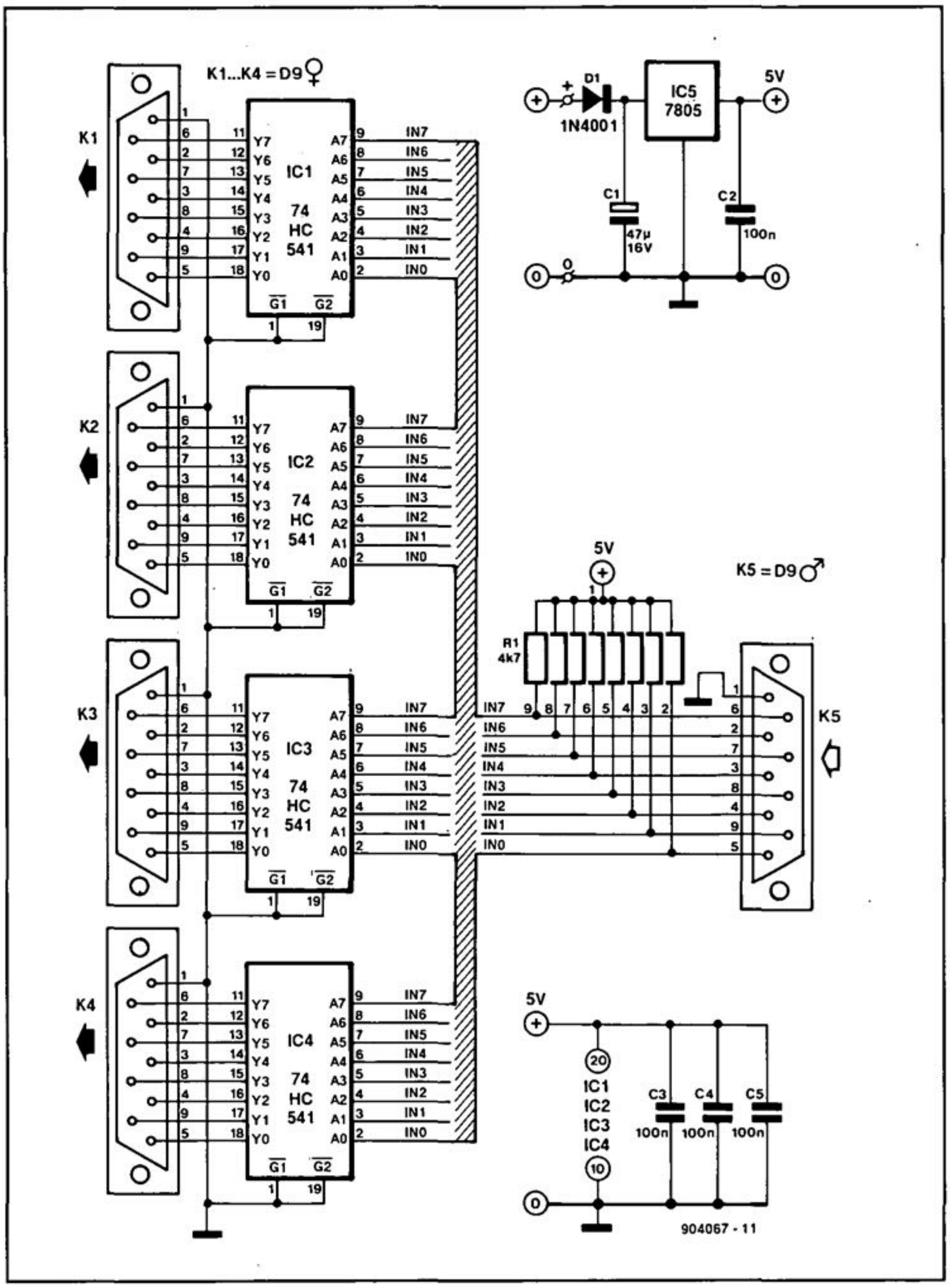
Pin	CGA	Hercules	EGA
1	ground	ground	ground
2	ground	ground	sec. red
3	red	not used	red
4	green	not used	green
5	blue	not used	blue
6	intensity	intensity	sec. green
7	not used	video	sec. blue
8	H sync.	H sync.	H sync.
9	V sync.	V sync.	V sync.

table. The monitors are connected to female 9-way sub-D connectors K1–K4.

The circuit is powered by an on-board 5 V supply based on the familiar Type 7805 voltage regulator, IC5. The unregulated input to the supply may lie between 9 V and 15 V. The total current drain is not greater than 10 mA. ■

(F. Tronchet)





TABLE			
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(F. Tronchet)

038 INVERSE RIAA NETWORK

Many people find that when they want to add a CD player to their audio system, there is no available input left on their power amplifier. It seems obvious to use the phono input, but that presents a couple of problems. This input is far too sensitive for a CD player and, moreover, it passes the signal via an RIAA frequency correction network to the amplifier. The present network can solve both problems without any internal changes to the audio system, since it is merely plugged into the phono input socket.

The circuit is not much more than a filter and an attenuator that reduces the output of the CD player (here assumed to be 200 mV) to 2 mV and has a frequency response that is exactly the opposite of that of the RIAA network. The resulting overall frequency characteristic is straight within ±1.5 dB.

It is recommended to fit the circuit (left- and right-hand channels separately) in a suitable metal enclosure, since it is susceptible to hum and noise. This is caused by the high-impedance input of about 1 MΩ at 50 Hz. The impedance decreases to 100 kΩ at 1 kHz, 10 kΩ at 10 kHz and about 1 kΩ for frequencies between 100 Hz and 500 Hz. Because of the variable input impedance, it is important that the internal impedance of the signal source is not much higher than about 2 kΩ in order to keep the –1 dB point above 30 kHz (provided the main amplifier is of reasonable to good quality).

Figure 2 shows the calculated frequency response of the circuit, which is identical to that of the recording amplifier. Figure 3 shows the overall frequency response of an (idealized) audio system with the present network connected to the phono input.

For good results, all capacitors should be polystyrene types, and metal film resistors should be used in the R1 and R2 positions. ■

(T. Giffard)

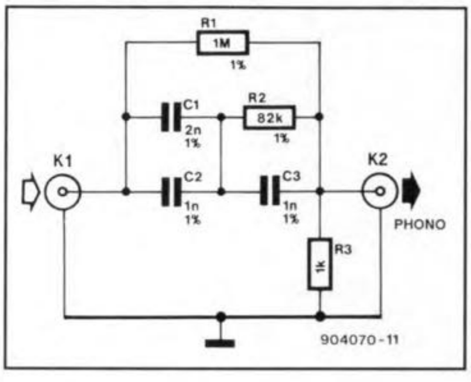


Fig. 1.

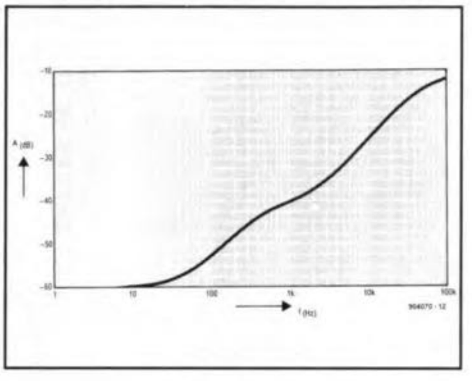


Fig.2.

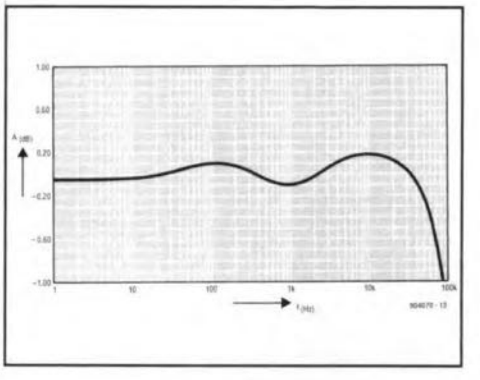


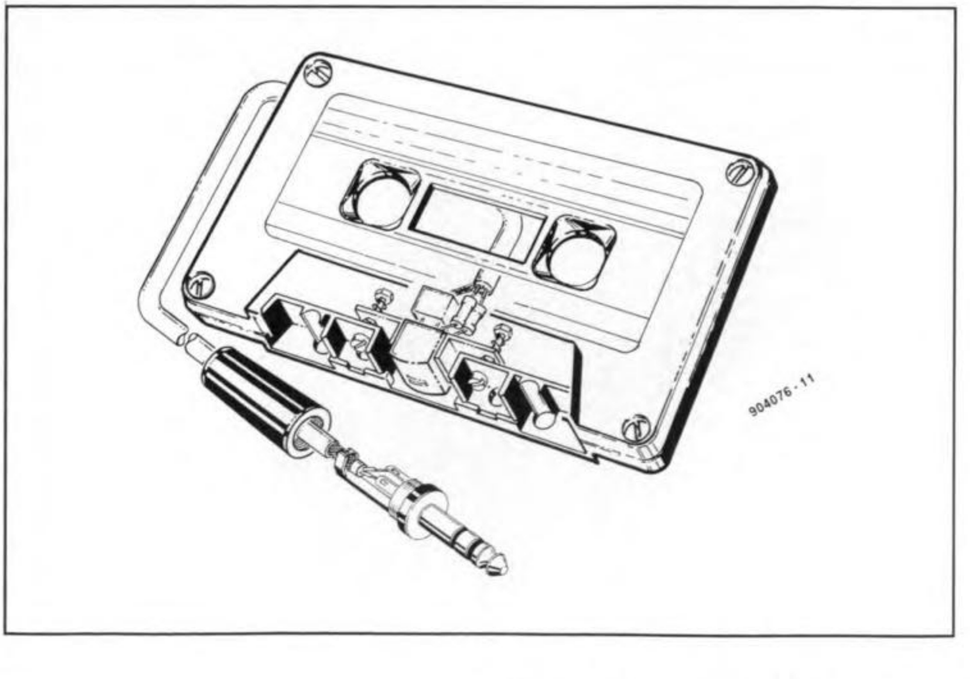
Fig. 3.

039 CD-TO-CASSETTE PLAYER ADAPTOR

With the availability of portable CD players, a need has arisen for a means of connecting such a machine to the radio-cassette player now found in most private cars. Unfortunately, few of these players have a suitable line input. However, a CD-to-cassette adaptor will solve the problem. Although such adaptors are freely available, it is fairly simple (and great fun) to make one yourself.

To build the adaptor, you need an old cassette case (screwed together, not glued), a stereo cassette recording head, a stereo 3.5 mm phono plug, some two-core screened cable, two 820 Ω resistors and two 15 nF capacitors (and some dexterity).

Remove all tape guides and fastening strips from the recording head. Cut two 20 mm long, 7 mm wide strips of thin tin plate and bend them at right angles about



5 mm from each end. Drill a 3 mm hole near one end of the long part of the bracket so formed. Solder the brackets to the sides of the recording head as shown in Fig. 1; gently fasten the head in a small vice to prevent it getting too hot.

Unscrew the cassette case and remove everything from the inside. With a small saw or sharp knife, remove the rib behind where the screening shield and pressure pad were.

Drill 2 mm holes in the remaining ribs at the same distance as the holes drilled in the tin strips.Fasten M2x7 screws in these holes and slide appropriate insulating sleeve over the protruding screw-thread. Next, slide the recording head assembly (slot at the underside) over the insulated screw-threads, fit light springs (such as those from an old ballpen) over the screw-thread and hold these in place with a nut (see Fig. 1). The springs will then push the recording head forward.

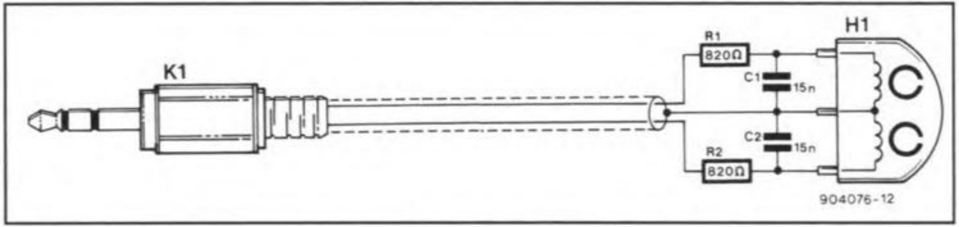


Fig. 2.

The electronics are very simple as may be seen in Fig. 2. One resistor and one capacitor form a low-pass filter for one channel that interfaces between the CD player output and the frequency-dependent recording head. These components may be soldered direct to the head. Finally, the short cable is connected to the components, taken out of the cassette case via a small hole in one of the short sides of the case and then connected to the stereo phono plug.

It pays to experiment with the setting of the volume control on the CD player to find which position gives the best reproduction. ■

(J. Ruffell)

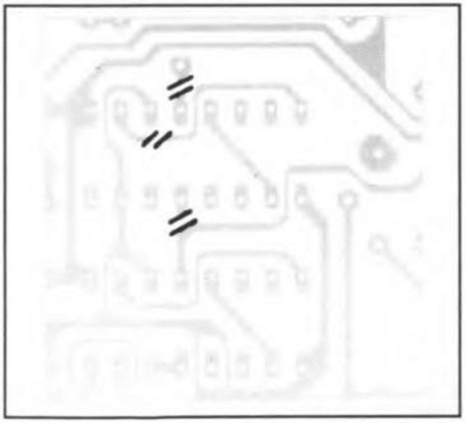
10 MHZ REFERENCE FROM FRANCE-INTER

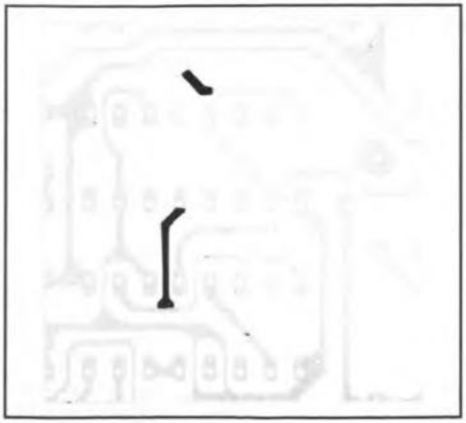
040

COMPONENT CHANGES

Capacitors:
C6, C9 = 2n7
C7, C10 = 15 n

Inductors:
L2, L3 = 10 mH



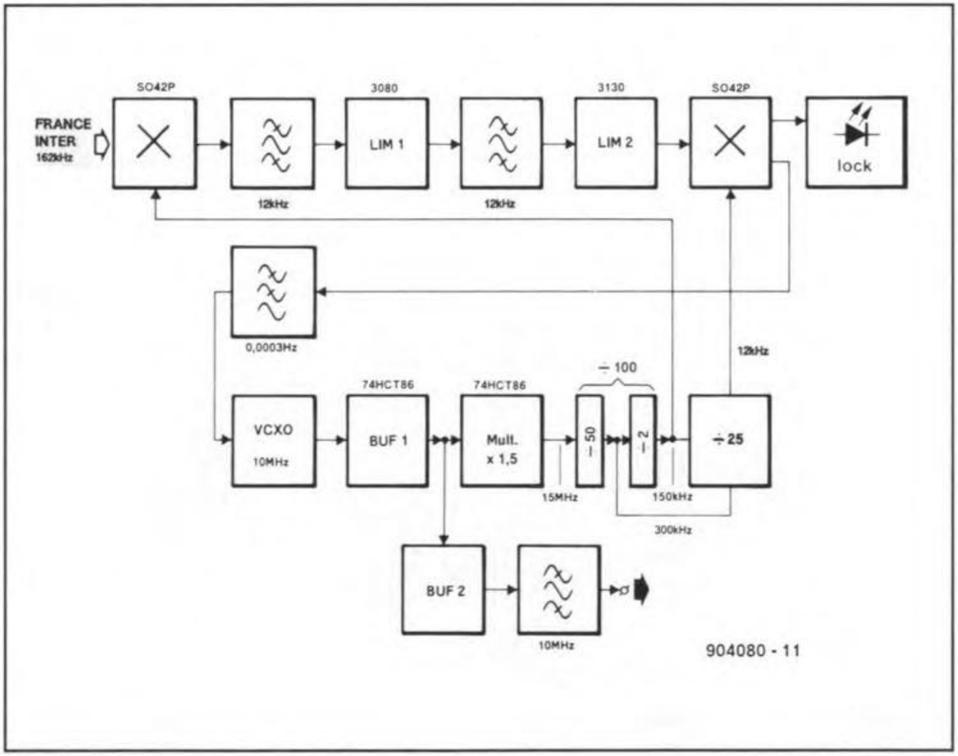


The "High-precision DLF-based locked frequency reference" we published in the April 1989 issue (p. 48), derived its stability from the Deutschland Funk (DLF) long-wave transmitter operating on 153 kHz. It is almost certain that in many areas the French broadcast transmitter France-Inter, operating on 162 kHz, will give a much stronger signal. That is because France-Inter radiates 2000 kW, whereas DLF radiates 500 kW by day and 250 kW by night.

The circuit published last year can easily be modified for reception of the French station. The shaded areas in the block diagram indicate what changes are needed.

The two figures showing sections of the original PCB indicate which connections at IC5 and IC6 must be cut and what new connections must be made. Furthermore, the table shows the changes in value of a number of original components. ■

(*A.N. Other*)



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041 ELECTRIC GUITAR ENHANCEMENT

An electric guitar with three single-coil elements, the well-known Fender model, can furnish more facilities than the manufacturers provided. A simple modification makes it possible to produce a real stereo signal or to connect one or more elements in anti-phase—see the circuit diagram.

If the guitar is fitted with a 5-position switch, this must be replaced by four in-line miniature toggle switches. Switch S4 determines the operating mode: upper position = stereo; lower position = normal or with elements in anti-phase.

Switches S1, S2 and S3 determine how the throat element, the central element and the bridge element are switched, although this is also dependent on the setting of S4.

With S4 set to 'normal/anti-phase', the other switches function as follows. In the upper position, the relevant element is switched into circuit, in the centre position, the relevant element is inactive, and in the lower position, the element is in anti-phase. The phase shift is varied with P2.

With S4 in position 'stereo', each element is connected to the left-hand channel if the relevant switch is in the upper position, and to the right-hand channel if that switch is in the lower position. If the associated switch is in the centre position, the relevant element is inactive. In this condition, P2 acts as the volume control for the right-hand channel.

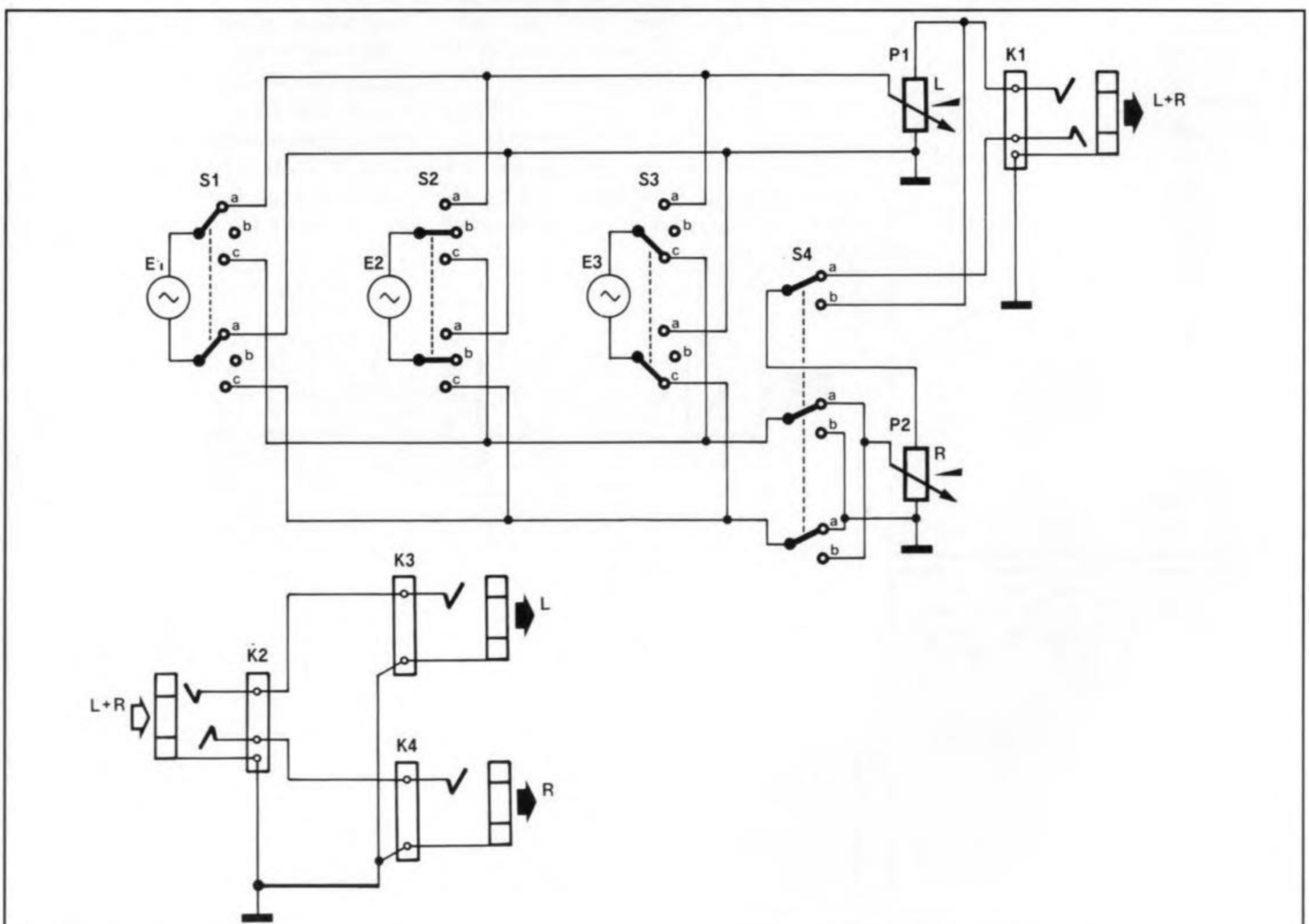
Potentiometer P1 is the volume control already fitted to the guitar (500 kΩ or 1 MΩ log). Potentiometer P2 should have the same value as P1 and be fitted in the hole previously occupied by the tone control. It is, of course, possible to retain the original tone control, but an identical one must then be fitted for the second channel (in exactly the same way as the original tone control).

E1 = throat element
E2 = central element
E3 = bridge element
S1, S2, S3: a = left-hand channel
b = off
c = right-hand channel or anti-phase
S4: a = stereo
b = anti-phase

The original jack connectors must be replaced by stereo versions. When the guitar is switched to stereo, it is connected via a good-quality 2-core screened microphone cable, terminated at each end in a stereo jack. The left-hand channel is then available at the tip of the plug, and the right-hand channel at the ring. The channels may be separated by a stereo splitter and can then be transmitted via mono jacks and cables. When the guitar is switched to normal/anti-phase, a standard guitar cable may, of course, be used.

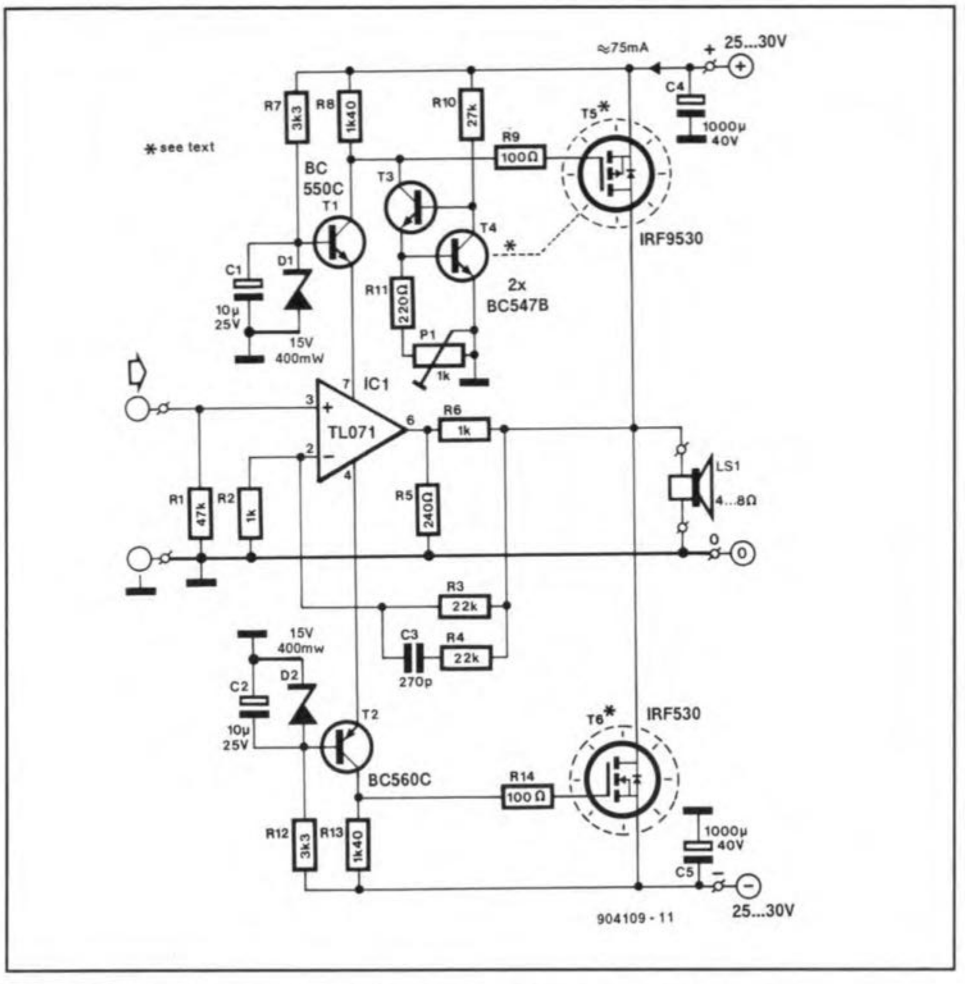
The modifications described here were carried out on a Yamaha SC1000 and gave excellent results. They are particularly recommended for good quality studio operation. ■

(A. Ferndown)



MOSFET AF POWER AMPLIFIER

042



This simple design, consisting of a TL071C driver and two MOSFET power amplifiers, can deliver up to 45 W into 8 Ω.

The design is basically a Siliconix application. The principle of this application is that the output transistors are driven by the voltage changes across two resistors inserted in the supply lines of the opamp driver. In other words, the currents drawn by the opamp determine the drive to the power amplifiers. Use is made of current feedback that is effected by connecting the output of the opamp to a potential divider across the amplifier output.

In the diagram, FETs T5 and T6 are driven by the potential differences across R8 and R13. Since the supply voltage is considerably higher than normal for a standard opamp, transistors T1 and T2 have been inserted in the supply lines to the IC. These transistors are provided with a fixed base potential of ±15 V by zener diodes D1 and D2. The supply voltage to the opamp, whatever its drive, is therefore always 14.4 V.

Setting of the quiescent current is effected with the aid of T3 and T4. It is essential that T4 is coupled thermally to T5 so as to ensure stability of the quiescent current.

Transistor T3 draws its current via R8, and this ensures a correct operating point for T5. When the temperature rises, the base-emitter voltage of T4 drops; the current through T3 decreases; and the gate-source voltage of T5 drops. The opamp ensures equilibrium of the circuit, so that the current through T6 is also adapted as appropriate. The total current drawn by the amplifier is adjusted to about 75 mA with P1. The current drawn by the FETs is then around 70 mA.

The presence of T1 and T2 makes the amplifier slightly slower than it would be without these transistors. However, the fairly large capacitors associated with the MOSFETs can only discharge via resistors R8 and R13. This results in an increase of the quiescent current at frequencies above 40 kHz. Because of this, the bandwidth is limited to 20 kHz by C3. A resistor is connected in series with this capacitor to further improve the stability.

The MOSFETs must be fitted on a heat sink of not less than 1 K/W. In contrast to the usual emitter or source follower, the configuration used here may be driven virtually up to the supply voltage, so that an efficiency of up to 70% may be attained. In the prototypes, cross-over distortion was not greater than 0.2% at 20 Hz (10 W into 8 Ω). With a stable supply of ±30 V, the amplifier can provide 45 W into 8 Ω or 70 W into 4 Ω.

Note that the amplifier is not protected against short circuits. Check therefore at all times what load is connected to it before the supply is switched on. ■

(*T. Giffard*)

TRIGGERABLE SAWTOOTH GENERATOR

043

The triggerable sawtooth generator presented here is intended to convert an old-fashioned oscilloscope with only a synchronized time base to a modern triggered instrument.

The circuit is straightforward: transistor T1 is a current source that charges one of capacitors C1–C4, depending on the position of switch S1. The linearly increasing voltage across the capacitor is compared with two reference voltages by IC1a and IC1b. The reference voltages are derived from potential dividers R3-R4 and R5-R6 respectively.

As soon as the voltage across the capacitor reaches 5 V—the toggle level of IC1a—bistable IC2a-IC2b is reset. Transistor T2 is then switched on, which causes the capacitor to discharge. Once the potential across the capacitor has dropped to the toggle level of IC1b, the output of IC1b goes high and the bistable is set via the trigger input. Transistor T2 is then switched off and the voltage across the capacitor increases.

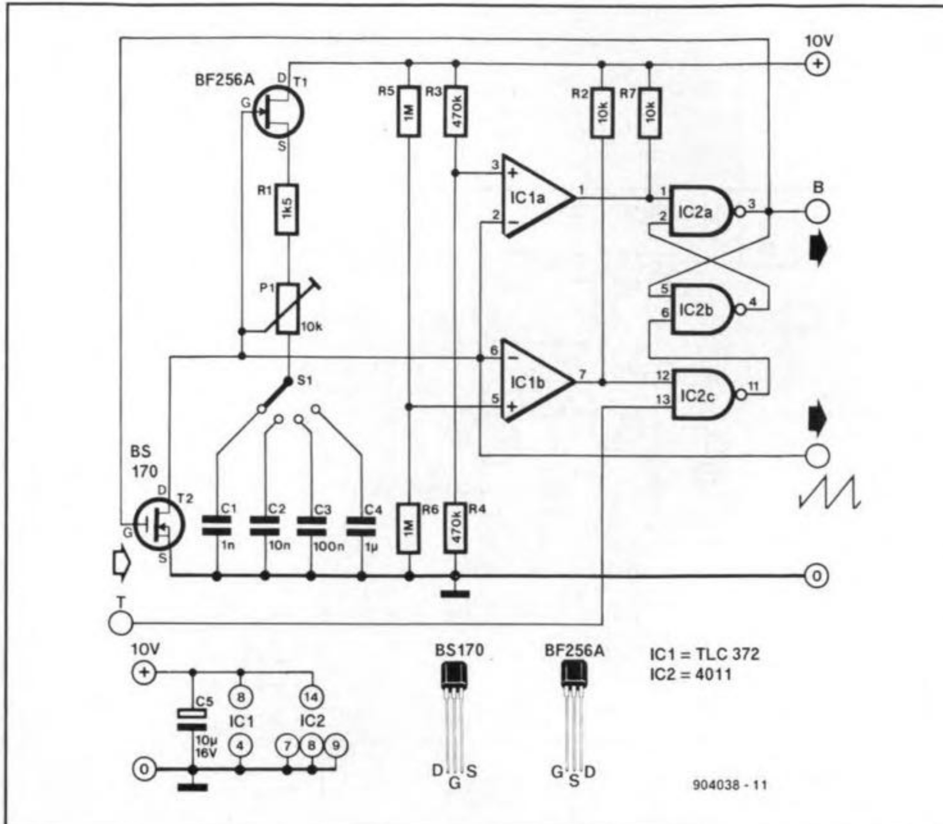
The pulse that switches on T2 may also be used as the blanking pulse for the oscilloscope and it is, therefore, available at a dedicated blanking output. For the duration of this pulse, the electron beam in the oscilloscope is suppresed.

Circuit IC2 is an AND gate that ensures that the trigger pulse can set the bistable only if the capacitor is really discharged. This prevents spurious triggering of the sawtooth generator.

The generator output must be terminated in a high-impedance load to prevent distortion of the sawtooth.

The generator draws a current of only 5 mA from a single 10 V supply. ■

(*P. Coster*)



044 SLAVE MAINS ON-OFF CONTROL

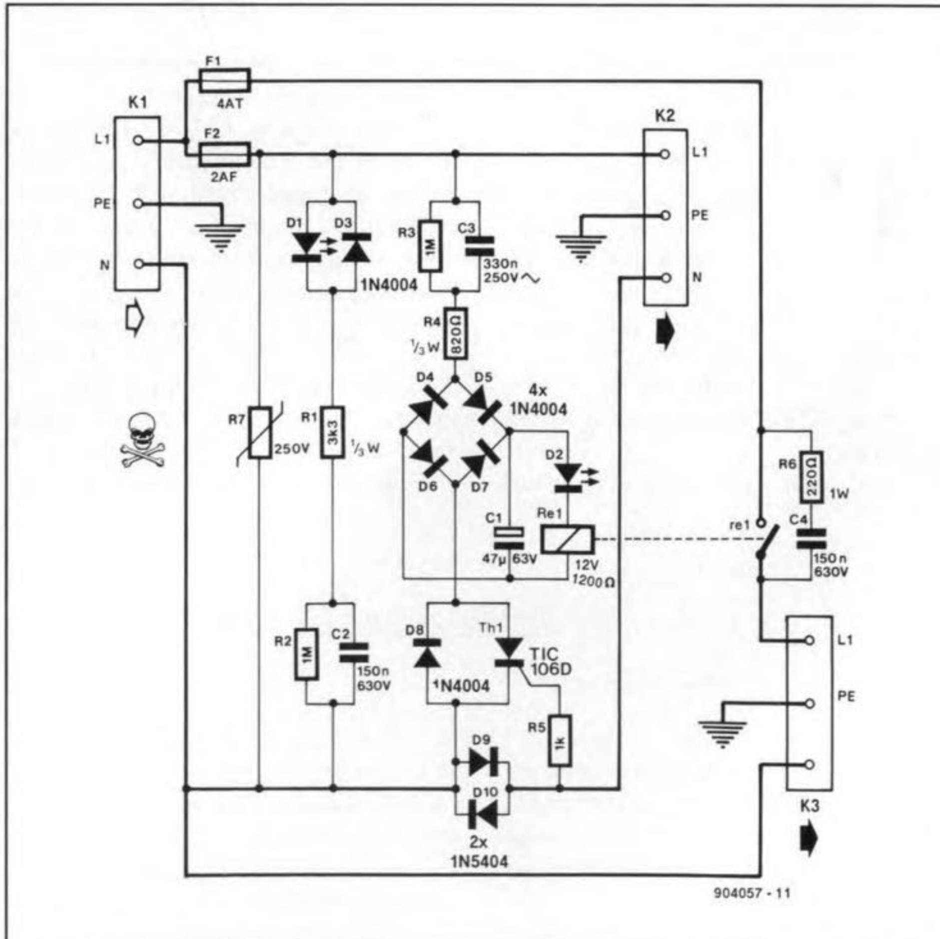
This circuit enables the automatic on/off switching of mains-powered equipment after a master unit has been switched on or off. One particularly useful application is in audio racks where a number of signal sources, such as the tuner, CD player, and tape recorder are to be switched on and off in unison with the power amplifier.

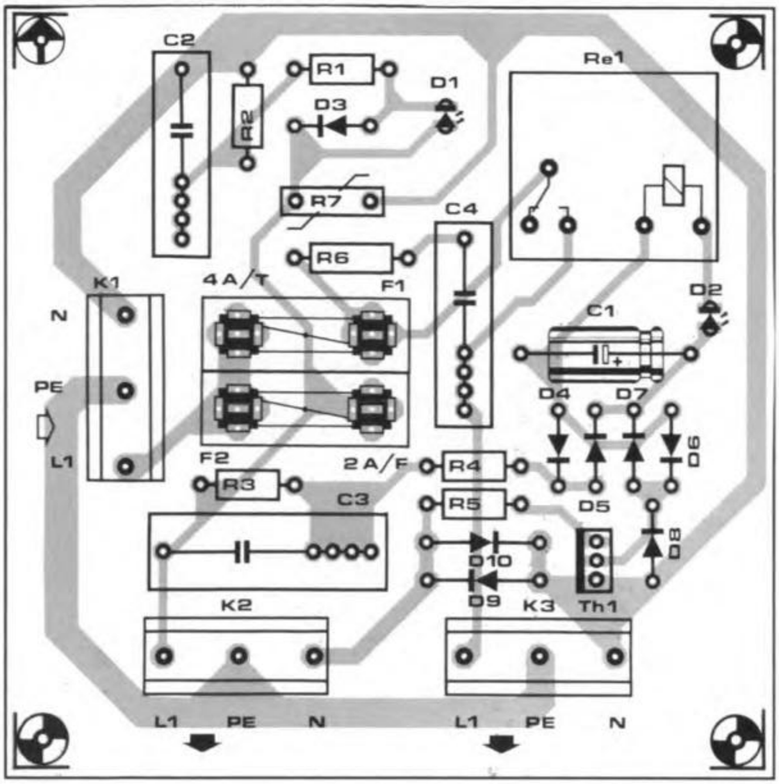
The circuit works on the principle of current sensing in the mains input lines to the master apparatus, which should be connected to K2. When the master unit is switched on, a voltage drop appears across diodes D9 and D10. This potential triggers thyristor Th1. When it conducts, the thyristor connects junction D6-D7 to the neutral line. The resultant output voltage of the bridge rectifier—about 25 V—is smoothed by C1 and then applied to the coil of relay Re1.

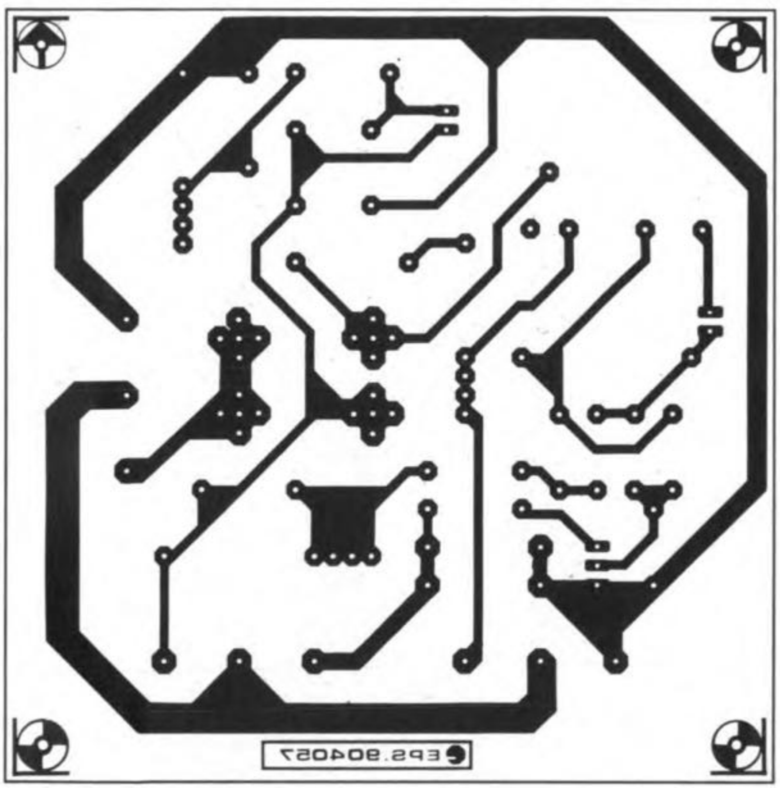
When the relay is energized, indicated by D2 lighting, its contact links the live line of the mains to K3 to which the slave units are connected.

When the master unit is switched off, the thyristor stops conducting so that the relay is de-energized and the mains live line is removed from K3.

The reactance of capacitor C2 limits the







current through on-off indicator D1 to a safe value. Resistor R2 enables the capacitor to discharge after the input has been disconnected from the mains. Similarly, R3 enables C3 to discharge after the power has been switched off.

Network R6-C4 across the relay contact prevents spikes and other noise signals generated when the relay is switched on or off being superimposed on to the output voltage.

Varistor R7 suppresses any voltage surges at the input that might cause spurious triggering of the thyristor.

For safety reasons, the circuit must be connected to the mains via a suitable mains isolating transformer and preferably be built on the PCB shown alongside. The completed assembly should be fitted in an ABS enclosure. Use an appropriate panel plug for the mains input (connected to K1 on the PCB) and panel sockets for the mains outputs (connected to K2 for the master and to K3 for the slave units).

The actuation current of the circuit is about 10 mA r.m.s. The current drawn by the master and slave units must not exceed 2 A and 4 A respectively.

WARNING. Since the circuit carries dangerous voltages at many points, it is essential that wherever possible proper electrical insulation is applied. Never work on the circuit when the mains is connected to it. Make sure that no part of the circuit can be touched when it is being adjusted or used. ■

(J. Ruffell)

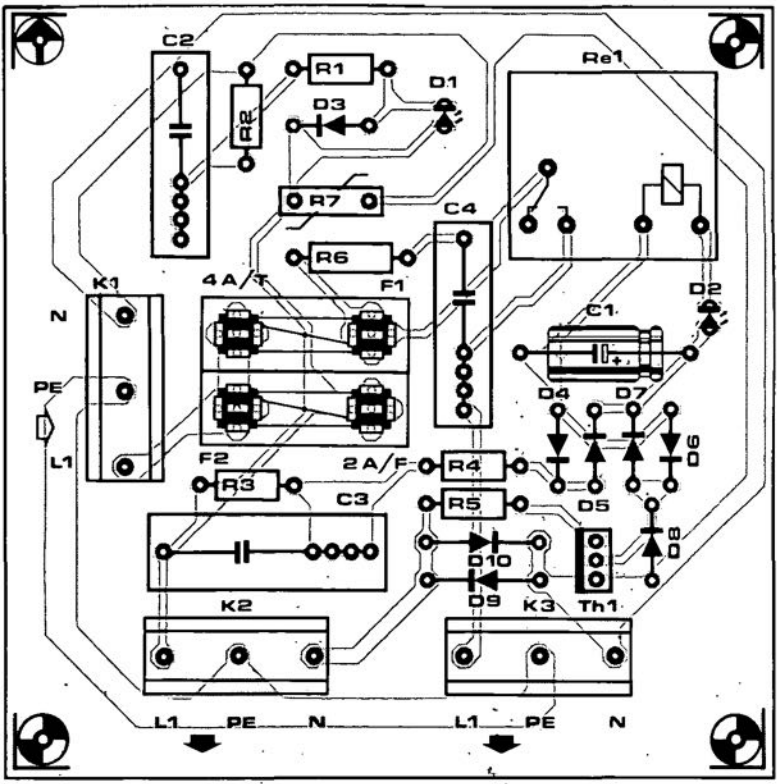
PARTS LIST

Resistors:
R1 = 3k3, 0.3 W
R2, R3 = 1 M
R4 = 820 Ω, 0.3 W
R5 = 1 k
R6 = 220 Ω, 1 W
R7 = SIOV-S10K250

Capacitors:
C1 = 47 μF, 63 V
C2, C4 = 150 nF, 630 V d.c.
C3 = 330 nF, 250 V a.c.

Semiconductors:
D1, D2 = LED, 5 mm, red
D9, D10 = 1N5404
D3–D8 = 1N4004
Th1 = TIC106D

Miscellaneous:
F1 = fuse, 4 A, slow
F2 = fuse, 2 A, fast
Re1 = SPST relay, 24 V, 1200 Ω, PCB mounting, e.g. Siemens V23127-A6-A201
K1, K2, K3 = 3-way terminal block for PCB mounting
2 PCB-mount fuseholders
1 mains panel plug
2 mains panel sockets
ABS enclosure, e.g. Retex Type RG4 (190×73×110 mm)


C2
R1
D1
Re1
D3
R2
R7
C4
R6
K1
N
4 A/T
F1
D2
C1
PE
D4
D7
L1
D6
F2
2 A/F
R4
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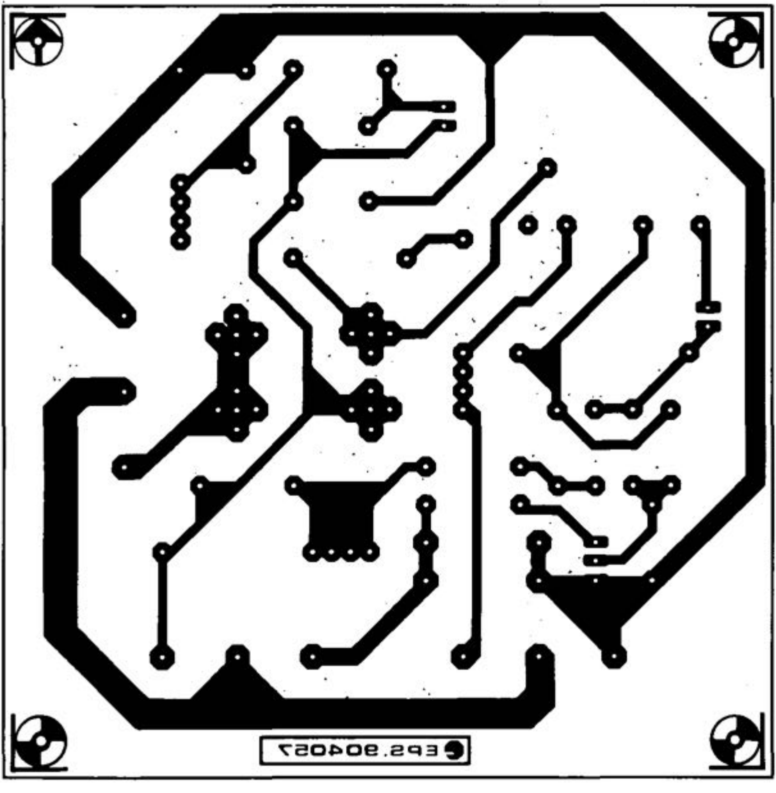
L1
PE
N



9-VOLT NICD BATTERY CHARGER

045

Nickel-cadmium batteries should normally be charged at 1/10 their capacity in Ah. The capacity of a 9 V NiCd battery is usually 110 mAh and this type of battery should therefore be charged in about 10 h. However, the efficiency of the charger is only around 70%, so that the real charging period should be of the order of 14 hours.

In the present charger, the charging time is measured by IC1, whose oscillator is set to a frequency of 1/6 Hz, ensuring that output Q13 goes high after 14 h.

When Q13 goes high, pin 1 of IC2 goes low, which causes bistable IC2a-IC2d to be reset. At the same time, pin 3 of IC2 goes high, resulting in the base voltage of current source T1 becoming equal to the supply voltage. This causes the transistor to be switched off, which interrupts the flow of charging current.

The charging cycle is started with S1 or the detection circuit based on IC3. This circuit ensures that batteries left in the charger are topped up after the batteries have lost some charge through self-discharge. This ensures that these batteries are always fully charged.

The voltage at which the battery is assumed flat is set with P2 to, say, 8.4 V. To this must be added the drop across D1. This diode prevents the battery from discharging through the charger in case the supply fails.

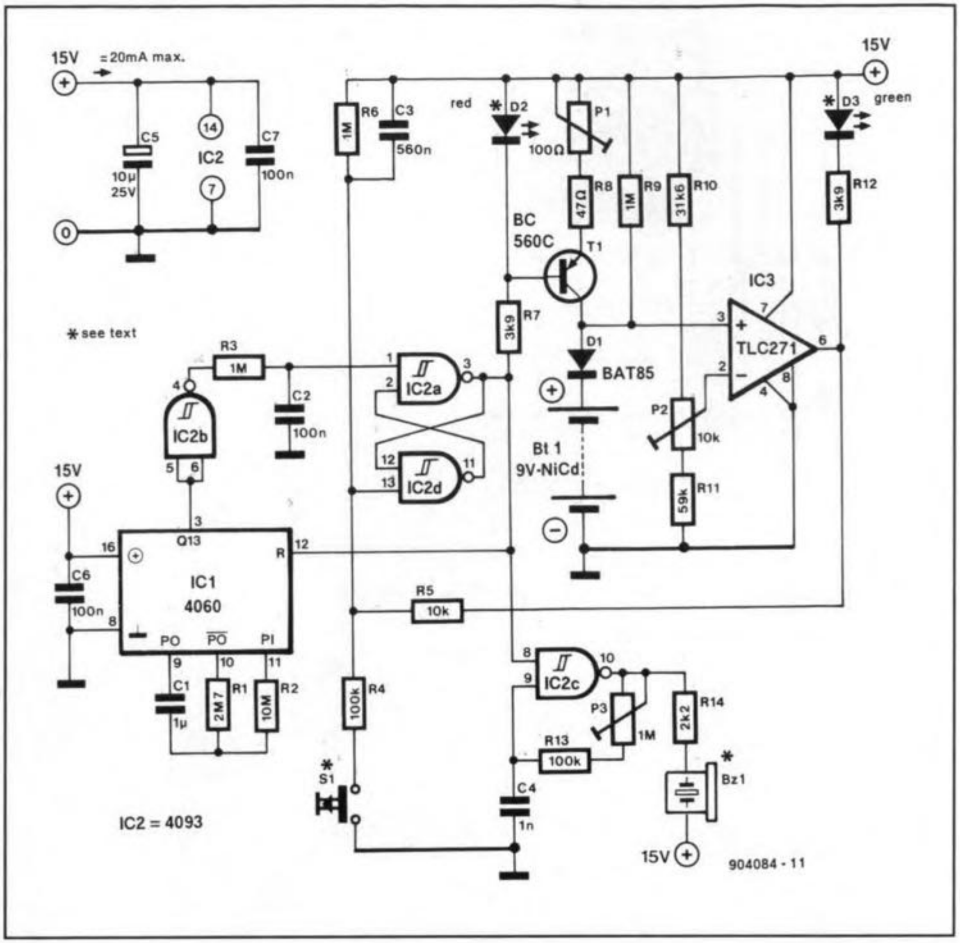
The detection circuit also ensures that as soon as a flat or partly discharged battery is connected to the charger, the charging cycle is started immediately.

Resistor R9 ensures that even when the current source is off, pin 3 of IC3 retains a certain positive voltage. In that condition, the resistor also maintains a small charging current (5–6 µA) through the battery. The drop across D1 then decreases to around 100 mV and this lowers the voltage at which the battery is assumed to be flat.

If two batteries are alternately used and charged, the value of R9 may be reduced to 1 kΩ. The self-discharge of the battery is then countered by a trickle charge at about 5 to 6 mA.

The charging current is adjusted to 11 mA with P1 as indicated by a milliammeter in series with the battery.

When the charging cycle is completed, buzzer Bz1 sounds. Square-wave generator IC2c may be set to ta suitable buzzer tone with P3. The buzzer may also be muted by strapping pin 8 of IC2 to earth



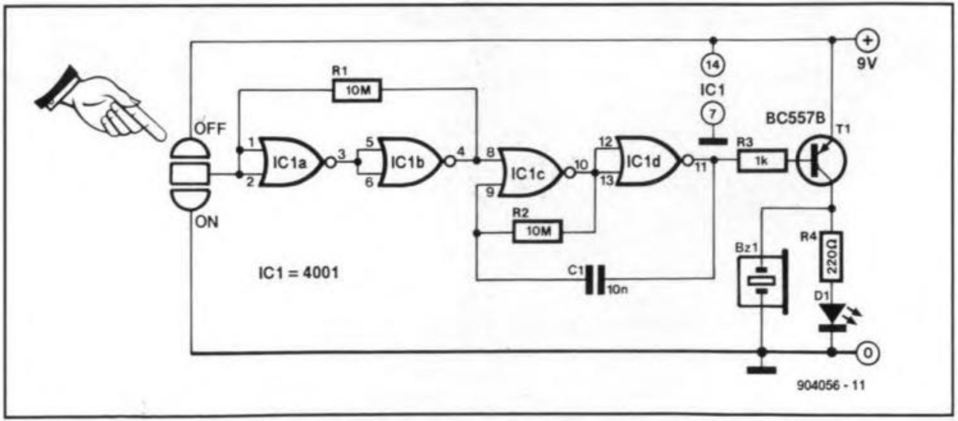
via a switch. The buzzer may, of course, also be replaced by a LED. The current through the buzzer or LED is limited by R14.

Diode D2 only lights when a charging current flows: when no battery is connected or if there is a bad contact, the low current through R7 will cause a drop of only about 1 V across the diode and this is not enough to switch it on.

When a connected battery is defect, D2 and D3 light and the buzzer is mute. With a good battery, there will come a time in the cycle when D3 goes out, while D2 remains on. ■

(T. Giffard)

046 MESSAGE PAD



This little circuit provides both a visual and an audible indication that a message has been left.

CMOS inverters IC1a and IC1b form a touch-sensitive on/off bistable in which R1 provides positive feedback so that the circuit will latch in whatever state it is left in.

The output of the bistable controls astable IC1b-IC1c, which is designed to oscillate at a rate of about 11 Hz. This rate is determined by R2 and C1.

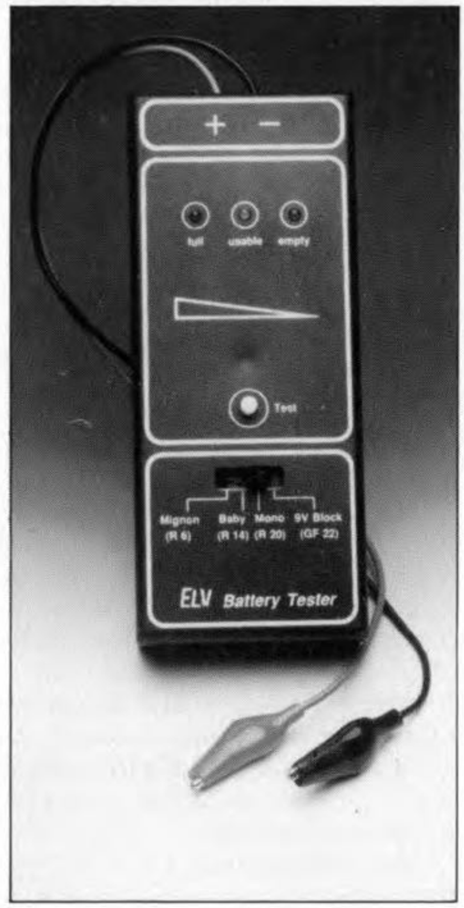
Although the astable could just about drive the LED and buzzer, a p-n-p driver transistor is used to make the circuit more versatile. When the 'on' pad is touched, the LED flashes and the buzzer bleeps intermittently.

Power is derived from a 9-V PP3 battery. Current drain in the 'off' condition is negligible. ■

(R.G. Evans)

BATTERY TESTER

This compact tester, designed and marketed as a kit by ELV, has three LEDs that indicate the condition of alkali-manganese, carbon-zinc, and alkali-zinc primary batteries of the mignon, mono, baby or power-block type.



The growing mass of battery-powered equipment brings with it the need of quick battery condition testing. The battery tester described here helps you to prevent getting stuck, at crucial moments, with a flat battery or battery pack in, for instance, a cassette recorder, a torch, a remote control unit or a personal radio.

The tester is simple to use: the primary battery to be tested is connected by two test clips and flexible wires. Three LEDs with different colours immediately indicate whether the battery is as good as new ('full'; green LED), usable (yellow LED), or exhausted ('empty'; red LED).

The tester is automatically actuated when a battery is connected to the test leads. This does not work, however, when the battery is completely exhausted. When this is suspected, the TEST button must be pressed to prove to the user that the battery is really exhausted, or that the internal 9-V battery of the tester itself is flat. When the red LED lights in this condition, the battery under test is completely empty.

Circuit description

The circuit diagram of the battery tester is given in Fig. 1. The circuit is powered by a 9-V (PP3) battery, of which the +terminal is connected to PCB terminal ST3, and the –terminal to PCB terminal ST4. The test circuit is powered only when transistor T1 conducts. Normally, T1 is held off by R25 so that the circuit does not receive a supply voltage.

When the TEST button, Ta1, is pressed, T1 receives base current via R24 and consequently starts to conduct. The test circuit then receives its supply voltage. The circuit is actuated in a similar manner when a voltage greater than about 0.65 V exists at the input terminals, ST1 and ST2. In that case, resistor R22 feeds a base current into transistor T2. The resultant base current of T1 causes the tester to be switched on automatically.

The load resistance of the battery under test is determined by a four-position slide switch, S1, and one of four potential dividers:

- 'Mignon' (IEC R6) batteries:R1-R7
- 'Baby' (IEC R14) batteries:R2-R8
- 'Mono' (IEC R20) batteries:R3-R9

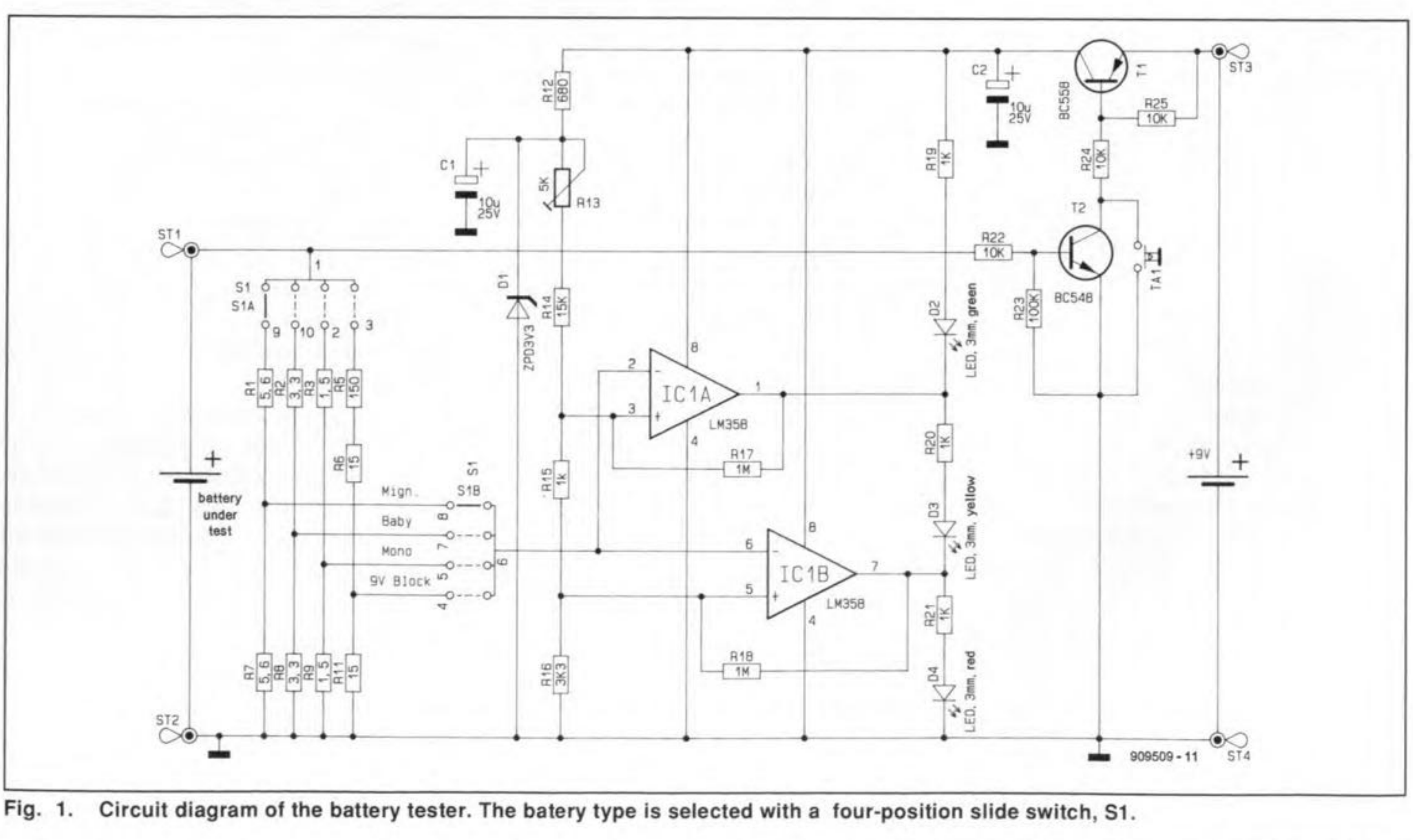
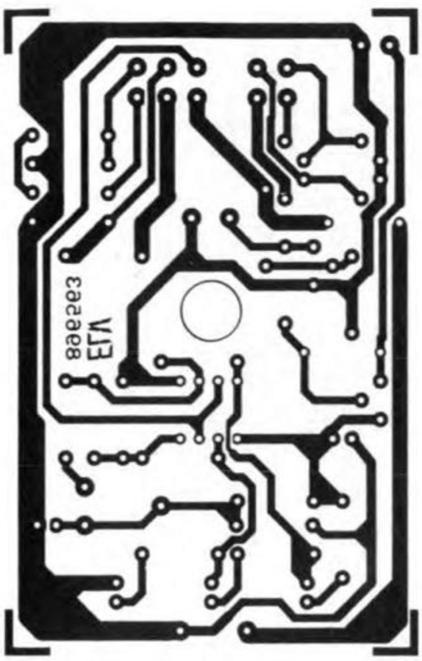
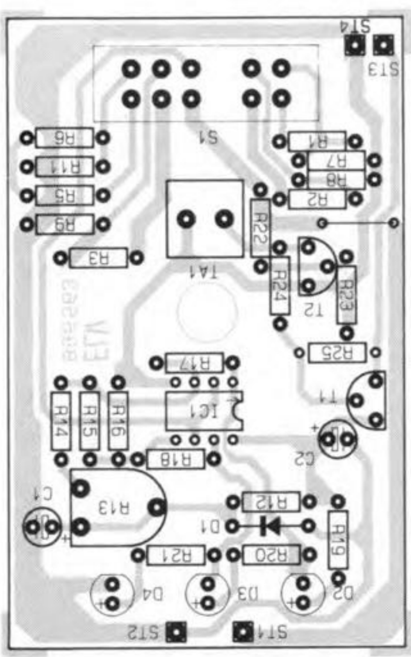
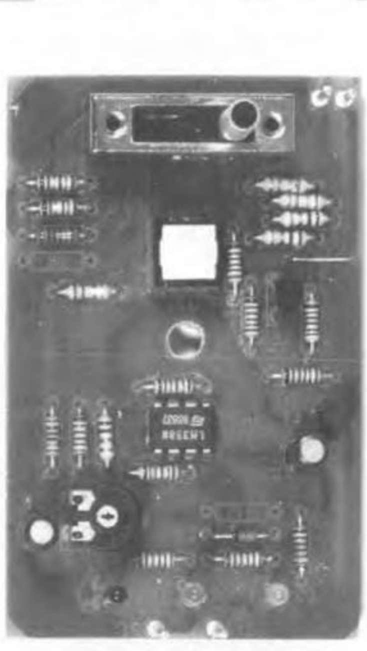


Fig. 1. Circuit diagram of the battery tester. The batery type is selected with a four-position slide switch, S1.







- '9-V power-pack' (IEC 6F22) batteries: R5-R6-R11

The voltage at the junction of the potential divider (R1–R11) selected with S1 is fed to the –inputs (pins 2 and 6) of comparators IC1A and IC1B. The +inputs of these comparators are held at a reference level created by zener diode D1 and potential divider R13–R16. Resistor R12 limits the current through this stabilizer, whose input voltage is decoupled and buffered by electrolytic capacitor C1. Feedback resistors R17 and R18 provide a certain hysteresis to ensure flicker-free operation of the LEDs.

When the battery voltage exceeds the minimum value of 1 V, both comparator outputs, pins 1 and 7 of IC1A and IC1B, are at a high potential, so that the red LED, D4, lights. When the battery voltage is between 1.0 V and 1.3 V, the output of IC1A remains high, but that of IC1B goes low. Consequently, the red LED goes out, and the yellow LED, D3, lights. When the battery voltage exceeds about 1.3 V, IC1A toggles, so that only the green LED, D2, lights to indicate that the battery is full.

The above comparator switching thresholds apply to the three types of 1.5-V battery that can be tested. For 9-V PP3 batteries, they lie at about 6.0 V and 7.8 V.

Preset R13 is adjusted to give a reference voltage of 0.65 V at pin 3 of IC1. This adjustment is made with a full battery connected to ST1-ST2, i.e., the input voltage must be 1.4 V or greater.

COMPONENTS LIST

content of kit supplied by ELV France

Resistors:

2	1Ω5	R3;R9
2	3Ω3	R2;R8
2	5Ω6	R1;R7
2	15Ω	R6;R11
1	150Ω	R5
1	680Ω	R12
4	1k0	R15;R19;R20;R21
1	3k3	R16
3	10k	R22;R24;R25
1	15k	R14
1	100k	R23
2	1M0	R17;R18
1	5k preset H	R13

Capacitors:

2	10μF 25V radial	C1;C2

Semiconductors:

1	LM358	IC1
1	ZPD3V3	D1
1	BC548	T2
1	BC558	T1
1	LED 3mm red	D4
1	LED 3mm yellow	D3
1	LED 3mm green	D2

Miscellaneous:

1	PCB-mount push button	Ta1
1	4-way 1-pole slide switch	S1
2	test lead with crocodile clip	
1	battery clip	
6	solder pin	
	15mm silver-plated wire	
1	printed-circuit board	
1	enclosure	

A complete kit of parts for the battery tester is available from the designers' exclusive worldwide distributors (regrettably not in the USA and Canada):

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82837213
Fax: +33 82838180

Construction

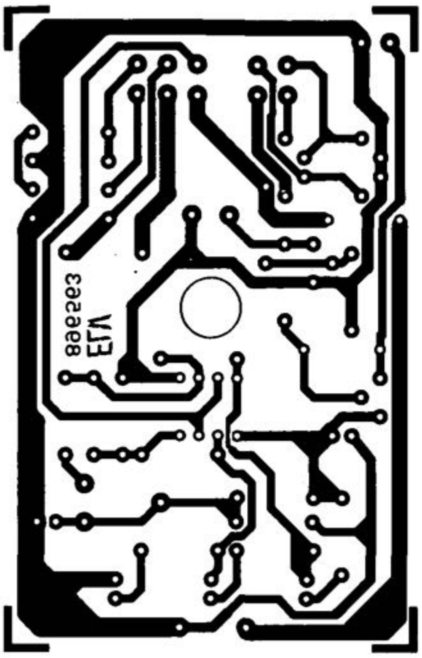
The construction of this small circuit is relatively simple. Start by fitting the single wire link on the board, followed by the resistors and the zener diode. This part has a coloured ring to mark the cathode.

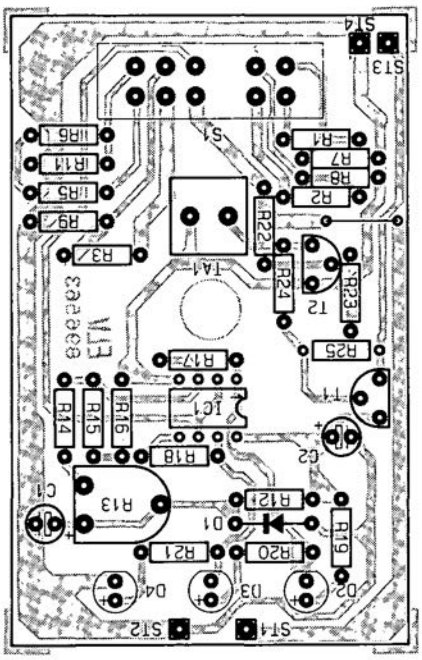
Next, fit the three LEDs so that their tops are about 15 mm above the board surface. The cathode of a LED is usually marked by the flat side of the plastic body. When the device is held against the light, the cathode is identified as the larger metal surface. The LEDs in this circuit are not normally damaged when fitted the wrong way around.

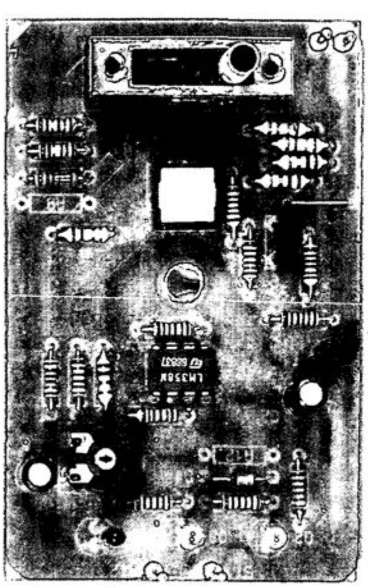
The last components to be fitted on the board are the capacitors, the transistors, the integrated circuit, the slide switch and the TEST button. Finally, connect the red wire of the battery clip to PCB terminal ST3, and the black wire to PCB terminal ST4.

Two flexible test leads with crocodile clips are supplied with the kit. The red lead (for the positive battery terminal) is connected to PCB terminal ST1, and the black lead (for the negative battery terminal) to PCB terminal ST2. The test leads pass through 2-mm dia. holes drilled in the short side of the top half of the enclosure. Make a knot in each wire, at about 20 mm from the free end, to provide some strain relief.

Fit the completed printed-circuit board into the top half of the enclosure, aligning its central hole over the moulded boss and making sure that the LEDs go into the respective holes. Then connect the 9-V battery and fit the other half of the enclosure with the self-tapping screw supplied. ■







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3	10k	R22;R24;R25
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1	100k	R23
2	1M0	R17;R18
1	5k preset H	R13

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Semiconductors:

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1	BC548	T2
1	BC558	T1
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1	LED 3mm yellow	D3
1	LED 3mm green	D2

Miscellaneous:

1	PCB-mount push button	Ta1
1	4-way 1-pole slide switch	S1
2	test lead with crocodile clip	
1	battery clip	
6	solder pin	
	15mm silver-plated wire	
1	printed-circuit board	
1	enclosure	

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Fit the completed printed-circuit board into the top half of the enclosure, aligning its central hole over the moulded boss and making sure that the LEDs go into the respective holes. Then connect the 9-V battery and fit the other half of the enclosure with the self-tapping screw supplied. ■

GLOW PLUG SWITCH FOR 4-STROKE MODEL ENGINES

A. Peperkamp

A notorious problem with four-stroke model engines is their tendency to stutter or even stall at low speeds with all the obvious risks of carefully wrought model aeroplanes fluttering helplessly about before crashing to the ground. This circuit gives ease of mind to aeroplane modellers by automatically switching on the glow-plug when required at low engine speeds to keep the combustion going.

Among the conditions for a reliably running model engine are a correctly adjusted carburettor, the use of the right fuel, and a suitably rated glow plug. The latter is fitted to ensure a sufficiently high temperature in the combustion chanber of the engine to enable this to be started. A glow plug is normally powered by a starting battery, which may be disconnected once the engine is running.

Most engines in model aeroplanes are 2-stroke types with glow ignition. Recently, however, 4-stroke types have become available with two main advantages: first, they produce less noise, and, second, their sound is more like that of a real aeroplane. Unfortunately, 4-stroke engines also have disadvantages with respect to 2-stroke types: their fuel/performance ratio is worse, their construction is more complex, they weigh more, and, importantly, their combustion is optimum at relatively high speeds only. When a 4-stroke engine runs at low speed for some time, its operating temperature drops to a level where combustion strokes fail, or the engine stalls altogether. Clearly, the additional heat of the glow-plug may help to prevent this happening. The present circuit connects the glow plug to a battery on board the plane. This connection is made automatically when the engine speed drops below a certain value.

Not the mechanical way...

A simple form of mechanical control is a lever switch coupled to the accelerator servo. In practical terms, this can take the form of a cam or notch in the servo disc operating a switch when a certain position (corresponding to a given engine speed) is reached. The switch, in turn, connects or breaks the supply voltage to the glow plug. The main problem of this approach is finding the right switching point by experimenting with a running engine. Also, taking into account that space is always at a premium in a model, a mechanical construction with a lever and a switch, however small, can become very complex indeed and take a lot of time to install and adjust.

...but with electronics!

The first 'electronics' idea that comes to mind is, of course, replacing the lever switch by a relay or a power transistor. The choice in favour of the latter will be obvious in view of reliability in a fairly hostile environment (vibration and shock caused by the model). The next step is to eliminate the mechanical coupling to the accelerator servo disc. Instead, the relevant pulses received from the remote control transmitter are accepted and processed in parallel by a special circuit, i.e., the glow plug switch is controlled by the accelerator servo pulses supplied by the receiver on board the model. When the pulse-width of the accelerator control signal reaches a certain (predefined) level corresponding to a relatively low engine speed, the glow-plug is automatically powered, and switched off when the engine is revved up again.

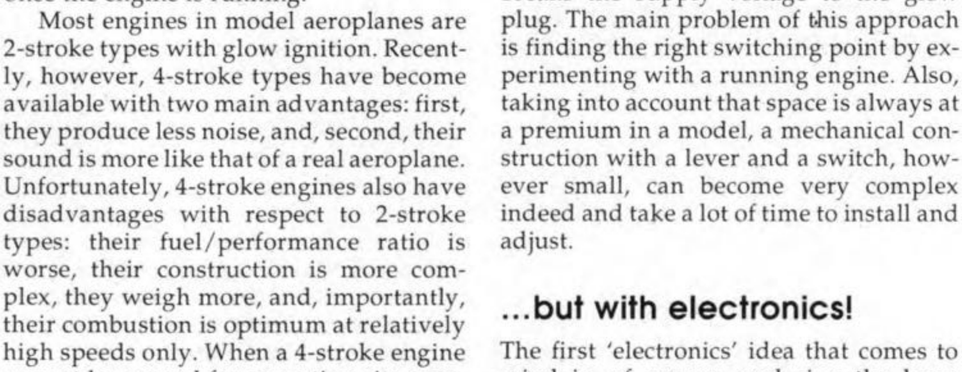

Fig. 1. Circuit diagram of the glow plug control.

Powering the glow plug

Glow plugs fitted in model engines generally operate from a 1.5-V supply. Depending on the type, the current consumption is usually between 2 A and 4 A. Provided the switch is virtually loss-free, a NiCd (nickel-cadmium) battery with a capacity of 1.2 Ah or 1.8 Ah may be used as a power source. Since the glow plug is switched on at low engine speeds only (e.g., while the model idles on the ground, or while descending or landing), this sort of battery capacity is sufficient for a number of flights.

The author has fitted one of his models with a 9-way sub-D plug for outboard current supply and charging of the receiver battery as well as the glow plug battery. A switch is added to enable the glow plug to be powered by an external (outboard) battery or power supply during starting on the ground. At the same time, this external power source charges the on-board battery. When the switch is set to the 'fly' position, the glow-plug battery is connected to the control board.

Circuit description

The circuit diagram of the glow plug switch is given in Fig. 1. The power FET, T4, used to switch the current to the plug, has a typical on-resistance of 0.04 Ω. This extremely low value is only achieved, however, at a sufficiently high gate voltage. Since the battery voltage is usually 4.8 V in models, voltage doubling is used to ensure that T4 can be driven into saturation.

The voltage doubler consists of T2, D1, D2, C4 and R4. The circuit supplies an out-

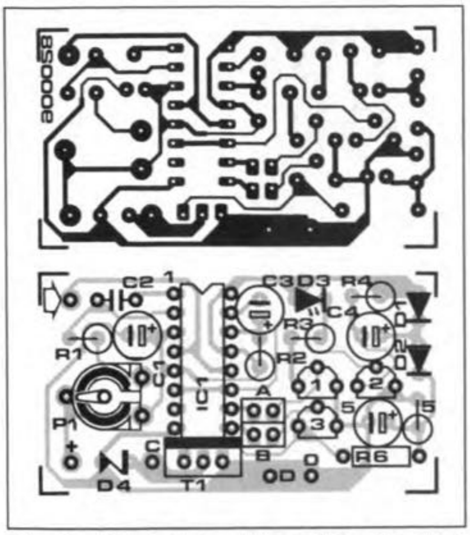

Fig. 2. Printed-circuit board for the glow plug switch.

COMPONENTS LIST

Resistors:

1	10k	R1
1	68k	R2
1	330Ω	R3
1	820Ω	R4
2	100k	R5;R6
1	50k preset H	P1

Capacitors:

1	10μF 25V	C1
1	47nF	C2
1	2μ2 25V radial	C3
2	47μF 25V radial	C4;C5

Semiconductors:

2	1N4148	D1;D2
1	LED	D3
1	39V 0.4W zener diode	D4
3	BS170	T1;T2;T3
1	BUZ11	T4
1	4538	IC1

put voltage of about 9 V across C5. This capacitor is discharged by R5 after the receiver is switched off to ensure that T4 is off no longer than 3 seconds afterwards. The value of R5 is a compromise between rapid turning off of T4 after the receiver is switched off, and a small load for the voltage doubler. The PWM (pulse-width modulated) signal taken from the accelerator servo terminals provides the clock for the voltage doubler, obviating a separate oscillator.

The drive signal for the power FET is generated by two monostable multivibrators, IC1A and IC1B. The former supplies an output pulse at the leading edge of the PWM input signal. The pulse width of this output signal (available at pin 6) may be set to a value between 0.5 ms and 3.0 ms by adjusting preset P1.

Monostable IC1B is wired to trigger at the trailing edge of the pulse supplied by IC1A, i.e., when the receiver output signal is at '0'. The control signal for T4 is taken from the $\overline{Q}$ output of IC1B and applied to the gate via a small-signal FET, T3.

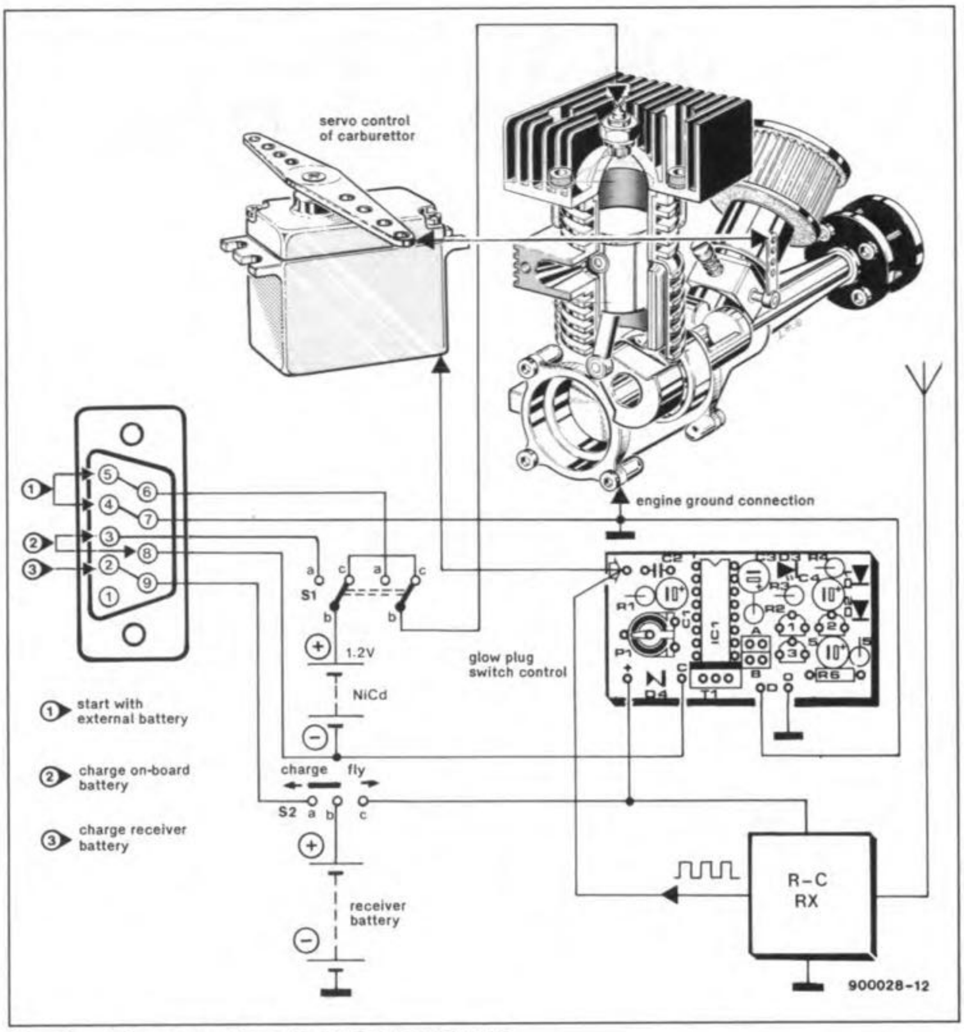

Fig. 3. Interconnections in a model aeroplane.

LED D3 forms the glow plug on/off indicator. Where appropriate, it may be replaced by an actuator such as a low-power piezo buzzer. The engine speed at which the glow plug is switched on may be adjusted with P1.

Construction and wiring

The printed-circuit board for this circuit is small (Fig. 2) in view of the restricted space in the model. Jumpers A and B allow the control to be geared to the polarity of the servo control signal supplied by the remote-control receiver.

The wiring between the control circuit, the two batteries, the glow plug, the 'charge/fly' switch and the 9-way sub-D connector is connected as shown in Fig. 3. Be sure to use heavy-duty wire for the connections that carry the glow plug current.

The current consumption of the circuit is negligible relative to that of the glow plug. In the off state, about 1 mA is drawn, in the on state, about 12 mA, mainly on account of the LED. ■

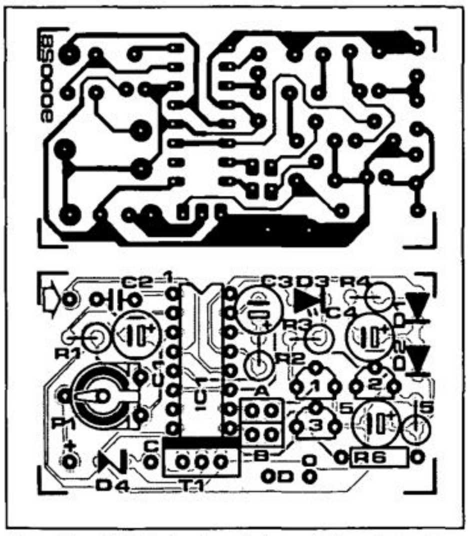

Fig. 2. Printed-circuit board for the glow plug switch.

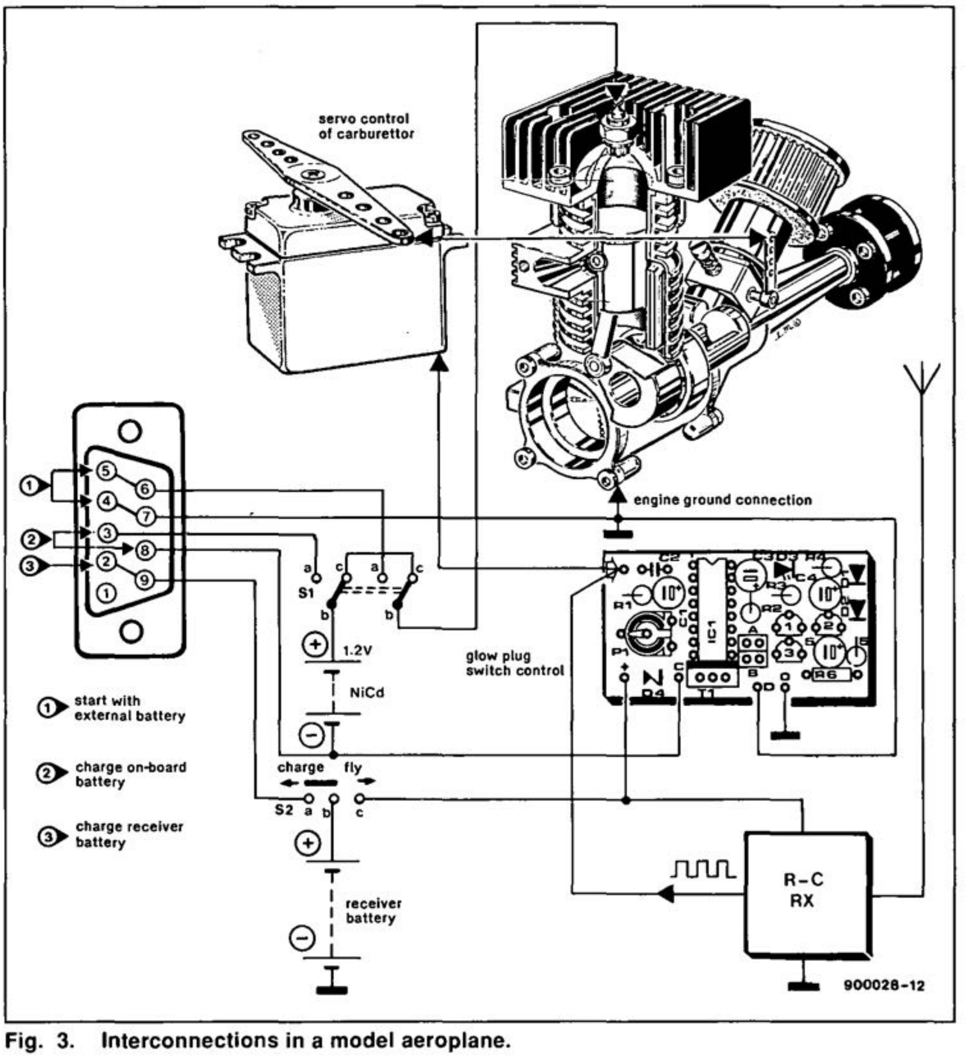

Fig. 3. Interconnections in a model aeroplane.

COMPONENTS LIST

Resistors:		
1	10k	R1
1	68k	R2
1	330Ω	R3
1	820Ω	R4
2	100k	R5;R6
1	50k preset H	P1
Capacitors:		
1	10µF 25V	C1
1	47nF	C2
1	2µ2 25V radial	C3
2	47µF 25V radial	C4;C5
Semiconductors:		
2	1N4148	D1;D2
1	LED	D3
1	39V 0.4W zener diode	D4
3	BS170	T1;T2;T3
1	BUZ11	T4
1	4538	IC1

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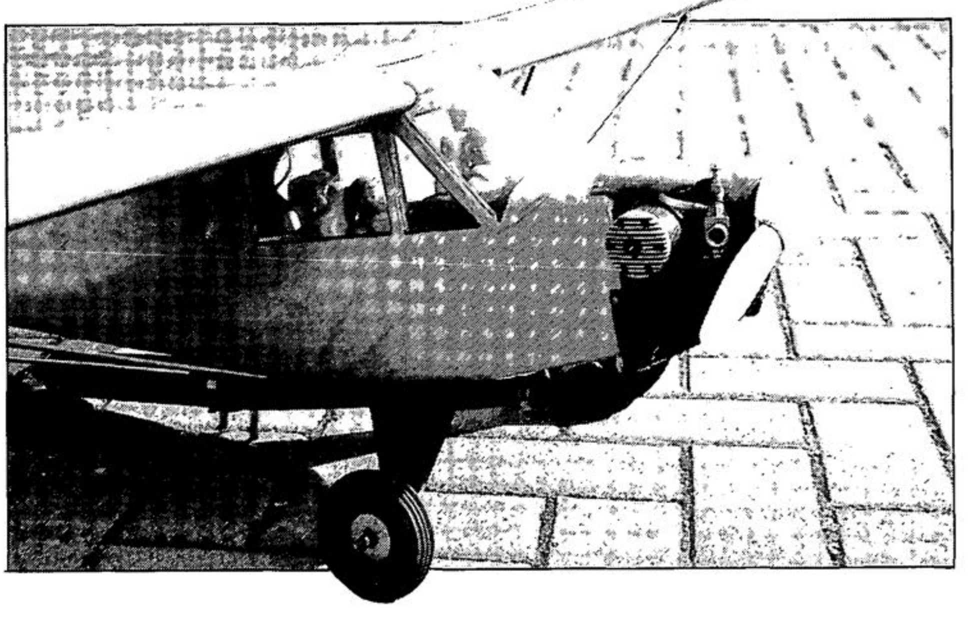
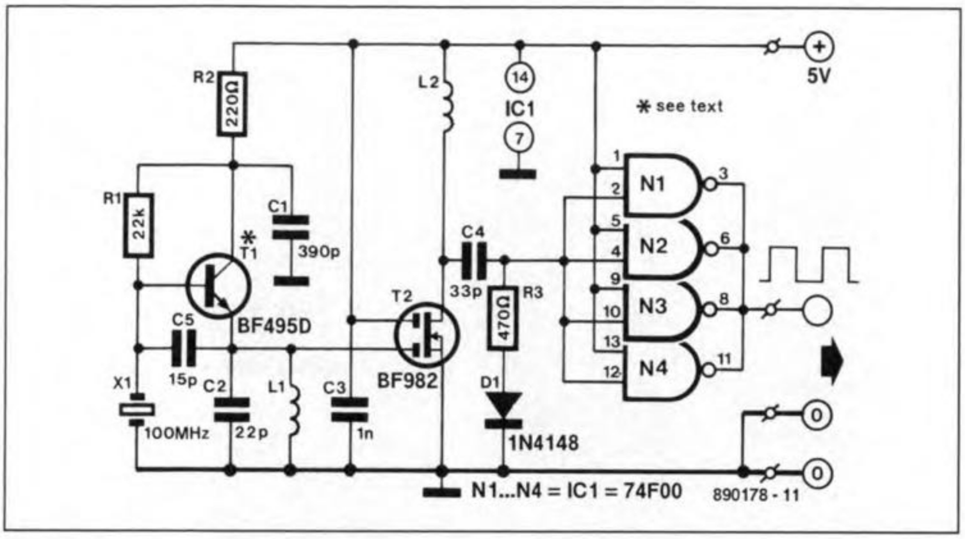


100 MHz TTL-COMPATIBLE CRYSTAL OSCILLATOR

J. Bareford



The use these days of clock frequencies of up to 100 MHz is not uncommon in digital signal processing (DSP) equipment, RF synthesizers, video storage circuits and logic analysers. The circuit described here is based on a quartz crystal, and supplies an output signal of 100 MHz with sufficient drive capacity for TTL circuits.

The circuit consists of three sections: a quartz-crystal controlled oscillator around X1 and T1; an impedance converter, T2, and an output buffer, IC1.

Tuned circuit L1-C2 is designed to make the quartz crystal, X1, operate at its fifth overtone (i.e., its fundamental, parallel-resonant, frequency is 20 MHz). Positive feedback in the oscillator is provided by C5 between the emitter and the base of T1.

Since the oscillator must be loaded as lightly as possible to prevent starting problems and instability, the output signal is applied direct to a dual-gate MOSFET, T2, of which gate 1 forms a very high impedance. The high transconductance of the MOSFET enables the oscillator signal to be taken from the drain at a relatively low impedance. Network R3-D1 raises the top level of the 100-MHz to above the TTL threshold for a logic 1 (approx. +2.42 V).

Four parallel-connected NAND gates in a 74F00 or 74AS00 package are used to digitize and boost the oscillator signal. The 5-V_{pp} logic swing of the output signal enables it to be applied direct to TTL-compatible clock inputs.

Construction

The oscillator is best constructed on the small, double-sided printed-circuit board shown here. The component side of this board has a ground plane of unetched copper to assist in decoupling the high-frequency signal.

Start the construction with winding the two inductors. L1 consists of 10 turns of 0.5 mm diameter enamelled copper wire (e.c.w.), and has an internal diameter of 3 mm. L2 consists of 25 turns of 0.3 mm e.c.w., and also has an internal diameter of 3 mm. Use the shaft of a 3-mm drill as a former to make these inductors, which are fitted just (<0.5 mm) above the board surface. Note that there is no ground hole for L1; after winding the inductor, bend the ground terminal at right angles and solder it straight to the ground plane.

All parts must be fitted with the shortest possible lead length. This means that the ceramic capacitors must be pushed as far as possible into the respective holes. Grounded component terminals are pretinned before they are soldered at the component side as well as at the track side of the board. On the component overlay, these terminals are indicated by the absence of a circle (look, for instance, at C3). The integrated circuit, IC1, must be fitted without a socket.

Finally, a note about the quartz crystal: this must be a fifth-overtone, parallel-resonance type with a fundamental frequency of 20 MHz. The loss resistance must be smaller than or equal to 20 Ω. The supply voltage range of the circuit is 4.75 V to 5.25 V. Current consumption at 5.0 V is smaller than 10 mA. ■

COMPONENTS LIST

Resistors:		
1	22k	R1
1	220Ω	R2
1	470Ω	R3
Capacitors:		
1	390pF	C1
1	22pF	C2
1	1nF ceramic	C3
1	33pF	C4
1	15pF	C5
Semiconductors:		
1	1N4148	D1
1	74F00 or 74AS00	IC1
1	BF495D or BF495C	T1
1	BF982	T2
Miscellaneous:		
1	100 MHz quartz crystal (5th overtone, parallel-resonance)	X1
	enamelled copper wire 0.3 mm diameter.	
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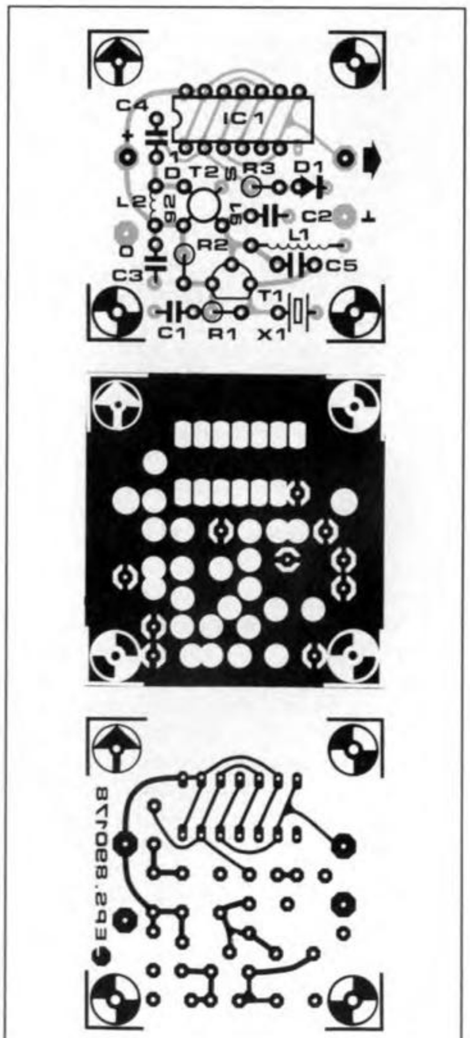

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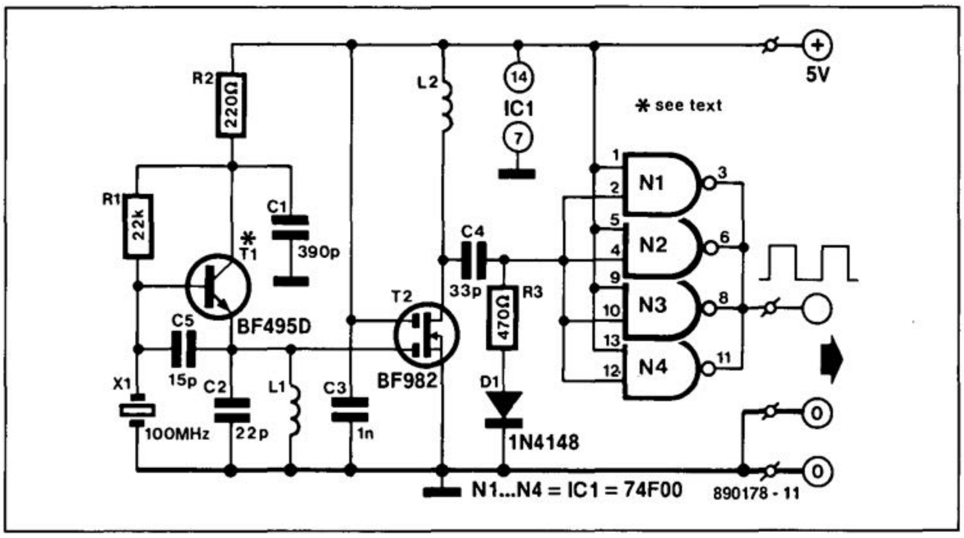

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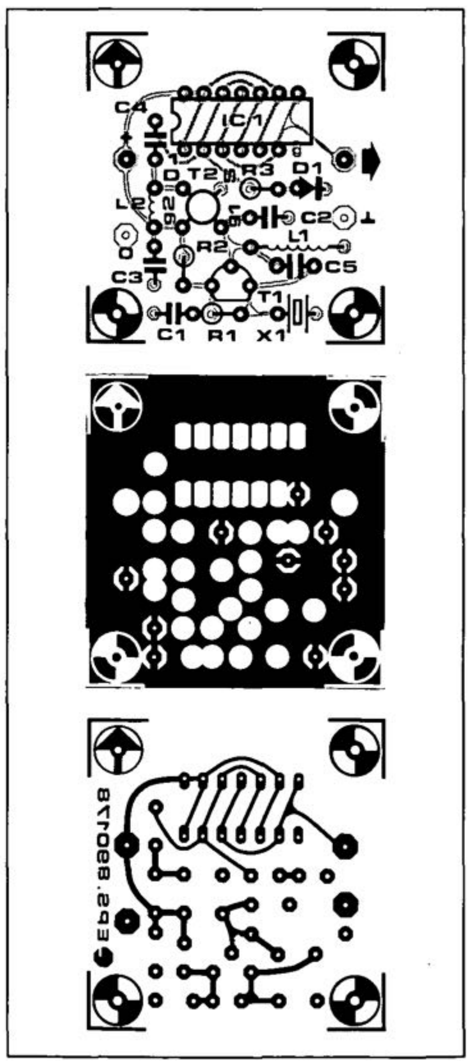


COMPACT 10-A POWER SUPPLY

G. Boddington

It has been some time since we published a 200+ Watt power supply like the one described here. Capable of supplying up to 10 A (and more if so configured) at an output voltage range of 4 to 20 V, this ultra-reliable PSU uses a minimum of components and is simple to build.

MAIN SPECIFICATIONS

- Output voltage: 4–20 V
- Output current: 10 A (basic version)
- Current limiting: internal
- Thermal protection: internal
- Ripple rejection: >80 dB
- Operating temperature: 0–50 °C
- Simple to extend for higher output currents
- Separate voltage and current indicators

If there is one instrument in the electronics workshop or laboratory you must be able to rely on at all times, it is the power supply. These days, a d.c. power supply with excellent regulation, low noise output and high output current can be built with relatively few components. The present design is based on the well-known Type LM317 integrated regulator. Since the LM317 is capable of supplying an output current of 'only' 1.5 A, a number of these devices is connected in parallel, under the control of a single voltage setting circuit. The result is a surprisingly simple PSU, which, in its basic version, is capable of providing an output current of up to 10 A.

The present PSU is ideal for charging batteries, for experiments with a wide range of electronic circuits, and for use with high-power transistorized RF amplifiers.

Design background

Before discussing the circuit in detail, it is worth while looking at the basic operation of the LM317.

The LM317 is a three-pin integrated high-power voltage regulator which may be used in a 'floating' circuit. As shown in Fig. 1, a voltage of 1.25 V exists between the output and the ADJUST input. Provided the output current is 5 mA or greater, and there exists a sufficiently high voltage difference between the input and the output, the internal control circuit will maintain the 1.25 V voltage difference between the output and the adjust input. Evidently, this voltage disappears when the chip switches off owing to a thermal overload condition.

Since the regulator may be used in a 'floating' circuit, it is possible to set the output voltage by raising the potential at the ADJUST input with the aid of a voltage divider between the output and ground as shown in Fig. 1. The constant voltage of 1.25 V across R_1 causes a constant current through R_1 and R_2 (disregarding, for the moment, the small current supplied by the ADJUST terminal). In this configuration, the output voltage, U_o, is determined by the ratio of R_1 and R_2. The output voltage rises when R2 is increased. When R2 is made 0 Ω, the output voltage equals the reference voltage of 1.25 V. In the form of an equation:

$$1.25 = U_o\,(R_1/(R_1 + R_2))$$

or

$$U_o = 1.25\,(R_1 + R_2)/R_1)$$

It will be clear that when a number of regulators are connected in parallel, the output current must be distributed equally. This may be achieved as shown in Fig. 2 by fitting a small resistance, R_s,

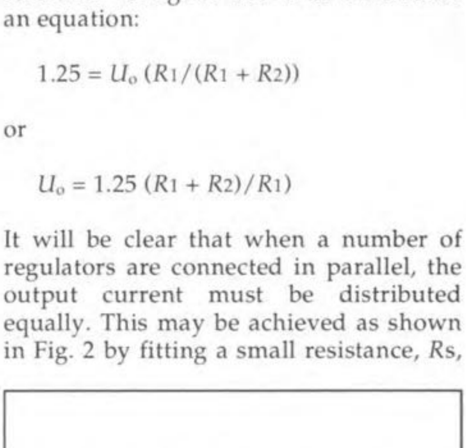


Fig. 1. LM317 in basic regulator circuit with adjustable output voltage.

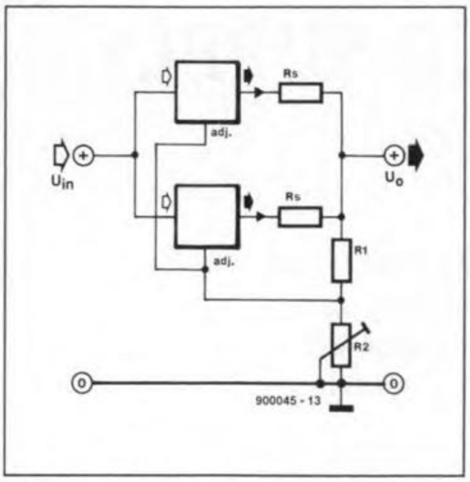


Fig. 2. Principle of parallel-connected LM317s with series output resistors to ensure equal current distribution.

in series with each regulator output. Assuming that $R_S << R_1$, the operation of the top voltage regulator in Fig. 2 is expressed by:

$$1.25 = U_o\,(R_1/(R_1+R_2)) + I_0\,R_s$$

and that of the one below by

$$1.25 = U_o\,(R_1/(R_1+R_2)) + I_1\,R_s$$

Since the equations are identical with the exception of the terms I_0 and I_1, these currents must be equal. The equations also suggest the use of more than two regulators in parallel, since the output current distribution is, in principle, determined by two tolerance factors only: first, the reference voltage of the LM317s used, and second, the values of resistors R_s.

Unfortunately, connecting LM317s in parallel is not so simple in practice, mainly for two reasons. First, the voltage drop across resistors R_s is dependent on the output current, so that the regulators can not keep the output voltage constant. This is so because they will attempt to keep the voltage across the series combination of R_1 and R_s at a constant level of 1.25 V. The upshot is that the voltage at the supply output (i.e., behind resistors R_s) will drop when the output current rises, since this means that the drop across resistors R_s rises. Second, small differences between the reference voltages of the regulators will result in unequal current distribution.

External voltage regulation

Figure 3 shows the basic circuit of an extended voltage control circuit connected to the LM317. For simplicity's sake, it is assumed that the power supply contains only one LM317.

The –input of the opamp (a Type 741) is held at half the supply output voltage, U_o, with the aid of R_4-R_5, while the +input is held at a reference potential. This reference is obtained from a constant current through resistor R_3 and preset P_1. This constant current is obtained from a resistor connected between the output and the ADJUST input of the regulator. Since the regulator maintains a constant voltage (1.25 V) between these terminals, R_3 and P_1 draw a constant current via a transistor, T_1.

Because the opamp is a difference amplifier, it will attempt to regulate its output voltage until its inputs are at equal levels. Transistor T_1 is, therefore, driven such that the voltage across R_3 and P_1 equals half the supply output voltage. This means that the changing resistance formed by the transistor causes the voltage at the ADJUST input to rise, and with it the output voltage of the supply. This closes the control loop. In the circuit shown in Fig. 3, preset P_1 forms the output voltage control because it determines the voltage at the +input of the opamp.

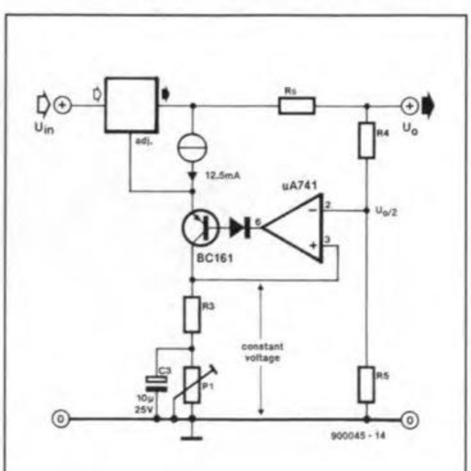


Fig. 3. LM317-based series regulator with active, external, voltage setting.

From theory to practice

After the above description of the basic operation of the power supply, little remains to be said about the circuit diagram in Fig. 5.

The mains transformer, Tr_1, is a toroid type with two secondary windings of 15 V/7.5 A each. The current specification is not obligatory, however, and may be geared to the anticipated loads. In any case, the transformer output current must be greater than or equal to 1.4 times the maximum anticipated load current of the power supply.

Although most high-power toroid transformers have two secondary windings which may be connected in parallel to boost the output current, it is better in practice to fit each secondary winding with its own bridge rectifier and associated reservoir capacitor.

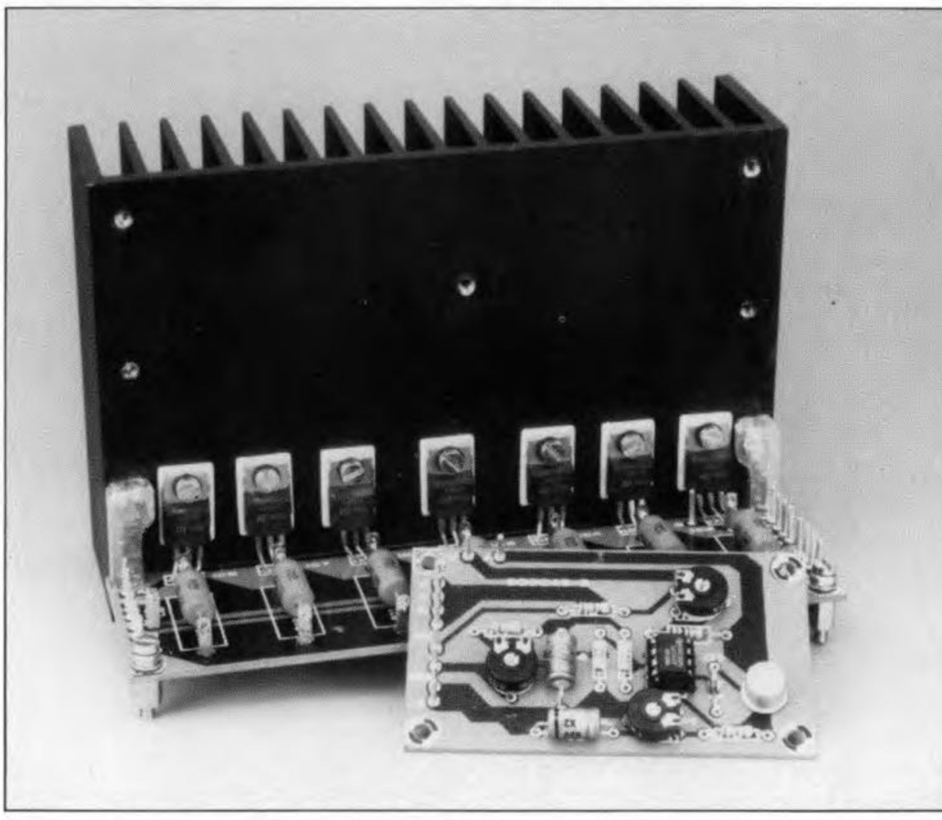


Fig. 4. Completed 10-A regulator board with heat-sink, and voltage control board.

The high currents that may have to be supplied by the bridge rectifiers force the use of adequately sized heat-sinks for these devices. As a rule of thumb, the reservoir capacitors must be 10,000 µF each per 10 A of output current. When the size of your enclosure allows it, this is best increased to 20,000 µF per 10 A.

Provided the reservoir capacitors are sufficiently large, a secondary voltage of 15 VAC will provide an output voltage of up to 12 VDC. Similarly, a transformer voltage of 18 VAC allows the circuit to supply up to 15 VDC. A maximum output voltage slightly lower than 28 V may be obtained by using a transformer with two 24-V secondary windings and reservoir capacitors rated at 63 V. Note that a 33-V transformer must **not** be used since it

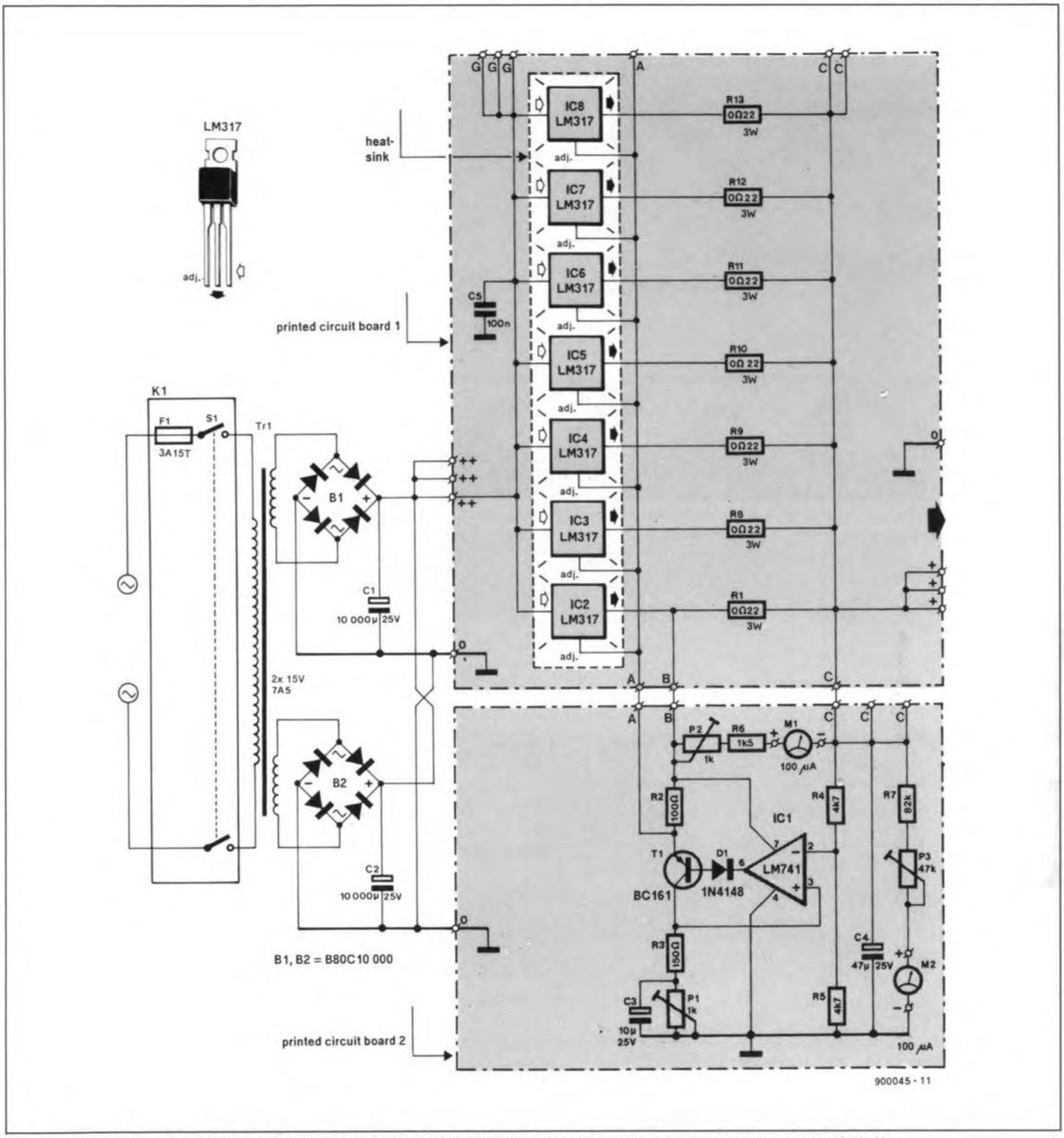


Fig. 5. Circuit diagram of the power supply (basic version with one regulator board for an output current of 10 A).

would cause the maximum input voltage of the circuit to be exceeded.

Two moving-coil meters are included in the circuit, one for voltage measurement and one for current measurement. These meters may be replaced by digital (LCD) read-outs, which are available as ready-made modules. Two such modules are used in the prototype of the power supply. Their one disadvantage is that they require a floating power supply. The simplest way to avoid problems with this supply is to power the modules from two separate 9-V batteries. To save battery power, an optocoupler circuit may be used to enable the power supply to switch the modules on and off. The basic circuit to realize this type of control is shown in Fig. 8. Alternatively, construct a small, separate, power supply for the LCD modules.

Most LCD modules are 200-mV voltmeters. Fortunately, they are easily converted into 100-µA ammeters as required here by shunting the input with a 1-kΩ resistor.

Construction

Figure 6 shows the two printed-circuit boards you need to build the power supply: one holds the seven regulators and their associated power resistors, the other, a much smaller type, the voltage control circuit.

All components, with the exception of the mains transformer, the bridge rectifiers and the reservoir capacitors, are accommodated on the two PCBs. The seven regulators are fitted on the heat-sink in a manner that allows them to be soldered direct to the printed-circuit board as shown in the photographs. Each regulator must be electrically insulated from the heat-sink. In the prototype, 1.5-mm thick ceramic insulators are used instead of the more common mica washers for reasons of safety and mechanical stability.

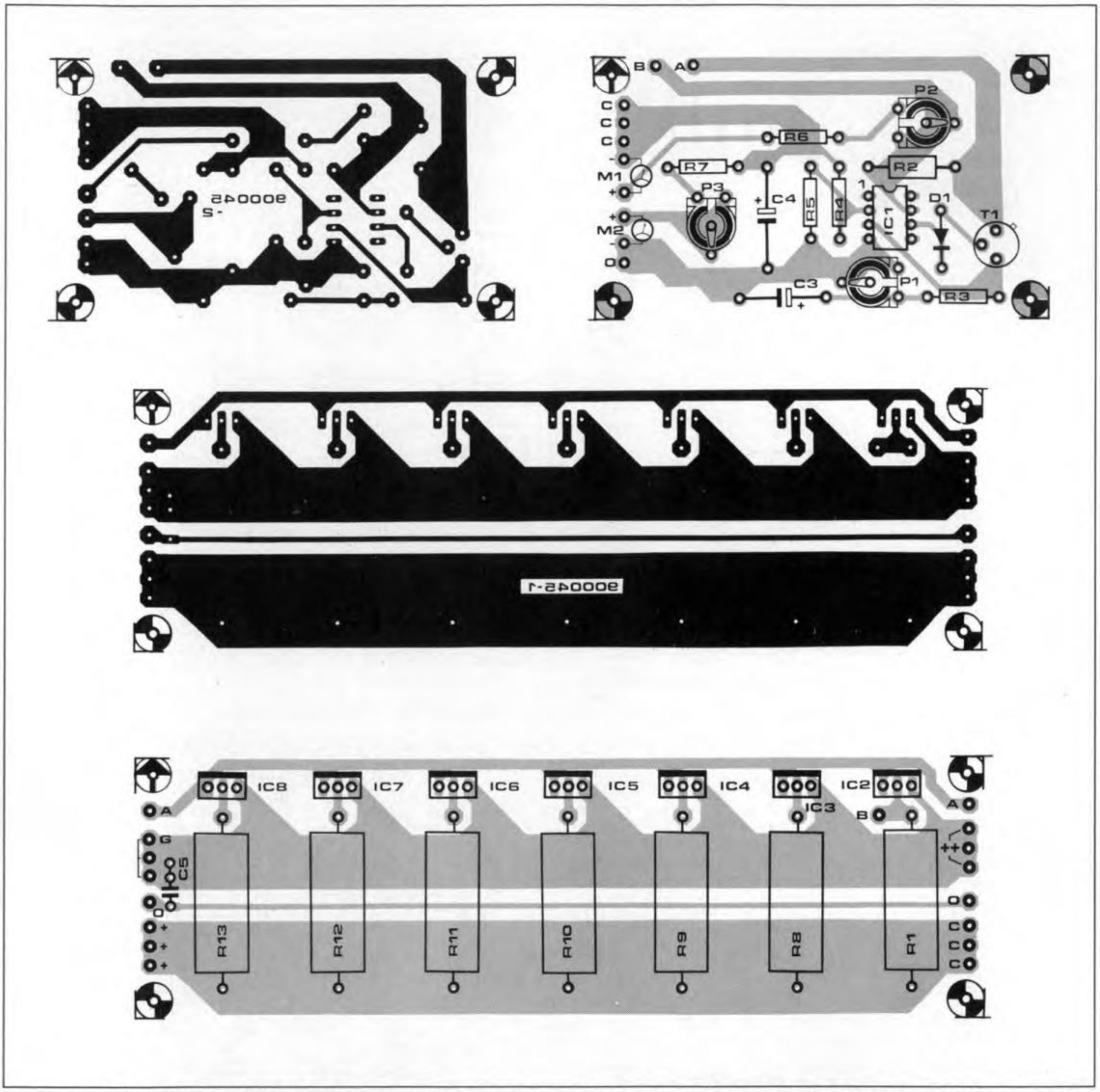


Fig. 6. Track layouts (mirror images) and component mounting plans of the two boards that go into the making of the 10-A power supply.

COMPONENTS LIST

Resistors:

7	0Ω22 3W	R1;R8–R13
1	100Ω	R2
1	150Ω	R3
2	4k7	R4;R5
1	1k5	R6
1	82k	R7
2	1k0 preset H	P1;P2
1	47k preset H	P3

Capacitors:

2	10,000μF 25V	C1;C2
1	10μF 25V	C3
1	47μF 25V	C4
1	100n	C5

Semiconductors:

1	1N4148	D1
1	BC161	T1
1	741	IC1
7	LM317 (SO-220 enclosure)	IC2–IC8

Miscellaneous:

2	100μA f.s.d. moving-coil meter, or two 200-mV LCD voltmeter modules	M1;M2
2	B80C10000	B1;B2
1	Mains appliance socket with built-in switch and fuseholder	K1
1	Fuse 3.15 A slow	F1
1	toroid mains transformer 2×15 V/7.5 A e.g. ILP Type 63013 (240V mains) or 61013 (220V mains)	Tr1
7	Insulating washer for LM317	
1	Heat-sink 1 K/W; e.g., Fischer SK120SA100	
1	printed-circuit board	900045-1
1	printed-circuit board	900045-2
1	enclosure e.g., Telet Type LC1050	

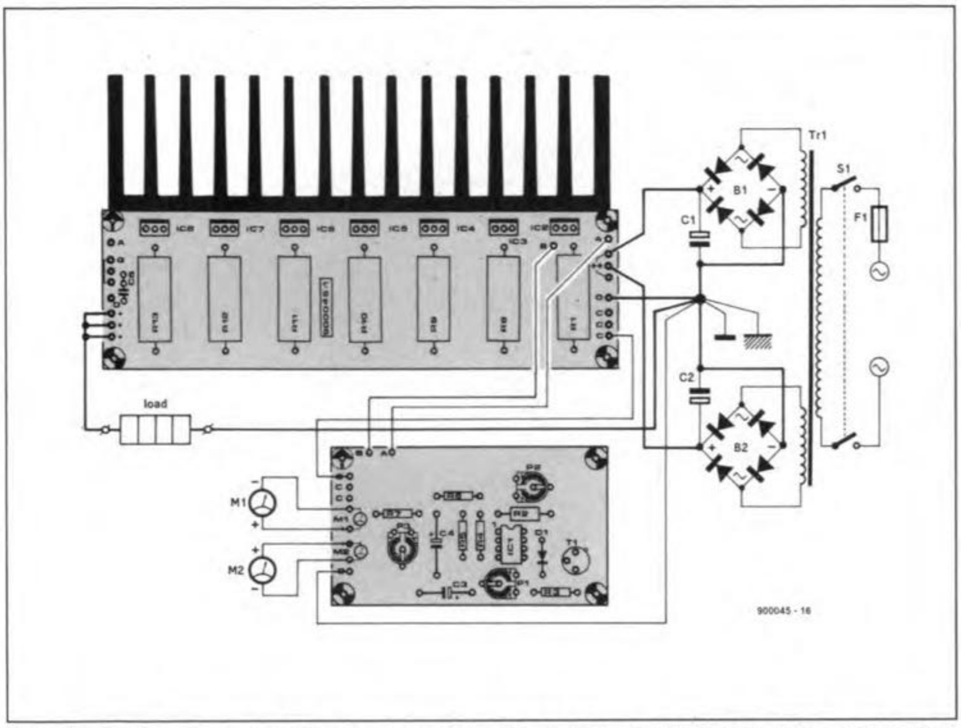
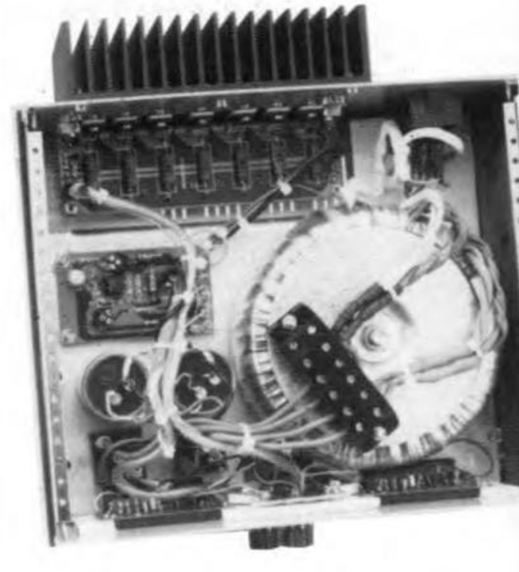


Fig. 7. Wiring diagram of the power supply. Use heavy-duty insulated wire as indicated by the heavy lines.



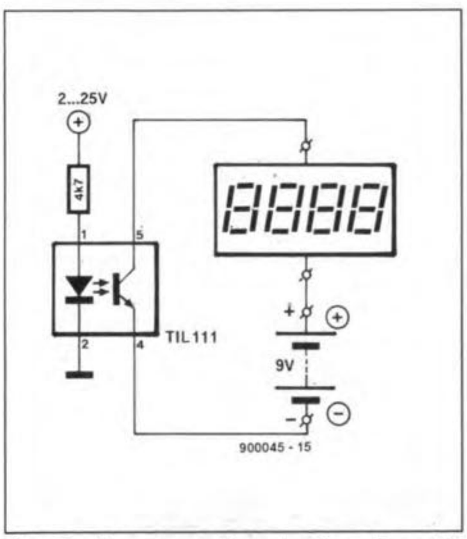


Fig. 8. Power supply and automatic on/off control for the LCD-based V/I read-outs.

The population of the two single-sided printed circuit boards is not expected to cause problems. Mount the power resistors on the regulator board at a height of about 3 mm above the PCB surface.

The completed boards are fitted in a sturdy metal enclosure as shown in the photographs of the prototype. A clearance is cut in the rear panel to enable the heat-sink with the attached regulators to be fitted.

The mains on/off switch is purposely not fitted on the front panel to avoid wires inside the enclosure that carry the mains voltage. Figure 9 illustrates the use of a mains socket with integral fuseholder and on/off switch fitted in the rear panel of the enclosure. The supply output terminals on the front panel are heavy-duty wander sockets.

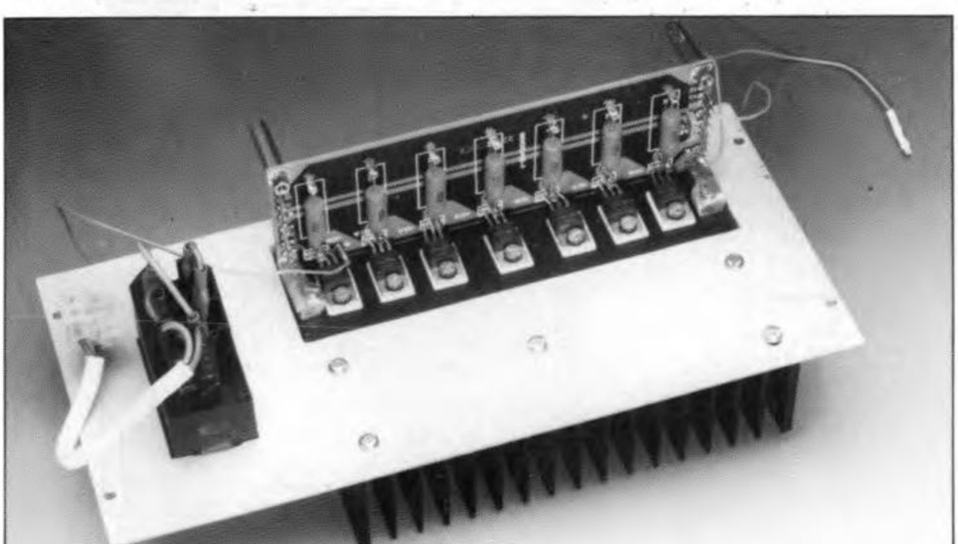


Fig. 9. Close-up of the rear panel assembly. Note the mains socket and the clearance cut for the heat-sink.

The thick lines in the wiring diagram in Fig. 7 show the connections that must be made in heavy-duty insulated wire. All other connections are made in medium-duty insulated wire. Do not forget to connect the '–' output terminal to the '0' terminals on the two boards. The heavy-duty wires are best connected to the PCBs via spade terminals and mating sockets as used in cars. When this is done, the output voltage of the supply will remain constant even at heavy loads. Our prototype was tested in this respect and found to degrade by only 60 mV when the load was increased from 0 to 10 A.

Need more than 10 A?

The power supply may, in principle, be extended with as many regulator boards Type 900045-1 as required for a particular maximum output current. Each additional regulator board increases the current by 10 A. If that is too much, simply fit one regulator per 1.4 A of additional current. The construction of the additional regulator boards is identical to that of the basic version described above. They are simply connected in parallel by interconnecting corresponding terminals, except terminal 'B', which is connected from one board 900045-1 only to the corresponding terminal on board 900045-2.

Adjustment

Switch the supply and on adjust P1 until the output voltage is 10 V. Next, connect a load (e.g., a 12-V car lamp). Use a digital multimeter to calibrate the current meter by adjusting P2, and the voltage meter by adjusting P3.

Note that when the supply is not loaded, the LM317 with the lowest reference voltage will supply the output voltage. In this condition, it may happen that current flows **into** the outputs of the other LM317s, causing the current meter to indicate a negative value. This is normal, however, and no cause for concern. ■

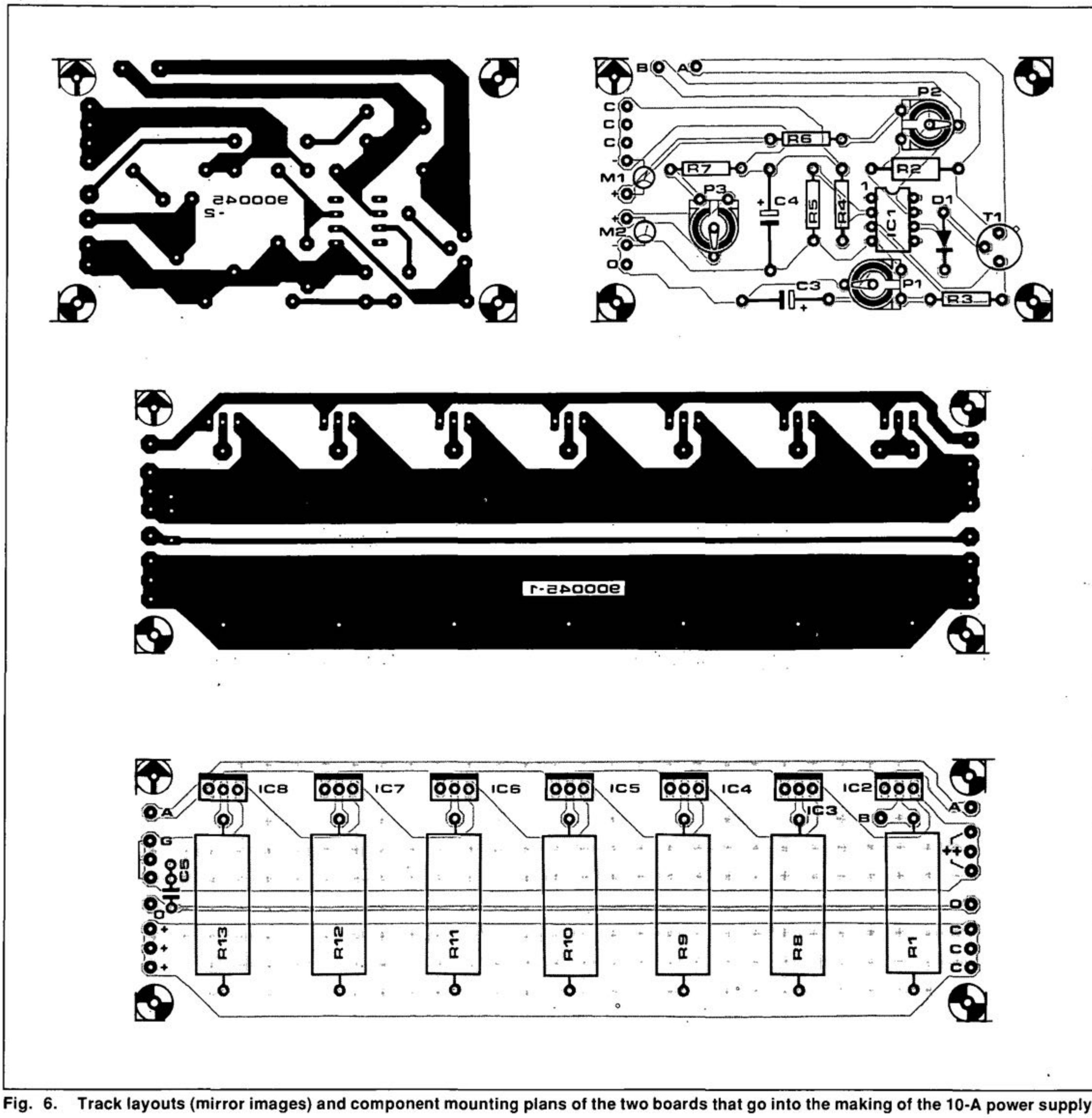


Fig. 6. Track layouts (mirror images) and component mounting plans of the two boards that go into the making of the 10-A power supply.

COMPONENTS LIST

Resistors:

7	0Ω22 3W	R1;R8–R13
1	100Ω	R2
1	150Ω	R3
2	4k7	R4;R5
1	1k5	R6
1	82k	R7
2	1k0 preset H	P1;P2
1	47k preset H	P3

Capacitors:

2	10,000µF 25V	C1;C2
1	10µF 25V	C3
1	47µF 25V	C4
1	100n	C5

Semiconductors:

1	1N4148	D1
1	BC161	T1
1	741	IC1
7	LM317 (SO-220 enclosure)	IC2–IC8

Miscellaneous:

2	100µA f.s.d. moving-coil meter, or two 200-mV LCD voltmeter modules	M1;M2
2	B80C10000	B1;B2
1	Mains appliance socket with built-in switch and fuseholder	K1
1	Fuse 3.15 A slow	F1
1	toroid mains transformer 2×15 V/7.5 A e.g. ILP Type 63013 (240V mains) or 61013 (220V mains)	Tr1
7	Insulating washer for LM317	
1	Heat-sink 1 K/W; e.g., Fischer SK120SA100	
1	printed-circuit board	900045-1
1	printed-circuit board	900045-2
1	enclosure e.g., Telet Type LC1050	

900089 900089

INMARSAT'S STANDARD-C

B. Higgins

International Maritime Satellite organisation (Inmarsat) are now marketing a package of satellite-based communications services to the business community generally. This article provides an introduction into the technical aspects of these services.

Inmarsat uses geostationary satellites in the three geographical regions of the Atlantic (AOR), Pacific (POR) and Indian Ocean (IOR) regions. For each region a working pair (known as a dual) and one spare satellite are employed. An overview of the currently used nine satellites is given in Table 1. It should be noted that Inmarsat plans to widen the Atlantic Ocean coverage towards the West, moving the Marecs B2 satellite from 26° W to 55.5° W.

What is Standard-C?

Standard-C is a satellite-based data communications system that operates at a speed of 600 bit per second. Data, in the form of computer output, or the output of various telemetry systems on board craft and vehicles, can be sent to a base station from almost any location on earth.

Standard-C is a message-based system, i.e., the sending equipment must 'pack' its data for each individual transmission. Once formatted, the data package ('packet') is stored in a message buffer until a suitable time slot is obtained on the satellite. The message buffer has a single packet .capacity of 32 KBytes.

Standard-C systems on board craft and vehicles are made with connections via an RS-232 port to allow for easy interfacing with data equipment such as personal computers, terminals, data logging systems, etc. functioning as DCE (data circuit terminating equipment) and/or DTE (data terminal equipment).

Two main categories of service are possible in the Standard-C system:

- store-and-forward message transfer
- end-to-end services

Store-and-forward message transfer involves the formatting of complete mess-

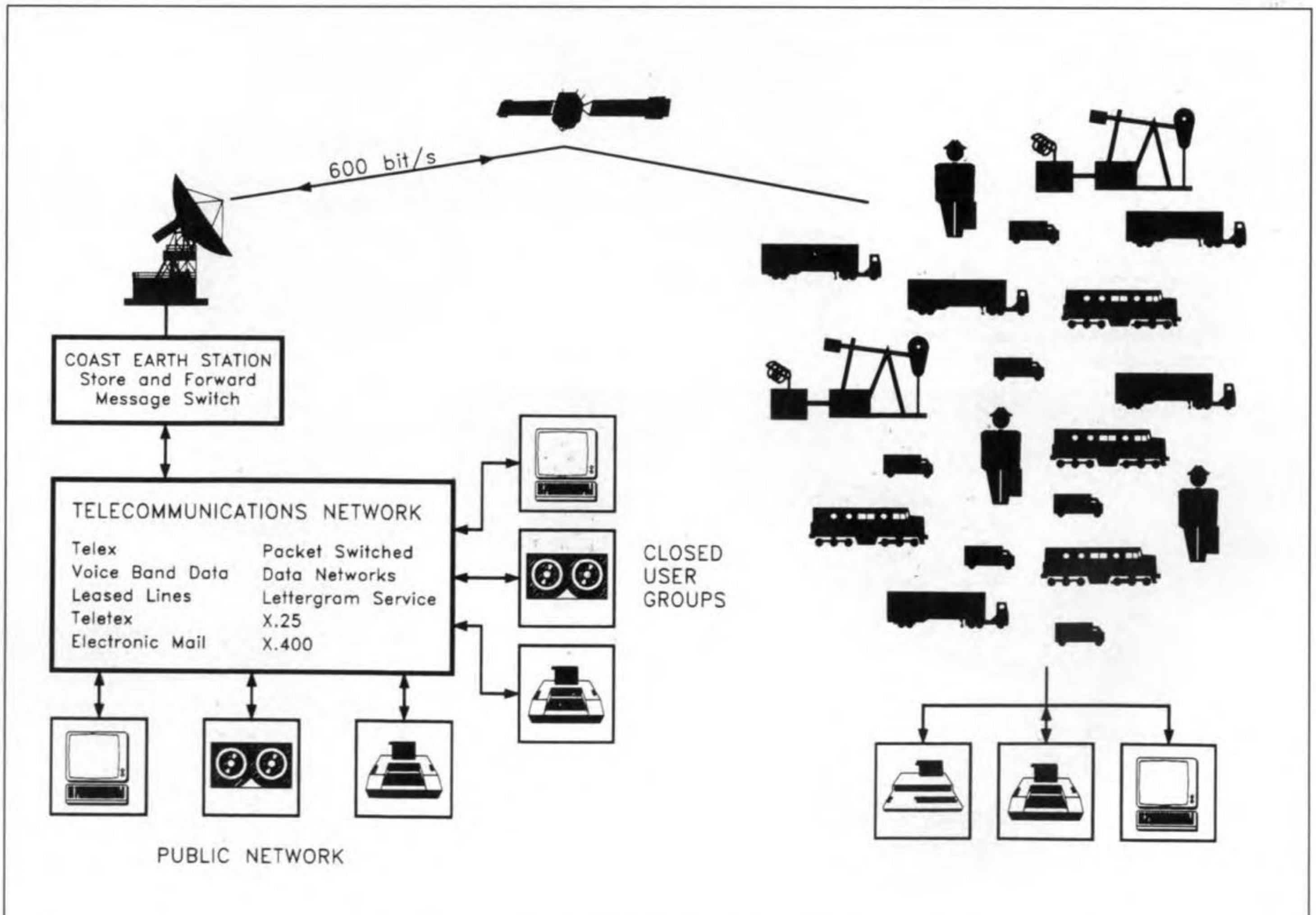


Fig. 1. Coast earth station (CES) acts as a store-and-forward message switch on to the satellite link (illustration courtesy of Inmarsat).

ages at the coast earth station or the ship earth station before transmission over the satellite channel. The (telex) messages are then transmitted to ship or to shore on a simplex basis when capacity is available.

End-to-end services require a permanent or semi-permanent circuit to be established from the coast earth station to the appropriate terrestrial facilities for the duration of the (telex) connection.

All other services provided by Standard-C are optional. These services include ship-to-shore half-duplex circuits, full-duplex circuits, polling, individually-directed, group-directed and area-directed calls, and automatic data reporting.

Standard-C uses radio frequencies in the L band and C band sections of the spectrum as shown in Table 2. Increments of 5 kHz are used throughout the bands.

Manufacturers

Around the world, interest is being shown by manufacturers of electronic equipment to produce Standard-C compatible communications sets for use at coast-earth stations and on board ships, vehicles and, recently, aircraft.

STC International Marine based at Mitchem in the UK produce equipment called Mascot C that is designed for all sizes of shipping vessels. Denmark-based Thrane & Thrane make systems for use in and on land-mobile vehicles. Table 3 shows the major equipment makers and the names of their products.

The SES: antennas and RF considerations

All manufacturers of ship earth stations (SES) use omnidirectional antennas in their systems, as this eliminates the need for the user to re-adjust for every transmission or reception.

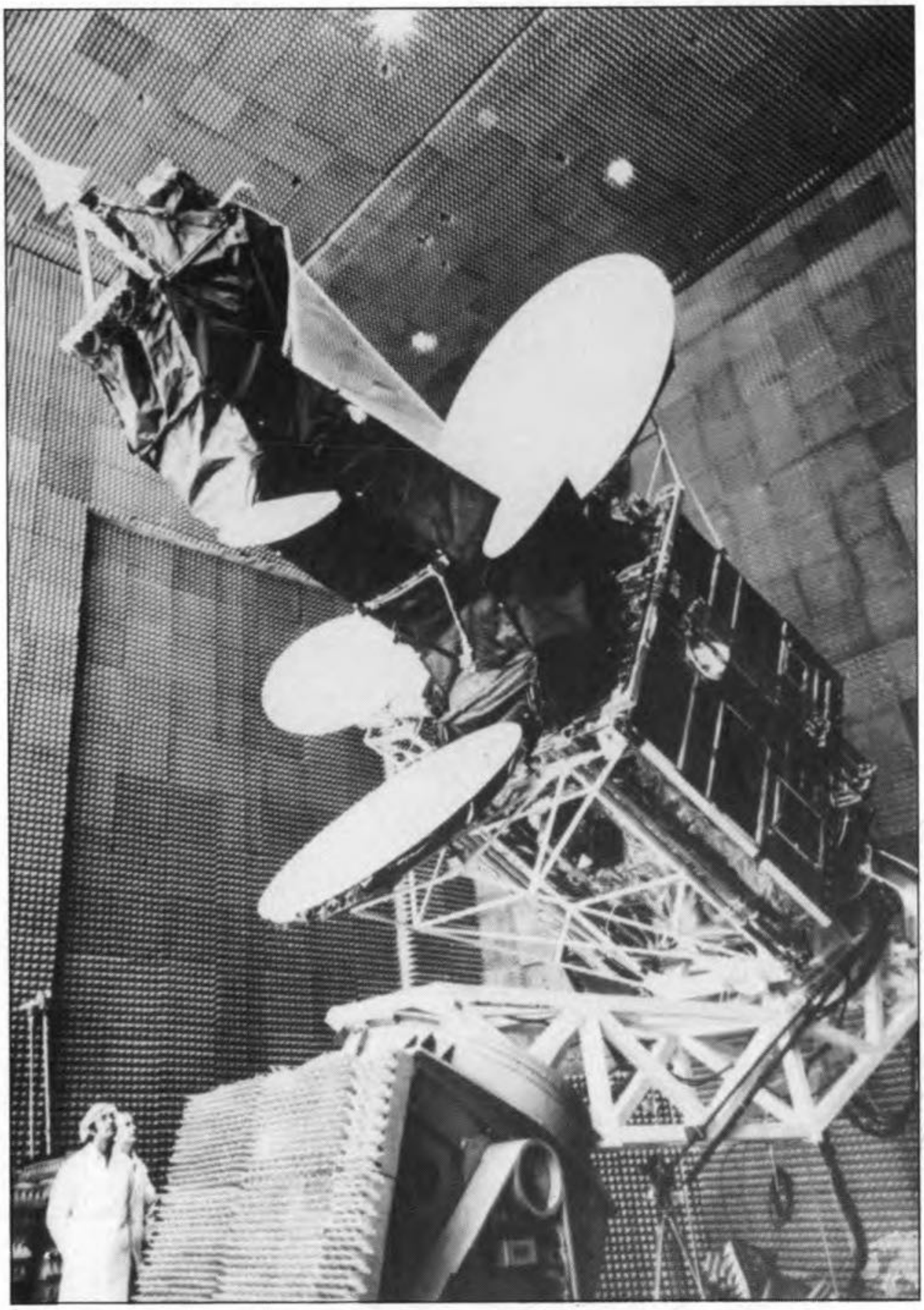


Fig. 2. Intelsat-V-F1 satellite undergoing tests in an anachoic chamber.

Antennas used in the Standard-C system must be installed at least 5 m away from C-band radar equipment. For X-band radar, the minimum distance is much smaller.

Antenna designs vary from manufacturer to manufacturer. Special designs have been adopted for use on different types of sea vessel and on land based vehicles. One innovative design stili to be tested is a flat circular antenna that can withstand an automatic car wash.

Power levels in the standard-C system are low compared to, say, satellite-TV. This is mainly because of the much smaller bandwidth requirement (typically, about 1 MHz instead of 27 MHz on a satellite-TV transponder). Circular polarisation is used to reduce the effects of Faraday rotation in the uplink and downlink paths to and from the satellite. Although the antenna gain pattern is not

Region	Longitude	Satellite	Use
AOR	26°W	Marecs B2	operational
AOR	18.5°W	Intelsat-V (MCS)	dual
AOR	15°W	Marisat F1	spare
IOR	63°E	Intelsat-V (MCS)	operational
IOR	60°E	Intelsat-V (MCS)	dual
IOR	73°E	Marisat F2	spare
POR	180°E	Intelsat-V (MCS)	operational
POR	177.5°E	Marecs A	dual

Table 1. Satellite positions and earth coverage.

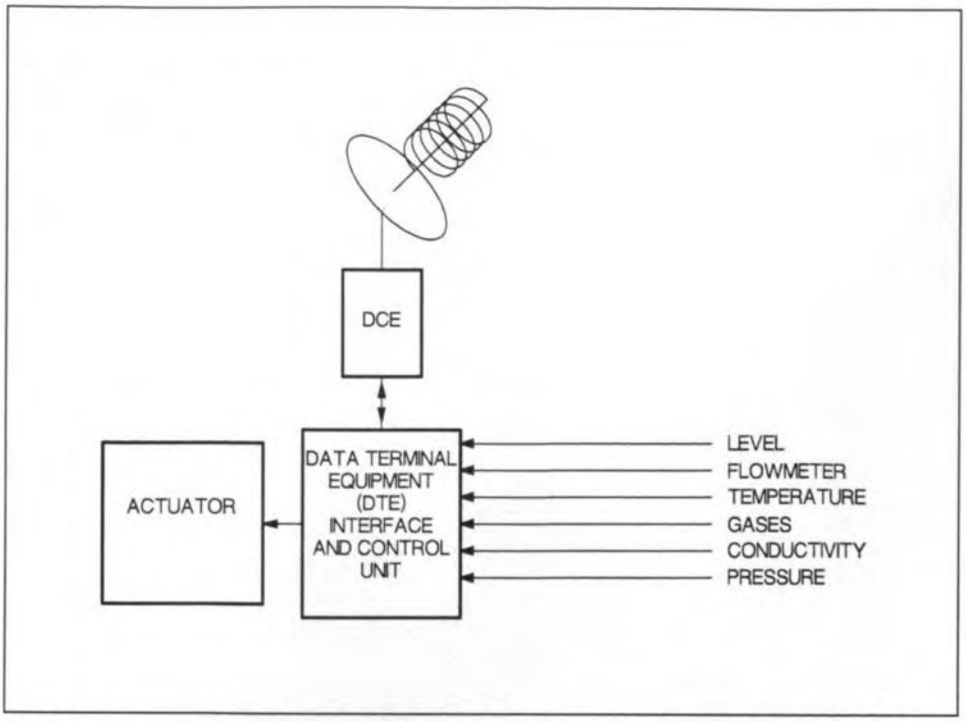


Fig. 3. Schematic diagram of Standard-C interfaces for remote data collection and control.

Use	Up/Down link	Frequency	Radio band
Coast earth station	up	6 GHz	C
Coast earth station	down	4 GHz	C
Ship earth station	up	1.6 GHz	L
Ship earth station	down	1.5 GHz	L

Table 2. Band/frequency assignment for Inmarsat Standard-C communication links.

Country	Manufacturer	Equipment name
Denmark	Thrane & Thrane	TT-3020A
France	SNEC	STC 01
Japan	Japan Radio Co.	JUE 65A
UK	STC International Marine	Mascot C

Table 3. Standard-C equipment manufacturers.

directly specified, provision must be made, e.g., with a gyroscope, to ensure that the minimum EIRP and G/T figures are met down to $-15°$ to cope with ship motion. Table 4 shows some typical RF equipment parameters.

Out at sea...

As Inmarsat is mainly concerned with maritime aspects of communication, equipment designed for the system has to conform to the rigours of operating at sea. Therefore, specifications always have a strong leaning to environmental conditions such as temperature, humidity and vibration—see Table 5.

Coast earth station

The link between the Inmarsat system and communications companies of vari-

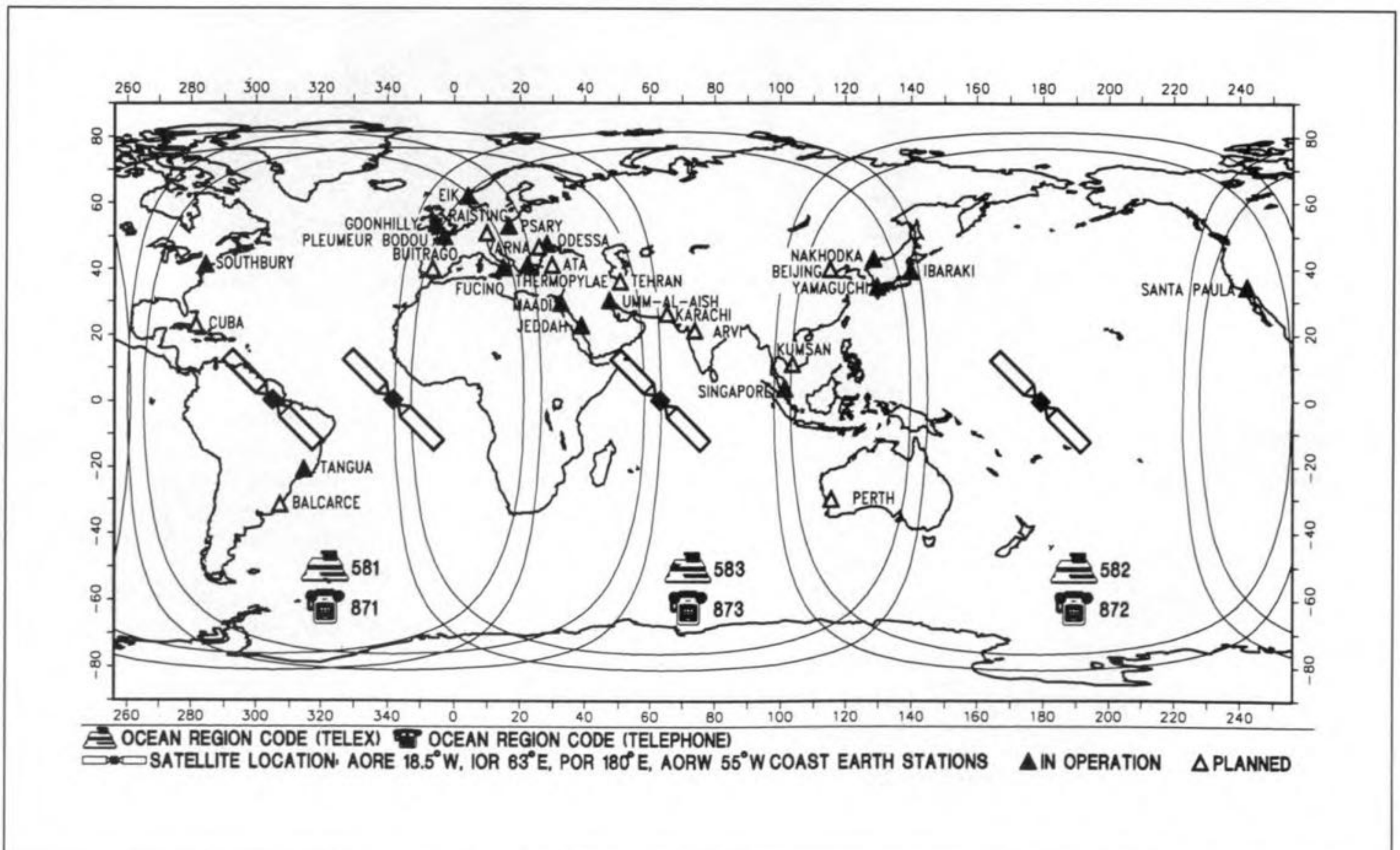


Fig. 4. Overview of frequencies used for Standard-C communications between vessels, satellites and coast earth stations.

Parameter	TT-3020A	JUE65A
Antenna polarisation	RHC	RHC
EIRP (at 5° elevation)	12 dBW (min.)	14 ±2 dBW
Figure of merit G/T (at 5° elevation)	−23 dB/K	−23 dB/K
TX frequency band	1.6265 – 1.6465 GHz	1.6265 – 1.6465 GHz
RX frequency band	1.530 – 1.545 GHz	1.530 – 1.545 GHz

Table 4. Some antenna unit specifications.

ous countries is all-important for the implementation of the complete Standard-C system.

Figure 3 illustrates how the coast earth station fits into the total system. It acts as a store-and-forward message switch to the satellite. Operating on C-band (i.e., 4/6 GHz), the link to the satellite is less susceptible to atmospheric interference as a result of problematic weather conditions.

Developments are continuously taking place in coast earth station technology. For example, Thrane and Thrane have recently been awarded contracts by Singapore Telecom and Telecom Denmark for the supply of equipment for coast earth stations in the respective countries. ■

Further information on the Standard-C communications system may be obtained from
International Maritime Satellite Organisation (Inmarsat) • 40 Melton Street • Euston Square • LONDON NW1 2EQ.

Parameter	TT-3020A	JUE-65A
Temperature (electronics)	0°C to 45°C (operating) −20°C to 70°C (storage)	0°C to 45°C
Temperature (antenna unit)	−35° to 55°C (operating) −40°C to 70°C (storage)	−35°C to 55°C
Humidity	95% non-condensing	up to 95% at 40°C
Ice	up to 2.5 cm	—
Precipitation	up to 10 cm/hr	—
Wind	up to 100 knots	—
Vibration (electronics unit)	2–15.8 Hz at 1 mm peak' 15.8–100 Hz at 1.0 g peak acceleration	2 to 15.8 Hz at 2.54 mm peak 15.8 to 100 Hz at 1.0 g peak acceleration
Vibration (antenna unit)	2–10 Hz at 2.54 mm peak 10–100 Hz at 1.0 g peak acceleration	2–10 Hz at 2.54 mm peak 10–100 Hz at 1.0 g peak acceleration

Table 5. Environmental specifications of two types of Inmarsat Standard-C SES (ship earth) stations.

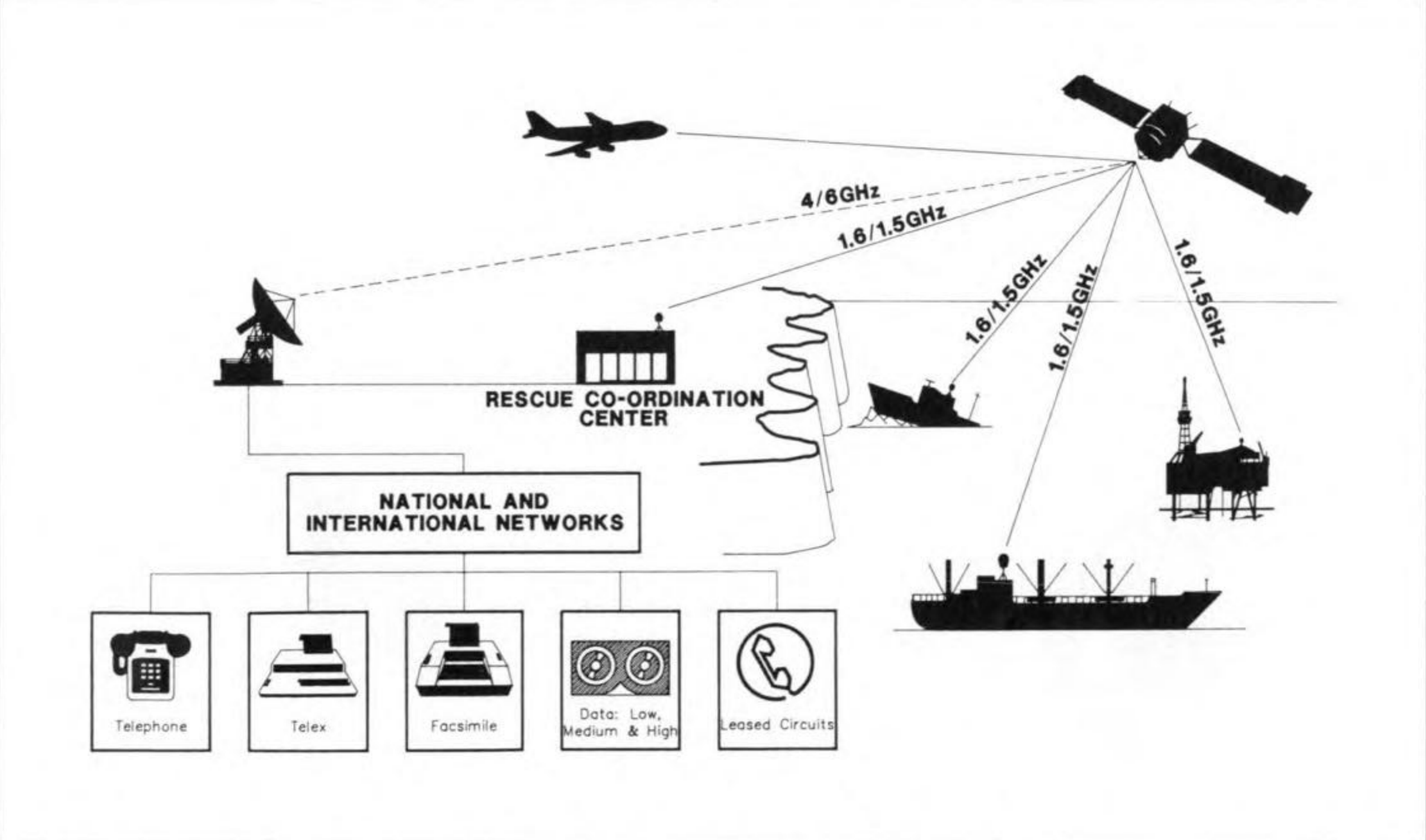


Fig. 5. Inmarsat satellite coverage at 0 and 5 degrees elevation above the horizon. The illustration shows the three ocean regions, AOR, POR and IOR, ocean codes and locations of coast earth stations (illustration courtesy of Inmarsat).

MIDI MASTER KEYBOARD

PART 2: TESTING AND PROGRAMMING

D. Doepfer

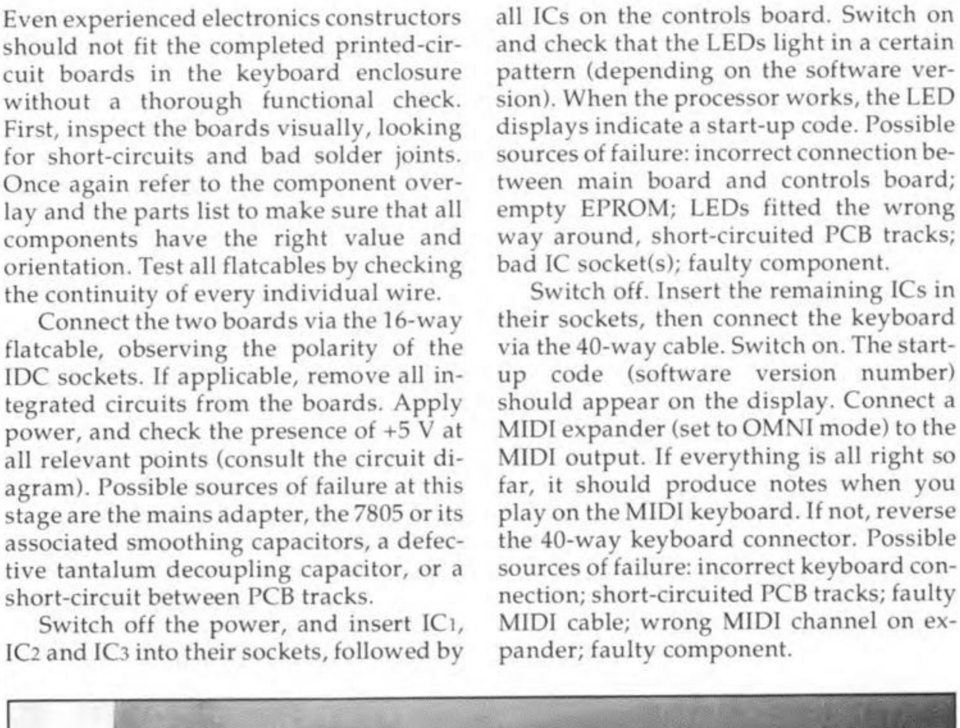
In this second and last instalment we check the operation of the two printed-circuit boards, help you on your way in faultfinding (if necessary) and last but not least discuss the various programming and menu options offered by the microprocessor-controlled operating system running on the keyboard.

Even experienced electronics constructors should not fit the completed printed-circuit boards in the keyboard enclosure without a thorough functional check. First, inspect the boards visually, looking for short-circuits and bad solder joints. Once again refer to the component overlay and the parts list to make sure that all components have the right value and orientation. Test all flatcables by checking the continuity of every individual wire.

Connect the two boards via the 16-way flatcable, observing the polarity of the IDC sockets. If applicable, remove all integrated circuits from the boards. Apply power, and check the presence of +5 V at all relevant points (consult the circuit diagram). Possible sources of failure at this stage are the mains adapter, the 7805 or its associated smoothing capacitors, a defective tantalum decoupling capacitor, or a short-circuit between PCB tracks.

Switch off the power, and insert IC_1, IC_2 and IC_3 into their sockets, followed by all ICs on the controls board. Switch on and check that the LEDs light in a certain pattern (depending on the software version). When the processor works, the LED displays indicate a start-up code. Possible sources of failure: incorrect connection between main board and controls board; empty EPROM; LEDs fitted the wrong way around, short-circuited PCB tracks; bad IC socket(s); faulty component.

Switch off. Insert the remaining ICs in their sockets, then connect the keyboard via the 40-way cable. Switch on. The start-up code (software version number) should appear on the display. Connect a MIDI expander (set to OMNI mode) to the MIDI output. If everything is all right so far, it should produce notes when you play on the MIDI keyboard. If not, reverse the 40-way keyboard connector. Possible sources of failure: incorrect keyboard connection; short-circuited PCB tracks; faulty MIDI cable; wrong MIDI channel on expander; faulty component.



When the previous tests check out, the hardware is likely to function correctly. Note, however, that errors like a stuck key may not come to light until later.

Testing the modulation wheel

As already noted in the description of the modulation wheel function, the mechanical end stops of the potentiometer do not allow this to cover the full turning range of 270°. Although this is taken into account by the control software, the period of the VCO signal (IC_4) should be made to lie in the range between 4.2 ms and 8 ms. The actual limits are 4096 μs and 8192 μs. Values outside this range produce erroneous control values.

The range limits are simple to adjust with the aid of a period meter or a frequency meter: alternately adjust P_1 (span) and operate the modulation wheel (period) until the above range is achieved.

In case such test equipment is not available, the modulation wheel may also be adjusted with the aid of routine 1 or routine 2 provided by the keyboard control software. These routines enable the microprocessor to send the MIDI control data obtained from the modulation wheel to the LED displays in two different ways.

Routine 2 is invoked by keeping the second key to the left in the upper row pressed while the keyboard is switched on. The displays indicate the measured control data in hexadecimal (00–FF). The correct range is achieved by alternately adjusting P_1 (parameter span) and operating the modulation wheel (actual parameter value). Set a sufficiently large range to allow for wear and tear, component tolerance, and drift owing to mechanical strain. Remember that the software caters for the reduction of the full range (00–FF) to 0F–F0 to cover the required MIDI range of 00–7F. This means that MIDI parameter value 00 is sent when the modulation wheel supplies a value smaller than 0F. Similarly, 7F is sent when the

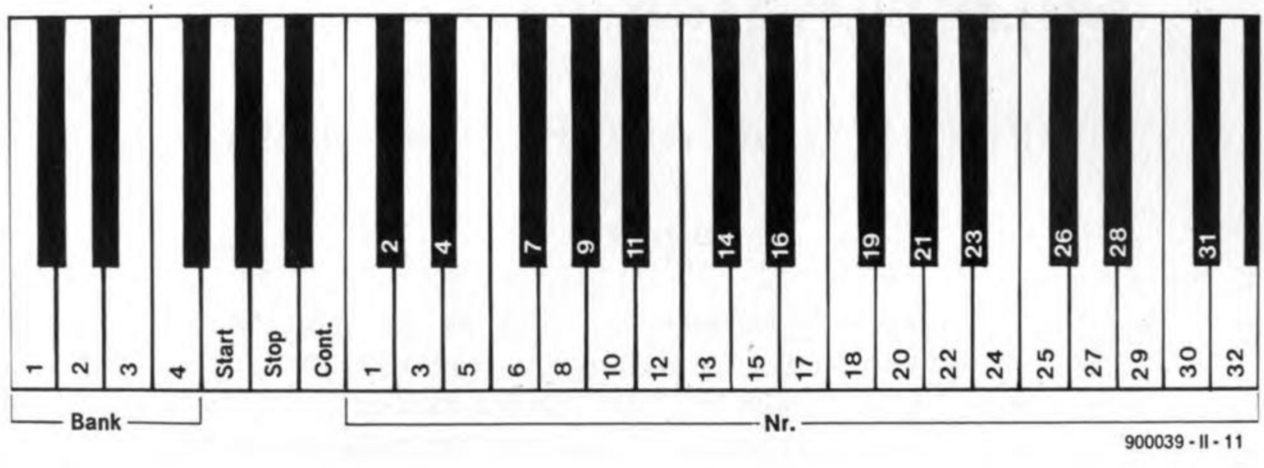


Fig. 7. A number of keys on the musical keyboard are used in conjunction with the control software to enter menu options and settings.

value exceeds F0. Hence, a range setting of, say, 0A–F2 is perfectly adequate.

Routine 1 is invoked by keeping the extreme leftmost key in the upper row pressed while the keyboard is switched on. It enables the processor to indicate the transmitted MIDI parameter value, i.e., the control data sent to the MIDI instrument after averaging and proportioning via a look-up table in the control program. Obviously, the range of this indication is 00–7F, corresponding to the permissible range of MIDI parameter values. When a self-adjusting modulation wheel is used, its rest position should produce a display reading of 40 (64 decimal), irrespective of the side from which it arrives at the rest position. If necessary ensure this centre value by carefully re-adjusting P_1 and checking the range limits with the aid of routine 2.

In the pitch-bend mode, the software provides an automatic correction function that subtracts or adds a certain value to give a range where only 40 is produced at the centre position of the modulation wheel. In case a steady value can not be achieved, fit a 2.2 µF tantalum capacitor (C_{10}) between the control voltage supplied by the potentiometer and ground. An instability of ±1 digit is allowed because it is compensated by averaging and threshold detection routines in the control program.

After the above adjustment, carefully tighten the nut on the modulation potentiometer, making sure its setting is not changed. Switch on the keyboard (without invoking either of the test routines) and check the correct range and function of the MIDI control parameters.

Programming functions

Most of the functions provided by the MIDI master keyboard are selected with the aid of the programming keys on the front panel. These keys work in conjunction with a menu, the control of which rests with the microprocessor and the software.

Connect either one of the MIDI outputs of the keyboard to the MIDI input of an expander, synthesizer, sequencer or sampler. Connect the foot switch to the keyboard via the jack socket. When so programmed and actuated in a particular keyboard zone, this foot switch forms a SUSTAIN pedal. Do not connect or disconnect the foot switch with the keyboard switched on.

Apply power. The LED displays indicate the software version number (e.g., 100), and the LEDs should light up briefly. Next, the display indicates 'PLA' for PLAY mode. At this stage, PRESET 1 is used.

The menu has the following eight options:

- preset
- program change/real time
- split
- channel
- transpose
- dynamic
- controllers
- panic

which are selected by the eight associated keys. The current option is indicated by a LED. In the PLAY mode, no LED lights. In some menus, the keyboard keys are used in addition to the control keys to set certain parameters. Also note that some menus return automatically to the PLAY mode after the parameter has been set. Most menus, however, can be left only by pressing another menu key.

In all menus, except PRESET and PANIC, the next keyboard zone is selected by pressing the same menu key once more (note: the boundaries of a zone are marked by splits). Therefore, press the menu key as many times as required to arrive at a certain zone. The selection is cyclic: i.e., zone 1 follows zone 4.

With some functions, the displays briefly shows a letter- or number-sequence before returning to 'PLA'. During these sequences, however, the play mode is active. The indication period before the display switches to 'PLA' allows set parameter values to be read off easily.

Figure 6 shows which keyboard keys are used in conjunction with the menu keys to set certain parameters. As shown in Fig. 7, the white keyboard keys in the lowest octave are assigned the following functions:

- Bank 1–4 (C; D; E; F)
- Start (G)
- Stop (A)
- Continue (B)

The number keys start at the next higher octave, i.e., C=1; C#=2, D=3, D#=4; etc. for all number entries required (program number, channel, dynamic response table, control code, etc.)

PRESET

The EPROM contains 16 presets, any one of which may be selected by the user as the basic keyboard setting. Each preset contains data on each of the four zones, the MIDI channel, transposition, assignment and actuation of control data, and the velocity table. The currently selected preset is indicated on the LED displays for a couple of seconds before the keyboard goes to PLAY mode as indicated by the letters 'PLA'.

PROGRAM CHANGE/REAL TIME

This menu allows the programs running on the connected MIDI equipment to be changed. It is used together with keyboard keys to send PROGRAM CHANGE commands and, with certain keys only, real-time MIDI commands START, STOP and CONTINUE.

The selected zone is shown to the left on the read-out, while the current bank appears to the right. The zone indication is important for the MIDI channel on which the PROGRAM CHANGE command is sent.

The programs are organized in the form of 4 banks of 32 numbers each. Bank 1 contains program numbers 1–32, bank 2 program numbers 33–64, bank 3 program numbers 65–96, and bank 4 program numbers 97–128. The banks are selected by the first four keys on the keyboard (see Fig. 7). When a bank number is changed, the new number is shown

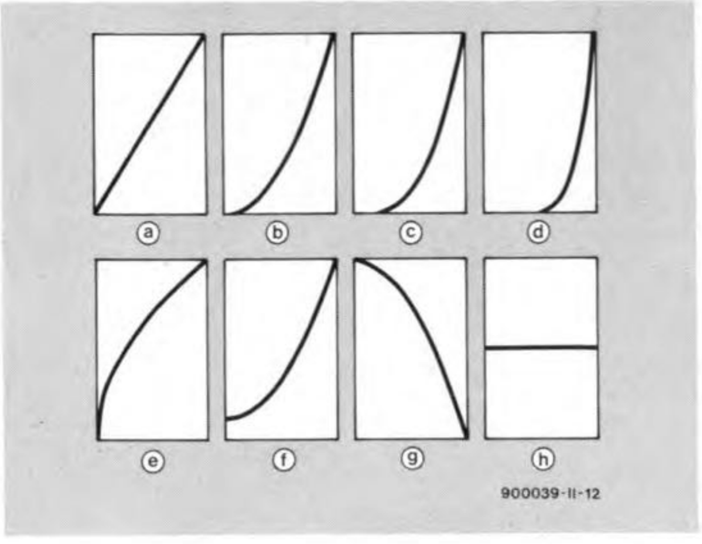


Fig. 8. Software-selected dynamic response curves. A linear response (curve 'a') often gives the impression that the dynamic characteristic can not be set with sufficient accuracy in the low-volume ranges. This may be compensated in three steps by response characteristics 'b', 'c' and 'd'. Curve 'e' provides the inverse effect, giving greater resolution in the high-volume ranges. Curve 'f' is exponential with a certain low-volume off-set. The inverse curve, 'g', gives a smaller dynamic range at greater key velocities. This response is particularly suited to velocity-controlled mixing effects: one zone is assigned a normal response (e.g., 'c'), and another zone, of equal size, the inverse response. These two zones are used to control two MIDI channels simultaneously, e.g., on two expanders, or on one expander capable of generating two sounds on two different channels. The key velocity then allows you to switch between these channels as you play. Curve 'h', finally, produces velocity parameter '64' irrespective of the key strike force. This response is particularly suited to non-dynamic sounds, e.g., those produced by organs.

on the displays. Note, however, that the PROGRAM CHANGE command is not sent until one of the number keys (1–32) is pressed on the keyboard. The transmitted program number appears briefly on the display before the keyboard returns to the PLAY mode. The program number is sent on the MIDI channel assigned to the currently selected keyboard zone.

On pressing one of the three white keys to the right of the bank selection keys (see Fig. 7), the associated real-time command START (G-key), STOP (A-key) or CONTINUE (B-key) is sent. The display shows the relevant code abbreviation, StP, StA or CON, for a couple of seconds.

SPLIT

The MIDI master keyboard is capable of transmitting on four channels at the same time. Each channel is assigned a particular keyboard zone, which may overlap to some extent depending on the size of the keyboard. This menu allows up to four zones to be defined by accepting their boundaries in the form of splits.

The current zone appears to the left on the LED read-out. The 'LO' indication to the right prompts the user to press the key that corresponds to the split at the LOw side of the zone. Next, the read-out changes to 'HI' for the HIgh split. After accepting the splits, the keyboard returns to PLAY mode.

CHANNEL

Use this menu to assign the MIDI channels to the zones. The current zone number appears to the left of the displays, the associated MIDI channel number to the right. Select the MIDI channel number with the aid of the number keys (1–16; starting with the C in the second octave). After accepting the channel selection, the keyboard returns to PLAY mode.

TRANSPOSE

On selecting this menu option, the current zone number appears to the left on the read-out, and an indication 'FI' to the right. The 'FI' indication prompts you to press the FIrst reference key for the transposition. Next, the indication changes to 'SE' for the SEcond reference key. To obtain, say, a transposition by one octave, press any keyboard key (LO), followed by the same key (i.e., note) one octave higher (HI). Similarly, a 'semi-tone down' transposition is set by pressing any key, followed by the next lower key.

After accepting the transpose reference key selections, the keyboard returns to PLAY mode.

DYNAMIC

This menu option allows the dynamic response curve for each of the four zones to be selected. The current zone number appears to the left of the read-out, and the associated dynamic response table (1–16) to the right. The basic response options are shown and explained in Fig. 8. They allow your personal keystroke to be geared to the (electronically determined) dynamic response of instruments.

Select the required dynamic response (Fig. 8) by pressing the relevant keyboard key (1–8; see Fig. 7).

After accepting the dynamic response selection, the keyboard returns to PLAY mode.

CONTROLLER

This menu option determines the MIDI control function of the modulation wheel in each of the four zones. The available MIDI functions include PITCH BEND, MODULATION, AFTER TOUCH and VOLUME. In addition, this menu determines whether or not the SUSTAIN pedal is actuated in a certain zone.

The current zone number appears to the left of the read-out, the CONTROLLER code to the right. Codes are selected by pressing the appropriate keyboard key (1–16). The codes belong with the following functions:

Wheel function	Code sustain off	Code sustain on
none	1	9
pitch bend	2	10
modulation	3	11
breath control	4	12
portamento	5	13
aftertouch	6	14
volume	7	15
pan	8	16

An example: code number '1' is selected when neither the modulation wheel nor the sustain pedal is required in a particular zone. When, however, the modulation wheel is to function as a pitch bend control, and the sustain pedal is to be used, code '10' is required. Similarly, select code '7' to make the wheel function as a volume control, while the sustain pedal is disabled.

The use of a self-adjusting potentiometer (i.e., one with a spring that retracts the spindle automatically to the rest position) is only recommended when the modulation wheel serves mainly as a pitch-bend control. In all cases, a normal potentiometer should be preferred.

PANIC

This is not really a menu, only a key that can be actuated in all modes to transmit ALL NOTES OFF commands on all MIDI channels. This function is mainly used to end hanging notes on MIDI equipment. ■

References:

1. Portable MIDI keyboard. *Elektor Electronics* November 1988.
2. Universal MIDI keyboard interface. *Elektor Electronics* June 1989 and July/August 1989.

Part 1 of this article appeared in the June 1990 issue.

MINI FM TRANSMITTER

J. Bareford

This low-power transmitter for the VHF FM band may be used as a wireless babysitter or a short-range repeater by virtue of its voice operated switch (VOX) and easy connection to a SCART outlet.

The block diagram in Fig. 1 shows that the microphone signal is amplified before it is applied to a VOX (voice-operated switch) circuit. The VOX consists of an amplifier, a rectifier, and a comparator with hysteresis. Its function is to switch the transmitter on when a predefined AF signal level is exceeded, and switch it off after a certain period, the length of which is determined by an *R-C* network.

The third block is a frequency-modulated (FM) oscillator operating at about 105 MHz.

Circuit description

With reference to the circuit diagram in Fig. 2, the electret microphone, M_1, receives its bias voltage via resistor R_1. The microphone signal is raised in an amplifier based on p-n-p transistor T_1. Choke L_1 prevents the microphone amplifier being blocked by the RF signal produced by the oscillator. The amplified microphone signal may be taken from terminal '1'. When the transmitter is used for line signals, the microphone preamplifier is not used since these signals are applied to terminals '3' (mono) or '4' and '5' (stereo).

VOX

The voice-operated switch receives the amplified microphone signal at its input, terminal '2'. Preset P_1 determines the input level and thus the switch-on threshold of the VOX. The signal at the wiper of P_1 is fed to CMOS inverter N_1, which functions as an amplifier. The next gate, N_2, functions as a limiter. The signal rectifier shown in the block diagram consists of diodes D_4-D_5, resistor R_7 and capacitor C_4. The gates that follow the rectifier, N_3, N_4 and N_5, raise the rectified signal to a level suitable for controlling switching transistor T_2. The switch-on period of the VOX is determined mainly by the value of C_4.

Depending on the signal level that exists at the input of the VOX, transistor T_2 is off or on. In that way, it controls the power supply to the RF oscillator, T_3. Choke L_2 at the collector of T_2 prevents the RF signal developed by the oscillator from being short-circuited by the power supply lines.

FM oscillator

LED D_3 lights when the VOX powers the RF oscillator based on p-n-p transistor T_3. The AF signal received at terminal '3' or terminals '4' and '5' is taken through a pre-emphasis network composed of C_5-C_6-R_{11} before it arrives on the modulation level control, preset P_2. Frequency modulation of the oscillator is achieved by C_7 superimposing the modulation signal on the bias voltage of dual varicap D_2, a Type BB204. Transistor T_3 oscillates at a frequency determined by tuned circuit L_3-C_{10}-D_2, with C_{11} providing positive feedback between the collector and the emitter.

The oscillator frequency can be set between 88 MHz and about 108 MHz by adjusting trimmer C_{10}. Capacitors C_8 and C_9 decouple the RF and the AF component respectively at the base of T_3. The transmitter has no aerial or aerial connection: its RF signal is radiated by L_3.

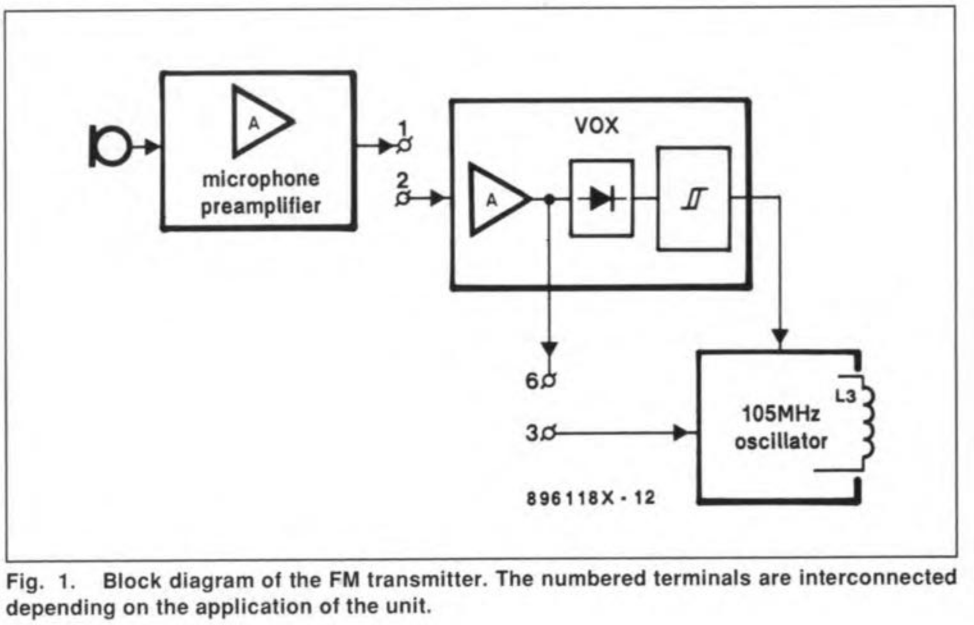


Fig. 1. Block diagram of the FM transmitter. The numbered terminals are interconnected depending on the application of the unit.

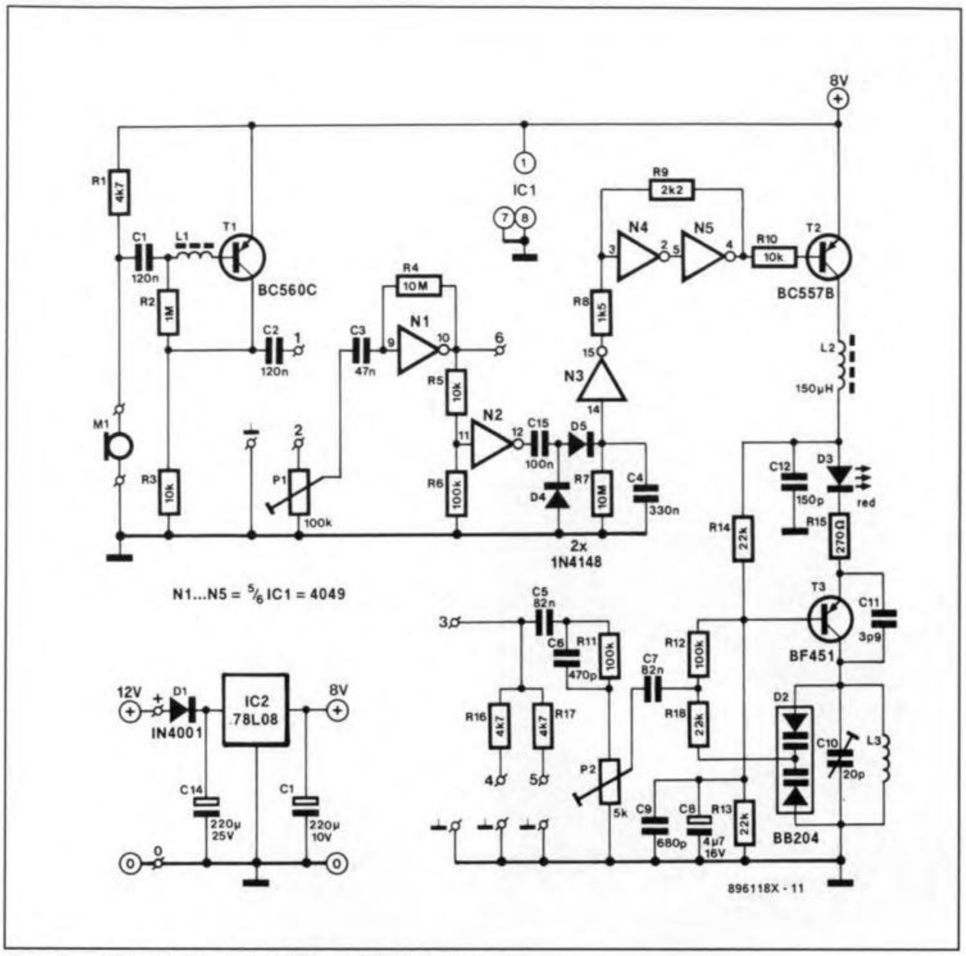


Fig. 2. Circuit diagram of the mini FM transmitter.

COMPONENTS LIST

Resistors:

3	4k7	R1;R16;R17
1	1M0	R2
3	10k	R3;R5;R10
2	10M	R4;R7
3	100k	R6;R11;R12
1	1k5	R8
1	2k2	R9
3	22k	R13;R14;R18
1	270Ω	R15
1	100k preset H	P1
1	5k preset H	P2

Capacitors:

2	120nF	C1;C2
1	47nF	C3
1	330nF	C4
2	82nF	C5;C7
1	470pF	C6
1	4µ7 16V radial	C8
1	680pF	C9
1	20pF foil trimmer	C10
1	3p9	C11
1	150pF	C12
1	220µF 10V radial	C13
1	220µF 25V radial	C14
1	100nF	C15

Semiconductors:

1	1N4001	D1
1	BB204	D2
1	red LED	D3
2	1N4148	D4;D5
1	BC560C	T1
1	BC557B	T2
1	BF451	T3
1	4049	IC1
1	78L08	IC2

Miscellaneous:

1	3-mm long ferrite bead	
1	150µH choke	L2
0.2-mm dia. enamelled copper wire		
1-mm dia. silver-plated wire		
1	printed-circuit board	896118

Resistors R16 and R17 sum stereo signals applied to terminals '4' and '5' to create a mono modulation signal.

Power supply

This is conventional, taking the form of a low-drop voltage regulator Type 78L08, IC2, with the usual decoupling capacitors, C14 and C13, at the input and output respectively. The input of the regulator is connected to a mains adapter with 12–15 VDC output.

Construction

The presence of an RF signal at about 100 MHz governs the use of the double-sided printed-circuit board (PCB) shown in Fig. 4. This board has a large unetched surface at the component side to assist in the decoupling of the RF signal. It is not through-plated: all contacts between the component side and the track side are made by component terminals.

The size of the circuit board allows it to be fitted into a compact ABS enclosure of about 8×5.5×3 cm. Start the construction by fitting the single wire link on the board. Next, mount the one IC socket and preset P1. Note that one terminal of P1 is soldered at both sides of the PCB. This is indicated by the absence of a circle on the component overlay and the fact that the copper is not removed around the hole at the component side. Do not fit P2 as yet.

Mount the remainder of the components. It will be noted that most of these are fitted vertically. Keep all component

Application	Wire(s)	AF signal to
TV sound	2 – 3	3
stereo sound	2 – 3	4 and 5 (stereo)
babyphone	1 – 2; 3 – 6	use microphone
doorbell extender	1 – 2; 3 – 6	use microphone

Table 1. The function of the transmitter is determined by wire connections as listed here.

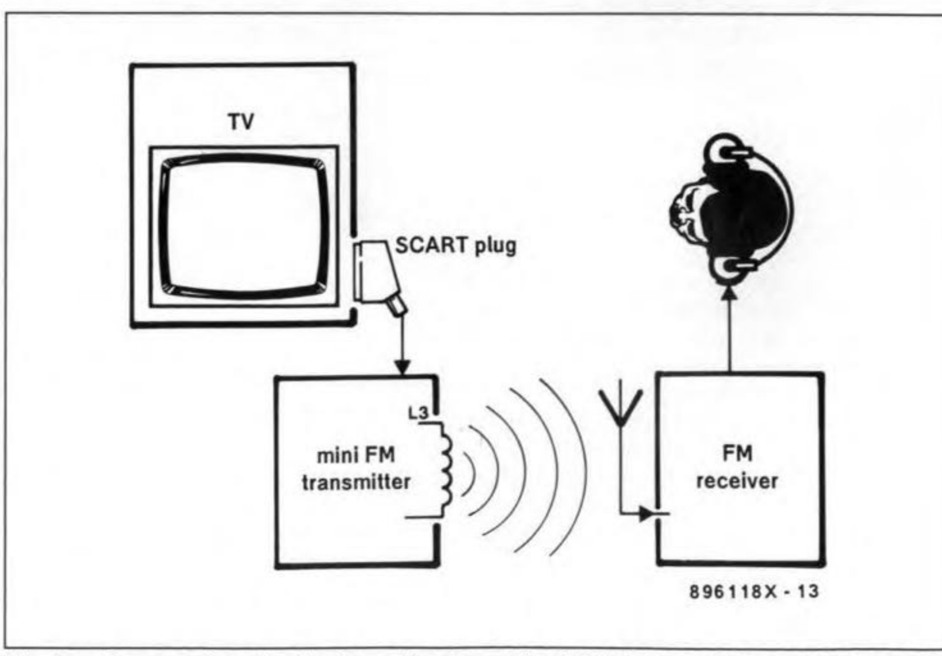


Fig. 3. Suggested application in conjunction with a SCART (Euro-AV) connection on a TV.

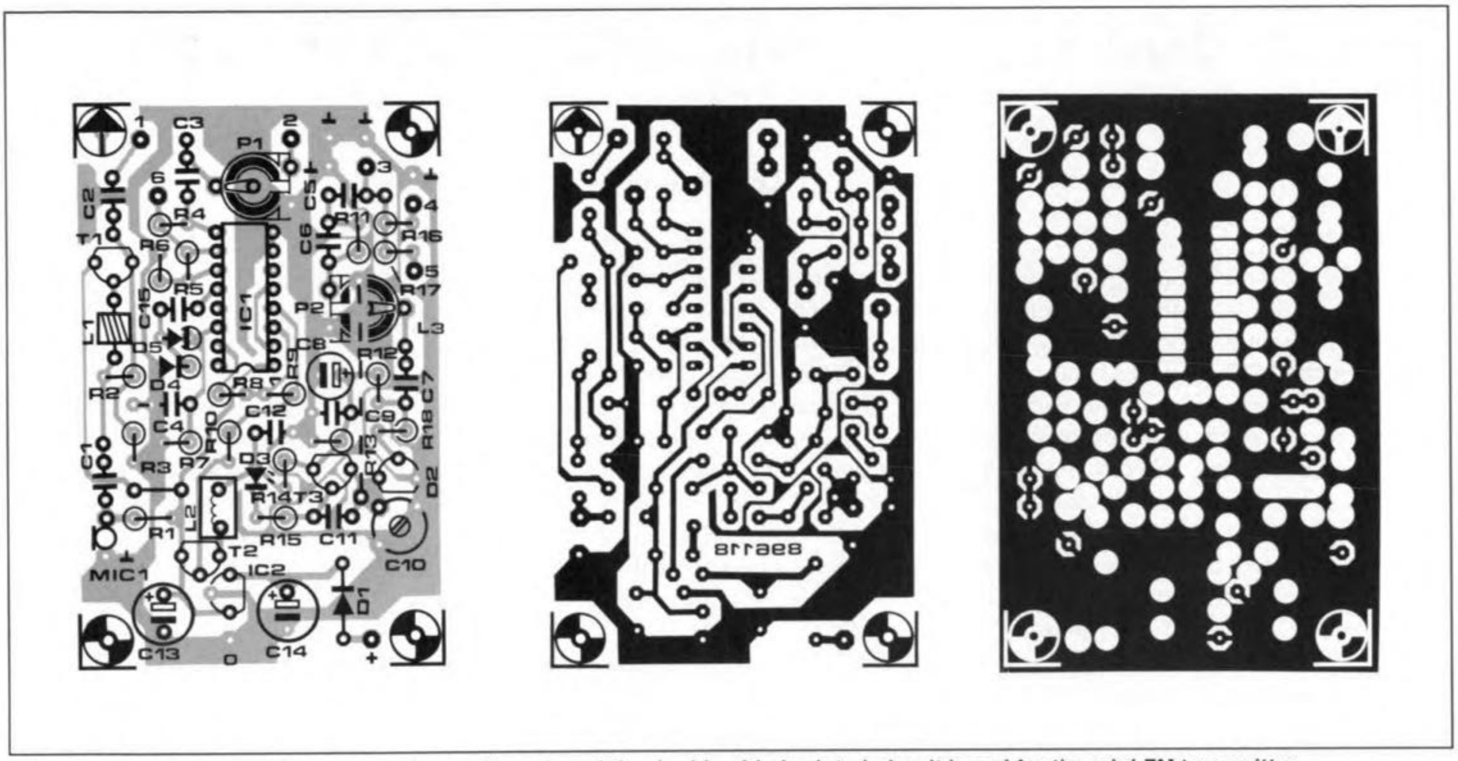


Fig. 4. Track layouts and component mounting plan of the double-sided printed-circuit board for the mini FM transmitter.

leads as short as possible. The two ground terminals of trimmer C10 must be soldered rapidly at both sides of the PCB to prevent the rotor package and the internal foil being damaged or deformed by overheating.

Home-made inductors

Choke L1 consists of 5 turns of 0.2 mm dia. (SWG36) enamelled copper wire through a 3-mm long ferrite bead. Carefully remove the enamel at the wire ends of this inductor by heating it with the solder iron while applying a little solder.

Inductor L3 appears on the component overlay (Fig. 3) as a dashed line between positions C11 and P2 to indicate that it is fitted at the track side of the PCB. The inductor consists of 4 turns of 1-mm dia. (SWG20) silver-plated wire drawn out to a length of about 3 cm. The 'cold' end of L3 is soldered to ground at both sides of the PCB. Preset P2 may then be fitted, followed by the wire links (consult Table 1).

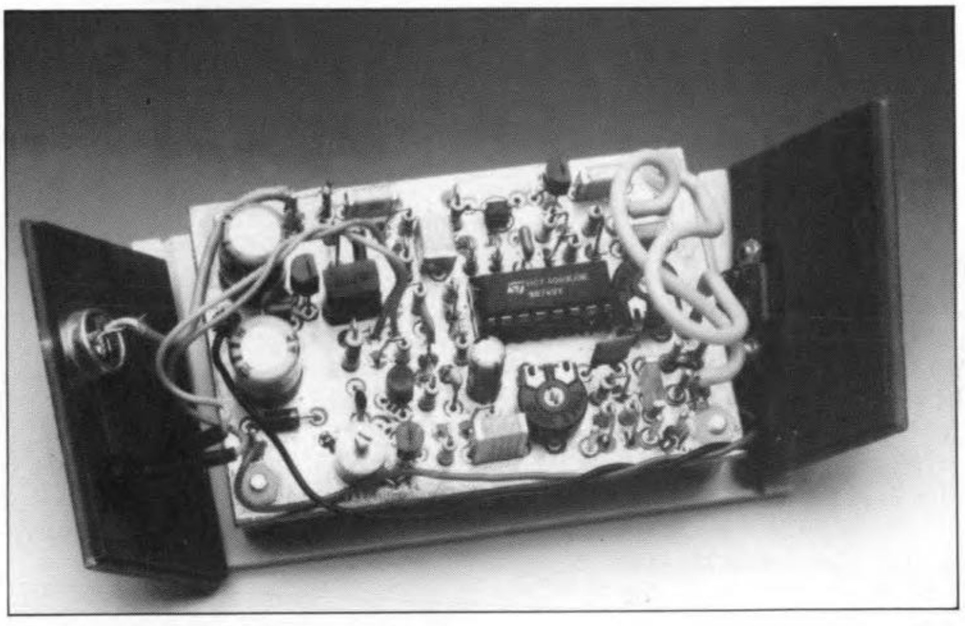
The LED and the electret microphone are fitted on the front panel of the enclosure, and the adaptor socket at the rear panel.

SCART connection

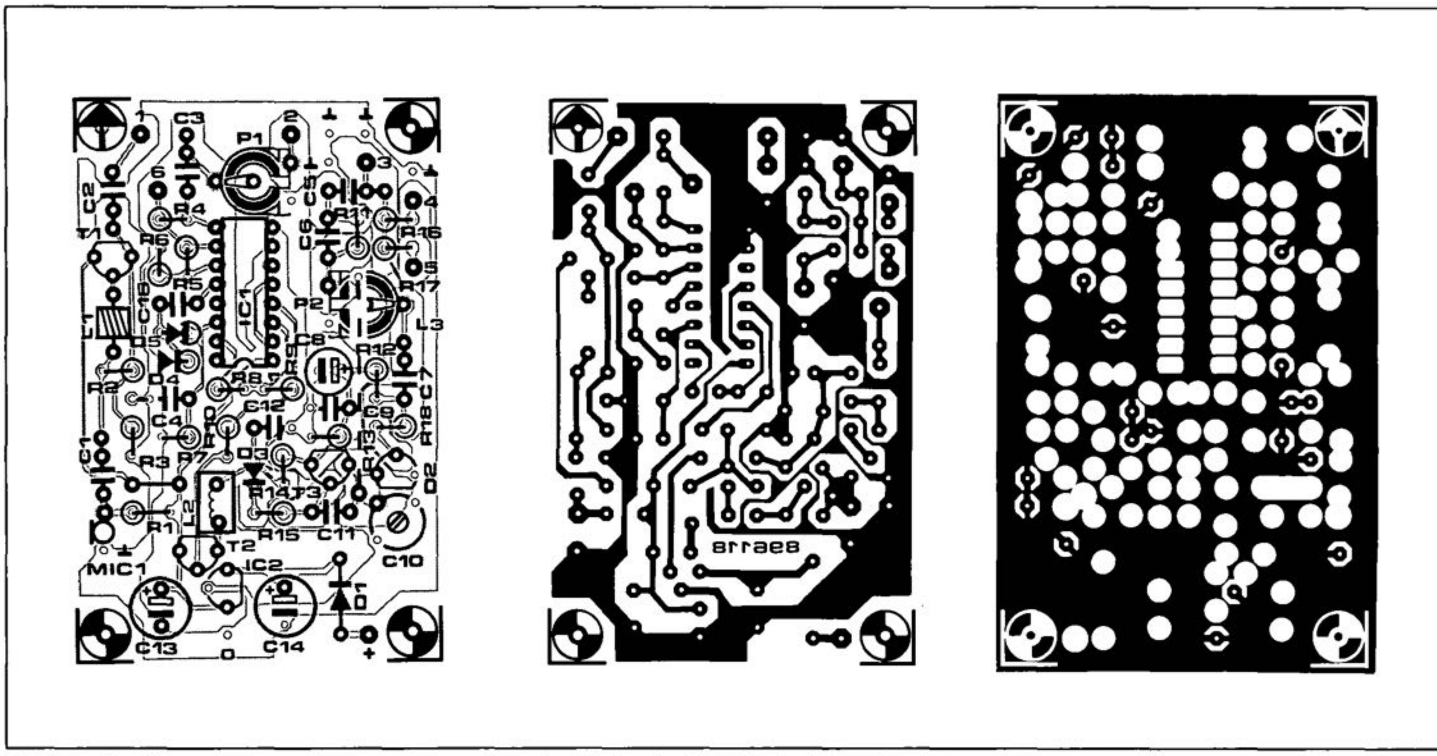
The FM transmitter may be used for 'wireless' listening to TV sound as illustrated in Fig. 3. Thanks to its small size, the unit may be fitted permanently near the TV set, connected to it via the SCART socket. The SCART connection is made via pin 3 (mono signal output) and pin 4 (audio ground).

Setting up

Set P2 to the centre of its travel and apply an audio signal to terminal '3' of the transmitter. Tune an FM receiver to a clear frequency near the upper end of the band, e.g., 105 MHz. Switch the transmitter on. Carefully adjust trimmer C10 until the signal is received. If necessary retune the receiver for minimum distortion. Next, increase the distance between the transmitter and the receiver. Adjust P2 until the optimum volume setting is obtained without running into distortion.



Change the wire link configuration to enable the microphone to be used (see Table 1), and test the VOX. Set the switch-on level to individual requirement by adjusting P1. ■



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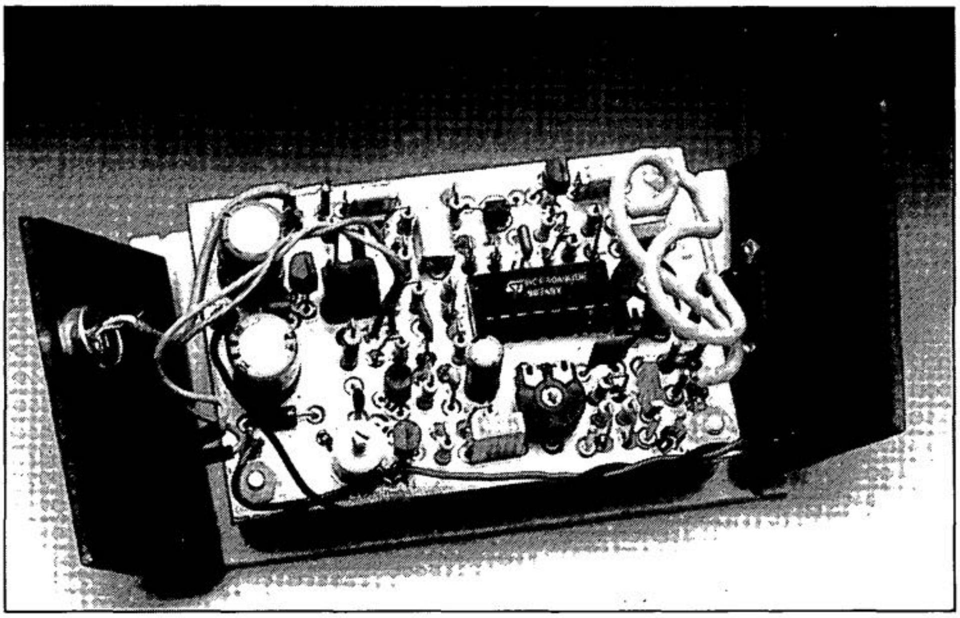

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THE 8031/8731 MICROCONTROLLER

7 geen nummer

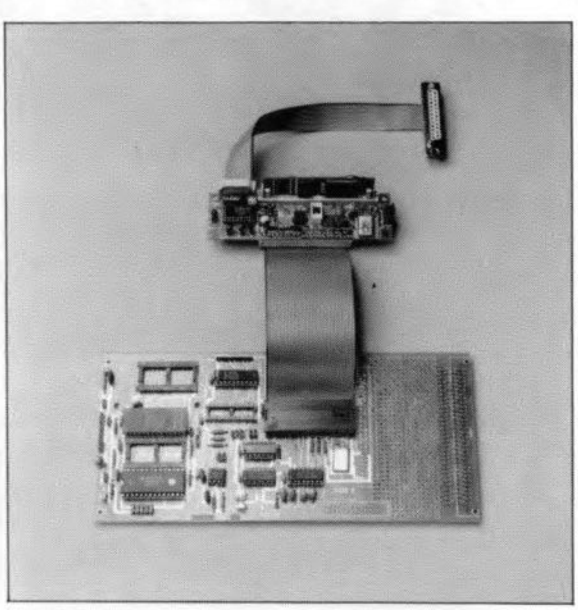
Dr David Kyte*

The single-chip micro seldom appears in hobbyist and amateur electronics projects in its true minimum form. The primary reason for this is the difficulty of developing code for the projects without specialist equipment, that is, the high cost of an ICE (In Circuit Emulator). Projects based on the 8031 microcontroller appear fairly regularly, but normally the 8031 is configured to run in its expanded memory mode. The amateur may not have encountered these terms before, and a brief description of the 8031 will, therefore, be given together with that of the EPROM version of the processor—the 8731.

The 8031 is a classic 8-bit microcontroller, with 64 Kbytes of external memory. The memory may be doubled to 128 K if the Harvard architecture is used. A Harvard architecture is defined as a separate data and program memory. The processor has an additional 128 bytes of internal memory and a variety of useful peripheral devices. The EPROM version of the processor incorporates some of the external program memory internally, starting at address 0, which is the power-up address.

If the on-chip EPROM memory (2 K, 4 K, 8 K, or 16 K, depending on part number and manufacturer) is adequate for the target application code and 128 bytes is sufficient for all the variables, program sub-routine stack, and the interrupt stack, the processor can be used in single-chip mode. As a consequence of this, the pins that were used to drive the address and data bus for controlling the external RAM and EPROM can now be used as parallel ports, that is, as general-purpose input/output. In single-chip mode, all of the parallel ports may be used, although some pins are multi-functional, i.e., timer inputs, serial receiver, serial transmitter, interrupt pins, etc. When the processor is used with external RAM or EPROM, the processor is said to be operating in expanded mode.

The average amateur application requires digital I/O, a timer, RAM and EPROM. Armed with these tools, he can tackle most controller applications. It therefore makes sense to consider the microcontroller with on-board EPROM. To build a minimum stand-alone system, a crystal, a PCB or hand-wired board, decoupling capacitors and some tie-up resistors are the only devices required. One Time Programmable—OTP—devices are standard EPROM-based microcontrollers with no quartz window, since that is used to erase the on-chip EPROM. The quartz package is expensive, but OTP devices offer low cost for low-volume production runs, prototypes and amateur applications.



Given all of the above advantages, why are there no single-chip micro projects? Simple: code development!

To develop code for single-chip micros, there are several budget related routes. The route opted by the professional is to rent or purchase an ICE. Typically, this may cost between £1,000 and £5,000, well beyond the means of the poor, old amateur.

The low-cost option is to purchase an EPROM device with quartz window. The code is written, compiled or assembled, downloaded to an EPROM programmer (additional investment may be required for the programmer) and the microcontroller EPROM programmed. The typical one-off cost of a quartz window EPROM microcontroller is £30 to £80, depending on the size of the EPROM. If a single device is purchased (minimum cost solution), when the code fails—as it surely will—the microcontroller EPROM must be erased. The EPROM erasure time (additional cost for the eraser!) will typically take 20 minutes, resulting in a minimum turn-around time of about half an hour at best.

The next problem is testing. To debug the minimum cost route, the processor is programmed, inserted into the target PCB and the power is applied. If the system works: eureka; otherwise go back to the software listing and start guessing. No real debugging information is available. The on-chip serial port can be used to provide diagnostic data, but only when the code to drive the serial port has been debugged. To configure the serial port, the baud rate, stop bits, start bits, et al, have to be initialized with the use of the 8031 special function registers, which are controlled by the timer rate register, timer mode register, serial mode register, and so on. To get all these configured correctly is for the newcomer an extremely daunting task, especially of no suitable tools are available.

Another route is to buy a version of the processor with the on-chip EPROM connected to a socket mounted piggy-back on the processor. An EPROM emulator can then be used to download the code into the PC. The cost of this route is, of course, higher, but the turn-around time is significantly shorter. The feasibility of this route is governed by the manufacturer and the company's policy on producing piggy-back devices.

The professional's life, however, is considerably simpler. To learn how to program the microcontroller, small example routines may be cobbled together, loaded into the ICE and executed. If the code fails, it can be re-executed one line at a time using a single-step instruction. By observing the memory and register contents, the action of the code may be monitored. The turn-around time for the professional is minutes. As such, learning by experimentation is positively encouraged, and the learning curve is reduced.

Dallas Semiconductor manufacture a range of 8031 microcontrollers with a difference. These devices have 8 K/32 K of

* Dr Kyte is joint managing director of MICRO AMPS Ltd.

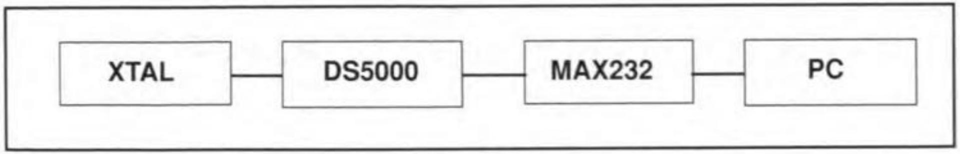


Fig. 1. Block diagram of a simple DS5000 system.

battery-backed RAM configured as internal memory replacing the internal EPROM and ROM provided by the 87C31 and 83C31 respectively. The Dallas family of devices have the part number DS5000.

The simplest circuit required to build a working DS5000 development system consists of a crystal, an RS232 voltage converter, a few capacitors, some resistors and a PCB—see Fig. 1.

The critical feature of the DS5000 is a bootstrap loader to initialize, load and configure the internal RAM. In parallel with the on-chip battery-backed RAM is a small ROM-based monitor program. The decision to boot from RAM or EPROM is determined by the PSEN line during reset. If PSEN is held low and the reset line is high when power is applied to the processor, the monitor program is selected.

The monitor offers a few simple functions. These include the capacity to download or upload Intel hex files through the serial port and configure the internal RAM as data or program memory. For those people that are sensitive about copyright protection, there is encryption hardware provided for programs stored in internal program memory. The encryption key is also defined using the bootstrap monitor. As it is unnecessary to program and erase EPROMS, the entire turn-around time for code development has been reduced significantly to a few minutes.

Unfortunately, the code development cycle is still a code, power-up and run scenario. No additional debugging information has been provided.

By executing a small monitor program residing in the DS5000 communicating with the PC through the serial port, the PC can interrogate the status of the microcontroller's internal and external RAM, special function registers, and so on. The data is presented to the user on PC; consequently, the size of the code in the microcontroller is kept to a minimum. However, sophisticated debugghing information (disassembly, memory dumping, single stepping, break points, etc.) can be generated as a sequence of transparent data transfers using the simple DS5000 monitor commands. If the DS5000 is configured in single-chip mode, and the port pins are attached to a ribbon cable terminated in a 40-pin DIL socket and an RS232 drive is connected to the serial port, an extremely low-cost ICE can be built.

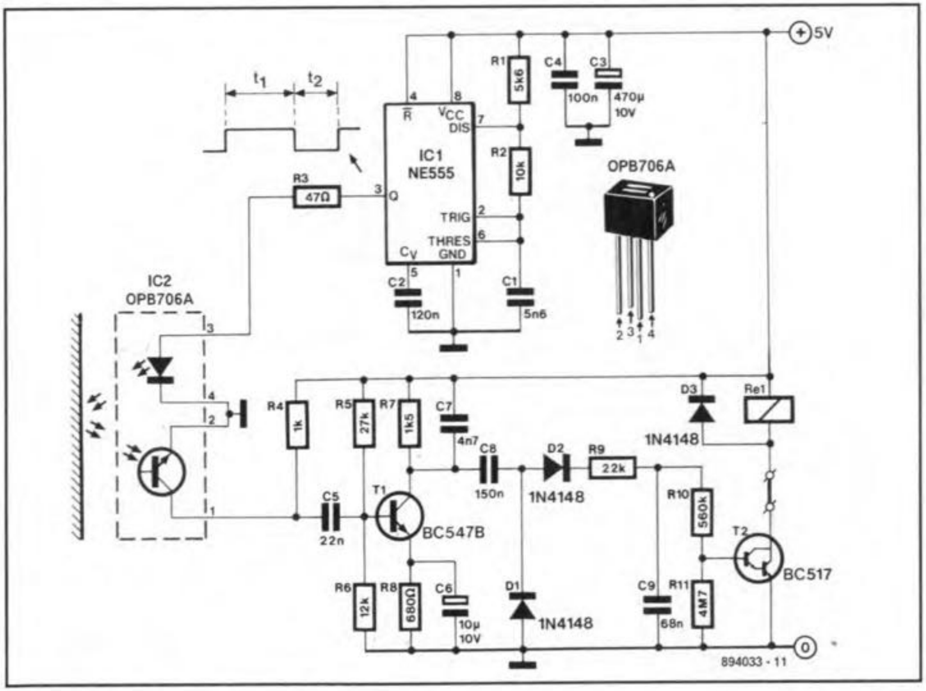
The hardware for the ICE (described in a forthcoming article) is less than £100. PC-based communications and debugging software is available for £75. Future articles will highlight features of the 8031 architecture during the description of applications. These applications and articles will probably include a dark-room timer with phonetic speech timing and a greenhouse watering system. ■

DAYLIGHT-RESISTANT OPTO-ISOLATOR

Many X–Y plotters, particularly the DIY type, have, for all sorts of reason, no protection against incident light, so that the photo-transistor in the opto-isolator can not differentiate between the light from the associated LED and daylight. The circuit shown here offers a solution to this problem.

A Type 555 timer pulses the LED in the opto-isolator at a rate of 10 kHz. If, at the receiver end, only the signal at that frequency is amplified, neither daylight nor bright artifical light can disturb the operation of the light barrier.

At a pulse rate of 10 kHz, the pulse spacing is 100 µs. If the duty factor is 6:4, the pulse width is 60 µs. At that rate, the LED can be pulsed at about 45 mA. The pulsating current is fully compensated in relation to the voltage by C3.



The on-time of the pulse signal at the Q output of IC1 is determined by R1-R2-C1 and the off-time by R2-C1:

pulse width $t_1 \approx 0.693C1(R1+R2) \approx 60$ µs;

pulse spacing $t_2 \approx 0.693C1R2 \approx 40$ µs.

The receiver in the opto-isolator, that is, the photo-transistor, is actuated by the reflected light from the light barrier and applies the consequent 10 kHz signal to a ×80 amplifier via C5. The capacitor and the input resistance of the amplifier form a a high-pass filter. The collector of T1 has a d.c. potential of 3 V. Capacitor C8 and diode D1 cause a d.c. shift, so that at the anode of D2 positive pulses with a width of 60 µs are present. These pulses charge C9 via D2 and R9. If sufficient pulsating light is reflected as, for instance, when the paper is less than 15 mm from the light barrier, the voltage across C9 is sufficient to switch on T2. The output signal is then taken direct from the collector or via the relay (do not forget D3).

The circuit draws a quiescent current of about 30 mA and an operational one of around 80 mA, excluding the relay current. ■

DECOUPLING POWER RAILS

Adequate decoupling of power rails of most circuits is a seriously underestimated necessity. Particularly in the design of printed-circuit boards, it often happens that at the last moment the thought occurs that there is no or insufficient space left for decoupling capacitors, small though these normally are. It is, of course, not surprising that such negligence often results in spontaneous oscillation in analogue circuits and unreliable operation of digital circuits. Especially sequential digital circuits, such as dividers, counters and bistables are prone to these problems, the causes of which are normally very difficult to find.

Power rails should be decoupled by a capacitor close to the relevant pins of the IC, since the rail has a certain amount of inductance. Variations in the current through this inductance cause a potential drop that manifests itself as a short pulse or spike. The capacitor serves to buffer (that is, minimize) these current transients.

Current transients arise, for example, during the switching of logic circuits, since all sorts of parasitic capacitance are charged or discharged during the change of output level. Also, just at the instant the change is taking place, the transistor in the output stage that switches to earth and the transistor that switches to the positive rail are conducting simultaneously. This means that for a very short time the power supply is short-circuited. The 50-Ω resistors in TTL-logic circuits limit the consequent short-circuit current, but 4000 and HC(T) series CMOS circuits have no such protection. It is for that reason that CMOS circuits need to be decoupled even more effectively than TTL circuits. Note that the static current, which is many times greater in TTL circuits than in CMOS circuits, has no bearing whatsoever on the degree of decoupling needed.

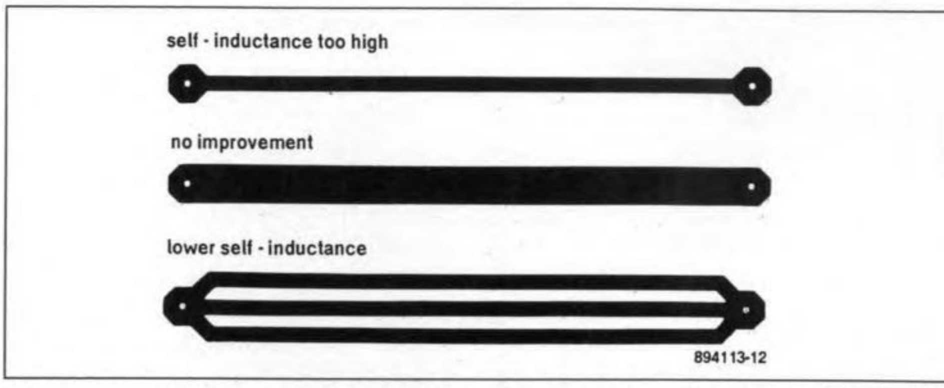


Fig. 2. The self-inductance of power rails may be reduced by using 2 or 3 rails in parallel.

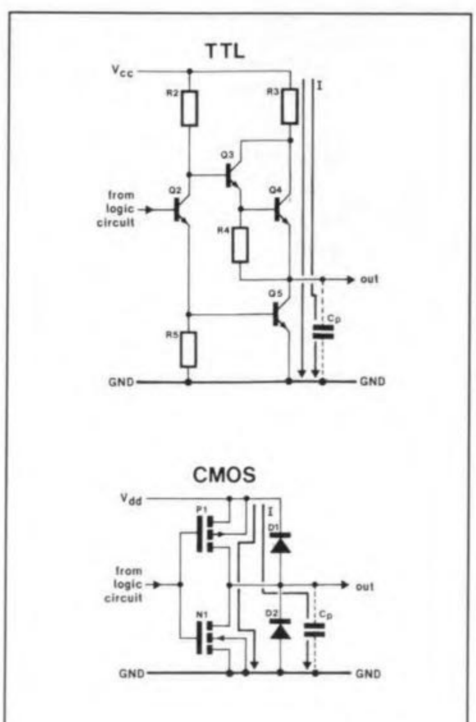


Fig. 1. Beacuse of the internal 50 Ω resistor, TTL circuits have some inherent short-circuit protection that CMOS circuits have not.

Decoupling capacitors must be connected with their terminals cut as short as feasible direct to the supply pins of the relevant IC.

The effectiveness of a number of standard types of capacitor is discussed below.

- The **wet aluminium electrolytic capacitor** has a fairly large self-inductance owing to its construction (rolled foil). Nevertheless, it performs very well as a decoupling device. Its value does not matter much: 1 μF to 10 μF are suitable values. Disadvantages are a relatively short life and fairly high leakage currents.
- The **dry aluminium electrolytic capacitor** is in the same league as the tantalum capacitor. Is life is considerably longer than that of the wet electrolytic type. Like tantalum capacitors, they are excellent for decoupling but, again like tantalum types, they are relatively expensive.
- The **tantalum capacitor**, although relatively expensive, is widely used for decoupling purposes, because of its excellent all-round properties.
- The **ceramic capacitor** is the decoupling capacitor *par excellence*. It is inexpensive, has good h.f. properties, so that relatively low values (22 nF to 100 nF) may be used, while its large tolerance and nonlinear temperature behaviour do not matter for decoupling.
- The **metallized film capacitor** (MKT, MKH, etc.) is, perhaps, too good for decoupling purposes. This is because the capacitor and the self-inductance of the power rail form an oscillatory circuit. The low losses of metallized capacitors caused underdamping of the occasional oscillations. It is interesting to note that the much higher losses of ceramic capacitors are, in this respect, a definite advantage.

Some guides for effective decoupling are given below.

- Provide each and every PCB with its own 47 μF to 100 μF electrolytic buffer capacitor.
- Both input and output of voltage regulators should be decoupled by a capacitor of at least 100 nF (positive regulators) or 220 nF (negative regulators).
- Simple logic gates should be decoupled by one capacitor of not less than 22 nF per four ICs if these are close together. More complex circuits like bistables need one capacitor for every two ICs, while counters and dividers should have one capacitor for each IC. Individual ICs should have a separate decoupling capacitor.

In addition to the use of decoupling capacitors, the self-inductance of the power rails can be reduced by two or three rails in parallel as shown in Fig. 2. Research has shown that increasing the diameter of the rails merely reduces the resistance, but not the self-inductance.

Another aspect is that the self-inductance of the rails is directly proportional to the enclosed surface. It is, therefore, better to place them close together than to separate them—see Fig. 3. ■

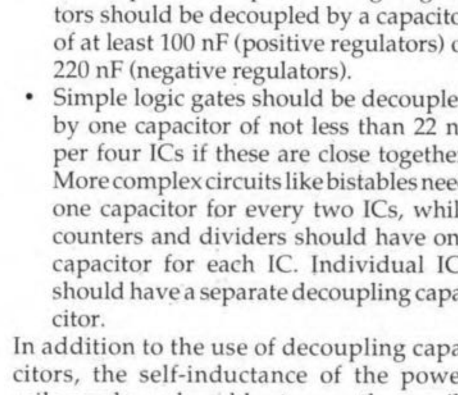


Fig. 3. Place the rails close together to minimize stray inductance.

SOUND DEMODULATOR FOR SATELLITE-TV RECEIVERS

R.G. Krijgsman PE1CHY

Since an Astra-compatible TVRO set is about the cheapest ticket to the satellite-TV arena, it is not surprising that many of you have at times aimed your 60- or 85-cm dish at TV satellites other than the Astra-1. At the same time, owners of larger dishes will agree that the price/performance ratio of most Astra-compatible indoor units (such as the Amstrad SRX200) makes them good enough for use as a second receiver. Unfortunately, problems with incompatible sound channels arise in both cases. Here's how to solve that little problem.

At the moment, there is at least one TV satellite other than Astra-1 that can be received with a 60- or 85-cm dish: the West-German DFS Kopernikus at orbital position 23.5° East. Astra-1 and DFS Kopernikus have roughly the same output power (approx. 55 dBW) and transmit in the same band. Although both TV satellites are generally classified as medium-power types, Astra has 16 channels and Kopernikus only four. More medium-power TV satellites are expected to be taken into use in the next few years.

Although the reception quality of DFS Kopernikus should be reasonable over most of the South-East of the UK, a typical Astra set will not reproduce the sound channels that accompany the four transmissions. Likewise, owners of 'full-size' TVRO equipment (1.2-, 1.5- or 1.8-metres dish with motor drive and a multi-band LNC) for reception of the low-power ECS and Intelsat TV satellites, may wonder why an Astra receiver fails to produce sound on most TV channels in spite of a strong input signal.

In the baseband

The signal recovered by demodulating a satellite-TV transmission is a frequency spectrum known as the baseband. The baseband signal is usually supplied direct by the FM demodulator in the satellite-TV receiver. It is an unfiltered, wideband signal that contains everything transmitted by the uplink via the transponder on board the satellite.

Figure 1 shows an example. The lower part of the baseband—say, up to 5.5 MHz—is reserved for the video signal. The part between 6 MHz and about 8.5 MHz contains a number of frequency-modulated subcarriers that provide the sound channels.

In the case of Astra, the frequency allocation in the baseband is fixed: the main (mono) sound channel is at 6.50 MHz, and the so-called secondary sound channels at 7.02 MHz, 7.20 MHz, 7.38 MHz and 7.56 MHz. Depending on the transponder user, these four narrow-band subcarriers may be used simultaneously for mono-, multi-lingual as well as stereo sound transmissions (an exception is Filmnet on channel 11, which uses the 7.56 MHz subcarrier for TV-sync pulses).

Unfortunately, the sound receiver and demodulator circuits in many Astra recei-

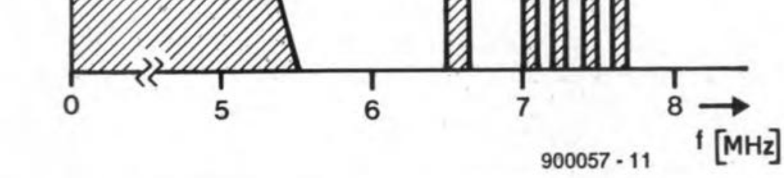


Fig. 1. Typical five-subcarrier frequency allocation in a satellite-TV baseband spectrum. The video information occupies the frequency range from 0 to about 5.5 MHz; the main (mono) sound channel is a wideband FM subcarrier at 6.5 MHz; narrow-band FM subcarriers for stereo transmissions and the like are assigned fixed frequencies of 7.02 MHz, 7.20 MHz, 7.38 MHz and 7.56 MHz.

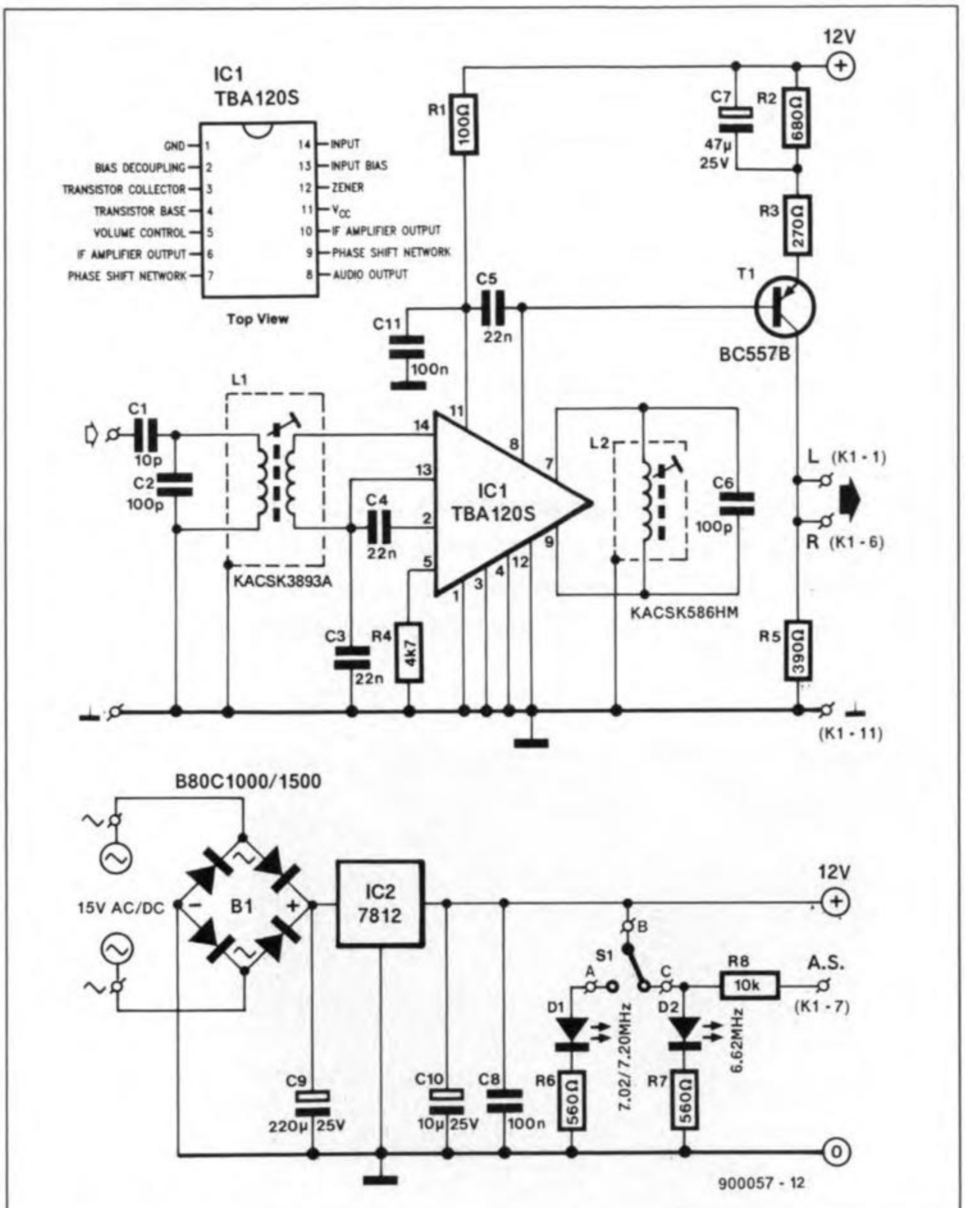


Fig. 2. Circuit diagram of the sound demodulator. The inductors, L1 and L2, are tuned to the relevant sound subcarrier frequency in the baseband (6.60 MHz, 6.65 MHz, or 6.62 MHz).

1	audio return L	500–700 mV into 10kΩ
2	video return	1 Vpp into 75Ω
3	video control voltage	0 V or +12 V
4	baseband output	25 Hz – 10.5 MHz (–3 dB)
5	video output	1 Vpp into 75Ω
6	audio return R	500–700 mV into 10kΩ
7	audio control voltage	0 V or +12 V
8	ground	
9	spare	
10	spare	
11	ground	
12	audio output L	500–700 mV into 10kΩ
13	audio output R	500–700 mV into 10kΩ
14	spare	
15	spare	

Fig. 3. Pinning and signal assignment of the INTERFACE connector fitted on many Astra-compatible indoor units.

vers use the secondary channels only, i.e., they can not be tuned to the main channel at 6.50 MHz. This seldom causes problems, however, because the main channel is usually duplicated (in mono or stereo) on one or more secondary channels.

A problem arises when an Astra set is used to receive, say, DFS Kopernikus, since all transmissions on that satellite have a single, mono, sound channel at 6.60 MHz in the baseband (6.65 MHz is also used on some ECS and Intelsat channels). Even if the Astra receiver does have a 6.50 MHz filter and demodulator, this does not produce the desired sound channel because of the frequency difference of about 100 kHz.

Upgrade kits: UNsound!

Alerted by a small opportunity in the already crowded satellite-TV market, some TVRO retailers supply upgrade kits claimed to make Astra receivers such as the popular Amstrad SRX200 compatible with ECS, Intelsat and Kopernikus transmissions. The performance of these kits is, in general, poor owing to an inherent design error.

Although these kits do provide sound reception and FM demodulation at 6.50 MHz in the baseband, the (wideband) main sound channel is passed through the filters originally fitted for the secondary channels. Obviously, these filters are too narrow for this application, since they were designed for a bandwidth of about 50 kHz, not 150 kHz and greater. To make things even worse, the designers of these upgrade kits, in an attempt to reduce the cost of their product, have 'forgotten' or overlooked that the audio signal is subsequently fed through a special de-emphasis section and a Panda-Wegener dynamic range expansion circuit, both of which are definitely inappropriate for the wideband signal. The result is a lot of distortion and a 'thumping' sound that lacks high-frequency components.

A better approach

There can be no question that the separate sound demodulator shown in Fig. 2 is a cut above the previously discussed upgrade kits. The circuit is simple to build, inexpensive and based on commonly available parts. Furthermore, it is connected as an external unit, so you do not even have to open your receiver (provided this has a standard 15-way interface socket).

Unlike many upgrade kits, the sound demodulator presented here is designed to handle wide-band FM subcarriers and provide the correct de-emphasis. The result is a clean, undistorted sound channel.

The circuit diagram shows a standard application of the excellent, nearly 20-year-old, TBA120S FM demodulator. The baseband signal is applied to the chip via inductor L1. The quadrature detector on board the TBA120S is tuned to 6.60 MHz (or 6.65 MHz) by L2-C6. Resistor R4 determines the output volume, and capacitor C5 the de-emphasis characteristic. A direct-coupled buffer, T1, feeds the demodulated signal to the receiver.

Set to the position indicated in the circuit diagram, switch S1 causes the receiver to feed the audio signal at the L and R terminals of the demodulator to the inputs of its internal AF amplifier, and from there to the UHF remodulator and the SCART socket. This selection is achieved by applying +12 V to pin 7 (AUDIO SWITCH; A.S.)

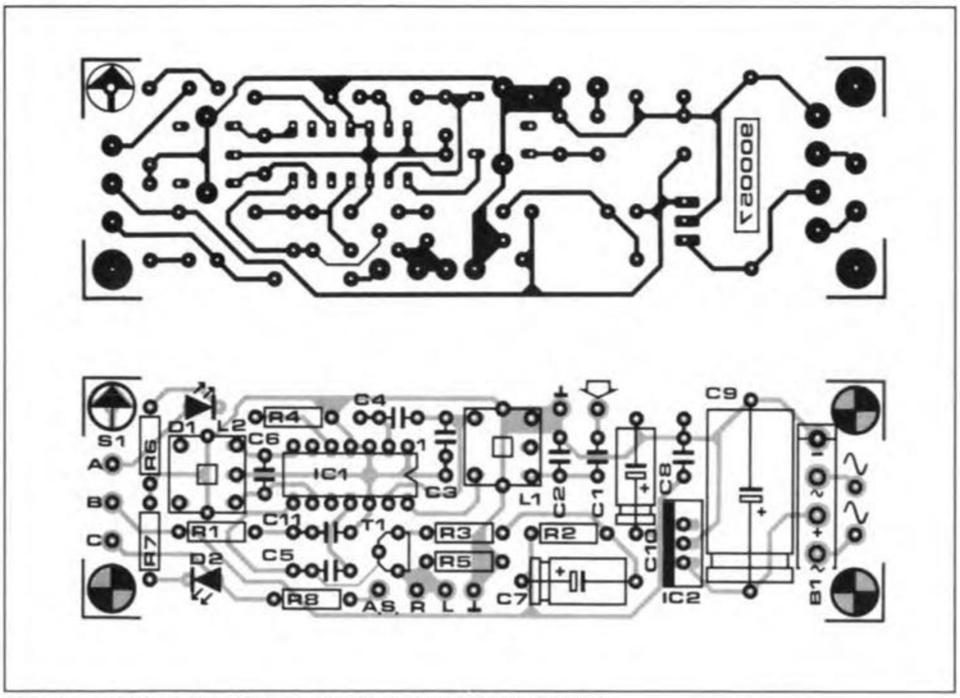


Fig. 4. Printed-circuit board for the sound demodulator.

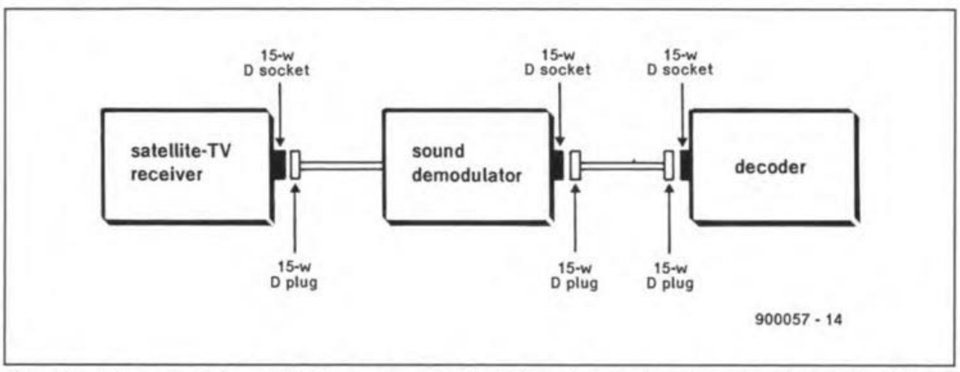


Fig. 5. The sound demodulator and a decoder, if used, may share the signals on the 15-way INTERFACE socket on the indoor unit as shown here.

COMPONENTS LIST

Resistors:		
1	100Ω	R1
1	680Ω	R2
1	270Ω	R3
1	4k7	R4
1	390Ω	R5
2	560Ω	R6;R7
1	10k	R8
Capacitors:		
1	10pF	C1
2	100pF	C2;C6
3	22nF	C3;C4;C5
1	47μF 25V axial	C7
2	100n	C8;C11
1	220μF 25V axial	C9
1	10μF 25V axial	C10
Semiconductors:		
1	B40C1000 or B80C1000	B1
2	LED 3mm red	D1;D2
1	TBA120S	IC1
1	7812	IC2
1	BC557B	T1
Miscellaneous:		
1	KACSK3893A	L1
1	KACS(K)586HM	L2
1	miniature SPDT switch	S1
1	15-pin high-density sub-D connector (male; not on PCB)	K1
1	printed-circuit board	900057

of the 15-way D-socket at the back of the receiver. The pinning of this (standardized) connector is shown in Fig. 3. LED D2 lights to indicate that the sound channel is produced by the external sound demodulator via the audio return inputs of the receiver.

When S2 is set to the other position, LED D1 lights, and the receiver uses its own audio signal.

A standard 12-V power supply based on bridge rectifier B1 and regulator IC2 completes the unit. The unregulated input voltage (15–18 V, a.c. or d.c) is supplied by a mains adapter.

Construction, alignment and connecting-up

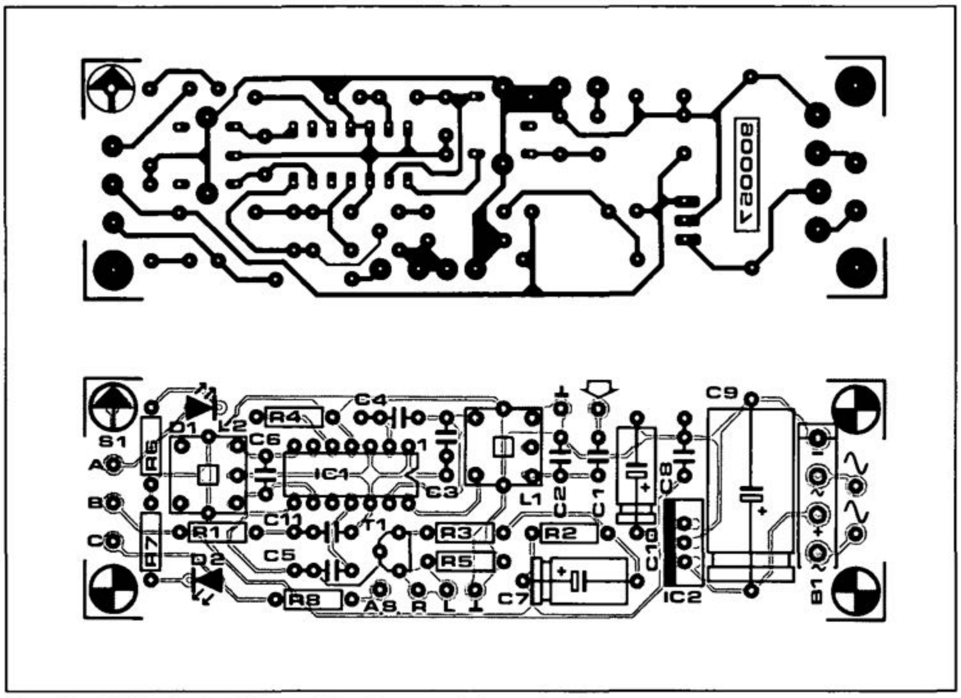
The circuit is best constructed on the printed-circuit board shown in Fig. 4. The completed board is fitted in a narrow ABS enclosure, with the mode switch and the LEDs protruding from the front, i.e., the short, side as shown in the photograph. The connection to the receiver is made via a 3- or 4-way screened cable (max. length: approx. 50 cm), the INTERFACE socket, and a mating plug (they do exist: ask your satellite-TV dealer).

The baseband signal, taken from pin 4 of the INTERFACE connector, is fed to the input of the demodulator via one of the screened conductors in the cable. Since the load formed by the sound demodulator is negligible, this connection may be made in parallel with a decoder. It is also possible to fit a 15-way INTERFACE socket on the rear panel of the sound demodulator enclosure. Since the audio return inputs are not (yet) used by decoders, this construction enables a loop-through connection between the receiver, the sound demodulator and a decoder as shown in Fig. 5.

The three (or two) outputs of the sound demodulator, L, R and A.S. (or just L and A.S.) are connected to the INTERFACE plug via separate conductors in the screened cable.

Alignment is fairly simple. Apply power and set S1 to external sound (D2 lights and the receiver probably produces noise). First adjust L2, then L1, until the desired audio channel is heard. Use an insulated trimming tool, never a common screwdriver, and remember that the tuning frequency of the inductors is **lowered** as you turn the cores **upward**. This is because the windings of both the KACSK586HM and the KACSK3839A inductors are located towards the top of the metal enclosure.

The demodulator may be tuned to 6.62 MHz to enable sound channels at 6.60 MHz as well as 6.65 MHz to be received without the need of re-adjustment. The tuning range of the prototype extended from 6.0 MHz to 8.5 MHz at an input sensitivity better than 1 mV$_{pp}$. Current consumption was about 40 mA at an input supply voltage of 15 V. ■



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2	100n	C8;C11
1	220µF 25V axial	C9
1	10µF 25V axial	C10
Semiconductors:		
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2	LED 3mm red	D1;D2
1	TBA120S	IC1
1	7812	IC2
1	BC557B	T1
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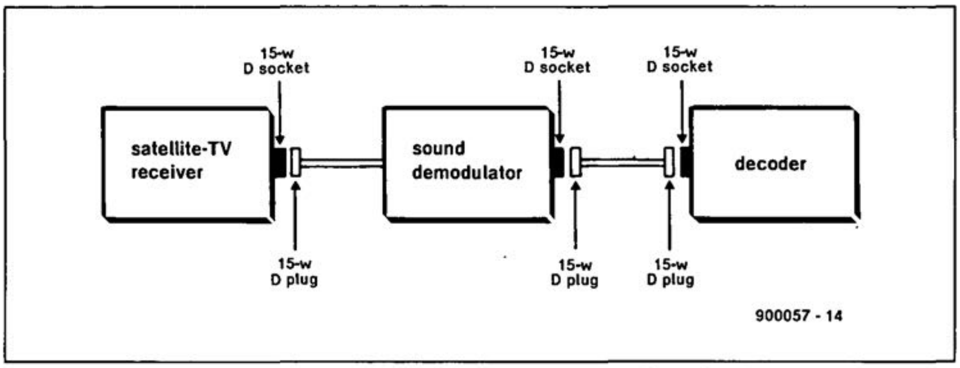

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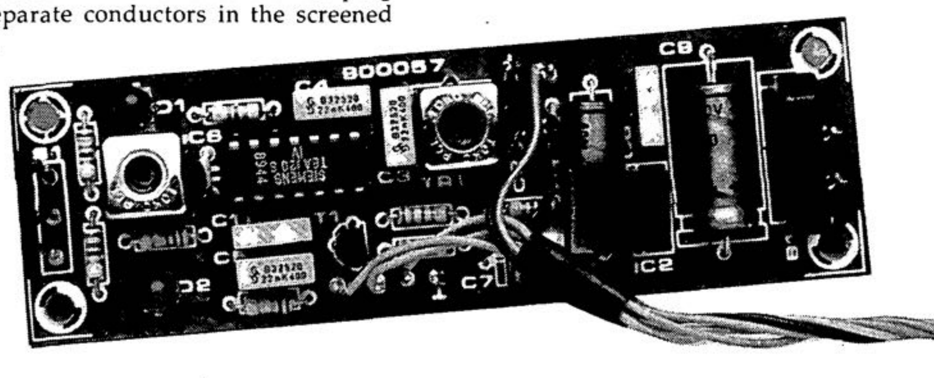

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PROFILE: HAPPY MEMORIES

by Bernard Hubbard

Electronics hobby enthusiast Geoffrey Greaves got tired of paying 'high prices' for components when building his own computer in 1978, so he decided to find an inexpensive source and sell the surplus to finance his hobby. And thus Happy Memories, now a highly successful electronics mail order business, was born.

Geoffrey told Elektor Electronics recently: "Initially, I was just buying components for my own projects and then selling the surplus via magazines."

However, very soon he had asked a friend to source cheap components in the USA for items which his customers were demanding, even though he was not interested in those components for his own projects. A hobby had become a true business.

"A turning point was reached with the 2114s," he recalls, "I was buying so many that I realized what had just been a hobby had become a business and I had better take it seriously."

Soon the nascent business had established a shop in a part of Southampton favoured by other electronic retailers. "Maplin bought a shop very close to us," he recalls.

The business grew so much that he had to employ a manager, while Geoffrey continued with his job in social services.

Long-term, however, Geoffrey and his wife, Diana, had nurtured the dream of raising a family in the country and they saw in the direct mail electronic components business a chance to finance a green lifestyle in the country far from the hustle and bustle of big-city life.

Ten years ago they took the chance of foresaking the city when they bought a run-down cottage in the Marcher country bordering Wales.

Today, they are living their green dream in another cottage on a hillside some five miles from the small market town of Kington in Hereford and Worcester. It is from here, with hens and ducks, two dogs and a collection of cats, spurred on by the demands of four children, soon to be five, that Geoffrey and Diana conduct their highly efficient, 24-hour dispatch electronics business.

Says Geoffrey: "The postman delivers the orders to the door one day and takes the components away with him the next, but if there is an urgent request we drive into Kington and the order goes that same day."

Despite a decade in the business, Geoffrey retains his hobbyist's enthusiasm and is often found talking to a customer on the phone for up to 20 minutes. He says: "This is the social services part of the job. The small customers normally take the most time and result in small orders, while some of the larger firms know exactly what they are looking for and we just get repeat orders."

Their individual customers range from all walks of life. "Many are middle-aged men," Geoffrey says, "but there are also a lot of calls from schoolboys. Often mum will ring up on their behalf with an order which she is reading from a piece of paper without having a clue what she is ordering. Often she pays, so the hobby doen't make too big a hole in the schoolboy's pocket money."

Orders come in by post, fax or phone, from all over the country, but there is also a growing amount of business from overseas, particularly from Scandinavia.

Despite a very commercial attitude to the Happy Memories business, Geoffrey and Diana have strict principles in not doing business with certain firms on ethical grounds, for instance, with arms manufacturers or oppressive regimes.

Asked how the mail order business first arrived at its name, Geoffrey said candidly: "I needed some more copy to fill up all the white space on one of my first magazine advertisements and it suddenly came to me. All those who know anything about computers immediately grasped the double meaning of 'Happy Memories'."

"The great advantage of having your business next to your home," says Geoffrey, "is 'going to work' in your bedroom 'slippers. ■

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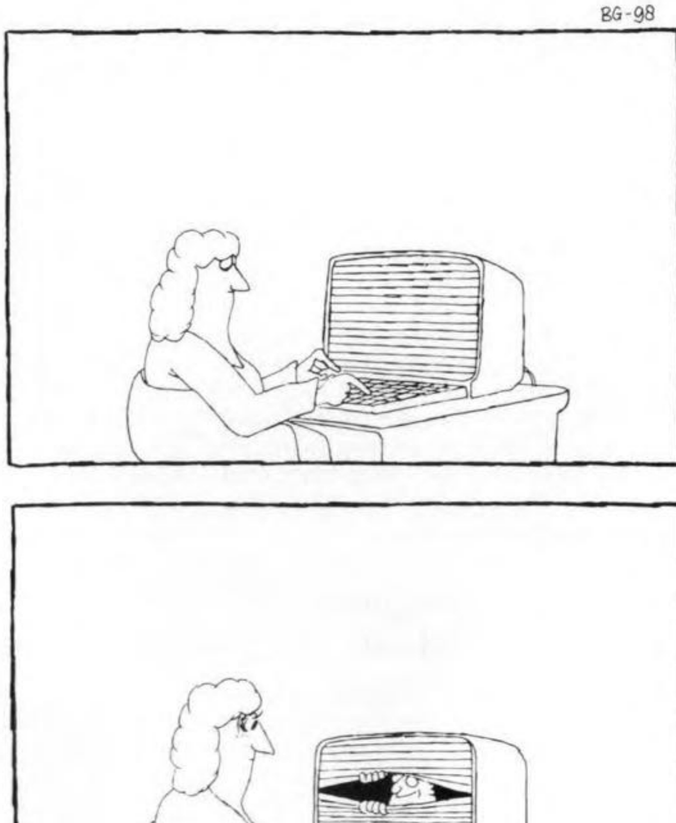


INTERMEDIATE PROJECTS

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.

VERSATILE Ni-Cd BATTERY CHARGER

M.E. Theaker, BSc, CEng, MIEE

As the price of primary batteries continues to rise, so the popularity of rechargeable batteries continues to grow. Of all rechargeable batteries, nickel-cadmium (Ni-Cd) types are justly the most popular; they are easily interchangeable with standard-size primary batteries and can be stored in any state of charge without damage.

The economics of rechargeable batteries are indisputable: a Ni-Cd battery will provide at least 500 cycles of charge and discharge. An AA-size Ni-Cd battery has a capacity of 600 mAh, so during its life will provide a minimum total capacity of 300,000 mAh, equivalent to 167 alkaline manganese batteries (each having a capacity of about 1800 mAh). The cost of an AA-size Ni-Cd battery is about £2 and that of an alkaline-manganese battery about 50p. To the cost of the rechargeable battery the cost of a battery charger (about £10) must be added, but since the charger will probably be used to charge, say, 4 batteries at a time, the cost per battery is only £2.50. So, overall, the cost of a Ni-Cd battery to provide 300 mAh is £4.50 whilst that of 167 alkaline batteries is £83.50. A clear saving of £79, excluding, of course, the electricity bill.

That is not to say that Ni-Cd batteries are better in every respect than primary batteries. Since their capacity is about one third of an equivalent-size primary battery, they need changing three times as often. Also, Ni-Cd batteries have a faster self-discharge than primary batteries. Typically, they self-discharge to 60% of their nominal capacity within 2 months, and so have to be recharged to regularly keep them topped up.

Notwithstanding these slight drawbacks, the advantages of rechargeable Ni-Cd batteries over primary batteries are obvious.

Chargers for Ni-Cd batteries are available from about £10. The circuit described here will not be significantly cheaper, but will have greater versatility than one ready-made and, in addition, will give the constructor an insight into the operation of constant-current battery chargers.

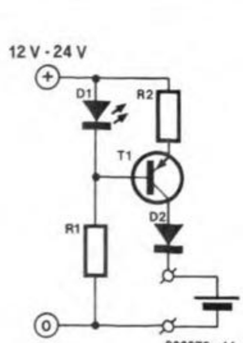
The charge process

Because Ni-Cd batteries show no significant rise in voltage when charged, it is not possible to use a constant-voltage charger as used for lead-acid batteries. Instead, a constant-current charger must be used.

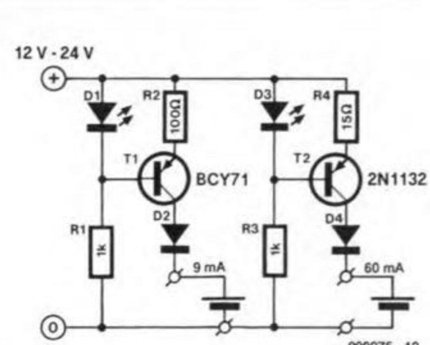
The choice of the charge rate is important. If the rate is too high, the battery can be damaged by the heat generated. If, however, the rate is limited to the 10 hour rate (i.e., the battery capacity divided by 10), the battery may be charged indefinitely without damage. To fully recharge a completely discharged battery will take about 12 to 14 hours at the 10-hour rate because of inefficiencies in the charging process.

How it works

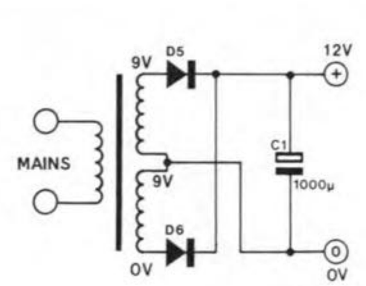
The charger described here is designed in two parts: a



Basic Circuit



Twin - Charger Circuit



12 V DC Supply

mains-to-12 V d.c. supply, and a 12–24 V constant-current battery charger with 60-mA and 9-mA outputs for AA- and PP3-size batteries respectively. The latter circuit can easily be adapted to charge larger batteries if required. The two circuits can be built into one case or into separate cases, which enables the charger proper to be used with any 12–24 V d.c. supply, such as a car battery via the cigarette lighter connection.

The circuits for the 60 mA and the 9 mA charger operate in the same manner. Both comprise an emitter follower based on a silicon p-n-p transistor and an emitter resistor. In the basic circuit, these parts are identified as T1 and R2 respectively. The batteries to be charged form the collector load. The base of the transistor is held at a reference voltage supplied by LED D1 and series resistor R1. The resistor limits the LED current to 10–22 mA depending on the input voltage (12–24 V). A LED is used in preference to a low-voltage zener diode because of its sharper voltage/current characteristic, and its temperature coefficient, which is roughly the same as that of the transistor.

The base of T1 is held at 1.6 V below the positive supply rail. The voltage drop across the transistor's base-emitter junction is 0.7 V, so that the voltage across R2 is 1.6–0.7=0.9 V. Hence, the battery current, I, is set by R2 in the following way:

$$I = U_{R2}/R2 = 0.9/R2\ [A]$$

For an AA-size battery with a capacity of 600-mAh, the continuous charge rate is 600/10 = 60 mA, so that

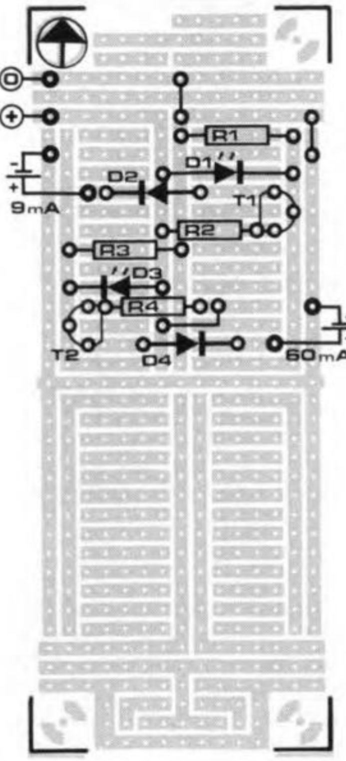
$$R2 = 0.9/(60\times10^{-3}) = 15\ \Omega$$

For a PP-3 battery with a capacity of 90 mAh and a maximum continuous charge rate of 9 mA, R2 works out at

$$R2 = 0.9/(9\times10^{-3}) = 1000\ \Omega$$

Using this same formula it is possible to calculate the resistor value for any required charge current. It is important to ensure that the dissipation of the transistor and the resistor (R2) are not exceeded.

Diode D2 prevents the batteries discharging through the circuit when the supply is switched off. Almost any medium-power silicon diode, like the 1N4001, will be suitable here.



900075-14

COMPONENTS LIST

Resistors (0.5W; 5%):

2	1k	R1;R3
1	100Ω	R2
1	15Ω	R4

Capacitor:

1	1000µF 15V	C1

Semiconductors:

2	red LED	D1;D3
4	1N4001	D2;D4;D5;D6
1	BCY71	T1
1	2N1132	T2

Miscellaneous:

1	2×9V/3VA mains transformer	Tr1
1	TO-5 heatsink	
4	AA-size battery holder	
1	PP3 battery clip	
1	printed-circuit board	UPBS-1
1	enclosure as required	

The minimum supply voltage to the charger is about 9 V to maintain the current through the LED. That voltage would allow sufficient headroom to charge 4 AA-size batteries in series, but not for a PP3-size battery.

The voltage drop across R2 is 0.9 V, that across the emitter-collector junction of T1 about 0.4 V, and that across D2 about 0.7 V, a total of 2 V. Therefore, in order to charge a PP3-size battery the minimum supply voltage is 9+2=11 V. For supply voltages greater than this the circuit maintains the voltage across the transistor in order to achieve a constant current through R2 and the batteries on charge.

The current requirement of the twin charger circuit is the sum of the LED currents and the charge currents, i.e.,

$$60 + 9 + 24 + 24 = 117\ \text{mA}.$$

The permissible dissipation of the transistors used in each circuit is such that they will accommodate a 24-V supply even with the charge terminals short-circuited. The overall requirement for the charger's d.c. supply is, therefore, 11 to 24 V at 120 mA minimum. A suggested circuit is shown for those who require a separate d.c. supply from the mains. It consists simply of a bi-phase full-wave rectifier followed by a smoothing capacitor. This circuit can be built into a separate housing and used as a general-purpose nominal 12-V d.c. supply, or incorporated into the same enclosure as the charger circuit. Alternatively, its function may be assumed by a mains adapter with 12-V d.c. output. In many cases, this will be preferred as it is a cheap as well as a safe solution.

Construction and practical use

The twin charger circuit may be constructed on universal prototyping board size-1 (UPBS-1) as shown by the accompanying component overlay. The 60-mA charger transistor is fitted with a small heat-sink to ensure it remains cool even when connected to a 24-V supply and with the charging output terminals short-circuited.

Upon completion of the construction, apply the input voltage and connect an ammeter across the charging terminals. Check that it reads 9 mA or 60 mA as appropriate. If necessary make small changes to R2 (9-mA output) or R4 (60-mA output) to achieve the correct charge current.

The maximum number of AA-size batteries that can be charged in series with a given supply voltage is given in the table below:

Supply voltage (V)	Max. number of cells
11	6
12.5	7
14	8
17	10
20	12
23	14

If it is required to charge C-size or D-size batteries, this can be done with the 60-mA charger but will require longer than the 12 to 14 hours to fully charge a completely discharged battery. The full charging time will be 25 hours and 58 hours for C-size and D-size batteries respectively.

Finally, a warning: always ensure that you observe the correct polarity of the batteries when connecting them to the charger. Failure to do so can lead to catastrophic and potentially dangerous failure. Never use the charger for recharging primary batteries. ■

REVERSING CAR ALARM

C. Brown

These days, an increasing number of lorries and buses is fitted with loud sirens and other actuators to warn pedestrians that the vehicle is reversing. Private cars have yet to be fitted with such devices, which seems strange, as they can be just as hazardous when reversing.

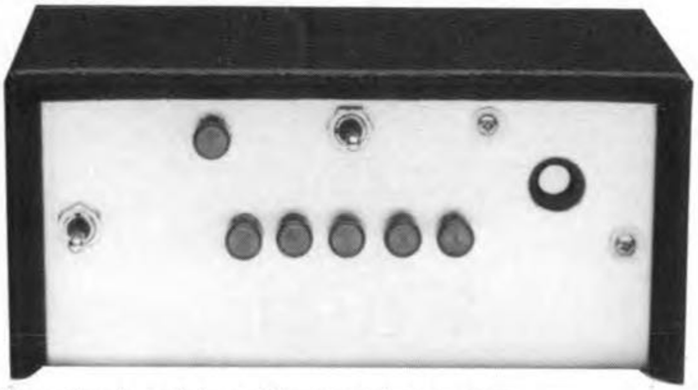
Few cars are fitted with any sort of alarm system; even a flashing LED would deter many thieves from risking setting off an alarm. This project aims at improving this situation somewhat by providing a KITT-like 'scanning eye' on the dashboard.

Furthermore, the circuit also provides a relay to switch two sets of indicator lamps on and off at a rate of about once a second. This is used to advise fellow motorists that your car is static, either having broken down, or parked temporarily.

The whole system operates from only three integrated circuits, two of which, IC1 and IC3, are the ubiquitous 555 timers. The first provides clock pulses, the second audio tones. In between these devices is a Type 4022 counter, IC2, whose eight outputs go high sequentially as each new clock pulse arrives. Each output drives a LED, and also supplies a small current to one of three presets P2, P3 and P4. The presets are adjusted to give an alternately ascending and descending sequence of three tones with short pauses between the sequences.

The CARRY OUT (CO) signal of the counter is used to drive a transistor, T1, which in turn operates a relay, Re1. The CO output remains high for four counts, then goes low for four counts. The on and off periods of the indicator lamps connected to the relay contacts are therefore equally long.

As well as powering a 5-watt weatherproof loudspeaker mounted on the rear of the car, the tones also operate a piezo sounder within the unit. With the internal sounder ringing out, the driver's attention is drawn to the fact that the hazard relay is still operating. However, having broken down on the M1, you do not want to spend a couple of hours listening to the piezo sounder whilst the relay operates the hazard indicators. Hence, a switch, S2, allows you to mute the sounder, and instead switch on a LED, D18, only.

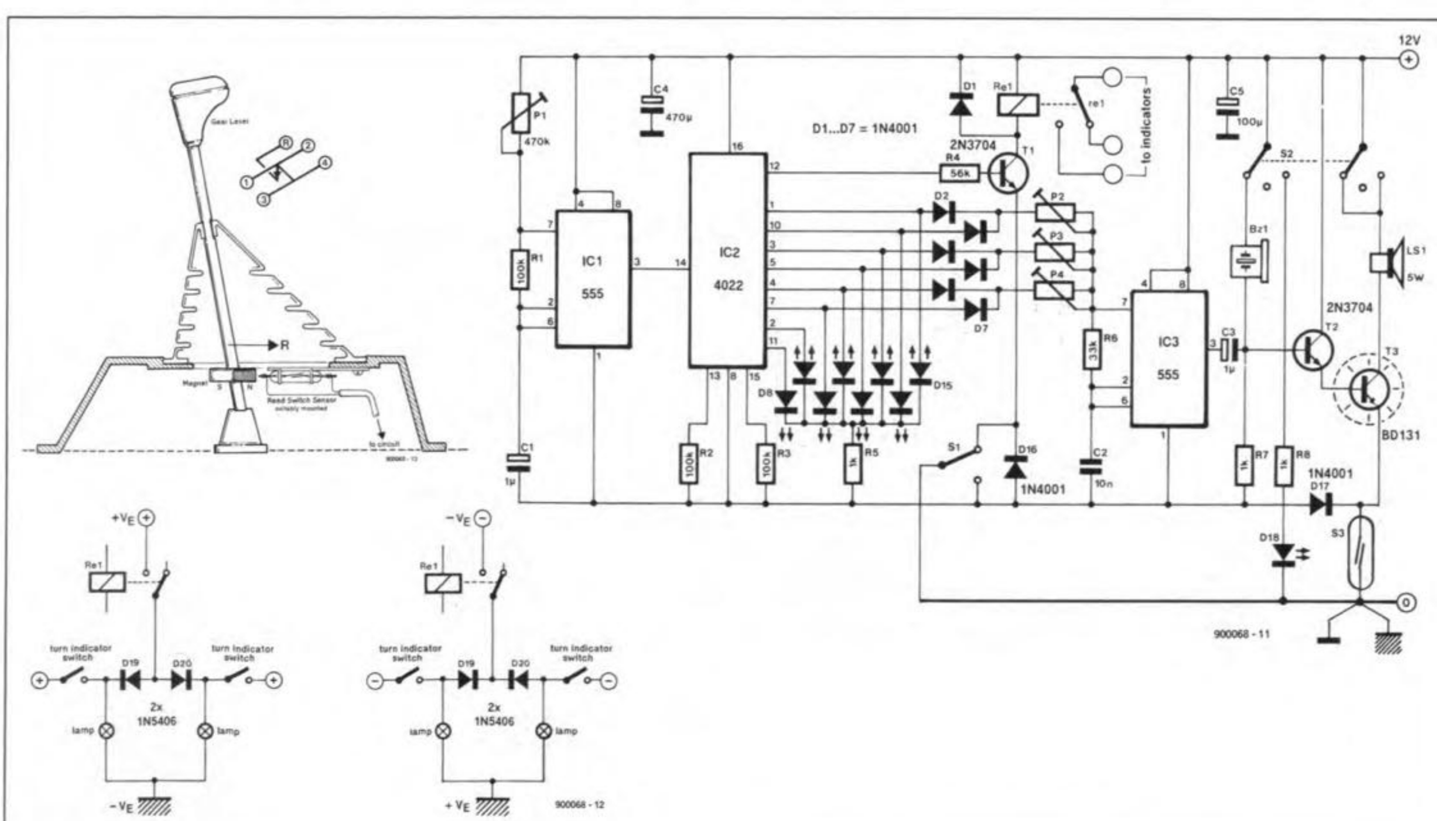


The author's prototype of the reversing car alarm.

Thanks to a magnet mounted on the gear lever operating a reed switch, the reversing alarm automatically sounds whenever reverse gear is chosen. Most neighbours, having just gone to bed, would be a bit annoyed if they were woken by you reversing your car into the garage at midnight. Setting S2 to the centre-off position disables the external sounder, although the LEDs still flash.

The other switch in the circuit, S1, is used to select either the hazard setting or the burglar deterrent. When the pole of S1 is connected to ground, the relay and the sounder are disabled, allowing just the LEDs to operate. These produce a KITT-like line of light which sweeps across the front panel. A short pause is simulated before the scanning line reverts by fitting only five of the eight LEDs on the front panel. Do not omit the three 'dummy' LEDs, however, as this causes a higher current to be fed into the tone generator, resulting in unwanted extra tones. Like S2, S1 has a centre-off position so that the unit may be turned off completely.

The indicator lamps must



be powered via diodes of appropriate power rating for the lamps. Without these, a loop would be set up causing both indicator lamps to flash when the normal left/right indicator switches were used.

Since the audio output transistor, T3, may run fairly warm, it is advisable to fit it on a heat-sink, or bolt it to the car chassis. Either way, do not forget to insulate the transistor electrically with the aid of a mica washer.

Setting up

Initially, set S1 to the burglar deterrent mode. Preset P1 is used to adjust the running speed of the LED scanner. The best adjustment at this stage is one clock pulse per three to five seconds. Switch S2 to the piezo sounder. When the far left LED lights, adjust P2 until you find the tone pleasing. When the second LED lights, adjust P3 for a higher tone. Do the same for the next preset, P4, when the third LED lights. This sets up the three ascending/descending tones. Next, re-adjust P1 until a complete cycle lasts about 2 seconds. The last stage is to move S2 to the hazard setting. The relay should now be heard clicking on and off at regular intervals, and the loudspeaker should produce loud tones.

Set S1 to its central-off position. This should turn off all functions. Connect the reed contact wires together to check that the circuit is actuated. If there is no sound from the loudspeaker, check that S2 is not in the central off position. ■

ELECTRONIC SIREN

D. Butler

This module was originally designed to be used in a plastic model kit of a police motorcycle. Other applications that come to mind are inclusion into a model railway layout, or as a burglar alarm.

The circuit diagram indicates that the module consists of two sections: a 555-based timer (IC1) and a voltage-controlled oscillator (IC2). The 555 is configured in astable mode. It produces a square-wave at its output, pin 3. The frequency of this square-wave, f_o, depends on components R1, R2, P1 and C2, as expressed by

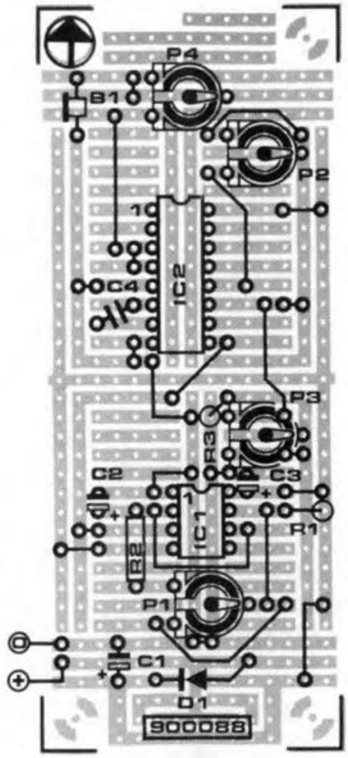
$$f_o = 1.44/[(R_1 + 2(P_1 + R_2))\,C_2]$$

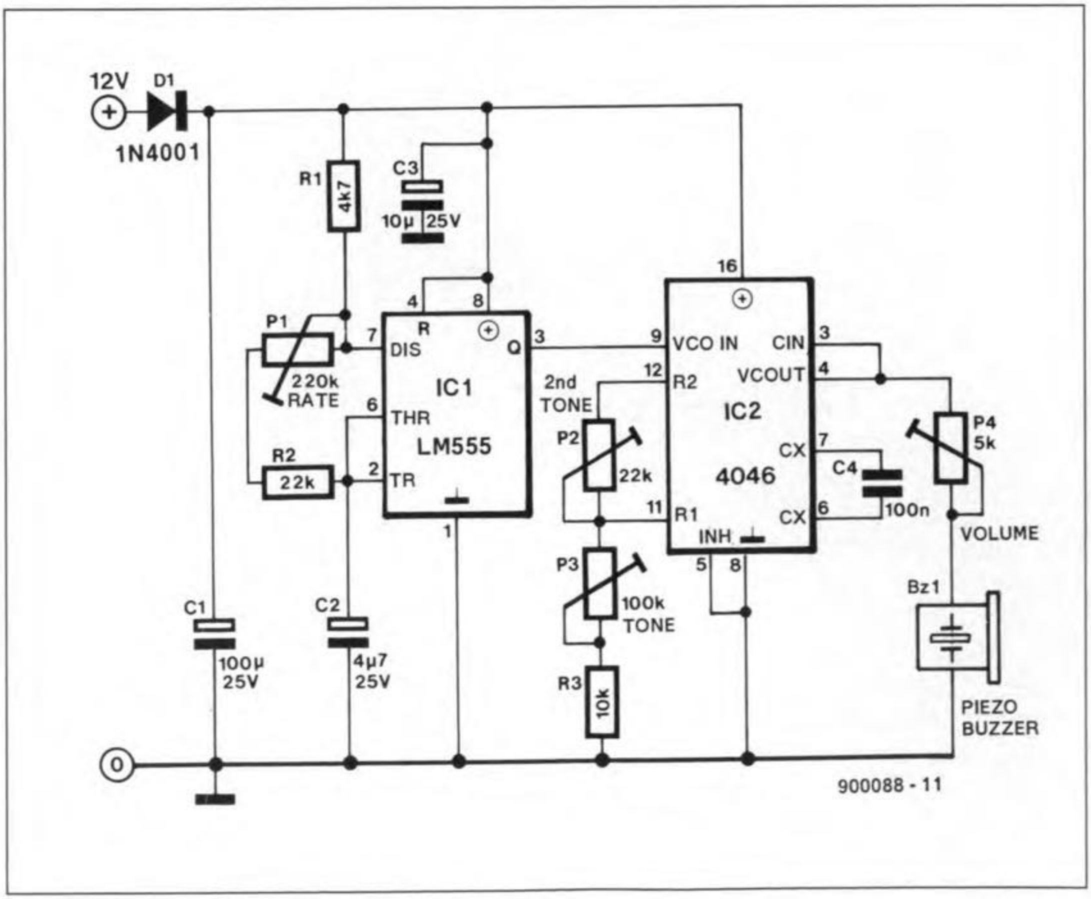
This square-wave is applied to IC2, a Type 4046 phase-locked loop (PLL). This IC contains a voltage-controlled oscillator (VCO), of which the frequency range is set by capacitor C4 connected between pins 6 and 7. The voltage level at the VCOIN terminal, pin 9, effectively controls the VCO section of IC2—in this case forming the two tones of the siren.

Presets P2 and P3 set the frequencies of the two tones generated by the siren, while P4 controls the output volume of the (passive) piezo buzzer. The rate at which the tone changes is controlled by P1.

Diode D1 affords protection against reversed supply voltages. Although a supply voltage of 12 V is shown, the module can operate over a range of 5 V to 15 V.

The accompanying component overlay shows a suggested construction on board UPBS-1. ■





COMPONENTS LIST

Resistors:

1	4k7	R1
1	22k	R2
1	10k	R3
1	220k preset H	P1
1	22k preset H	P2
1	100k preset H	P3
1	5k preset H	P4

Capacitors:

1	100µF 25V	C1
1	4µ7 25V	C2
1	10µF 25V	C3
1	100n	C4

Semiconductors:

1	1N4001	D1
1	555	IC1
1	4046	IC2

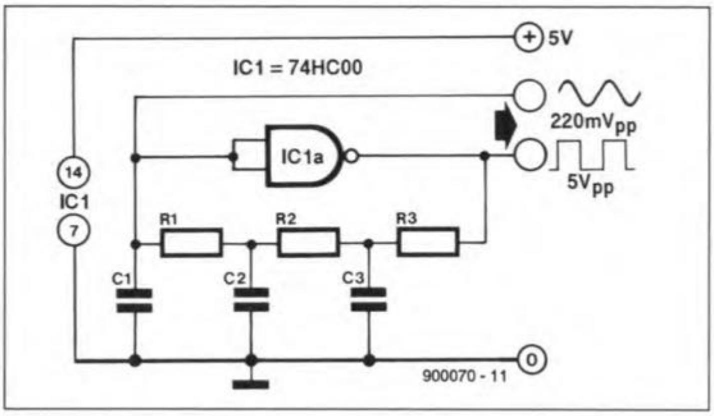
Miscellaneous:

1	passive piezo buzzer	Bz1
1	printed-circuit board	UPBS-1

SIMPLE SQUARE-WAVE GENERATOR

C.M. Clarkson

This circuit provides the perfect solution in all those cases where you are stuck for a clock generator, yet have one ordinary logic gate left in your design.



The circuit works because of the high gain of the inverting logic gate and the 180° phase shift introduced by an *R-C* network at the oscillating frequency, f_o, calculated from

$$f_o = \sqrt{6}/(2\pi RC).$$

At first you may be a little bewildered by the use of a logic gate where you would expect a linear amplifier. Most logic gates however will work as linear amplifiers provided they are biased correctly. The resistors in the network supply the necessary bias for the logic gate. The 74HC00 NAND gate used here should provide a gain well in excess of 1,000. In some cases, however, its hysteresis may cause starting problems of the oscillator. This may be resolved however by replacing the 74HC00 by a 74HCU00, which has unbuffered outputs.

The *R-C* network attenuates the output signal of the logic gate by a factor of 1/29. Provided the gain of the logic gate is greater than the attenuation of the *R-C* network, and the logic gate inverts the signal, the output of the gate will be in phase with its input. Any small change in the input signal will cause a large change in the output, which is added to the input change, thereby reinforcing it. This is called positive feedback, and the effect can only occur at the oscillating frequency.

The output of the logic gate is a square-wave because the output saturates owing to the high gain of the gate. The input to the gate is a sine-wave of amplitude $(4U_{cc})/(29\pi)$ V. The reason for the term $4/\pi$ is explained by the Fourier analysis of a squarewave. This analysis shows that the amplitude of the fundamental frequency of a squarewave is 4π larger than the original squarewave amplitude.

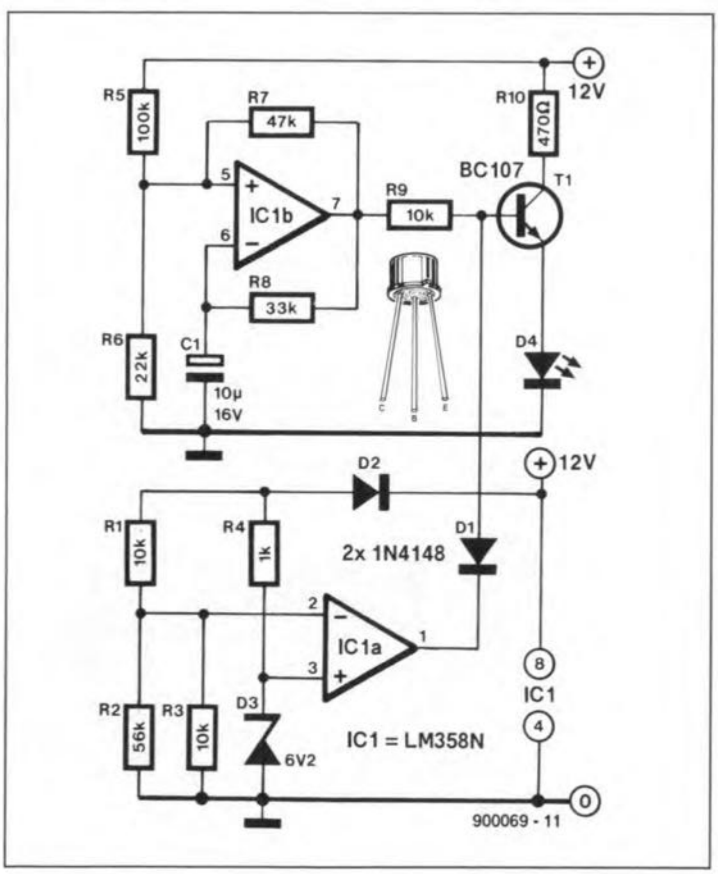
The input to the logic gate is a sinewave because the squarewave at the output is fed through three low-pass filters connected in series. These filters attenuate the harmonics of the squarewave, leaving a slightly distorted sinewave at the input. If the gain of the HC00 gate were exactly 29, the output would be a pure sinewave, since any non-linearities owing to output saturation would not occur.

The oscillator works fine at frequencies between 100 Hz and 10 MHz. It is very useful because it provides both a squarewave and a sinewave at the frequency of oscillation. ■

SIMPLE EFFECTIVE CAR THEFT DETERRENT

K. Hebborn

The purpose of this simple circuit is to deter potential car burglars simply by a brightly flashing LED, implying that the car is fitted with a sophisticated alarm system.



The circuit is remarkable because it is connected to the car's electrical system by two wires only. It senses the car battery voltage and automatically switches itself off after the engine is started. This prevents distraction while driving and avoids the need of an on/off switch.

The circuit uses a Type LM358 dual opamp. One opamp, IC1a, is used to sense the battery voltage, and the other, IC1b, to pulse the LED driver transistor, T1.

When the battery voltage exceeds about 13.5 V, which happens when the engine is running and the battery is being charged, the voltage across R2-R3 exceeds that across zener diode D3. This results in the output of IC1a going low, so holding the base of T1 off via diode D1. Thus, when the engine is running, the LED can not flash.

When the engine is off, the output of IC1a is high because the voltage across R2-R3 is lower than that across D3. This results in the base of T1 being driven from the output of IC1b. This opamp is configured as a simple square-wave oscillator whose frequency is set by R8 and C1. The relevant component values indicated in the circuit diagram give a flash rate of about 100 per minute.

The average current consumption of the circuit is about 20 mA so that it causes negligible drain on the car battery. ■

DOOR CHIME/ALARM SIREN

T. Giffard

The Type U450B integrated circuit from Telefunken is a tone sequence generator for applications in toys and electronic household utensils.

In the circuit shown here, the output signal of the U4050B is boosted by a one-chip amplifier from Philips Components, the TDA7052. The combination of the two chips results in a relatively simple tone generator that may be used for at least two applications (as indicated by the title).

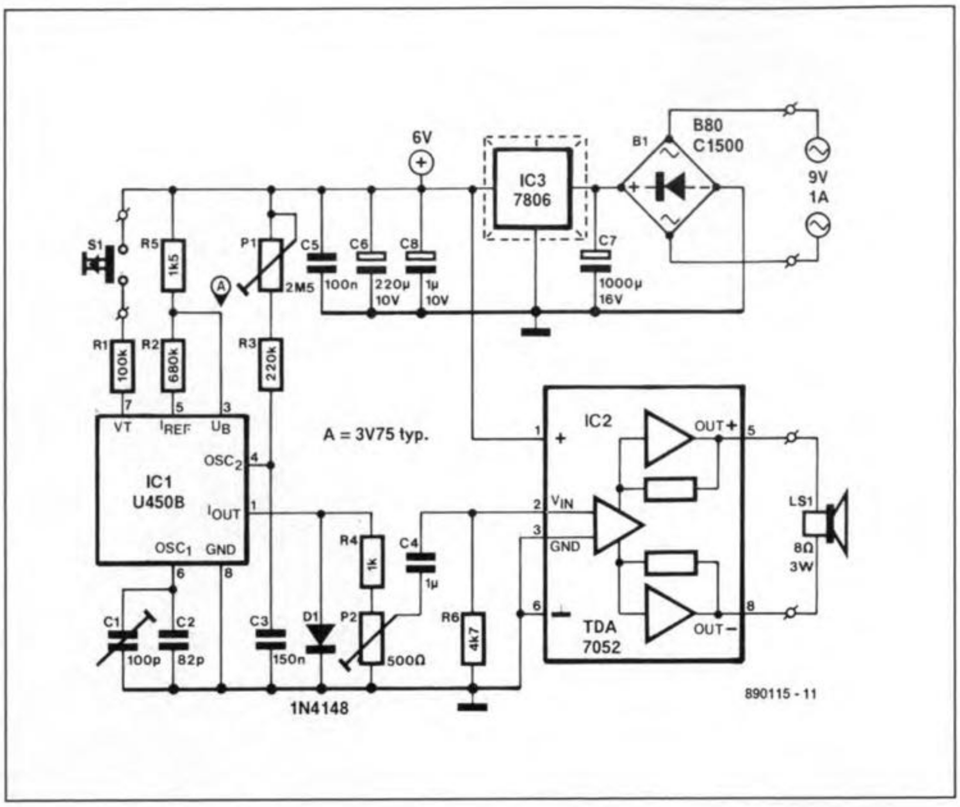
The TDA7052 is capable of supplying a nominal sine-wave output power of up to 1.2 W when loaded with 8 Ω. Since in this application the IC handles rectangular signals only, an output power of 3 W may be supplied—enough for quite high sound levels.

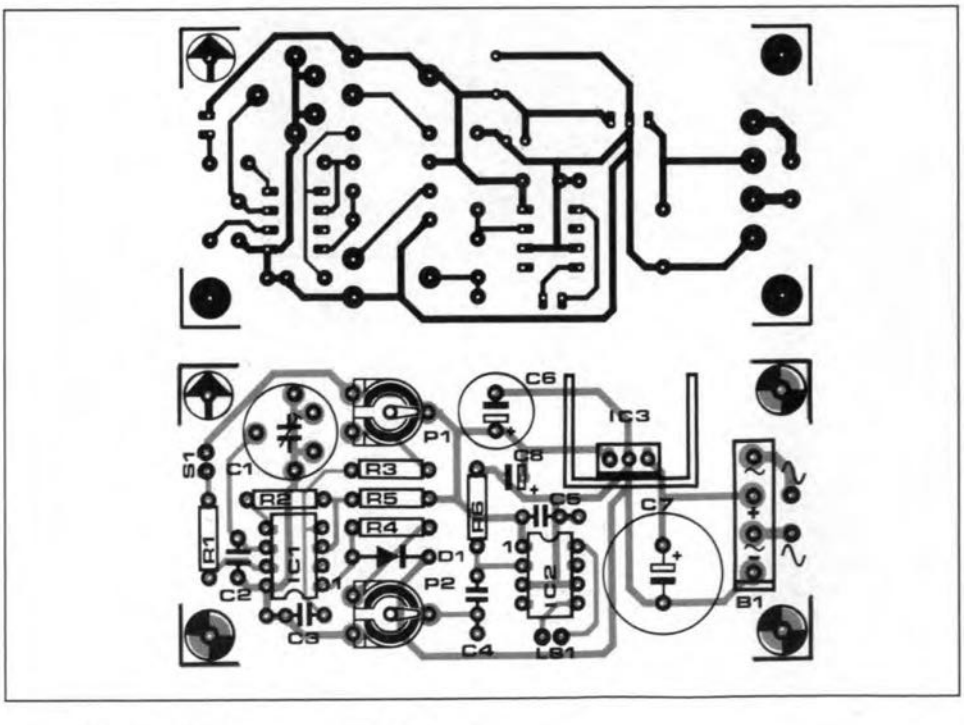
When S_1 is pressed, the sound generator, IC_1, produces a sequence of three tones whose frequency is determined by the setting of trimmer C_1. The total duration of the tones is determined by P_1. The output signal is supplied in the form of a current at pin 1 of the chip. This current is converted into a corresponding alternating voltage by R_4, P_2 and C_4. The AF voltage so obtained is fed to the power amplifier via input pin 2. The maximum output voltage of IC_1 is limited to about 0.6 V by diode D_1.

The U450B receives its supply voltage via resistor R_5, which forms the current limiter for the on-board 3.75-V voltage regulator. Here, the value of R_5 causes a current of about 1 mA to flow into pin 3.

The AF power amplifier, IC_2, contains a phase splitter and two output stages which are used in a bridge configuration to drive an 8-Ω loudspeaker.

The 9-VAC power supply must be capable of delivering up to 0.5 A. Circuit IC_3, a 7806, provides the required voltage regulation. In the off state, the circuit consumes about 6.5 mA. When the sound generator replaces the existing doorbell, the bell transformer may be conveniently used to provide the a.c. supply voltage. ■





COMPONENTS LIST

Resistors:

1	100k	R1
1	680k	R2
1	220k	R3
1	1k	R4
1	1k5	R5
1	4k7	R6
1	2M5 preset H	P1
1	500Ω preset H	P2

Capacitors:

1	100pF trimmer	C1
1	82pF	C2
1	150nF	C3
1	1µF	C4
1	100nF	C5
1	220µF 10V radial	C6
1	1000µF 16V radial	C7
1	1µF 10V radial	C8

Semiconductors:

1	1N4148	D1
1	B80C1500	B1
1	U450B	IC1
1	TDA7052	IC2
1	7806	IC3

Miscellaneous:

1	SPDT push-button	S1
1	8Ω/3W loudspeaker	LS1
1	heat-sink for IC3	

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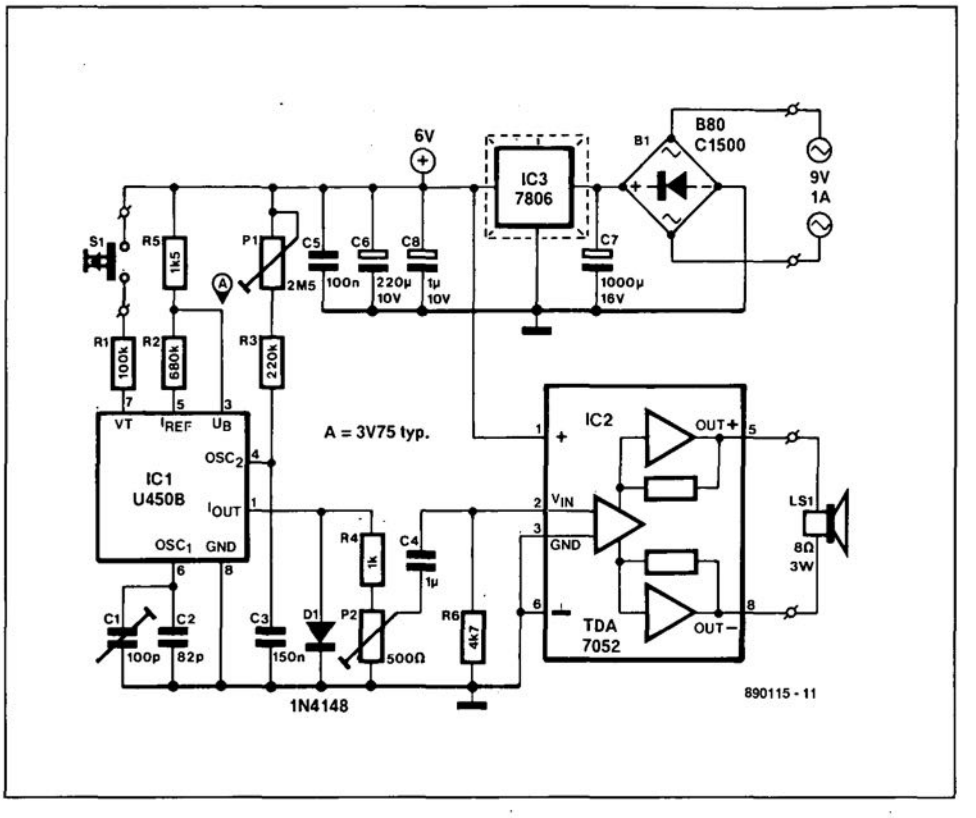

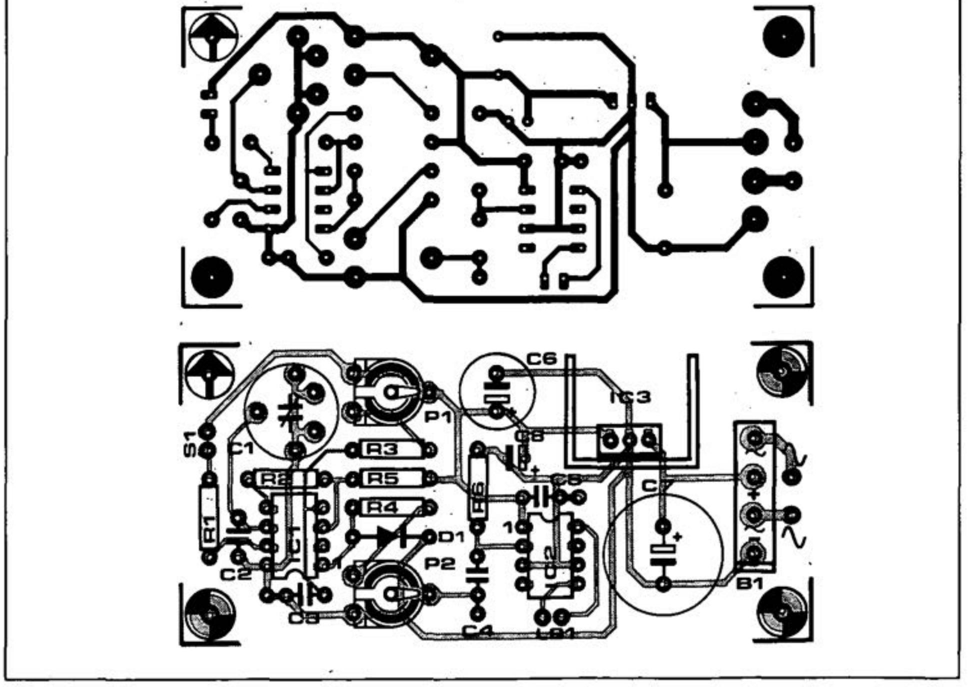

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Miscellaneous:

1	SPDT push-button	S1
1	8Ω/3W loudspeaker	LS1
1	heat-sink for IC3	

SCIENCE & TECHNOLOGY

COMPUTERS LEARN FROM HUMAN MISTAKES

by R.A.J. Arthur

When a system fault develops on an earth satellite, a human controller may find it easy to correct. On an unmanned space vehicle, one fault exceeding the limited liability of an automatic control correction system can write off an asset valued in billions of pounds.

Now, however, the Turing Research Institute in Glasgow has produced a remarkable remedy, and it could also have applications in general industry. Professor Donald Michie, chief scientist at the institute, said: "At last we have opened the door to transferring cognitive skills from humans to machines."

He described the method as "like being able to take an X-ray photograph of a cognitive skill". A satellite equipped with this transplanted human expertise would be able to get itself out of trouble.

The institute is named after the late Alan M. Turing, the British mathematician and logician whose work had a lasting influence on the foundations of modern computing. It shares premises, and is closely linked, with the Scottish Human Computer Interaction (HCI) Centre, part of the University of Strathclyde. The executive director of both is Professor James Alty.

Human computer interface

The institute's unsurpassed on-line Artificial Intelligence (AI) library of 55,000 documents is the only one of its kind available electronically from remote sites.

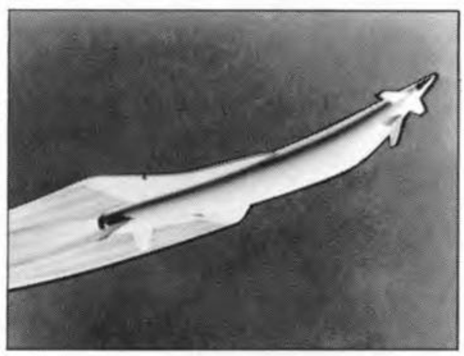
The objective of the HCI Centre is to develop the greatly improved human computer interfaces that alone can make AI systems acceptable.

Professor Michie's approach aims to overcome a potentially dangerous impasse. Conventional automatic systems, as widely used in industry, may still encounter the unexpected but lack the adaptive ability to deal with it.

If a pedal comes off a bicycle, a human cyclist will reorganize a strategy for keeping going, but a robot cyclist will almost certainly fall off. Although human beings are not good at any one task, they are very good at re-learning on the job. The ability of machines to study human trial-and-error learning and extract a set of rules defining a successful strategy is the basis of the new technique. Machine learning is used as a tool for digging law-governed regularities out of data.

Learning process recorded

A machine's ability to learn from human examples is demonstrated in a computer-simulated exercise—balancing a pole on a cart free to move along a short piece of track without letting the pole fall over or the cart run off either end.



HOTOL (Horizontal Take-off and Landing Aircraft), the unmanned orbital craft for which the Turing Institute has researched a new method of control if a fault occurs.

Various human operators built up skill at this game while the machine recorded the entire learning process. From the record of these attempts it extracted a series of generalized 'rules' for success. This experiment was used simply to demonstrate the absolute reliability of the method.

The difficulty in designing an automatic attitude controller for systems on satellites is formidable. Machine simulation falls far short of handling so complex a task. Simulators always represent a large simplification. When pilots are trained on them for unstable craft such as helicopters, it is impossible to turn the trainee loose on the real thing without a period of conventional human instruction.

In the case of an unmanned orbital craft such as the HOTOL, there is no question of human control, nor can a perfect automatic controller be designed by mathematical means. However, the new method opens a possible route.

Behaviour examples

First, a human operator is trained as far as he can go on the simulator. Then, the machine's learning programs are allowed to ruminate over the record of his behaviour. From that perhaps messy and inconsistent record they distil the essence of what his strategy is likely to be.

The machine learns from examples of behaviour and this overcomes a big problem. A top practitioner carries so much of his skill in his subconscious memory that his verbal accounts of method are wholly inadequate. Extracting general rules from actual examples, the machine constructs its own standard strategy far better than the human one because human inconsistencies have been smoothed out.

The aim is to generate a streamlined strategy far superior to what a human being would develop on a satellite. Observation shows that the streamlined strategy of the machine is strikingly better than the original human example.

The close cooperation of the Turing Institute with the Scottish HCI Centre has already produced remarkable results. As the era of Artificial Intelligence develops, this kind of expertise will make its impact in most fields of human activity. ■

NEW PUBLIC CORDLESS TELEPHONE SYSTEMS

by John Williamson

In the 1980s, mobile communication has become one of the most dynamic sectors of the telecommunications industry and, as is widely accepted, the United Kingdom has established an international lead in the technological development and commercial exploitation of a number of mobile services.

Thanks to the introduction of private sector competition in the supply of service and equipment, Britain's two cellular telephone networks are individually among the world's largest.

The operators Cellnet and Racal Vodafone had a combined subscriber base of about 800,000 early this year, and conservative estimates put the total growth rate at 30,000 new customers a month.

Local market volumes and competitive

supply have already driven the price of analogue cellular telephones down. Projected demand and the use of an open specification for the future pan-European digital system are expected to have a similar effect on the digital terminal market.

However, the high cost of building and maintaining cellular infrastructures dicates that service costs will remain relatively high for both analogue and digital systems.

With this in mind, a number of companies are beginning trials of what is termed telepoint. This is a limited distance, cordless public telephone service. Since it has no roaming capability and can only make outgoing calls, the telepoint infrastructure cost is much lower and, accordingly, calls are much cheaper.

Phonepoint, one of four consortia licensed to provide telepoint services in the United Kingdom, estimates that total annual average usage costs for a telepoint payphone service could be between £200 and £300 as compared to £1000 for a cellular service.

Public locations

Telepoint terminals are initially expected to retail at about £180, falling to around the price of a convential domestic cordless telephone during the early 1990s.

Telepoint in Britain is based on the second generation of a cordless domestic or business telephone specification known as CT2. As such, the same terminal can be used in the home, in the office, and when in proximity of an access point, in a public place.

CT2 telepoint uses a mixed frequency division multiple access (FDMA) and time division duplex (TDD) mode of operation. This allows a single radio channel to support two-way or duplex speech through the use of a single block of radio spectrum rather than the paired frequency arrangements found in conventional duplex mobile radio systems.

With TDD, the handset transmits for a short period while the base unit at the public access point receives, and vice versa. This involves both the handset and the base station being synchronized, and at either end of the link the input speech is digitized, stored for one burst period and transmitted in the next.

Accordingly, transmission is time-compressed by a factor of two. At the receive end of a link, packages of burst speech are recombined and converted from digital to analogue form.

In the United Kingdom, 40 channels of 10 kHz have been allocated to CT2 in the range 864.1–868.1 MHz. Base units scan the 40 channels using a channel selection algorithm.

If the handset and base station are within working range of each other (1–200 metres), the algorithm dictates that both select the same vacant radio channel for communication.

This self-tuning characteristic and the relatively short transmission range produce very efficient spectrum utilization. It is estimated, for example, that the 40×100 kHz space allocated to CT2 telepoint in Britain can support over 5000 handsets per square kilometre.

A second advantage of CT2 over some other forms of mobile communication is that the required infrastructure is neither costly nor complex. In fact, most of it already exists. Transmitters and receivers are hooked direct into the public switched telephone network.

Using Phonepoint, one of the world's first cordless public telephone systems, at Euston railway station in London.

A third consideration is that telepoints are mounted on poles, walls or other structures physically inaccessible to the general public. This means that the possibility of vandalism and misuse associated with conventional telephone call boxes is much reduced.

International calls

There are two approaches to handling the necessary subscriber verification and billing processes—local and centralized.

In the former, the public access point stores the necessary information for verification and also has a metering capability. During quiet periods, data collected is downloaded to a central computer following interrogation.

In the alternative, the telepoint needs to seek verification from the central control processor on the receipt of each call request, and passes billing information back on call completion.

In operation, users approach to within 200 metres or so of the nearest telepoint access station, switch on their cordless telephones, and enter a personal identification number.

This is checked by the system and in moments the user is presented with a dial tone and can call into the national or international network in the normal way. Each telepoint service provider will eventually have access stations in all major public places—including railway stations, main streets, shopping centres and motor service stations.

Telepoint phones can also be used for two-way communication as domestic or public branch automatic exchange (PABX) extension terminals.

Of the four British telepoint consortia, Phonepoint, in which British Telecom has a major stake, was the first off the mark with a pilot commercial trial in London in August 1989.

Joint ventures

Ferranti Creditphone launched a larger scale commercial telepoint service in October last year under the Zonephone banner. After spending some £20 million on development, Zonephone was put over on a network of 300 base stations. The company aims to have 5000 base stations all over Britain by the end of this year.

Zonephone is a proprietary system, but Ferranti's base stations have compatibility with the common air interface (CAI) that all four British operators will be required to implement by 1991.

Ferranti, which has already sold equipment for a trial system in France, has been talking to six other countries with a view to setting up local joint venture telepoint operating companies. Two of these are thought to be Japan and South Korea.

Another British operator, the BYPS Communications consortium, made up of Barclays Bank, Philips and the Shell Oil Company, has said its service will use CAI from day one. BYPS has just started operations and has spent £30 million on 100,000 handsets, 7,000 base stations and a quantity of network equipment.

BYPS subscribers will have the option of buying their pocket phones from Orbitel, the Plessey/Racal joint venture, which last October unveiled its own range of CAI products.

Outside Britain, Finland's Helsinki Telephone Company began CT2 trials last November, using terminal equipment provided by Shaye Communications, and infrastructure jointly developed with Shaye shareholder, Nokia. ■

Phonepoint Ltd, BTMC, Mobile House, 1 Eversholt Street, LONDON NW1 2DW.

Ferranti Creditphone Ltd, St Mary's Road, MANCHESTER M10 0BE.

BYPS, Westbrook Centre, Milton Road, CAMBRIDGE CB4 1YH.B

Orbitel Mobile Communications Ltd, Keytech Centre, Ashwood Way, BASINGSTOKE RG23 8BG

Shaye Communications Ltd, Capital House, 48-52 Andover Road, WINCHESTER SO23 7BM.

THE WORLD'S FIRST DESIGN MUSEUM

by Tony Aldous

Contrasting telephones from 1895 and 1989; a one-key typewriter of 1924 beside a computer keyboard and video display unit from the late 1980s; and a Kodak camera of 1889 together with a Ricoh Mirai of 1988—these are just some of the exhibits charting changes in design and technology at London's new Design Museum, the first of its kind in the world.

The brainchild of Sir Terence Conran, leading designer and head of the Storehouse retail chain, the museum stands on the south bank of the Thames just below Tower Bridge.

Connected by a ferry shuttle to the Tower of London on the opposite bank, it is well placed for tourists. In its first four months, it drew more than 48,000 visitors from all over the world.

According to Helen Rees, the museum's founding curator and now its acting director, its prime aim is "to increase public awareness of the benefits of design to society, to culture and to the economy". She and her colleagues are not trying to elevate the idea of design nor to promote the design profession, but to show how the mass produced artefacts of our day work and why they look as they do. "Design decisions," she says, "have a profound effect on everybody".

The Boilerhouse

Set up by Sir Terence's charitable Conran Foundation, the museum is housed in a sturdy 1950s warehouse, refurbished and enlarged at a cost of £7 million. The result, painted white inside and out and with large windows looking out on the river, has a 1920s look but provides a spacious and appropriately neutral showcase for all manner of exhibits.

The conversion was managed by Sir Terence's own architecture/development firm, Conrain Roche, but the two most important interiors are the work of the leading exhibition and museum designers, Alan Stanton and Paul Williams.

From a ground floor entrance hall containing a cafe/bar, the visitor moves to the first floor gallery, The Boilerhouse. This is named after the project from which the Design Museum has grown, and which between 1982 and 1988 mounted a series of exhibitions at the Victoria and Albert Museum in the courtyard of its boilerhouse.

The new Boilerhouse fulfils a similar function as a showcase for a frequently changing programme of exhibitions. Its first, 'Commerce & Culture' looked at the relationship between mass produced goods and the ethos of the museum world. It revealed some remarkable similarities between today's commercially orientated museums and art and design orientated department stores. This was followed by an exhibition on French design between 1960 and 1990, imported from the Pompidou Centre in Paris.

State-of-the-art

The Boilerhouse is a white-walled, windowless but very flexible space. It can be divided in a variety of ways and its false ceiling can be removed to create a lofty space two storeys high.

On the same level is the Review, a smaller gallery devoted to a constantly changing international survey of new and speculative products and processes. These have included credit cards in colours chosen to match their holders' dresses, an elegant French bus shelter, and a push chair designed for the Mothercare chain.

On the newly added top floor is another gallery space, known as the Study Collection, which is flooded by light.

Windows on three sides give views of the Thames and the adjacent Butler's Wharf development that lies between the museum and Tower Bridge. If the sunlight becomes too bright, photosensitive blinds automatically protect the exhibits.

Visitor participation

The Study Collection contains some 400 objects chosen to illustrate three themes: stylistic or technological innovation; success or failure in the market; and 'the context of consumption', that is, how products are used and understood.

There are sections on items used in the home, at work and in transport, and the display follows the development of particular products, such as typewriters, telephones, radios and cameras. Video films projected alongside exhibits show how they were initially marketed and what the public perception of them was. A tape-recorded audio-guide and study notes are also available.

Visitors to this section of the museum can try their hand at computer-aided design (CAD) and, using a computer terminal, can also view short biographies of the world's most famous designers, accounts of companies famous for their design quality, and products that have had a significant effect on people's lives.

There is, for example, a 120-year old Singer New Family sewing machine, other domestic appliances of the time, and later sewing machines whose use and impact on women's lives the study notes explore with the help of contemporary advertisements and user manuals.

Educational charity

All this reflects the long-standing wish of Sir Terence Conran to create a research and resource centre for professional designers, design students and members of the public. He wanted to provide the kind of facility he had vainly sought as a young design student himself.

The building also includes a lecture theatre/cinema, a library with 5,000 books, journals and other reference items, a restaurant and a poster gallery presenting shows that change regularly.

Run as an educational charity, the museum received government grants to help it get under way but, with the aid of private sponsorship, is expected to be largely independent of public subsidy within two to three years. ■

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DESIGN NOTES

The contents of this article are based on information obtained from manufacturers in the electrical/electronics industry and do not imply practical experience by *Elektor Electronics* or its consultants.

QUAD POWER FAULT MONITOR

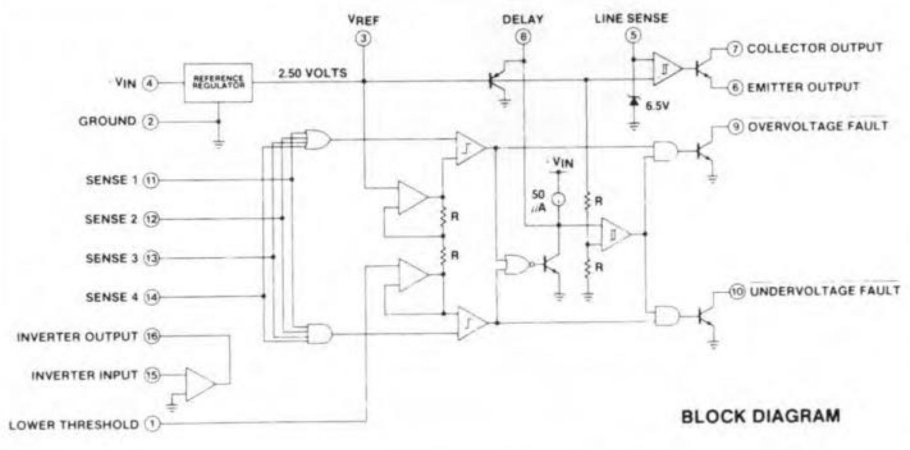
Silicon General's Type SG1548 is an integrated circuit capable of monitoring up to four positive DC supply voltages simultaneously for overvoltage and undervoltage fault conditions. An on-chip inverting opamp also allows monitoring one negative DC voltage. The fault tolerance window is accurately programmable from ±2% to 40% using a simple divider network on the +2.5-V reference.

A single external capacitor sets the fault indication delay, eliminating false outputs caused by switching noise, logic transition current spikes, and short-term AC line interruptions. An additional comparator referenced to +2.5 V allows the AC line to be monitored for undervoltage conditions or for generation of a line clock. The comparator can also be used for programable undervoltage lockout in a switching power supply.

Uncommitted collector and emitter outputs permit both inverting and non-inverting operation. External availability of the precision +2.5-V reference and open-collector logic outputs permit expansion to monitor additional voltages using available open-collector quad comparators.

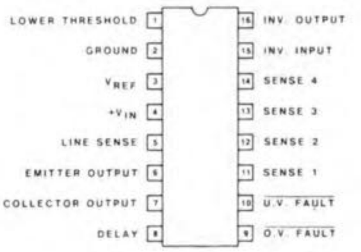
FEATURES

- **Monitors four DC voltages and the AC line**
- **Precision +2.5 volt ±1% low-drift reference**
- **Fault tolerance adjustable from ±2% to ±40%**
- **±3% trip threshold tolerance over temperature**
- **Separate 10 mA, 40 volt overvoltage, undervoltage and AC line fault outputs**
- **Fault delay programmable with a single capacitor**
- **Comparator hysteresis to prevent oscillations**
- **On-chip inverting op amp for negative voltage**
- **Open-collector output logic for expandability**
- **Operation from +4.5 volt to +40 volt supply**
- **Standard 16-pin dual-in-line package**



BLOCK DIAGRAM

CONNECTION DIAGRAM (TOP VIEW)



J or N PACKAGE TO-116 STYLE

ABSOLUTE MAXIMUM RATINGS (Note 1)

Supply Voltage ($+V_{IN}$)	+40V	Power Dissipation at $T_A = +25°C$ (Note 2)	1000mW
Fault Output Collector Voltage	+40V	Thermal Resistance: junction to ambient	100°C/W
Sense Input Voltage Range	−0.3V to +6.0V	Power Dissipation at $T_C = +25°C$ (Note 3)	2000mW
Fault Output Sink Current	20mA	Thermal Resistance: junction to case	60°C/W
Line Sense Input Current	±1mA	Operating Junction Temperature	+150°C
Inverting Op Amp Input Current	−5mA	Storage Temperature Range	−65°C to +150°C
Inverting Op Amp Output Current	25mA	Lead Temperature (soldering, 10 seconds)	+300°C

Note 1. Values beyond which damage may occur.
Note 2. Derate at 10 mW/°C for ambient temperatures above +50°C
Note 3. Derate at 16 mW/°C for case temperatures above +25°C.

RECOMMENDED OPERATING CONDITIONS (Note 4)

Supply Voltage Range		Line Sense Output Current Range	0 to 10mA
±25% Maximum Fault Window (Note 5)	+4.5V to +35V	Voltage Reference Output Current	0 to 10mA
±40% Maximum Fault Window	+5.0V to +35V	Operating Ambient Temperature Range	
Lower Threshold Input Range	+1.50V to +2.45V	SG1548	−55°C to +125°C
Fault Tolerance Window Range	±2% to ±40%	SG2548	−25°C to +85°C
Fault Output Sink Current Range	0 to 10mA	SG3548	0°C to +70°C

Note 4. Range over which the device is functional and parameter limits are guaranteed.
Note 5. Limited by inverter amplifier positive swing at −55°C.

ELECTRICAL CHARACTERISTICS

($+V_{IN}$ = 15 volts, and over operating temperature, unless otherwise specified)

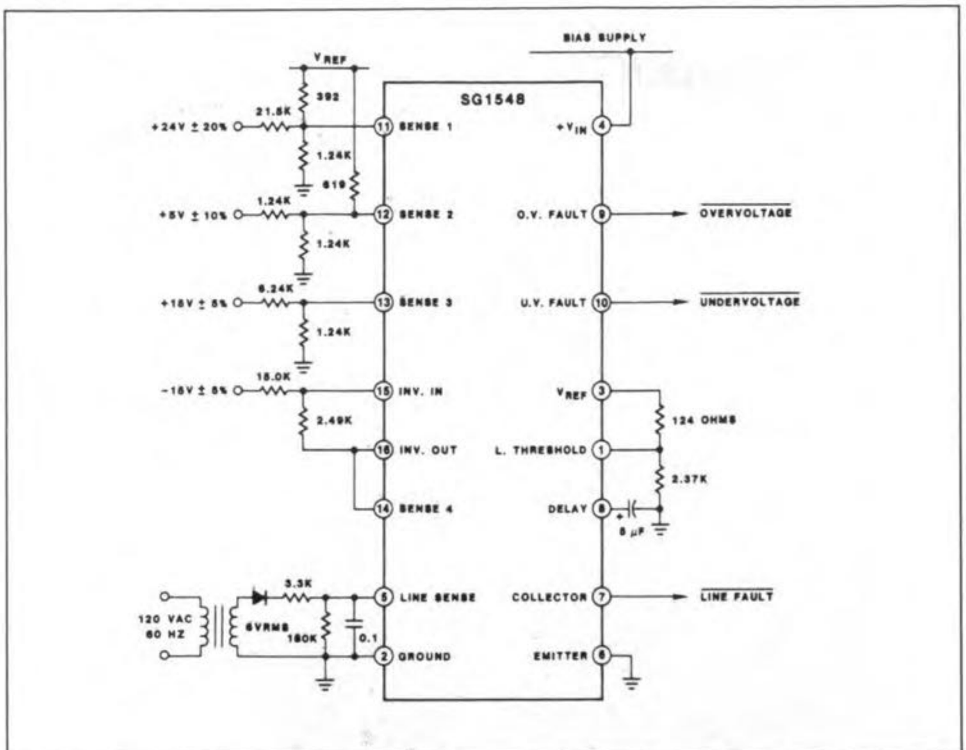
PARAMETER	CONDITIONS	SG1548/2548			SG3548			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	
SUPPLY								
Supply Current	$+V_{IN} = 40V$		4.8	10		4.8	10	mA
REFERENCE SECTION (Note 6)								
Output Voltage	$T_J = +25°C$	2.475	2.500	2.525	2.475	2.500	2.525	V
	Over Temperature	2.450		2.550	2.450		2.550	V
Line Regulation	$+V_{IN} = 4.5$ to 35V		1	5		1	5	mV
Load Regulation	$I_L = 0$ to 10mA		3	10		3	10	mV
Short Circuit Current	$V_{REF} = 0V$	10	25	40	10	25	40	mA
FAULT WINDOW GENERATOR								
Input Bias Current	$V_{Pin\ 1} = +1.50$ to $+2.45V$		−0.4	−1.0		−0.4	−1.0	μA
DC SENSE INPUTS								
Overvoltage Threshold	$V_{Pin\ 1} = 0.98 \times V_{REF}$	1.010	1.020	1.030	1.010	1.020	1.030	V_{TH}/V_{REF}
	$V_{Pin\ 1} = 0.60 \times V_{REF}$	1.386	1.400	1.414	1.386	1.400	1.414	V_{TH}/V_{REF}
Undervoltage Threshold	$V_{Pin\ 1} = 0.98 \times V_{REF}$	0.970	0.980	0.990	0.970	0.980	0.990	V_{TH}/V_{REF}
	$V_{Pin\ 1} = 0.60 \times V_{REF}$	0.594	0.600	0.606	0.594	0.600	0.606	V_{TH}/V_{REF}
Input Bias Current	$V_{SENSE} = +1.5$ to $+3.5V$		±0.6	±2.0		±0.6	±2.0	μA
Threshold Supply Rejection	$+V_{IN} = 4.5$ to 35V	72	100		72	100		dB
FAULT DELAY SECTION								
Comparator Threshold		1.200	1.250	1.300	1.200	1.250	1.300	V
Comparator Hysteresis			25			25		mV
Delay Charging Current	$V_{Pin\ 8} = 0V$	40	50	60	40	50	60	μA
ON Saturation Voltage	$I_{Pin\ 8} = 0mA$		0.1	0.2		0.1	0.2	V
OFF Clamp Voltage	$I_{Pin\ 8} = 0mA$		+3.2	+3.6		+3.2	+3.6	V

Note 6. $I_L = 0$ mA.

PARAMETER	CONDITIONS	SG1548/2548			SG3548			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	
INVERTING OP AMP								
Input Offset Voltage			2	10		2	10	mV
Input Bias Current			−0.3	−1.0		−0.3	−1.0	μA
Output High Voltage	$I_{SOURCE} = 5mA$	13	14		13	14		V
Output Low Voltage	$I_{SINK} = 5mA$		1.0	1.9		1.0	1.9	V
Large Signal Voltage Gain	$R_L = 10K$	72	100		72	100		dB
Output Source Current		5	15	25	5	15	25	mA
Output Sink Current		5	15		5	15		mA
Power Supply Rejection Ratio	$+V_{IN} = 4.5$ to 35V	72	100		72	100		dB
AC LINE SENSE SECTION								
Comparator Threshold	$V_{Pin\ 5}$ = Low to High	2.440	2.500	2.560	2.440	2.500	2.560	V
Comparator Hysteresis			25			25		mV
Input Bias Current	$V_{Pin\ 5} = +2.5V$		1	2		1	2	μA
Collector Leakage Current	$V_{CE} = +40V$		1	10		1	10	μA
Collector Saturation Voltage	$I_C = 10mA$		0.2	0.5		0.2	0.5	V
Emitter Output Voltage	$I_E = 10mA$	12	13		12	13		V
Diode Clamp Voltage	$I_{Pin\ 5} = +1mA$	6.0		7.5	6.0		7.5	V
	$I_{Pin\ 5} = -1mA$	−0.3		−1.0	−0.3		−1.0	V
FAULT LOGIC OUTPUTS (Each output)								
Collector Leakage Current	$V_C = +40V$		1	10		1	10	μA
Collector Saturation Voltage	$I_C = 10mA$		0.2	0.5		0.2	0.5	V

ORDERING INFORMATION

Part Number	T_A Operating Range	Package
SG1548J	−55°C to +125°C	16 Pin Ceramic DIP
SG2548J	−25°C to +85°C	16 Pin Ceramic DIP
SG3548J	0°C to +70°C	16 Pin Ceramic DIP
SG3548N	0°C to +70°C	16 Pin Plastic DIP



In this application example, the SG1548 simultaneously monitors four DC voltages: +5 V, +24 V, and ±15 V. Three different fault tolerances are programmed: ±5% on the two 15-V supplies, ±10% on the +5-V suply, and ±20% on the +24-V supply. The 5-µF capacitor provides 125 milliseconds of fault delay.

APPLICATION INFORMATION

Setting the fault tolerance window

The fault tolerance window is set by applying a voltage less than the +2.50-V reference to the Lower Threshold input (pin 1). The voltage is obtained by a resistor divider from the reference (pin 3) to ground. If a 2% tolerance is desired, then 98% of the reference (+2.45 V) is applied to pin 1. If 40% is wanted, then 60% of the reference (+1.50 V) is applied. In the application circuit, the tolerance is 5% since the divider produces 95% of the reference (2.375 V) at the Lower Threshold pin. The nominal overvoltage and undervoltage thresholds are centred about the reference at 2.625 V and 2.375 V (+2.500 V 0.125 V).

Scaling the monitored supply voltages

Each positive voltage to be monitored is divided down to +2.50 V with a resistor network, and connected to one of the Sense inputs. Unused Sense inputs should be connected to the reference. This will not increase the bias current. A variation of the monitored voltages out of the programmed tolerance range will cause the appropriate overvoltage or undervoltage fault output to switch low.

The effective tolerance on any input may be broadened with an additional resistor to the voltage reference. The application circuit shows a 10% tolerance on the +5-V supply although the SG1548 is programmed for a 5% tolerance.

Monitoring a negative voltage

A negative voltage can be converted to a positive one and simultaneously scaled to +2.50 V by using the internal operational amplifier as an inverter. Only an input resistor and feedback resistor are required.

Setting the fault delay

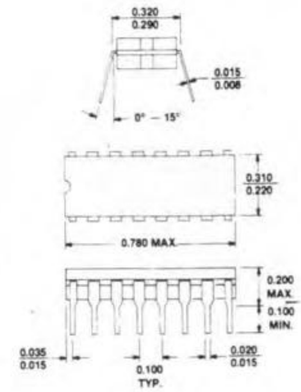
A single capacitor at the Delay pin sets the time an out-of-tolerance fault must persist before a fault is actually declared. This feature allows switching noise on the supplies to be rejected. The delay time is given by:

delay = 25 milliseconds/microfarad

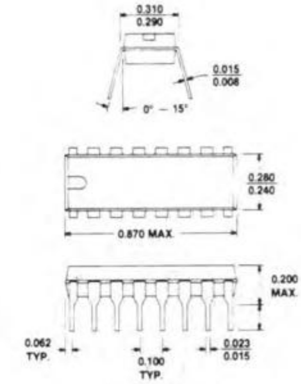
AC line monitoring

The AC line voltage can be monitored for single-cycle dropouts with the few components shown in the application circuit. A half-wave rectifier charges the capacitor on positive line cycles. After the positive peak and during the negative line cycle the capacitor discharges from a fixed voltage controlled by the internal zener diode. If a positive cycle is missing, the capacitor discharges to below the +2.5-V trip point of the comparator, causing the output transistor to turn on.

PACKAGE DIMENSIONS

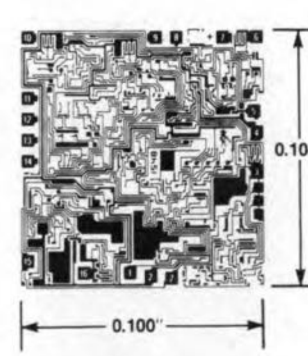


J PACKAGE
16 PIN CERDIP
ALLOY 42 LEADFRAME



N PACKAGE
16 PIN PLASTIC DIP
COPPER ALLOY LEADFRAME

CHIP DIMENSIONS



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IN QUEST OF A PANGRAM

900103-I

by Lee C.F. Sallows

The pangram problem

Some years ago, a Dutch newspaper, the *Nieuwe Rotterdamse Courant*, carried an astonishing translation of a rather tongue-in-cheek sentence of mine that had previously appeared in one of Douglas Hofstadter's *Scientific American* columns ("Metamagical Themas", January 1982). Both the translation and an article describing its genesis were by Rudy Kousbroek, a well-known writer and journalist in Holland. Here is the original sentence:

Only the fool would take trouble to verify that his sentence was composed of ten *a*'s, three *b*'s, four *c*'s, four *d*'s, forty-six *e*'s, sixteen *f* 's, four *g*'s, thirteen *h*'s, fifteen *i*'s, two *k*'s, nine *l*'s, four *m*'s, twenty-five *n*'s, twenty-four *o*'s, five *p*'s, sixteen *r*'s, forty-one *s*'s, thirty-seven *t*'s, ten *u*'s, eight *v*'s, eight *w*'s, four *x*'s, eleven *y*'s, twenty-seven commas, twenty-three apostrophes, seven hyphens and, last but not least, a single *!*

Complete verification is a tedious task: unsceptical readers may like to take my word for it that the number of letters and signs used in the sentence do indeed correspond with the listed totals. A text that inventories its own typography in this fashion is what I call an *autogram* (*auto* = self, *gramma* = letter). Strict definition is unnecessary, different conventions giving rise to variant forms; it is the use of cardinal number-words written out in full that is the essential feature. Below we shall be looking at some in which the self-enumeration restricts itself to the letters employed and ignores the punctuation.

Composing autograms can be an exacting task, to say the least. The process has points in common with playing a diabolically conceived game of patience. How does one begin? My approach is to decide first what the sentence is going to say and then make a flying guess at the number of occurrences of each sign. Writing out this provisional version, the real totals can be counted up and the initial guess updated into an improved estimate. The process is repeated, trials and error leading to successively closer approximations. This opening soon shades into the middle game. By now all of the putative totals ought to have been corrected to within two or three of the true sums. There are, say, nine *f* 's in fact but only *seven* being claimed, and 27 real *t*'s where *twenty-nine* are declared.

An English explorer's self-referent account of his hybrid machine for solving a challenging word puzzle.

Switching *seven* with the *nine* in *twenty-nine* to produce *nine f* 's and *twenty-seven t*'s corrects both totals at a single stroke. Introducing further cautious changes among the number-words with a view to bringing off this sort of mutual cancellation of errors should eventually carry one through to the final phase.

The end game is reached when the number of discrepancies has been brought down to about four or less. The goal is in sight but, as in a maze, proximity is an unreliable guide. Suppose, for instance, a few days' painstaking labour have at last yielded a near-perfect specimen: only the *x*'s are wrong. Instead of the *five* claimed, in reality there are six. Writing *six* in place of *five* will not merely invalidate the totals for *e*, *f*, *s*, and *v*, the *x* in *six* means that their number has now become seven. Yet, replacing *six* by *seven* will only return the total to six. What now?

Paradoxical situations of this kind are a commonplace of autogram construction. Interlocking feedback loops magnify tiny displacement into far-reaching upheavals; harmless truths cannot be stated without disconfirming themselves. Clearly, the only hope of dehydrating this Hydra and getting every snake-head to eat its own tail lies in doctoring the text accompanying the listed items. In looking at the above case, for example, only a fool will fail to spot instances where style has been compromised in deference to arithmetic. Short of a miracle, it is only the flexibility granted through choice of alternative forms of expression that would seem to offer any chance of escape from such a labyrinth of mirrors.

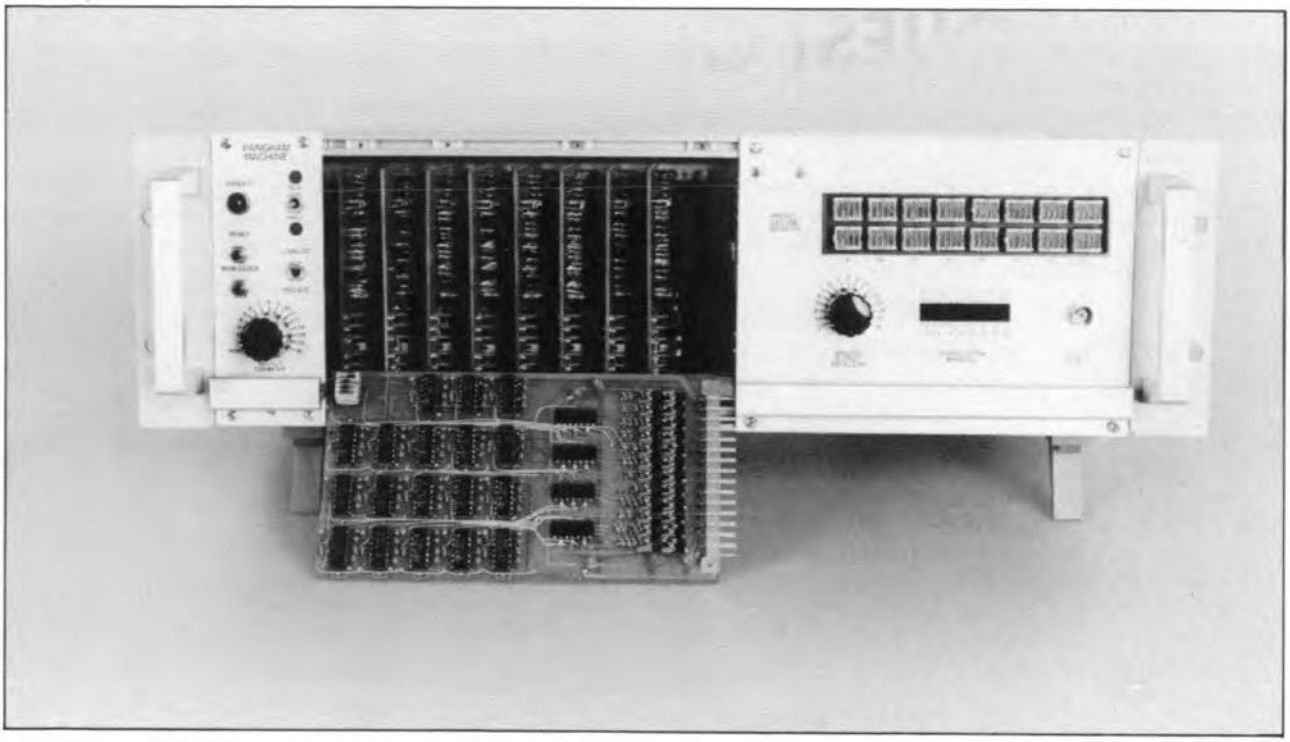
This is what made Kousbroek's translation of my sentence so stunning. Number-words excepted, his rendering not only adhered closely to the original in meaning, it was simultaneously an autogram in Dutch!

Or at least, so it appeared at first sight. Counting up, I was amused to find that three of the sums quoted in his sentence did not in fact tally with the real totals. So I wrote to the author pointing out these discrepancies. This resulted a month later in a second article in the same newspaper. Kousbroek wrote of his surprise and dismay in being caught out by the author of the original sentence, "specially come over from America, it seems, to put me right." The disparities I had pointed to, however, were nothing new to him. A single flaw had been spotted in the supposedly finished translation on the very morning of submitting his manuscript. But a happy flash revealed a way to rectify the error in the nick of time. Later, a more careful check revealed that this 'brainwave' had in fact introduced even more errors elsewhere. He'd been awaiting 'the dreaded letter with its merciless arithmetic' ever since. The account went on to tell of his titanic struggle in getting the translation straight. The new version was included; it is a spectacular achievement.

The tail concealed a subtle sting, however. At the end of his story, Kousbroek threw out a new (letter-only) autogram of his own:

Dit pangram bevat vijf *a*'s, twee *b*'s, twee *c*'s, drie *d*'s, zesenveertig *e*'s, vijf *f* 's, vier *g*'s, twee *h*'s, vijftien *i*'s, vier *j*'s, een *k*, twee *l* 's, twee *m*'s, zeventien *n*'s, een *o*, twee *p*'s, een *q*, zeven *r*'s, vierentwintig *s*'s, zestien *t*'s, een *u*, elf *v*'s, acht *w*'s, een *x*, een *y* en zes *z*'s.

Lee Sallows is an English electronics engineer employed at the Psychology Laboratory of the University of Nijmegen. Besides the design and construction of electronic instruments associated with psychological experiments, he does a good deal of translation work, mostly of scientific and technical papers. A self-confessed dilettante, his interests have included ham radio, psychoanalysis, classical guitar, recreational mathematics and linguistics, impossible figures, logical paradoxes, Sherlockian studies, runology, mountain walking, and writing.



The automatic number-word selector board that transformed the original pangram machine into the Mark II version. On the left, 18 window-detector chips determine the number of *g*'s, *l*'s, *x*'s, and *y*'s. At right, four more integrated circuits and 24 transistors switch in the appropriate PROFILES on the resistor-bearing cards above.

A finer specimen of logological elegance is scarcely conceivable. The sentence is written in flawless Dutch and couldn't possibly be expressed in a crisper or more natural form. In ordinary translation, it says, "This pangram contains five *a*'s, two *b*'s, two *c*'s ... one *y*, and six *z*'s." [A *pangram*, it is necessary to explain, is simply a phrase or sentence containing every letter of the alphabet at least once (*pan* = all, *gramma* = letter). This article is about self-enumerating pangrams, that is, pangrams that are simultaneously autograms. In such pangrams, some letters will occur only at the point where they themselves are listed (look at *k*, *o*, *q*, *u*, *x*, *y*).] Following this pangram came a devilish quip in my direction: "Lee Sallows will doubtless find little difficulty in producing a magic English translation of this sentence," wrote Kousbroek.

Needless to say, I didn't manage to find any errors in *this* sentence of his!

Autograms by computer

Rudy's playful taunt came along at a time when I had already been looking into the possibility of computer-aided autogram construction. Anyone who has tried his hand at composition will know the drudgery of keeping careful track of letter totals. One small undetected slip in counting can later result in days of wasted work. At first I had envisaged no more than an aid to hand-composition: a program that would count letters and provide continuous feedback on the results of keyboard-mediated surgery performed on a sentence displayed on screen. Later I began to wonder what would happen with a program that cycled through the list of number-words, checking each against its corresponding real total and making automatic replacements where necessary. Could autograms be evolved through a repetitive process of selection and mutation? Several such LISP programs were in fact written and tested; the results were not unpredictable. In every case, processing would soon become trapped in an endless loop of repeated exchanges. Increasing refinements in the criteria to be satisfied before a number-word was replaced would win only temporary respite from these vicious circles.

What seemed to be needed was a program that could look ahead to examine the ramifications of replacing *nineteen* by *twenty*, say, before actually doing so. But how is such a program to evaluate or rank prospective substitutions? Goal-directed problem solving converges on a solution by using differences between intermediate results and the final objective so as to steer processing in the direction of minimizing them. The reflexive character of autograms frustrates this approach. As we have seen, proximity is a false index. 'Near-perfect' solutions may be anything but near in terms of the number of changes needed to correct them, while a sentence with as many as eight discrepant totals might be perfected through replacing a single number-word. If hand-composition is obliged to rely on a mixture of guesswork, word-chopping, prayer, and luck, how can a more intelligent strategy be incorporated into a program?

I was pondering this impasse when Rudy Kousbroek's challenge presented itself, distracted my attention, and sent me off on a different tack. The sheer hopelessness of the undertaking caught my imagination. But was it actually impossible? What a comeback if it could really be pulled off! The task was to complete a letter-only autogram beginning, "This pangram contains ...". A solution, were it discoverable, must in a sense already exist 'out there' in the abstract realm of logological space. It was like seeking a number that has to satisfy certain predetermined mathematical conditions. And nobody—least of all Kousbroek—knew whether it existed or not. The thought of finding it was a tantalizing possibility. Reckless of long odds, I put aside programs and launched into a resolute attempt to discover it by hand-trial.

It was a foolhardy quest, a search for a needle in a haystack without even the reassurance of knowing that a needle had been concealed there in the first place. Two weeks' intermittent effort won only the consolation prize of a near-perfect solution: all totals correct save one; there were 21 *t*'s instead of the 29 claimed. With a

The SUMPROFILE

LABEL	e	f	g	h	i	l	n	o	r	s	t	u	v	w	x	y	NUMBER-WORD	LETTER
27	3	0	0	0	0	0	2	0	0	1	2	0	1	1	0	1	twenty-seven	E
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	F
3	2	0	0	1	0	0	0	0	1	0	1	0	0	0	0	0	three	G
5	1	1	0	0	1	0	0	0	0	0	0	0	1	0	0	0	five	H
11	3	0	0	0	0	1	1	0	0	0	0	0	1	0	0	0	eleven	I
2	0	0	0	0	0	0	0	1	0	0	1	0	0	1	0	0	two	L
20	1	0	0	0	0	0	1	0	0	0	2	0	0	1	0	1	twenty	N
14	2	1	0	0	0	0	1	1	1	0	1	1	0	0	0	0	fourteen	O
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	R
28	2	0	1	1	1	0	1	0	0	0	3	0	0	1	0	1	twenty-eight	S
29	2	0	0	0	1	0	3	0	0	0	2	0	0	1	0	1	twenty-nine	T
3	2	0	0	1	0	0	0	0	1	0	1	0	0	0	0	0	three	U
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	V
10	1	0	0	0	0	0	1	0	0	0	1	0	0	0	0	0	ten	W
4	0	1	0	0	0	0	0	1	1	0	0	1	0	0	0	0	four	X
5	1	1	0	0	1	0	0	0	0	0	0	0	1	0	0	0	five	Y
	7	2	2	2	4	1	10	11	2	24	7	1	2	5	1	1	INITIAL TEXT CONSTANTS	
+	27	6	3	5	11	2	20	14	6	28	21	3	6	10	4	5	SUMPROFILE	

Fig. 1. A stack of PROFILES and initial text constants are added to produce a SUMPROFILE. The example shown is the hand-produced near-perfect pangram. All SUMPROFILES and label numbers coincide except that for *T*.

small fudge, it could even be brought to a shaky sort of resolution:

t t t t t
t
t
t

this pangram contains five *a*'s, one *b*, two *c*'s, two *d*'s, twenty-seven *e*'s, six *f*'s, three *g*'s, five *h*'s, eleven *i*'s, one *j*, one *k*, two *l*'s, two *m*'s, twenty *n*'s, fourteen *o*'s, two *p*'s, one *q*, six *r*'s, twenty-eight *s*'s, twenty-nine *t*'s, three *u*'s, six *v*'s, ten *w*'s, four *x*'s, five *y*'s, and one *z*.

To the purist in me, that single imperfection was a hideous fracture in an otherwise flawless crystal. Luckily, however, a promising new idea now suggested itself. The totals in the near-solution must represent a pretty realistic approach to what they would be in the perfect solution, assuming it existed. Why not use it as the basis for a systematic *computer search* through neighbouring combinations of number-words? Each of the near-solution totals could be seen as centred in a short range of consecutive possibilities within which the perfect total was likely to fall. The number of *f*'s, say, would probably turn out to lie somewhere between two and ten, a band of nine candidates clustered about 'six'. With these ranges defined, a program could be written to generate and test every combination of twenty-six number-words constructible by taking one from each. The test would consist in comparing these sets of potential totals with the computed letter frequencies they gave rise to, until an exact match was found, or until all cases had been examined. Blind searching might succeed where cunning was defeated.

Profiles

It isn't actually necessary to deal with all twenty-six totals. In English there are just ten letters of the alphabet that never occur in any number-word between *zero* and *hundred*, the one too low and the other too high to appear in the pangram. These are *a*, *b*, *c*, *d*, *j*, *k*, *m*, *p*, *q*, and *z*. The totals for these letters can thus be determined from the initial text and filled in directly:

This pangram contains five *a*'s, one *b*, two *c*'s, two *d*'s, ? e's, ? *f*s, ? *g*'s, ? *h*'s, ? *i*'s, one *j*, one *k*, ? *l*'s, two *m*'s, ? *n*'s, ? *o*'s, two *p*'s, one *q*, ? *r*'s, ? *s*'s, ? *t*'s, ? *u*'s, ? *v*'s, ? *w*'s, ? *x*'s, ? *y*'s, and one *z*.

This leaves exactly sixteen critical totals. Counting up shows that there are already 7 *e*'s, 2 *f*'s, 2 *g*'s, 2 *h*'s, 4 *i*'s, 1 *l*, 10 *n*'s, 11 *o*'s, 2 *r*'s, 24 *s*'s, 7 *t*'s, 1 *u*, 2 *v*'s, 5 *w*'s, 1 *x*, and 1 *y*: sixteen constants that

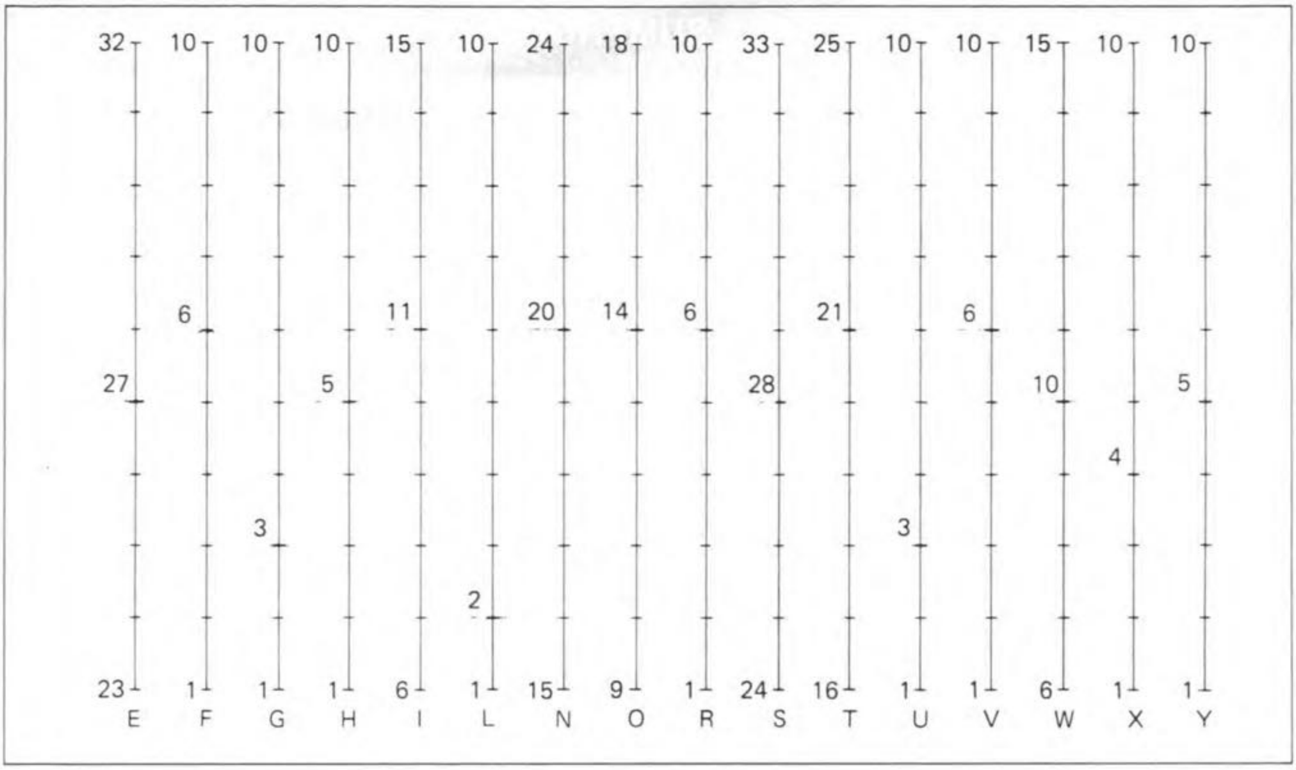


Fig. 2. The range of frequency values to be considered for each letter that appears in number-words.

must be added to those letters occurring in the trial list of sixteen number-words.

Though straightforward in principle, the program I then set out to write carried its practical complications. Number-words lack the regularity of numerals (in whatever base notation), still less the harmony of the numbers both stand for. An obvious step was to replace number-words by PROFILES: alphabetically ordered sixteen-element lists representing their letter content. The PROFILE for *twenty-seven*, for instance, would be:

e f g h i l n o r s t u v w x y
(3 0 0 0 0 0 2 0 0 1 2 0 1 1 0 1)

The letters above the list are for guidance only, and form no part of the PROFILE itself. A special case was the PROFILE for *one*, which provided for the disappearance of plural *s* ('one *x*, two *x*'s') by including −1 in the *s* position. PROFILES for all number-words up to *fifty* (anything higher than *forty* was unlikely ever to be needed) were stored in memory, and a label associated with each. These labels were chosen to coincide with the number represented. The label for the PROFILE of *twenty-seven*, for example, would be the decimal number 27.

Starting with the lowest, a simple algorithm could now generate successive combinations of labels, that is, numbers, drawn from the 16 pre-defined ranges. We shall return to these shortly. Each set of lables would be used to call up the associated set of PROFILES. These 16 PROFILES would be added together element for element, and the resulting sums in turn added to the above-mentioned constants so as to form a SUMPROFILE—see Fig. 1. The SUMPROFILE would thus contain the true letter frequencies for the presently activated sentence (the 16 number-words represented by the current combination of labels plus residual text). All that remained was for the program to check whether the numbers in the SUMPROFILE coincided with the present set of PROFILE labels. If so, the candidate combination of number-words agreed with the real totals and the pangram had been found. If not, generate the next combinations and try again ... .

The simplicity of this design conveys no hint of the uncounted alternatives reconnoitered before reaching it. The 'obvious' PROFILES were not quite so conspicuous as suggested, being in fact a later improvement over a previous look-up table. Weeks were spent in exploring a quite different approach that sought to exploit the mutual-cancelling technique formerly used in hand-composition. By the time the final version of the program had come into focus, half a dozen prototypes lay behind and several months had slipped by. In the mean time, cheerful enthusiasm had given way to single-minded intensity as the problem wormed its way under my skin. Neither was I working entirely alone. Word of the pangram puzzle had spread among colleagues, discussion sprang up and contending design philosophies were urged. At one stage, complaint of "excessive CPU-time devoted to word games" came in from the University of Nijmegen Computing Centre, whose facilities had been shamelessly pressed into service. This was when rival programs were running simultaneously. It was bad enough to be in search of a Holy Grail that might not even exist; the thought of someone else finding it first added a sticky sense of urgency to the hunt.

The question of determining the exact ranges of number-words to be examined seemed to me an essentially trivial one, and I put it off until last. The important thing was to get the program running. For the time being it was enough to decide what the lowest combination was going to be, and to let the algorithm generate all possibilities up to, say, ten higher for each number-word. In terms of software it was convenient for ranges to be of equal length; ten might be unnecessarily high, but better the net be too large than that the fish should escape. Since the totals in the near-solution were to define the midpoint of these ranges, their lower limits would commence at about five less. 'Fourteen *o*'s,' for instance, implied a range running from nine up to eighteen (or perhaps ten up to nineteen). The values actually settled upon—on the basis of pencil-and-paper trials with near-autograms—may be seen in Fig. 2. Ranges for each of the sixteen critical letters are represented as vertical scales with numbers (standing for number-words) indicating their starting and finishing totals. Within these ranges fall the hand-produced near-solution sums tracing out a histogram silhouette. In most cases these are, by definition, situated roughly in the middle of the range. For the low totals l, *g*, and *u*, however, this is impossible: in a pangram all letters must occur at least once; the range cannot extend below *one* (see Fig. 2.).

The second part of this article, reproduced by kind permission of Springer-Verlag, Heidelberg and New York, will apear in the September issue of *Elektor Electronics.*

NEW PRODUCTS

INFRA-RED SENSOR SEES THROUGH DIRT

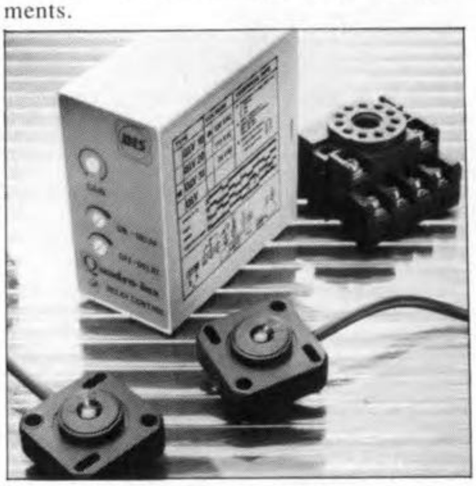
Quiller Ltd. have introduced an infra-red proximity sensor with significant advantages over currently used optical sensing devices.

The Quad-Lux sensor is small, rugged and reliable. Because it has no lenses, the device is claimed to be able to see through condensation, dirt, cloudy atmosphere and even fingers. In such unfriendly environments, the units have the ability to sense target objects up to 15 metres distant.

The operational integrity of each pair of sensors is enhanced by incorporating phase-locked loop electronics in the switching circuitry, eliminating problems associated with signal confusion from other sensors or ambient lighting irregularities.

The sensors are easy to fit and adjust. Signal cabling is at right angles with the detection surface, for loop-free installation.

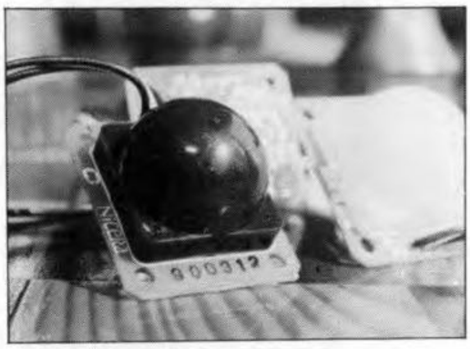
The new sensor is said to have an operational lifetime in excess of three times that of present-day sensors. The device is seen as ideal for applications where dust and dirt build up have traditionally caused problems. These conditions typically arise in outdoor applications, woodworking areas, foundries, mining, farms and similarly 'dirty' environments.



A wide range of mounting arrangements is offered. Dimensions are 15×30×30 mm. Operating temperature range is from −25°C to +65°C. Typical operational lifetime is 100,000 hours. Control amplifiers are also available.

Quiller Ltd. • 2 Paisley Road • BOURNEMOUTH BH6 5EU. Telephone: (0202) 417744. Fax: (0202) 421255.

NEW FRONT END PIR MODULE



The SGM5910 passive infra-red (PIR) module is intended for the detection of human body movements. The unit measures 35×30×30 mm, and has an in-built lens providing coverage of 100 degrees in the horizontal plane and 60 degrees in the vertical plane.

Output is of the open-collector (active-low) type, and the SGM5910 will operate on connection of 5 to 12 VDC at 1 mA.

The use of a moulded semi-hemispheric lens (black or white) provides a high degree of protection and should enable the unit to be added to equipment by the simple expedient of drilling a 25-mm hole through the face and inserting from behind. Versions for wall mounting and ceiling mounting are available with different pyrosensors.

Chartland Electronics • Chartland House • Twinoaks • COBHAM KT11 2QW. Telephone: (0372) 842553/843976. Fax: (0372) 844264.

WIDE RANGE OF MINIATURE CRYSTALS

Euroquartz have introduced new miniature quartz crystals to its wide range held in stock.

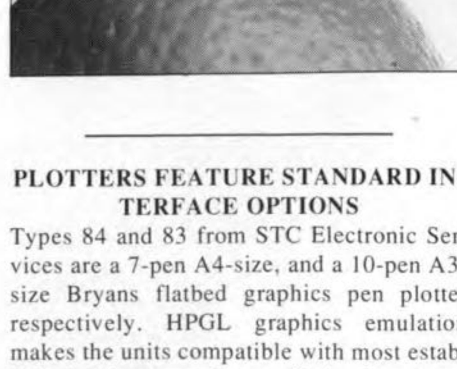
A new 32.768 kHz device is available in two package styles: 3×8 mm and 2×6 mm. Frequency tolerances available include ±15 ppm and ±20 ppm and typical temperature stability is −0.035 ppm/°C^2. Operating temperature is −10 to +60°C whilst ageing is specified as 3 ppm maximum at 25°C over the first year.

HC-45/UM-1 type crystals are available in frequencies from 1.0 MHz to 100.0 MHz with calibration tolerances of ±50 ppm at 25°C for devices from 1.0 to 1.048576 MHz, and ±30 ppm for devices from 3.579545 MHz to 100.0 MHz. These crystals are housed in a package measuring 7.1×8.0×2.4 mm.

Euroquartz is also stocking a wide range of miniature cylindrical crystals covering various sizes and frequency ranges including 2×6 mm, 25.6 kHz to 200 kHz, 3×8 mm, 16 kHz to 200.0 kHz, 3×9 mm, 3.579545 MHz to 26 MHz, and 3×10 mm, 3.20 MHz to 3.50 MHz.

As well as the wide range of stock crystals, Euroquartz can also manufacture or supply devices to order.

Euroquartz Ltd. • Blacknell Lane Industrial Estate • Crewkerne • Somerset TA18 7HE. Telephone: (0460) 76477. Fax: (0460) 77229.



PLOTTERS FEATURE STANDARD INTERFACE OPTIONS

Types 84 and 83 from STC Electronic Services are a 7-pen A4-size, and a 10-pen A3-size Bryans flatbed graphics pen plotter respectively. HPGL graphics emulation makes the units compatible with most established software packages. Both plotters are suitable for ordinary paper and transparency film, and are backed with a wide variety of consumable options.

Further details from

STC instrument Services • Dewar House • Central Road • HARLOW CM20 2TA. Telephone: (0279) 641641.