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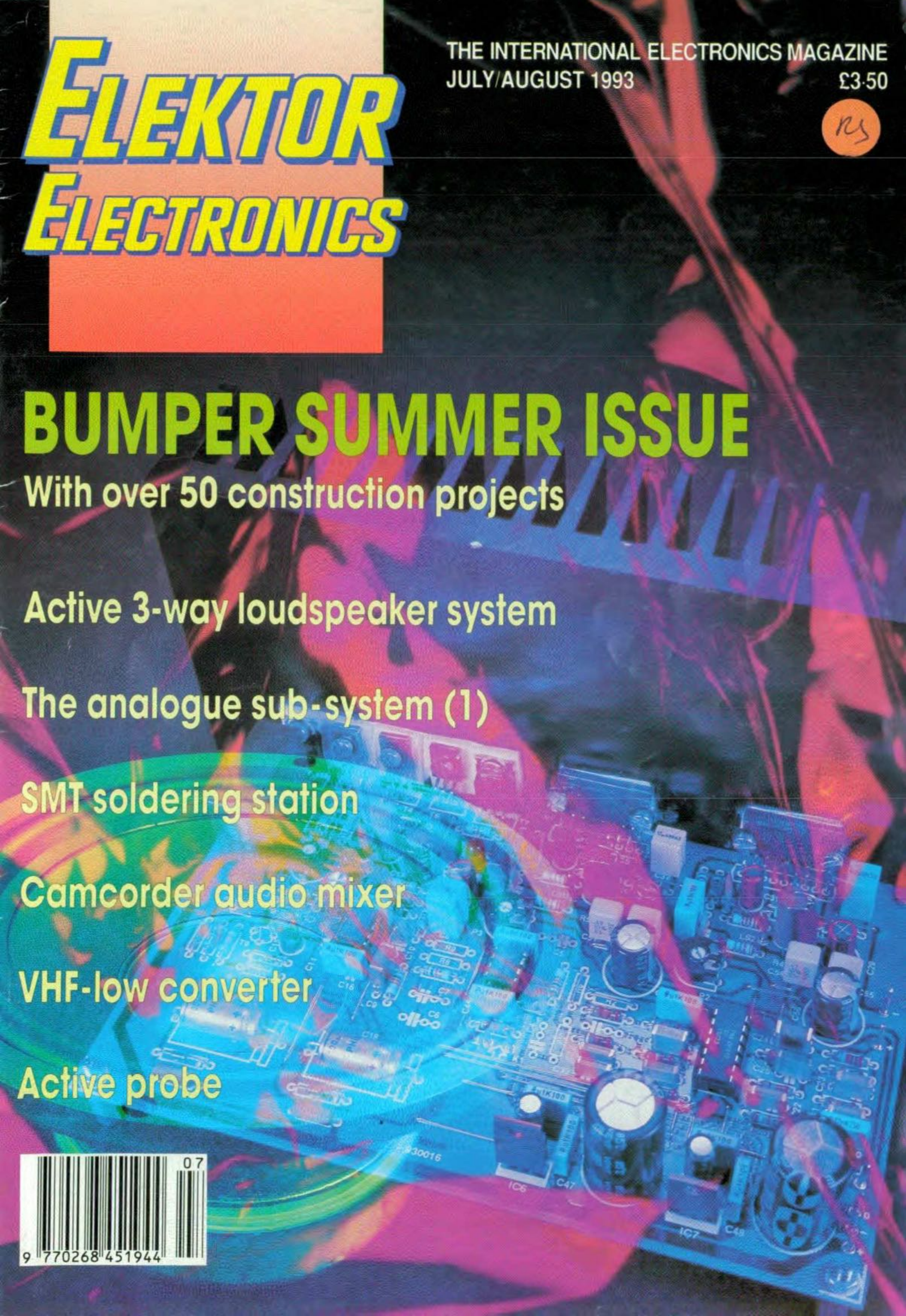

BUMPER SUMMER ISSUE

With over 50 construction projects

Active 3-way loudspeaker system

The analogue sub-system (1)

SMT soldering station

Camcorder audio mixer

VHF-low converter

Active probe

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SMD SOLDERING STATION

Design by T. Will

Surface mount technology (SMT) was developed in the 1980s to make possible the use of very small components that have no wire leads or pins. These components are now termed surface mount devices (SMDs). For many years, these new components were thought to be too difficult to use outside large-scale electronics manufacturing plant. However, more and more people are now losing their reluctance to working with SMDs and these components are, therefore, becoming more readily available from electronics retailers. This development brings with it a growing need for tools suitable for working with these tiny components. One of these is a temperature-controlled soldering outfit. The one described in this article is appreciable less expensive than commercially available models.

The circuit

The circuit consists of a variable direct voltage source and an LED display.

Transformer Tr_1 steps down the mains voltage to 15 V, which is subsequently rectified by diodes D_{12}–D_{15}, buffered by capacitors C_3 and C_4, and regulated by IC_2. The level of the output voltage is determined by potential divider R_5-R_6-R_7-P_1. The drop across R_7 is constant at 1.2 V. With the values of these components shown in the diagram, the output voltage may be varied with P_1 over the range 6.8–11.8 V.

The function of R_6 is twofold. On the one hand it ensures that should the preset become faulty the output voltage does not rise unduly, and on the other it changes the operating range of P_1 into a non-linear one, which is more convenient in practice.

The LED display indicates the level of the direct voltage applied to the soldering iron. Since the soldering iron has no temperature sensor, it is not possible to measure the temperature of the soldering tip. This is not too much of a handicap because the exact temperature is not very important in the first place, and in the second, the manufacturer's data show that with an applied voltage of 12 V the temperature of the tip rises to a maximum of 330 °C. As soon as the hot tip is applied to the point to be soldered, its temperature drops in proportion with the thermal load. This results in a maximum, practical temperature of some 300 °C. The minimum temperature may be assumed to be 200 °C. Temperatures in this range are indicated by D_2–D_{11}. Since each LED draws a current of 10 mA, the dissipation of IC_1 rises rapidly with increasing tip temperature. If no precautions were taken, the total current would rise to 100 mA. Since each LED drops about 2 V, IC_1 would dissipate 1 W, which, since the manufacturer's data indicate a maximum permissible dissipation of 625 mW, would mean its immediate demise. This all means that the dissipation must be reduced and this is brought about by R_4. The more diodes light, the larger the drop across this resistor and the more electrical energy it converts into heat. This is very effective in reducing the dissipation of IC_1.

Resistors R_1 and R_2 set the operating range of IC_1, which operates as an analogue-to-digital (A-D) converter and display driver. This is done in conjunction with the internal reference source. With the specified values of the resistors, the display driver range lies between 1.8 V and 3.0 V. At a voltage of 1.8 V at the signal input (pin 5), the first LED begins to light and at 3.0 V all LEDs light. Diode D_1 lights as soon as the power is switched on (and supply voltage is available).

The (miniature) soldering iron has two terminals, of which one is connected to earth and the other to the variable direct voltage. The iron is shunted by potential divider R_8-R_9-P_2. This network reduces the direct voltage to about a quarter of its value. This means that the monitoring circuit is provided with a variable direct voltage of 1.75–3.0 V. This is exactly the same as the voltage range for the display driver.

The earth connection is made via R_{10}. This arrangements prevents any build-up of static charge on the iron, which might damage the electronic circuit on

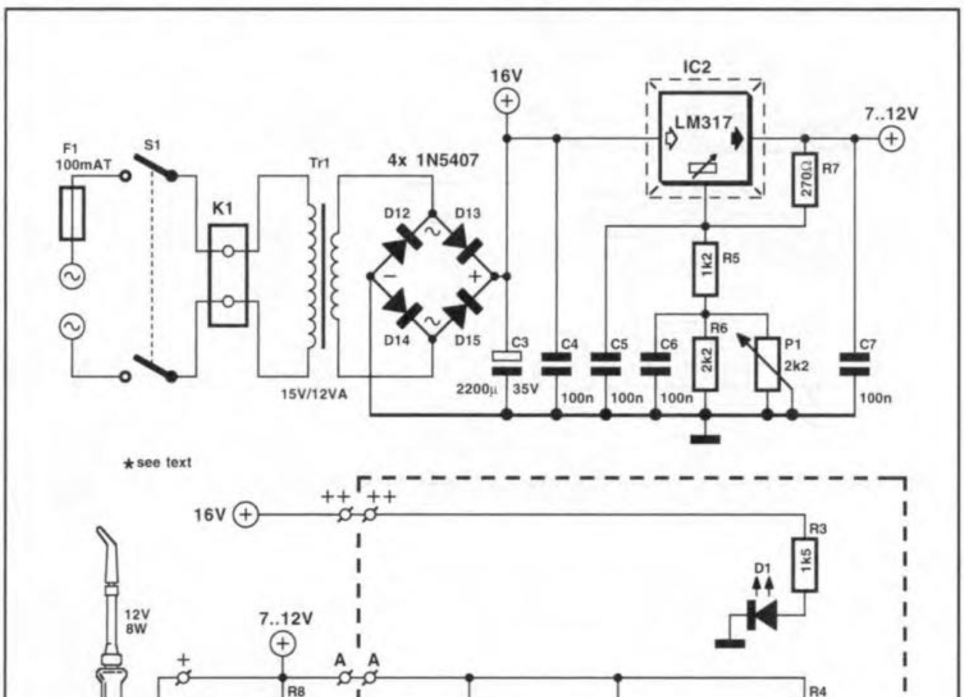


Fig. 1. Circuit diagram of the temperature-controlled SMD soldering station.

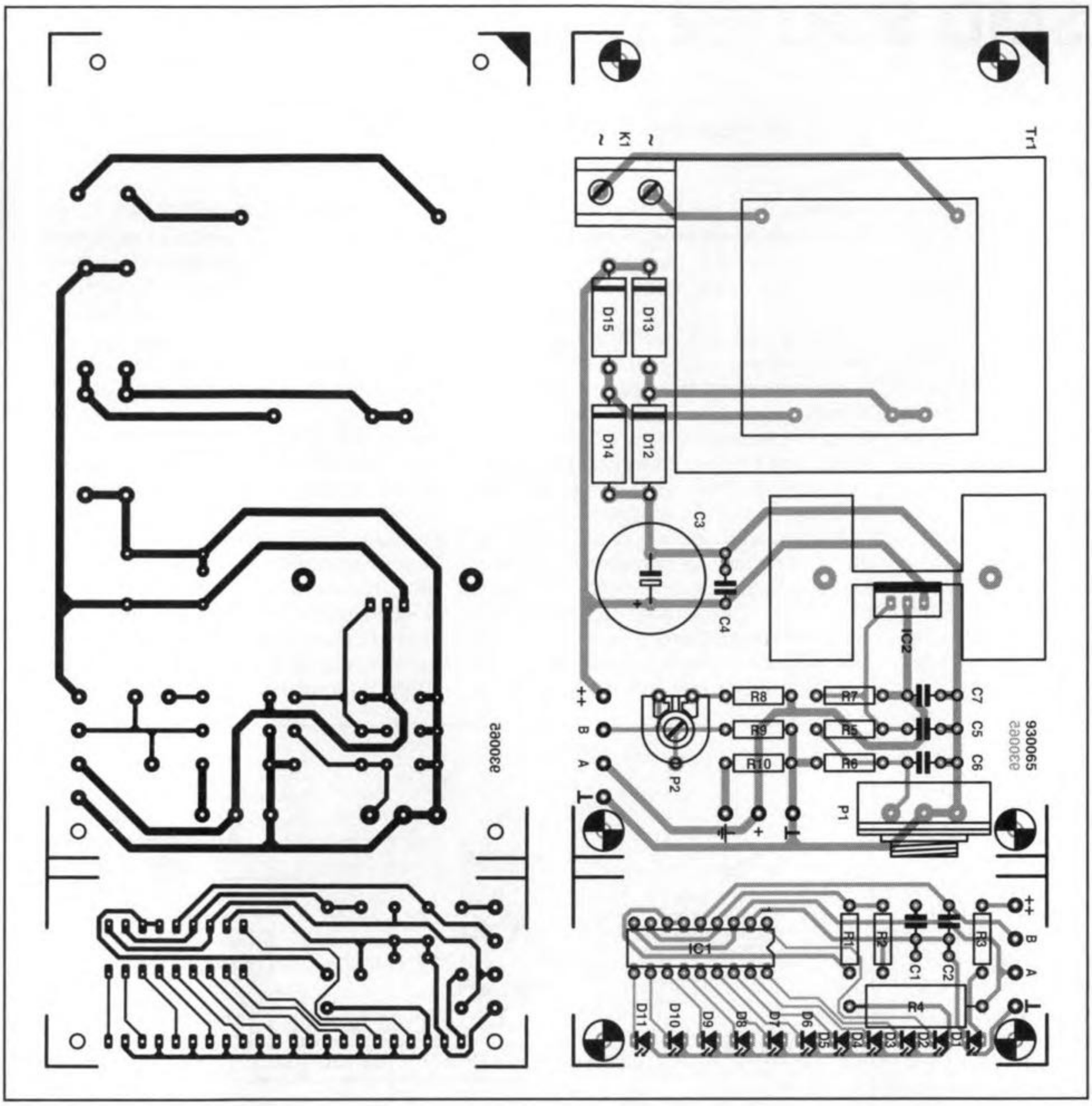


Fig. 2. The printed-circuit board for the soldering station should be cut into two.

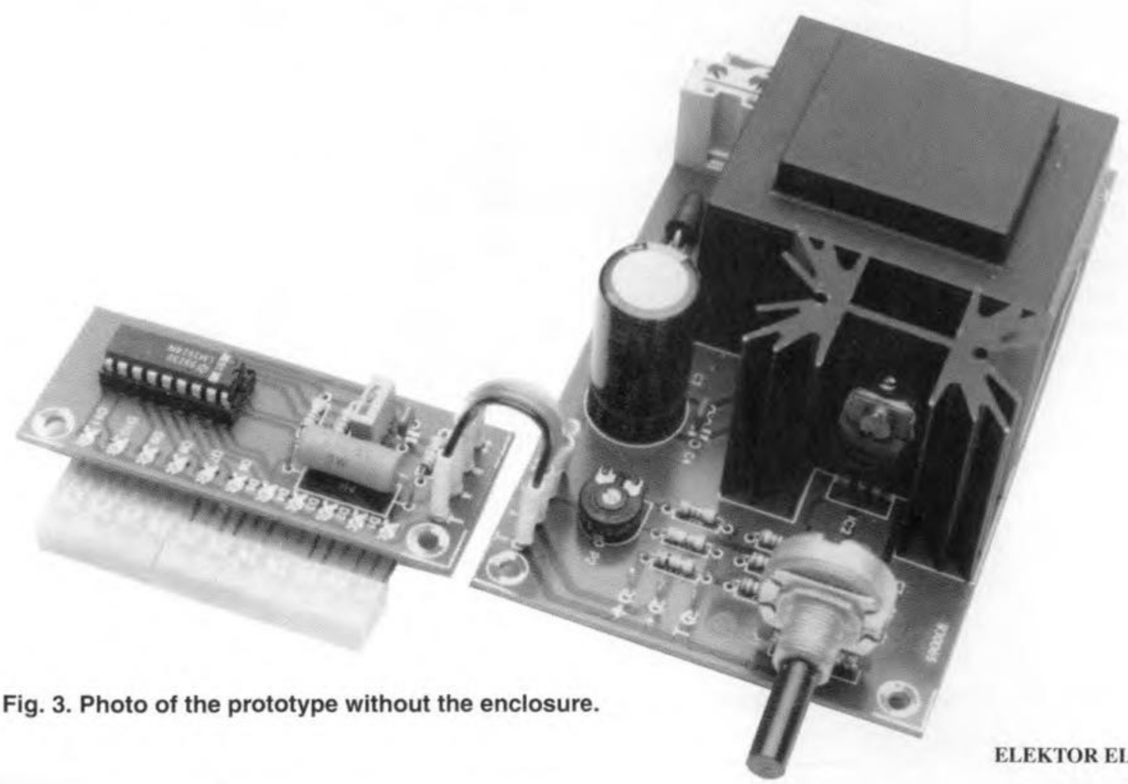


Fig. 3. Photo of the prototype without the enclosure.

PARTS LIST

Resistors:
R1 = 1.2 kΩ
R2 = 1.8 kΩ
R3 = 1.5 kΩ
R4 = 68 Ω, 5 W
R5 = 1.2 kΩ
R6 = 2.2 kΩ
R7 = 270 Ω
R8 = 8.2 kΩ
R9 = 3.9 kΩ
R10 = 100 kΩ
P1 = 2.2 kΩ preset (linear)
P2 = 2.5 kΩ preset

Capacitors:
C1 = 220 nF
C2, C4–C7 = 100 nF
C3 = 2200 µF, 35 V, radial

Semiconductors:
D1–D3 = LED, square 5×5 mm, green
D4–D7 = LED, square 5×5 mm, orange
D8–D11 = LED, square 5×5 mm, red
D12–D15 = 1N5407
IC1 = LM3914
IC2 = LM317

Miscellaneous:
K1 = 2-way terminal block, pitch 7.5 mm
K2 = 3-way terminal block, pitch 5 mm
S1 = mains on/off switch, 3 A
Tr1 = mains transformer, 15 V, 12 VA (e.g., Monacor Type VTR12115)
Heat sink SK129/35/SA*
3 banana sockets (red, blue, black)
Mains entry with integral switch and fuse (100 mA delayed action)
Enclosure 150×80×55 mm ($5^{15}/_{16}$×$3^{3}/_{16}$×$2^{3}/_{16}$ in) (e.g. OKW 9030087)
PCB Type 930065 (see p.110)
Front panel foil Type 930065F (see p. 110)

*Dau Electronics Phone (0243) 553031

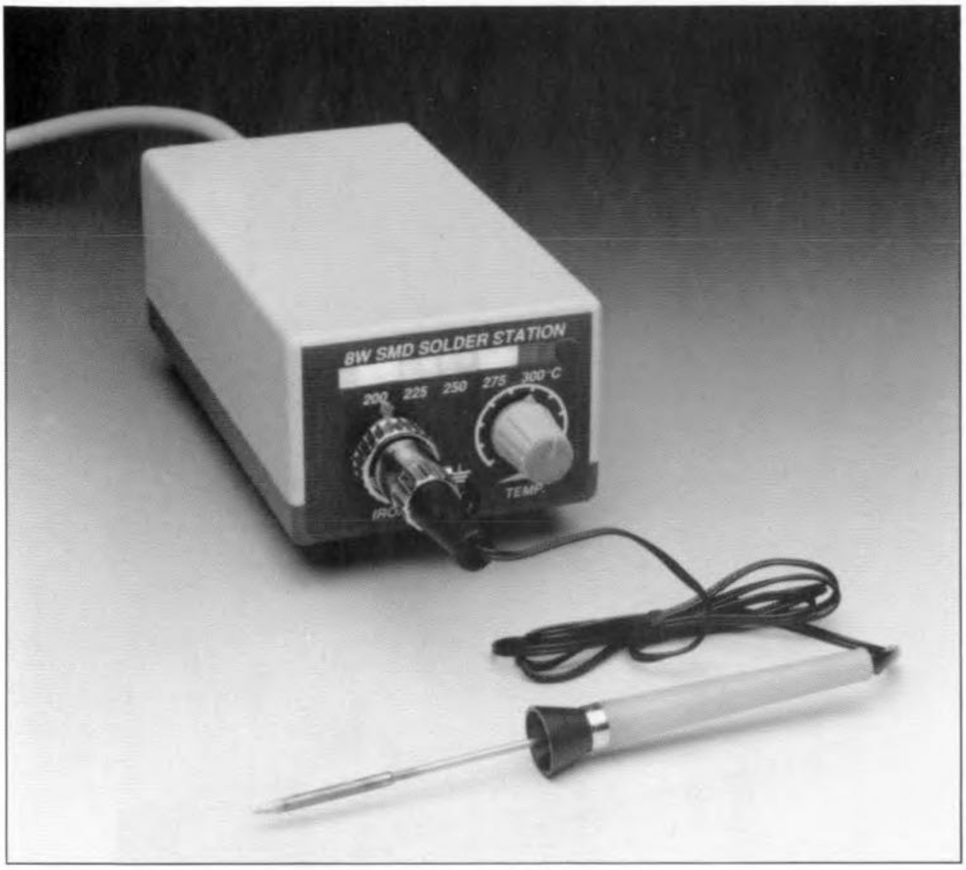


Fig. 4. The completed soldering station in its enclosure.

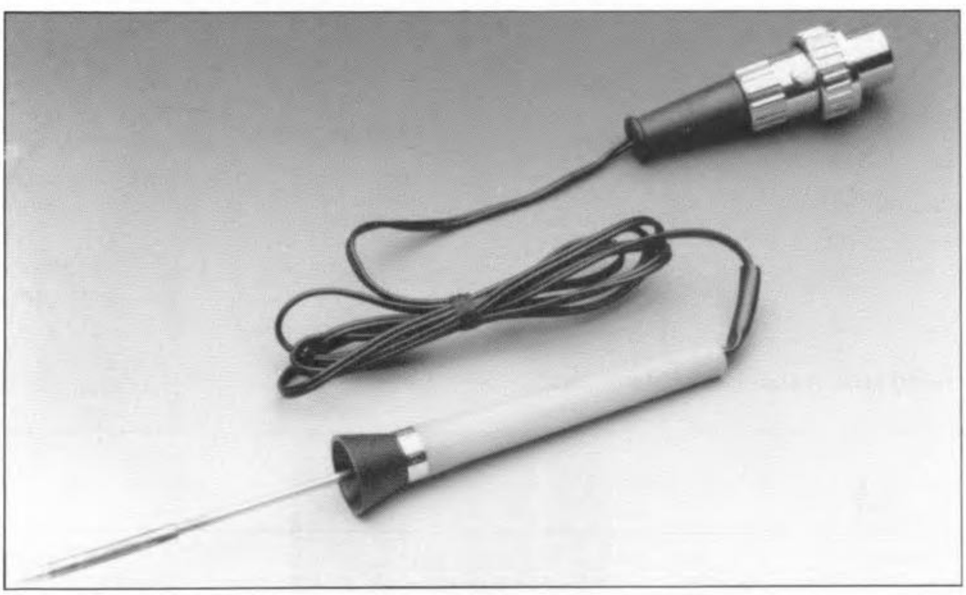


Fig. 5. This miniature 8 W soldering iron is ideal forsoldering SMDs.

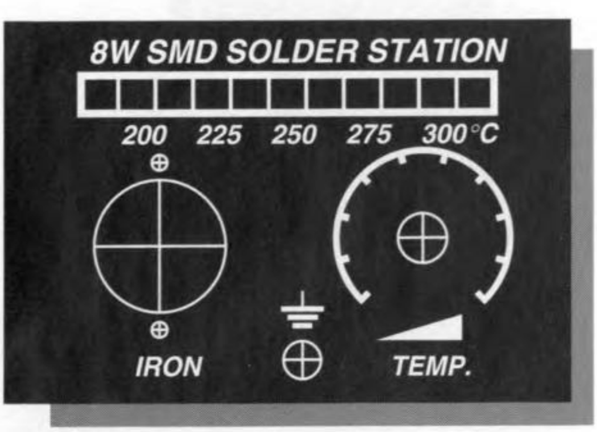


Fig. 6. A front panel foil is available through our Readers' Services (see p. 110).

which soldering is carried out.

Construction

The circuit is best built on the printed-circuit board shown in Fig. 2. Before any construction takes place, however, the board must be cut into two. Populating the resulting supply board an display board is straightforward. Commence with fitting the connectors and transformer to the supply board, followed by the passive components and IC_2. Make sure of good thermal contact between the IC and its heat sink: use a generous amount of heat conducting paste.

Although P_1 may be soldered direct to the board, it is far better to solder it to three solder pins, which gives a much more solid (and more reliable) construction.

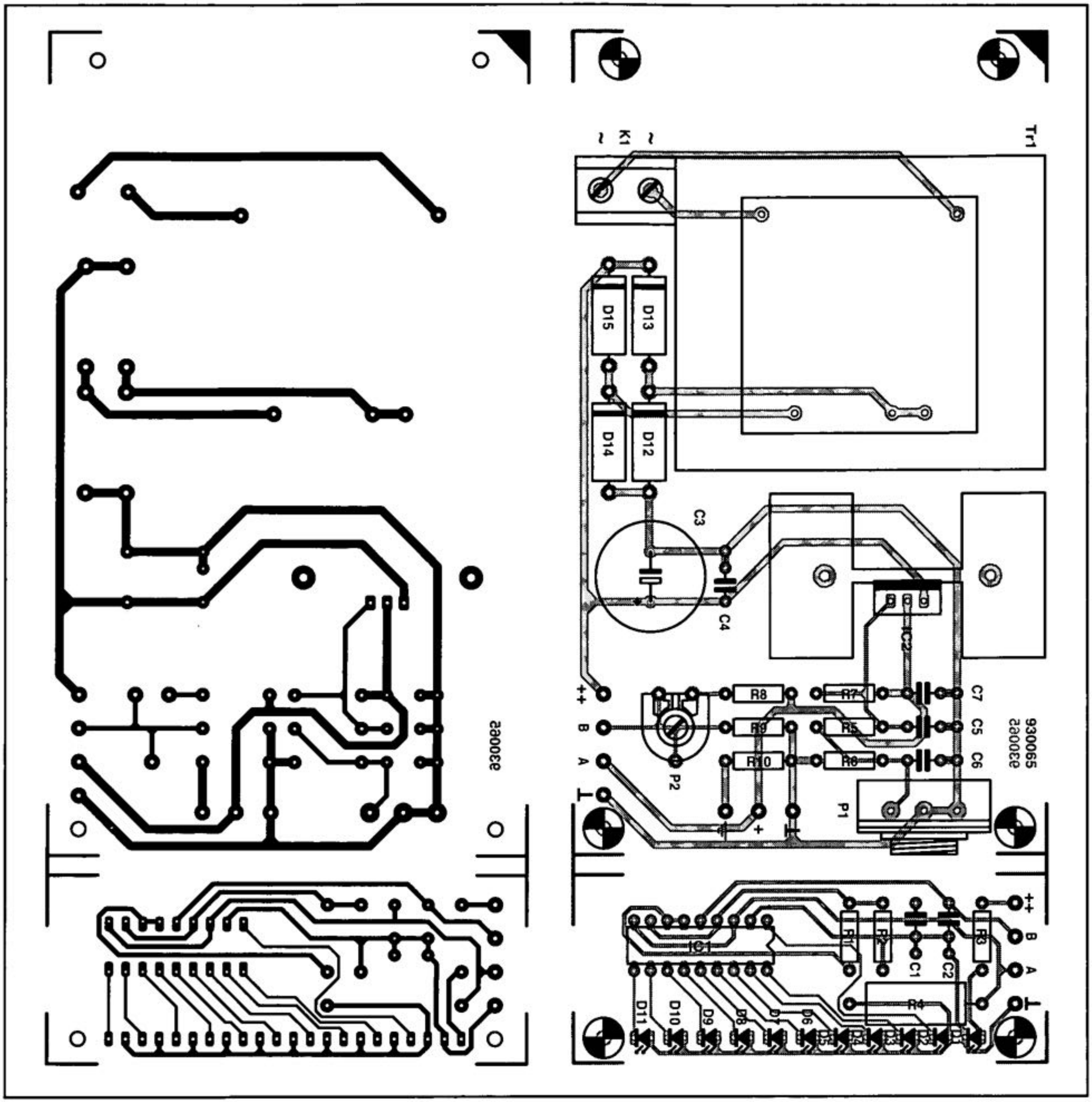
Use square LEDs on the display board: when these are fitted side by side, they result in a pleasing indicator bar.

When both boards have been completed, they should be housed in a suitable enclosure. For the prototype, an OKW box Type 9030087 (150×80×55 mm; $5^{5}/_{16}\times3^{3}/_{16}\times2^{3}/_{16}$ in) was used, but any non-metal box of the same dimensions will do nicely. A pleasing front panel foil is available through our Readers' services (see Fig. 5).

Before the boards can be fitted into the enclosure, the necessary holes and slots must be made for the LED display, the mains cable, the three banana sockets and the on/off switch.

The display board is mounted upside down on to the supply board with the aid of suitable spacers.

When the mains voltage is switched on, supply on indicator D_1 should light. Set temperature controller P_1 fully anticlockwise and adjust P_2 until D_2 just begins to light. Next turn P_1 fully clockwise, whereupon all LEDs should light. The soldering station is then ready for use. ■



PARTS LIST

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K2 = 3-way terminal block, pitch 5 mm
S1 = mains on/off switch, 3 A
Tr1 = mains trans-

ACTIVE 3-WAY LOUDSPEAKER - PART 1

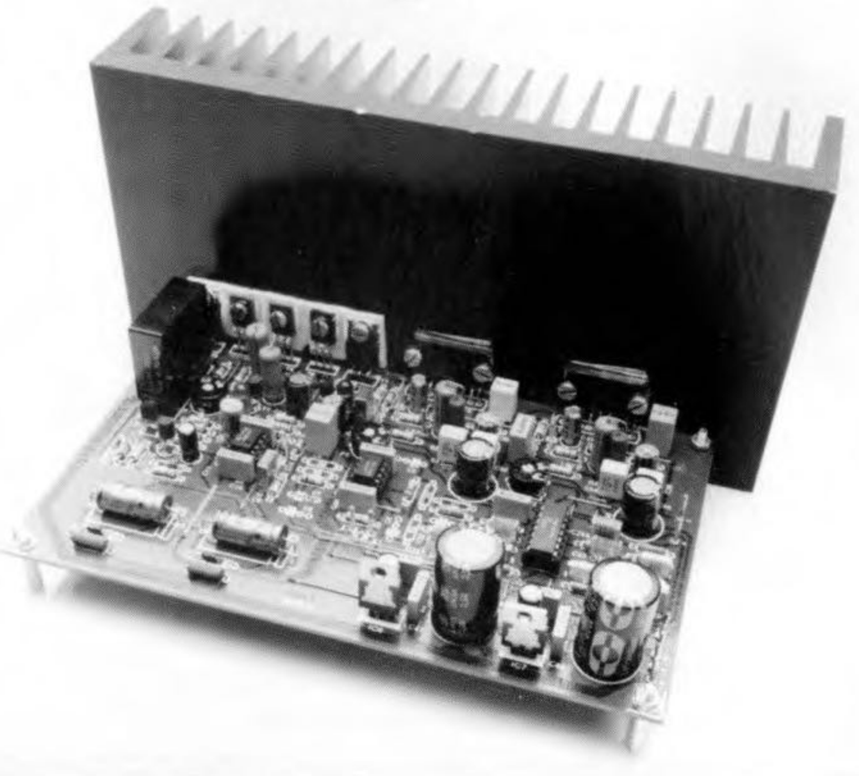
Design by T. Giesberts and H. Baggen

Designing an active loudspeaker system is a complex (but fascinating) matter because each drive unit needs its own output amplifier and cross-over filter. But, of course, an active system has several advantages over a passive one. The present circuit, comprising three filter sections, active bass corrective network, and three output stages, fits on just one printed-circuit board.

Two of the advantages of an active loudspeaker system over a passive one are that neither loudspeaker cables nor a passive cross-over filter are required. The lack of loudspeaker cables saves money,too, since good quality ones are expensive. A passive filter dissipates energy in the inductors and capacitors, and this causes some deterioration in the quality of sound reproduction. An active filter uses no inductors and the capacitors have a much lower value, so that their quality can be of better quality. Moreover, in direct coupling there are no capacitors and inductors with loss resistances, and this means that the loudspeakers are under more direct control of the output stages.

Two of the drawbacks of an active system are its cost and higher complexity. These two go together, because it is the complexity of the electronic circuits (each loudspeaker needs it own output stage, for instance) that costs the money.

In the present design, the output stages have been kept small through the use of modules for the medium and high frequency sections.



Parameters

Three filters and three output stages on one PCB
Filters may be 1st order, 2nd order, or 3rd order as required
Optional facility for Linkwitz correction network

Power rating output stages	- low	70 W into 4 Ω (U_b=±25 V)
		70 W into 8 Ω (U_b=±35 V)
	- middle	30 W into 8 Ω
	- high	30 W into 8 Ω
Nominal sensitivity	1.1 V r.m.s.	
Input impedance	47 kΩ	

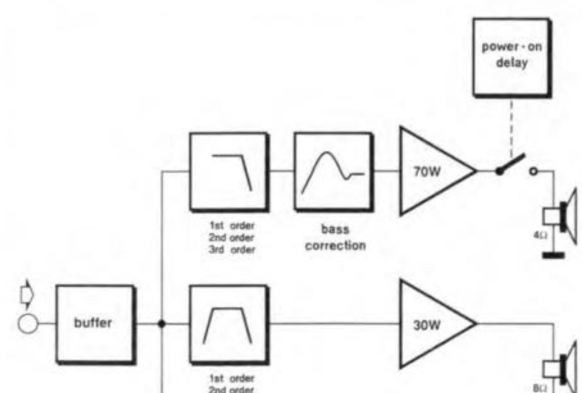


Fig. 1. Block diagram of the active 3-way loudspeaker system.

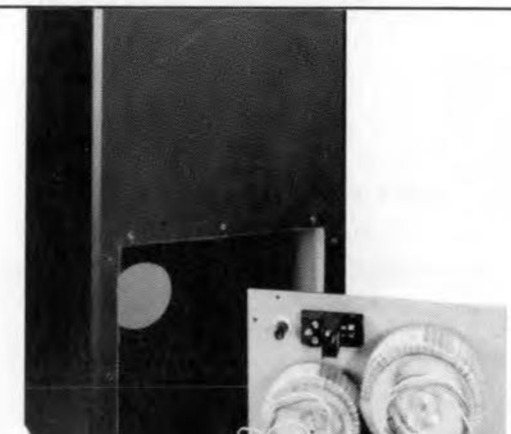


Fig. 2. Detail of prototype.

In this first part of the article, the electronic circuits will be described in general, while in the second part the design proper, including a three-way bass reflex enclosure, will be discussed.

Some design considerations

The design (see block diagram) allows for various configurations of the cross-over networks. The audio signal from the preamplifier is applied to the three filters (bass, medium and treble) via a buffer stage. Each of the filters may be given a rolloff of 6 dB, 12 dB or 18 dB per octave depending on the value of certain components.

The low-pass filter is followed by a bass correction network. The original design of this network is due to Linkwitz. It is particularly useful to lower the response of the woofer in a closed box.

The formulas for the calculations of the correct component values and the rolloffs will be given later.

The outputs of the filters are applied to amplifiers. Note that the LF stage has more than twice the power output of the middle and high frequency one. Since the LF stage has been designed with discrete components, its output is applied to the drive unit via a power-on delay. This delay is an integral part of the medium and high frequency modules.

Practical design

Opamp IC_{2b} (see Fig. 3) buffers the applied audio signal to prevent this being loaded unnecessarily. The input impedance is determined by R_1. The output of IC_{2b} is split threeway to IC_{1a}, IC_{1c}, and IC_{1b}. Each of these circuits forms a third order filter, which may be converted into a first or second order type by omitting certain

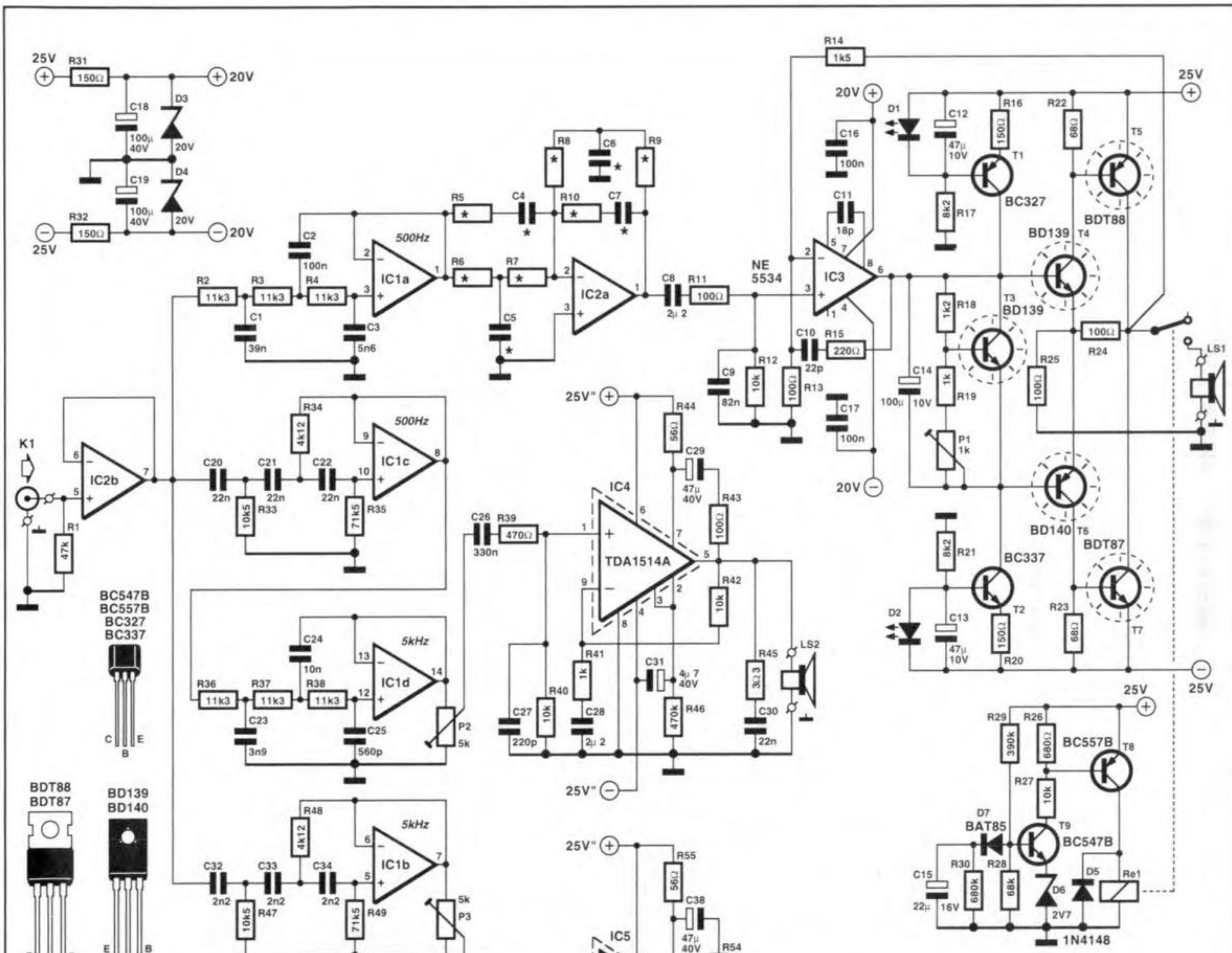


Fig. 3. Circuit diagram of the active 3-way loudspeaker system.

components. The middle-frequency section has two filters, IC_{1c} and IC_{1d}, because it needs a rolloff at the low frequency end and one at the high frequency end. With values as shown, the cut-off frequencies are at 500 Hz and 500 Hz respectively. The response is a Butterworth type.

It is possible to convert the present system into an active two-way one by omitting the entire middle-frequency section and the associated output stage (IC_4).

The low-pass filter is followed by the Linkwitz correcting network that matches the frequency response to the low cut-off point of the box. In this way, an octave is added to the lower portion of the response of the enclosure. It is, however, only possible to use the arrangement with boxes whose Q_{tc} and f_c are known. The calculations of the values of the network components are given later.

The supply voltages for the opamps are stabilized by IC_6 and IC_7. These voltages are derived from the ±25 V supply for the LF output amplifier (and for the other output stages if this is desired—more about this later).

Output modules IC_4 and IC_5 need only a few external passive components and a feedback loop. If the supply to them is ±25 V, they deliver an output of up to 30 W into 8 Ω. Noteworthy in the diagram of their internal circuitry, shown in Fig. 2, are the many protection circuits. Their input has a mute stage to prevent on

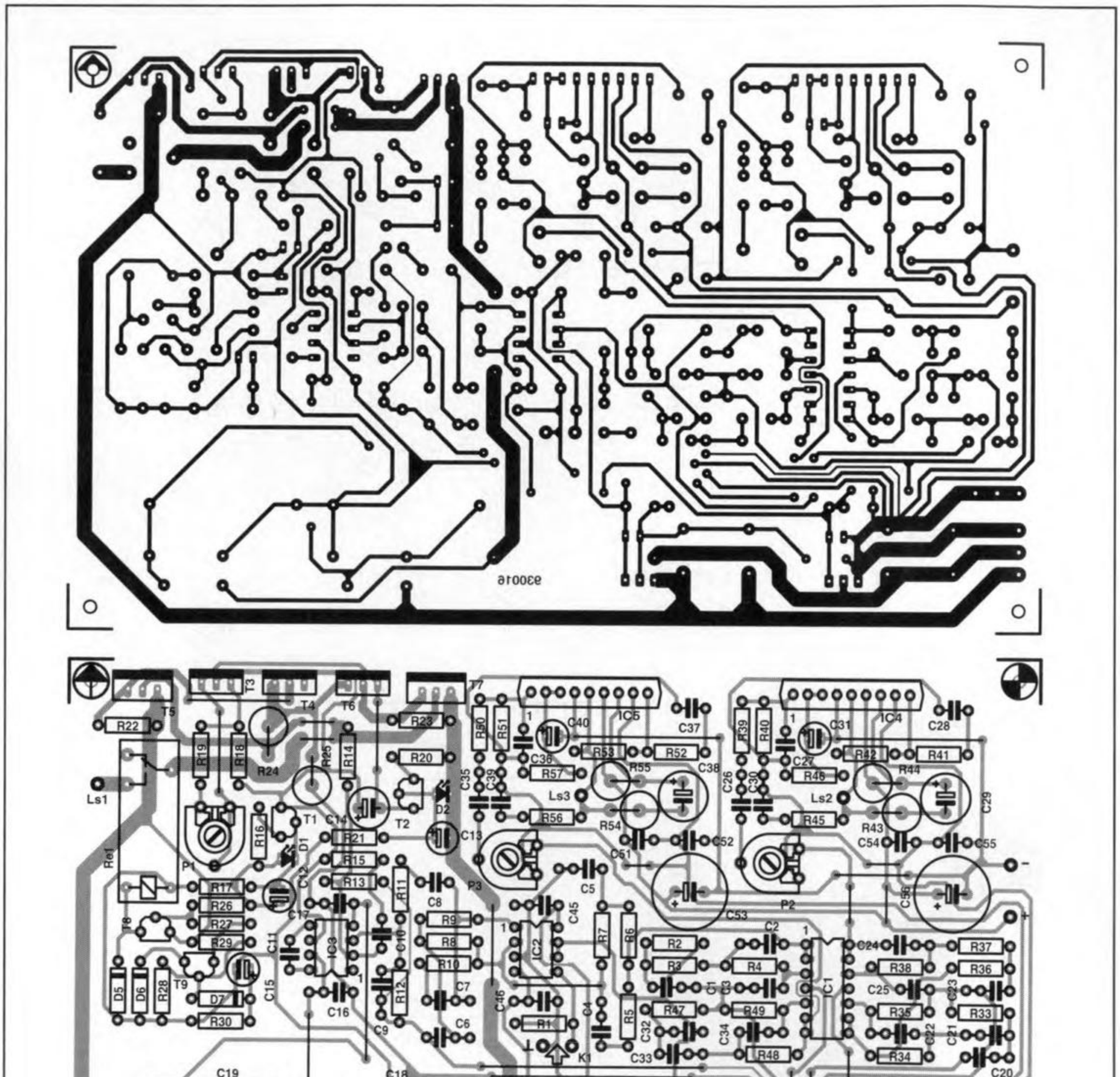


Fig. 4. The printed-circuit board for the active 3-way loudspeaker system.

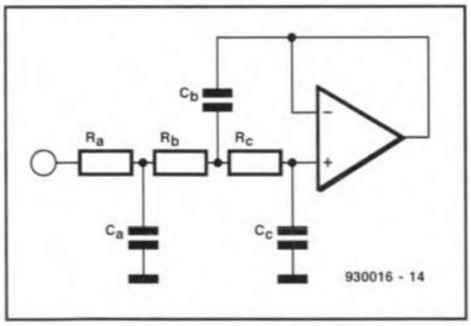
PARTS LIST

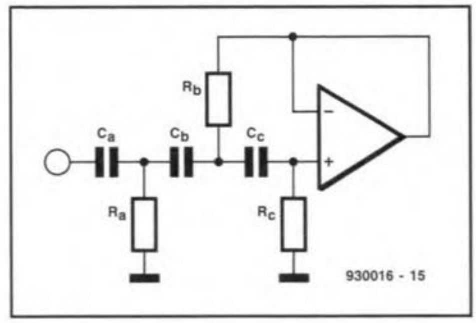
Resistors:
R1 = 47 kΩ
R2–R4, R36–R38 = 11.3 kΩ, 1%
R5–R10 = see text
R11, R13 = 100 Ω
R12, R40, R42, R51, R53 = 10 kΩ
R14 = 1.5 kΩ
R15 = 200 Ω
R16, R20, R31, R32 = 150 Ω
R17, R21 = 8.2 kΩ
R18 = 1.2 kΩ
R19, R41, R52 = 1 kΩ
R22, R23 = 68 Ω
R24, R25 = 100 Ω, 2.5 W
R26 = 680 Ω
R27 = 10 kΩ
R28 = 68 kΩ
R29 = 390 Ω
R30 = 680 kΩ
R33, R47 = 10.5 kΩ, 1%
R34, R48 = 4.12 kΩ, 1%
R35, R49 = 71.5 kΩ, 1%
R39, R50 = 470 Ω
R43, R54 = 100 Ω, 1.5 W
R44, R55 = 56 Ω, 1.5 W
R45, R56 = 3.3 Ω
R46, R57 = 470 kΩ
P1 = 1 kΩ preset
P2, P3 = 4.7 kΩ (5 kΩ) preset

Capacitors:
C1 = 39 nF
C2, C16, C17, C43–C48 = 100 nF
C3 = 5.6 nF
C4–C7 = see text
C8, C28 = 2.2 μF, 50 V, MKT
C9, C35 = 82 nF
C10 = 22 pF
C11 = 18 pF
C12, C13 = 47 μF, 10 V, radial
C14 = 100 μF, 10 V, radial
C15 = 22 μF, 16 V, radial
C18, C19 = 100 μF, 40 V
C20–C22, C30, C39 = 22 nF
C23 = 3.9 nF
C24 = 10 nF
C25 = 560 pF, polystyrene
C26 = 330 nF
C27, C36 = 220 pF
C29, C38 = 47 μF, 40 V, radial
C31, C40 = 4.7 μF, 40 V, radial
C32–C34 = 2.2 nF
C37 = 680 nF
C41, C42 = 10 μF, 25 V, radial
C49, C50 = 1000 μF, 40 V, radial
C51, C52, C54, C55 = 470 nF
C53, C56 = 47 μF, 63 V, radial

Semiconductors:
D1, D2 = LED, 3 mm, red
D3, D4 = zener diode, 20 V, 1.5 mW
D5 = 1N4148
D6 = zener, 2.7 V, 400 mW
D7 = BAT85
T1 = BC327
T2 = BC337
T3, T4 = BD139
T5 = BDT88
T6 = BD140
T7 = BDT87
T8 = BC557B
T9 = BC547B
IC1 = TL074
IC2 = NE5532
IC3 = NE5534
IC4, IC5 = TDA1514A
IC6 = 7815
IC7 = 7915

Miscellaneous:
K1 = audio plug for PCB mounting
Re1 = relay with make contact, 24 V coil
Heat sink, 0.5 K W^{-1} (SK47)
PCB Type 930016





	Low-pass		High-pass	
	Butterworth	Bessel	Butterworth	Bessel
1st order	R_a=10 kΩ R_b=R_c=wire bridge C_b=C_c=not used C_a=0.1592/fR		C_a=4.7 nF C_b=C_c=wire bridge R_b=R_c=not used R_a=0.1592/fR	
2nd order	R_a=wire bridge C_a=not used R_b=R_c=10 kΩ		C_a=wire bridge R_a=not used C_b=C_c=4.7 nF	
	C_b=0.2251/fR_b C_c=0.1125/fR_b	C_b=0.1443/fR_b C_c=0.1082/fR_b	R_b=0.1125/fC_b R_c=0.2251/fC_b	R_b=0.1755/fC_b R_c=0.234/fC_b
3rd order	R_a=R_b=R_c=10 kΩ		C_a=C_b=C_c=4.7 nF	
	C_a=0.2215/fR_a C_b=0.5644/fR_a C_c=0.03221/fR_a	C_a=0.1572/fR_a C_b=0.2265/fR_a C_c=0.04039/fR_a	R_a=0.1125/fC_a R_b=0.04488/fC_a R_c=0.7864/fC_a	R_a=0.1611/fC_a R_b=0.1116/fC_a R_c=0.6272/fC_a

Table 1. Computations for low-pass and high-pass filters.

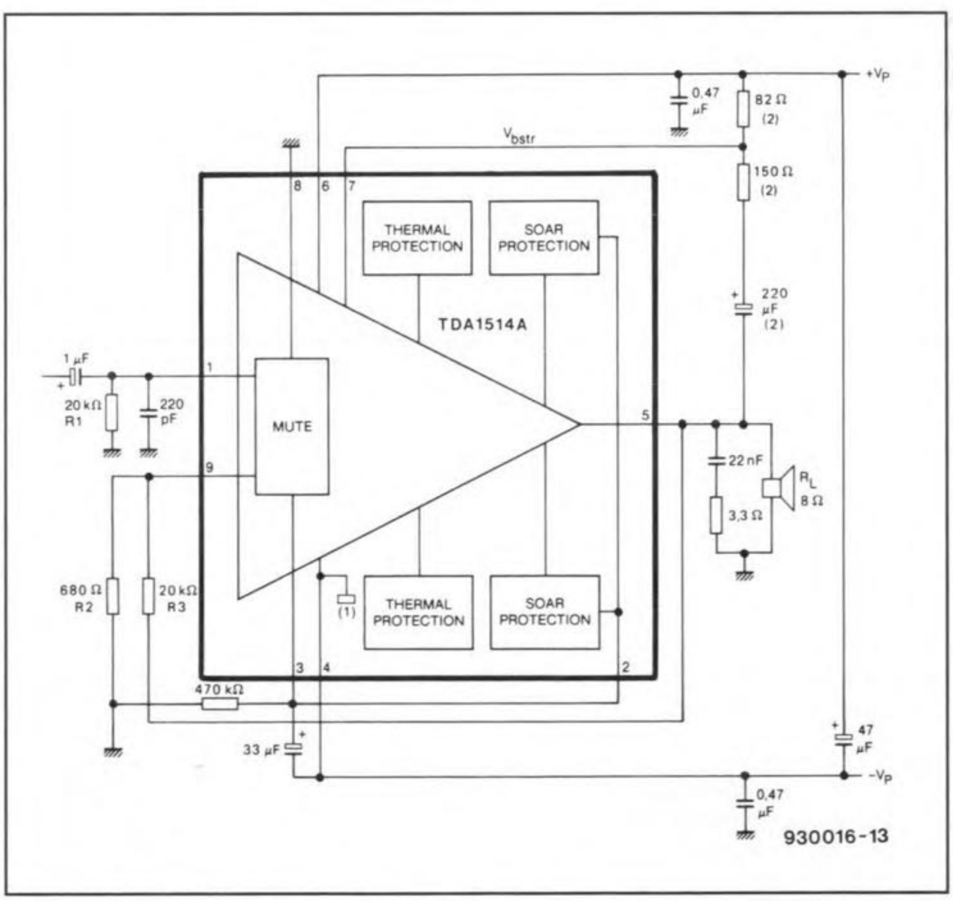
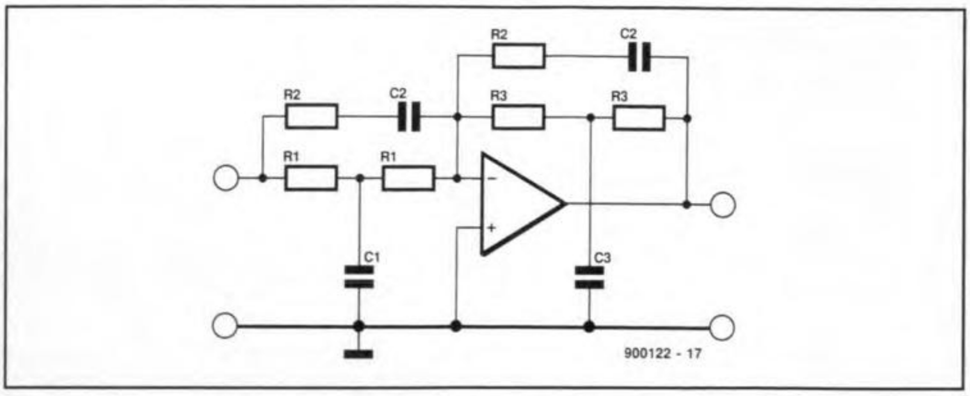


Fig. 4. Circuit diagram of the TDA1514A.



Parameters required to be known
Q_{tc} (Q factor of loudspeaker in closed box)
f_c (resonance frequency of loudspeaker in closed box

Required new parameters
Q_{tc} '(new Q_{tc} with correction network)
f_c' (new f_c with correction network)

The new parameters should be chosen so that the necessary correction is not too large to prevent the loudspeaker operating outside its linear performance.

Condition for chosen Q_{tc}' and f_c':

$$k = \frac{\dfrac{f_c}{f_c'} - \dfrac{Q_{tc}}{Q_{tc}'}}{\dfrac{Q_{tc}}{Q_{tc}'} - \dfrac{f_c'}{f_c}} > 0$$

where k is the pole shifting factor

Computation
Choose a value for R_1 and compute the remaining components as follows

$$R_2 = 2kR_1$$

$$R_3 = \left(\frac{f_c}{f_c'}\right)^2 \cdot R_1$$

$$C_1 = \frac{2Q_{tc}(1+k)}{2\pi f_c R_1}$$

$$C_2 = \frac{1}{4\pi f_c Q_{tc} \cdot (1+k)R_1}$$

$$C_3 = \left(\frac{f_c'}{f_c}\right)^2 \cdot C_1$$

Table 2. Computations for Linkwitz correction network.

and off switching becoming audible.

Concentrating on IC_4, capacitor C_{26} prevents any d.c. components reaching the opamp. Low-pass filter R_{39}-C_{27} limits the bandwidth of the input signal to a degree that is suitable for the output amplifier. The input impedance of that amplifier is determined by R_{40}. The a.c. amplification of the module is set by feedback network R_{42}-R_{41}. The d.c. amplification is limited to unity by C_{28}. Network R_{43}-R_{44}-C_{29} forms a bootstrap that increases the full power available from the output stage slightly. Boucherot network R_{45}-C_{30} in parallel with the loudspeaker serves as load at high frequencies when the loudspeaker becomes inductive. Power-on delay is provided by R_{46}-C_{31}.

The presets between the filters and the loudspeakers serve to match the speakers as far as efficiency is concerned.

The LF output stage is a semi-discrete design to obtain a higher power since human hearing is less sensitive to low frequencies. The board has been designed to enable this amplifier operating from a power supply different from that for the middle and high frequency stages.

The supply for the differential input opamp, IC_3, is derived from the main supply via networks R_{31}-D_3 and R_{32}-D_4. The input impedance of the opamp is determined mainly by R_{12}. The bandwidth of the input to the opamp is limited by R_{11}-C_9. R_{15}-C_{10} is a compensating network.

The output of IC_3 is applied to a compound output stage, which provides not only current amplification, but also voltage amplification. This allows the opamp, although its output is limited by the ±20 V supply, to fully drive the power stage, whose supply is ±25 V.

The output stage consists of T_4-T_5 for the positive half of the signal and T_6-T_7 for the negative half. The power stage is driven by IC_3 via two current sources, T_1-D_1 and T_2-D_2. The quiescent current through the output stage is determined by T_3, which functions as a variable zener diode. The overall feedback is provided by R_{13} and R_{14}.

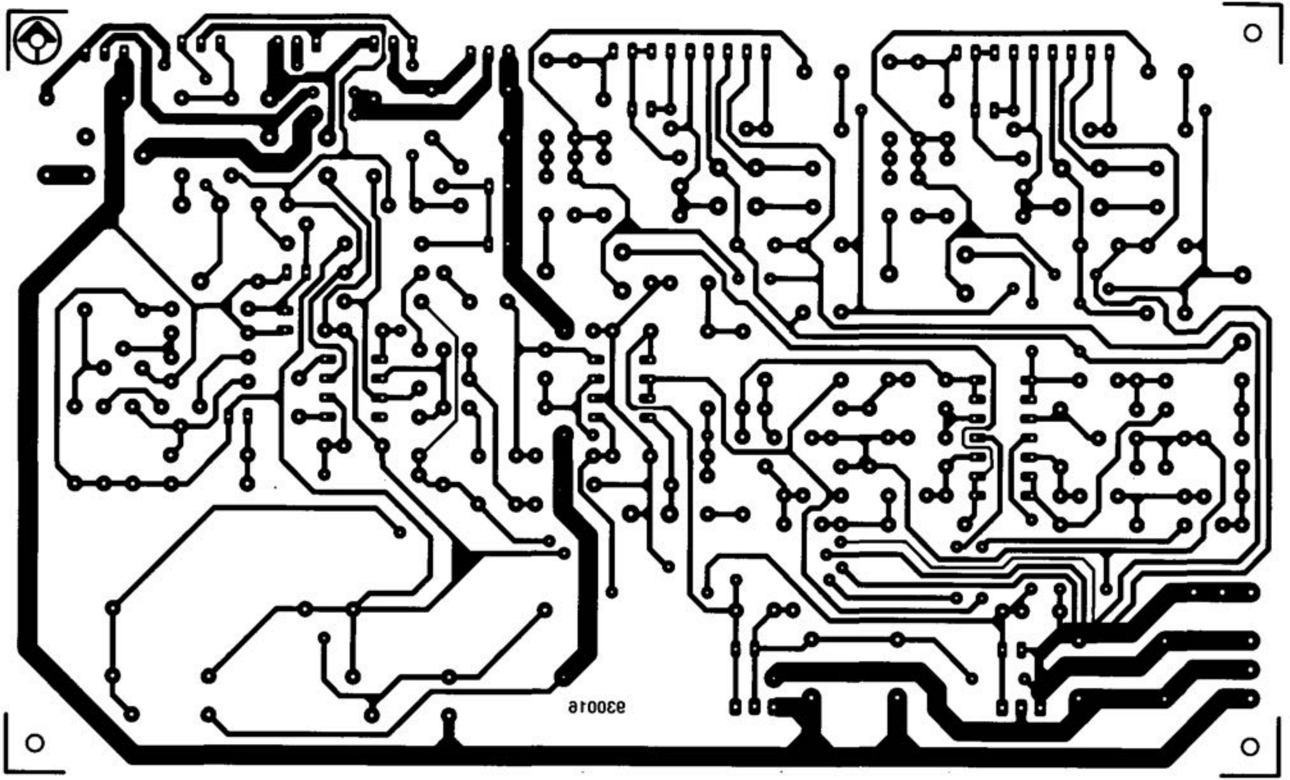
The relay contact at the output of the low-frequency power stage prevents on/off clicks becoming audible. At power on, the relay is energized after a delay by T_8 and T_9. Before T_9 can conduct, C_{15} needs to be charged via R_{29} and D_7 to a voltage of 2.7 V (D_6) plus 0.6 V (base-emitter junction of T_9). Diode D_7 ensures that the relay is deenergized immediately the supply voltage is switched off.

Pre-construction notes

As already stated, the entire electronics part of the system is fitted on a single printed-circuit board—see Fig. 3. Before construction can be started, the gradients and cut-off points of the filters need to be known, and these are dependent on the loudspeaker enclosure. Component values of the bass correction network are also dependent on the bass enclosure used. Formulas for the computation of the component values for the filters and the correction network are given in Tables 1 and 2. The design of the circuit in Fig. 3 is based on third-order filters and cut-off frequencies of 500 Hz and 5 kHz. Different values may be needed, or it may be desired to imitate the performance of a passive filter (in an existing enclosure) in active form. All this is possible and will be discussed in more detail in Part 2.

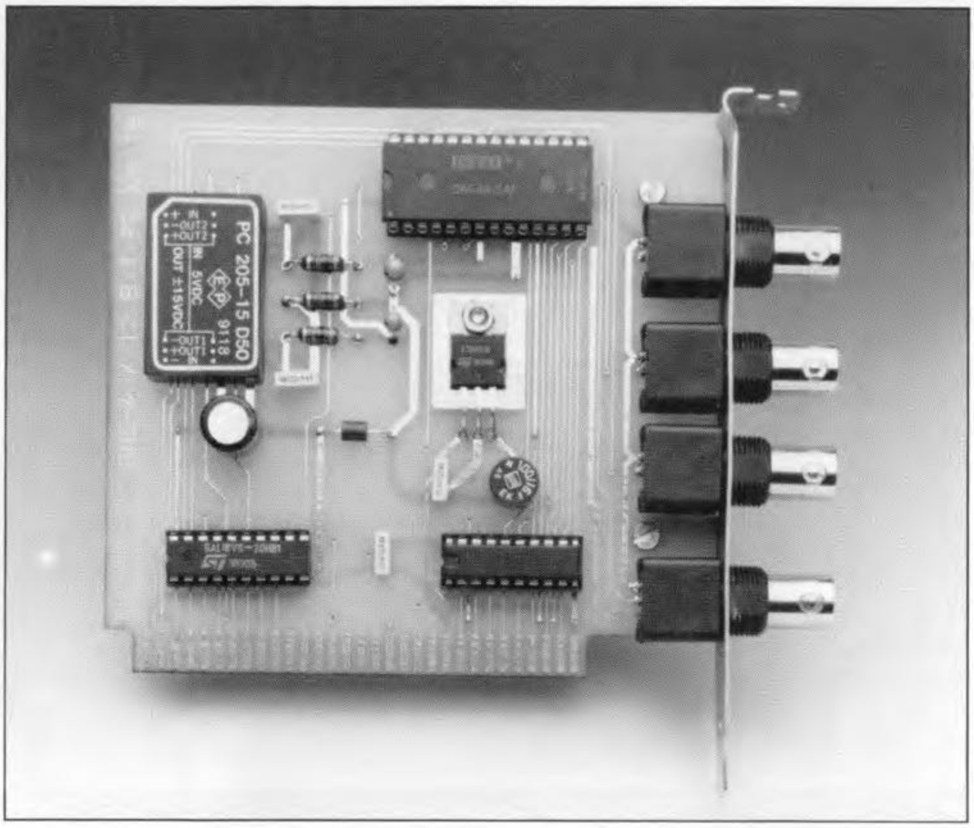
The construction of the output stages depends to some extent on the loudspeakers used. The power distribution over the audio spectrum shows that in a multi-way system the woofer needs at least twice as much power as the other units combined. This requirement may be met by using a 4 Ω drive unit for the woofers and 8 Ω speakers for the other units. If that is not possible, for instance, because the low-frequency drive unit is not available in a 4 Ω version, the low-frequency output stage may be powered by a higher supply voltage. In the present design, this is achieved by connecting a separate 25 V supply to terminals + and – adjacent to capacitor C_{56}. The supply must not exceed ±35 V since regulators IC_6 and IC_7 cannot handle higher input voltages. A drawback of this arrangement is the higher price since two separate power supplies are needed.

Another aspect that needs bearing in mind is the polarity of the drive units. Owing to the bass correction network, the bass unit is out of phase with the other two units and its polarity must therefore be reversed. If the correction network is not used, all components around IC_{2a} with an asterisk must be omitted, except R_7 and R_{10}, whose value must be altered to 10 kΩ. Resistor R_6 and capacitors C_7 must be replaced by a wire bridge. ■


930016

FOUR-FOLD DAC CARD FOR PCs

Measurement cards of the PC insertion type to digitize analogue voltages are available in countless versions. By contrast, digital to analogue converters are few and far between. The card presented here is fast, and offers four independent analogue channels with a resolution of 12 bits.



Design by H. Kolter

ALTHOUGH it often appears that everything around us goes digital, the realm of analogue control is still wide, and has much in store for the electronics designer. The measurement and control field, in particular, has a traditional requirement for PC-driven direct voltage sources. Unfortunately, most I/O cards have no analogue output at all, or one with a resolution of only 8 bits. To meet today's requirements for accuracy coupled with speed, a D-A (digital-to-analogue) card has been designed, which is capable of supplying direct voltages at a resolution of 12 bits, on four independent channels.

Main specification

- 4 independent direct output voltages between -10V and +10V
- 12-bit resolution
- 3.5-μs conversion time
- ±0.5 LSB max. non-linearity
- short 8-bit slot insertion card
- no jumpers; card address fixed at 0300H
- on-board decoupled supply voltage

Circuit description

The structure of the DAC card is so simple that a separate block diagram is superfluous. The circuit diagram is given in Fig. 1. The left-hand part of the circuit shows the connector that forms the link between the PC motherboard and the DAC card. Since it should be possible to use the present card in an XT machine also, the databus width is limited to 8 bits. The PC's datalines are fed to the DAC proper, U4, via a bidirectional buffer/transceiver, U1. Although the LS245 (U1) can also transfer data from the DAC to the PC (by appropriate control of the direction control input, pin 1), this ability is not used here because it is not required, the DAC being a write-only device in this application. The direction control input of the bus buffer is connected to the PC's IORD\ (input/output read) line. The PC can access the I/O memory range by read or write operations only. When a write operation is performed, for instance, copying new data to the DAC, IORD\ is not active, i.e., logic high. This logic level is used to control the direction input of the databus buffer.

The address decoding circuitry is simple because it has to select only eight addresses in the PC's I/O memory range. A GAL (generic array logic), U2, turns the levels on address lines A2 through A9, plus control lines IORD\ and IOWR\, into three control signals for the DAC. The 'write' command issued by the PC is also fed directly to the DAC via the IOWR\ line. The same goes for address lines A0, A1 and A2, which allow eight different registers in the DAC to be accessed.

The ready-programmed GAL (supplied through the Readers Services) fixes the DAC card base address at 300H. If you have a GAL programmer, other addresses may be defined by appropriate programming of the GAL, making use of the programming information given in Table 1. The listing shown is compatible with National Semiconductor's 'OPAL' GAL programming utilities contained on diskette 1791 (available through the Readers Services).

Power supply problems

The remainder of the circuit consists essentially of relatively extensive stabilizers on the DAC supply voltages. A

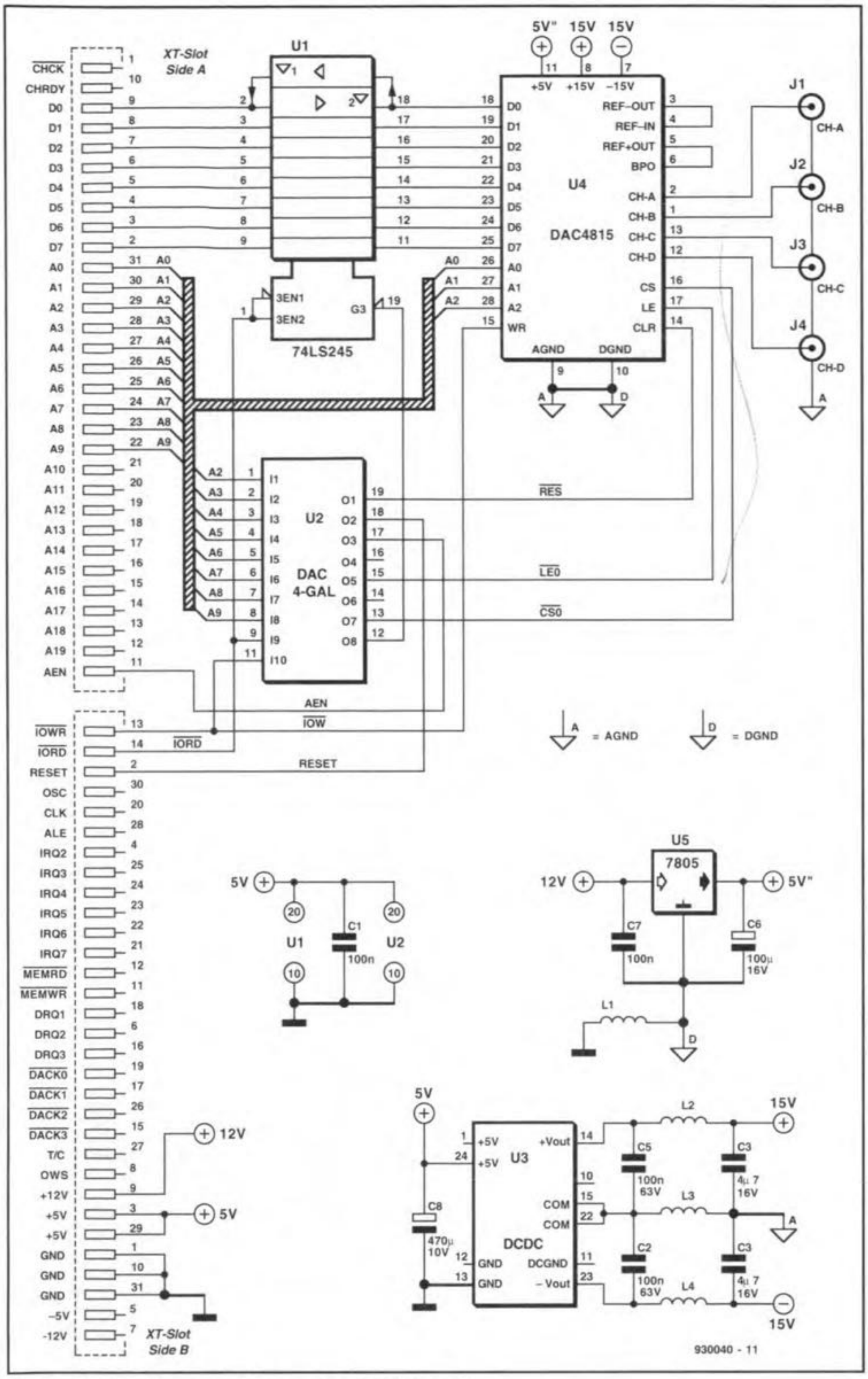


Fig. 1. Circuit diagram of the quadruple DAC card.

three-pin voltage regulator, U5, converts the PC's 12-V supply line into 5 V (circuit denotation: 5V"), which already affords better noise immunity than the 5-V output of the PC's internal switch-mode power supply.
The PC's 5-V line is used to power a DC-DC converter module which provides the symmetrical 15-V supply for the DAC4815. The pre-regulated 5-V (5V") and ±15-V supply lines ensure that the DAC can reach its full resolution of 12 bits, which are impossible to achieve lacking clean supply voltages. According to the datasheets, the DAC4815 requires a minimum supply voltage of ±11.4 V for an output voltage swing of ±10 V. With 15 V applied to the DAC, we are, therefore, on the safe side.

Also part of the measures against supply noise are the *L-C* filter behind the DC-DC converter, and the strict separation between the analogue and digital ground lines. These ground lines are connected at one point only on the PCB. The connection is made via a wire taken through a ferrite bead, L1.

DAC: internal

The internal architecture of Burr-Brown's DAC4815 is given in Fig. 2. The IC contains four identical converters, which are controlled by a central logic circuit. The IC is completed with a reference voltage source.

The main function of the control logic block is the distribution of byte-wise organized data at inputs pins 18 through 25 over the 4-bit and 8-bit input registers associated with each of the four channels. The data stored in the registers are captured in a 12-bit wide latch, and fed to the DAC proper as required. Each DAC has an output opamp to provide the necessary drive capacity, and afford protection against (short) output short-circuits. The outputs are not protected against continuous short-circuit conditions. Their output current capability is 10 mA continuously, and 20 mA briefly.

```
title Address Decoder for DAC-4
pattern GATES
revision A
author Elektor
Date 05.01.93

chip GATES GAL16V8

;pin 1  2  3  4  5  6  7  8  9    10
     A2 A3 A4 A5 A6 A7 A8 A9 /IORD GND
;pin 11    12 13  14  15  16  17  18    19     20
     /IOWR EN CS0 NC1 LE0 NC2 AEN RESIN RESOUT VCC

@UES MELEKTOR

     @define BASE_ADDRESS "/A3*/A4*/A5*/A6*/A7*A8*A9*/AEN" ; 0300H
;    @define BASE_ADDRESS " A3*/A4*/A5*/A6*/A7*A8*A9*/AEN" ; 0308H
;    @define BASE_ADDRESS "/A3* A4*/A5*/A6*/A7*A8*A9*/AEN" ; 0310H
;    @define BASE_ADDRESS " A3* A4*/A5*/A6*/A7*A8*A9*/AEN" ; 0318H


equations

/LE0   = /A2* BASE_ADDRESS * IORD
/CS0   = BASE_ADDRESS*  IORD + BASE_ADDRESS *  IOWR
/EN    = BASE_ADDRESS *  IORD + BASE_ADDRESS *  IOWR
RESOUT = /RESIN

; end of GATES
```

Table 1. GAL programming data written for National Semiconductor's 'OPAL' GAL programming package.

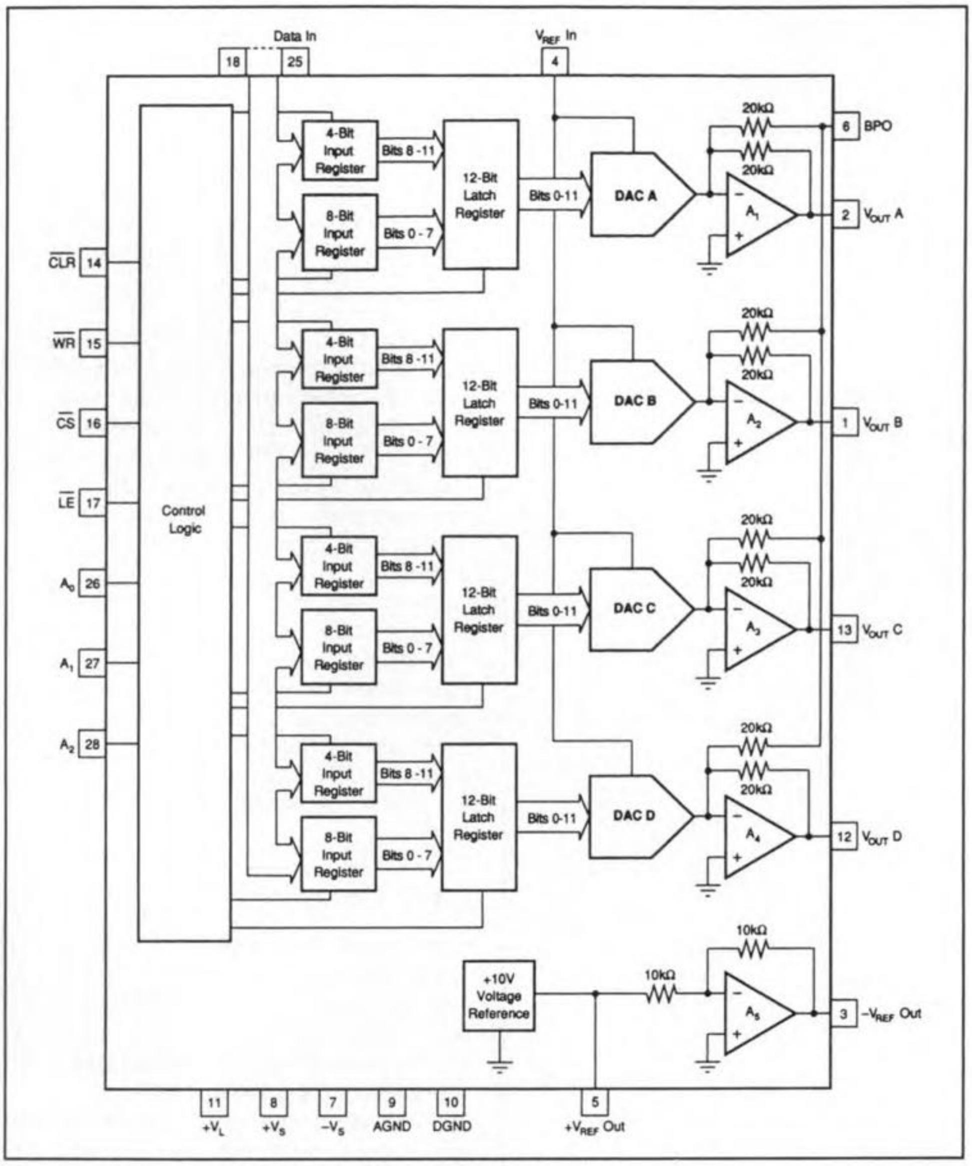


Fig. 2. Internal structure of Burr-Brown's DAC4815.

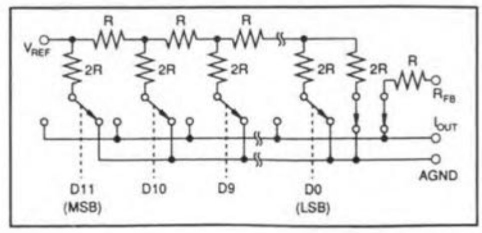


Fig. 3. R-2R ladder network used in the DAC4815.

All outputs can be switched to 0-V at any time via an asynchronous CLEAR input.

The data and command transfer between the PC and the DAC is shown in Table 2. All DAC connections are level sensitive rather than edge sensitive.

The output voltage range can be adjusted via the REFIN pin of the DAC. Since the REFIN voltage is applied to all four DACs, they all have the same output voltage range.

The block marked '+10V voltage reference' supplies a stable reference voltage which is bonded out to pin 5 of the IC. The voltage source is followed by an inverting operational amplifier which supplies –10 V. This auxiliary voltage allows the DAC output voltage to be made negative also.

Finally, the structure of the *R-2R* network contained in the DAC4815 is shown in Fig. 3.

Construction

The artwork for the double-sided printed circuit board designed for the DAC card is shown in Fig. 4. The board

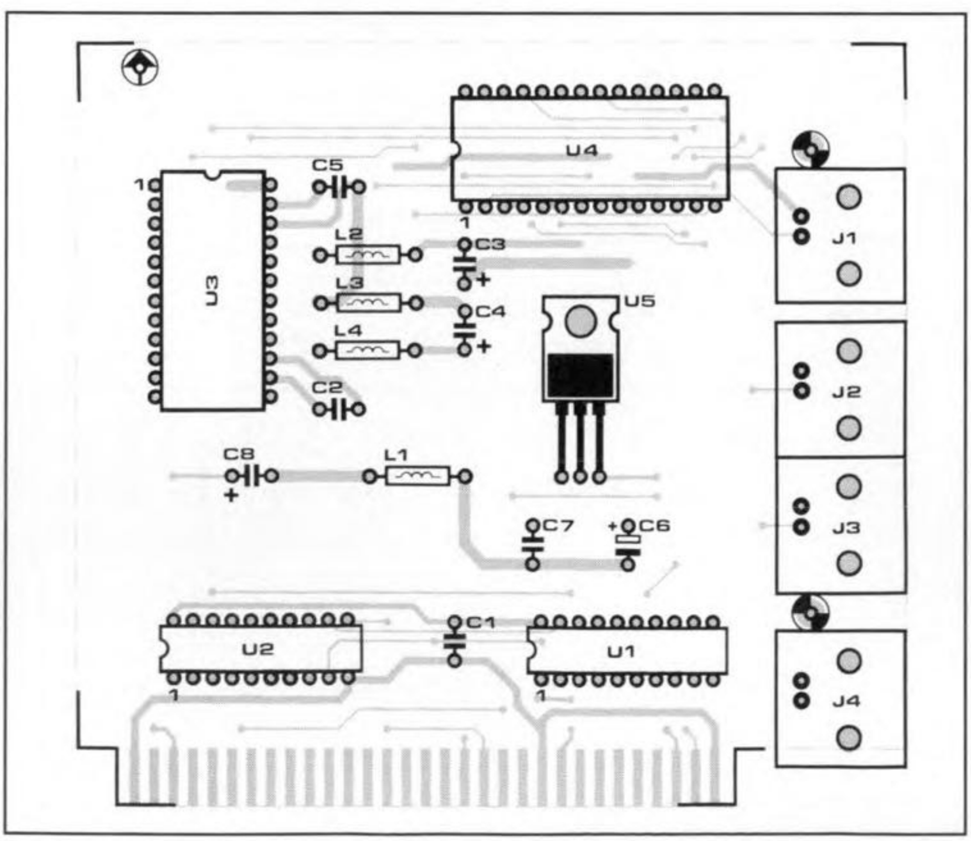


Fig. 4a. Component mounting plan of the double-sided PCB designed for the DAC card.

COMPONENTS LIST

Capacitors:

4	100nF	C1;C2;C5;C7
2	4μF7 16V tantalum	C3;C4
1	100μF 16V radial	C6
1	470μF 10V radial	C8

Semiconductors:

1	74LS245	U1
1	GAL 16V8, order code 6251 (see page 110)	U2
1	PWR1105 (Burr-Brown)	U3
1	DAC4815 (Burr-Brown)	U4
1	7805	U5

Miscellaneous:

4	Isolated BNC socket, PCB mount, angled pins	J1-J4
1	Ferrite bead	L1
3	100μH choke	L2;L3;L4
1	Printed circuit board 930040 (see page 110)	

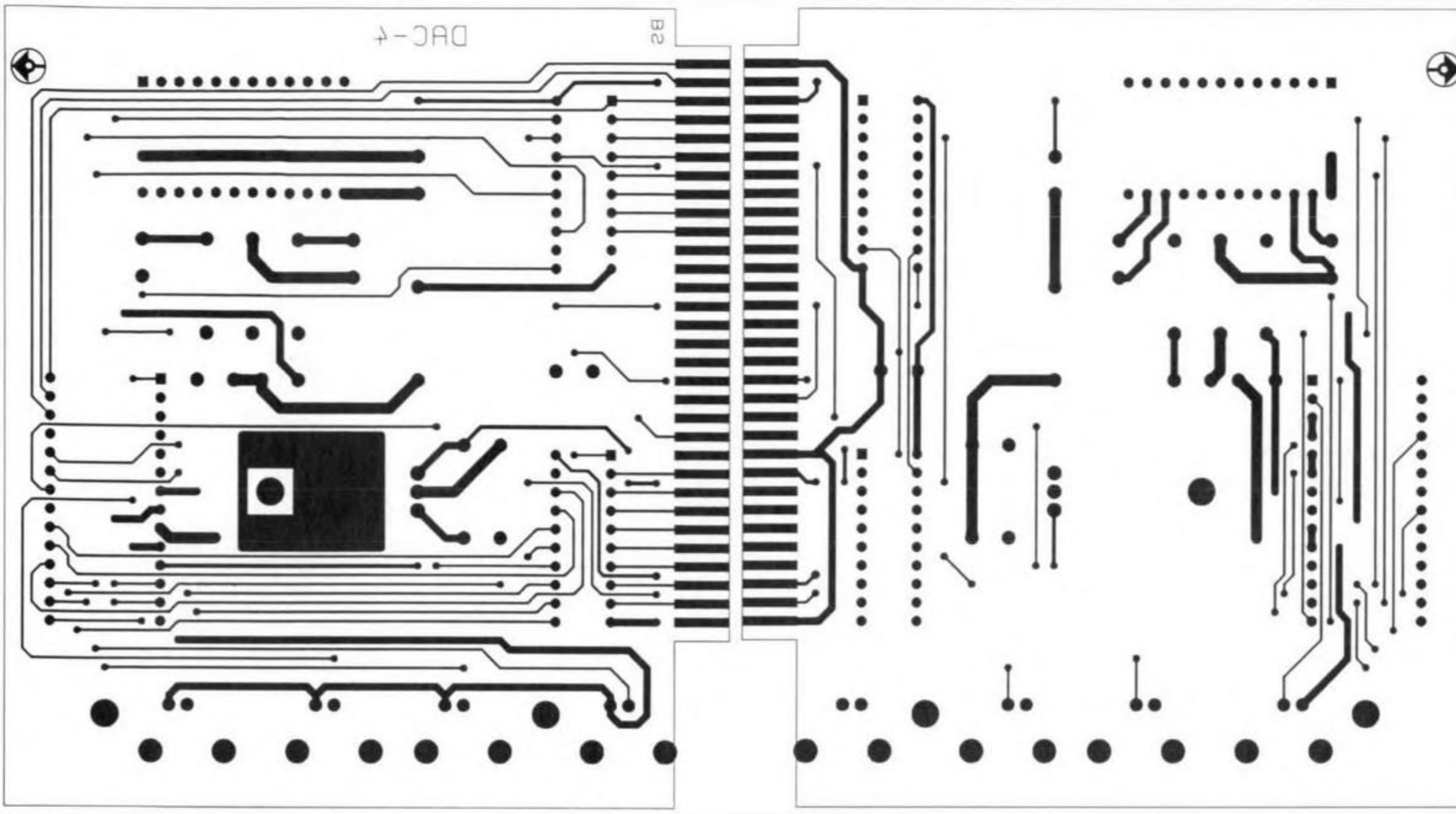


Fig. 4b. Component side and solder side track layouts (mirror images).

```
program DAC4;

uses crt,graph,dos,printer;

const       CH1  =  $0300; { channels 1-4 }
            CH2  =  $0302;
            CH3  =  $0304;
            CH4  =  $0306;
            EN   =  $0300; { enable all channels }

var A  :integer;

procedure init; { initialize and load DAC4815 }

begin
   port [CH1]    :=0  ;     { channel 1 LSB (8-bit) 0-255 }
   port [CH1+1] :=0  ;      { channel 1 MSB (4-bit) 0-15 }
   port [CH2]    :=255;     { channel 2 LSB (8-bit) 0-255 }
   port [CH2+1] :=15 ;      { channel 2 MSB (4-bit) 0-15 }
   port [CH3]    :=0  ;     { channel 3 LSB (8-bit) 0-255 }
   port [CH3+1] :=8  ;      { channel 3 MSB (4-bit) 0-15 }
   port [CH4]    :=0  ;     { channel 4 LSB (8-bit) 0-255 }
   port [CH4+1] :=12  ;     { channel 4 MSB (4-bit) 0-15 }
   A := port [CH1];         { convey all channels using RD command }
end;

begin
   writeln      (' OUTPUT MODE ACTUATED          ');
   writeln      (' Channel 1 = -10.000 volt      ');
   writeln      (' Channel 2 = +10.000 volt      ');
   writeln      (' Channel 3 = 0.000 volt        ');
   writeln      (' Channel 4 = +5.000 volt       ');

repeat
       init;
       until Keypressed;
 clrScr;
 TextMode(80);
end.
```

Fig. 5. Turbo Pascal version of the test program.

is not densely populated, and construction is straightforward.

The card is secured in the PC by an aluminium bracket drilled to pass the four BNC output sockets. Voltage regulator U5 does not require a heat-sink — the copper area underneath it is sufficient to assist in its cooling.

Software

As a matter of course, the card does nothing at all if it is not programmed. Fortunately, programming is relatively simple, and can be done in GWBASIC, Turbo Pascal, or other higher programming languages. The GWBASIC example program listed in Fig. 6 shows you how to have the card generate +10 V, −10 V, 0 V and +5 V. The Turbo Pascal program listed in Fig. 5 has the same effect. Both programs indicate that two bytes are to be copied to each of the DACs. Of the second byte, only the 4 most significant bits (high nibble) is of interest.

With the DAC wired as in the present circuit, the following relation exists between the data written to the DAC, and the analogue output voltage:

```
100 REM Test program for 4-fold DAC using BurrBrown DAC4815. DAC4.BAS
110 S=&H300 : REM set card address to 0300 HEX.
120 REM
130 REM --------- set channel 1 to +10.000V ---------
140 OUT S+0,255 : REM D0..D7 LSB = 0..255 ( 8-Bit )
150 OUT S+1,15 : REM D8..D11 MSB = 0..15 ( 4-Bit )
160 REM --------- set channel 2 to -10.000V ---------
170 OUT S+2,0
180 OUT S+3,0
190 REM --------- set channel 3 to 0.000V ---------
200 OUT S+4,0
210 OUT S+5,8
220 REM --------- set channel 4 to +5.000V ---------
230 OUT S+6,255
240 OUT S+7,11
250 A=INP(S) : Read channel 1-4 in one go!
260 REM done!
```

930040 - 15

Fig. 6. Listing of a simple GWBASIC program used to test the card.

CLR	LE	CS	WR	A2	A1	A0	Function
1	1	0	0	0	0	0	Load D0 - D7 into DAC-A LS input register
1	1	0	0	0	0	1	Load D3 (MSB) - D0 into DAC-A MS input register
1	1	0	0	0	1	0	Load D0-D7 into DAC-B LS input register
1	1	0	0	0	1	1	Load D3 (MSB) - D0 into DAC-B-MS input register
1	1	0	0	1	0	0	Load D0 - D7 into DAC-C LS input register
1	1	0	0	1	0	1	Load D3 (MSB) - D0 into DAC-C MS input register
1	1	0	0	1	1	0	Load D0 - D7 into DAC-D LS input register
1	1	0	0	1	1	1	Load D3 (MSB) - D0 into DAC-D MS register
1	0	0	1	x	x	x	Load all DAC registers simultaneously
1	0	0	0	x	x	x	All DAC registers transparent
1	x	1	x	x	x	x	No data transfer
1	1	x	1	x	x	x	No data transfer
0	x	x	x	x	x	x	Clear input registers, set DAC registers to 800H

x = don't care

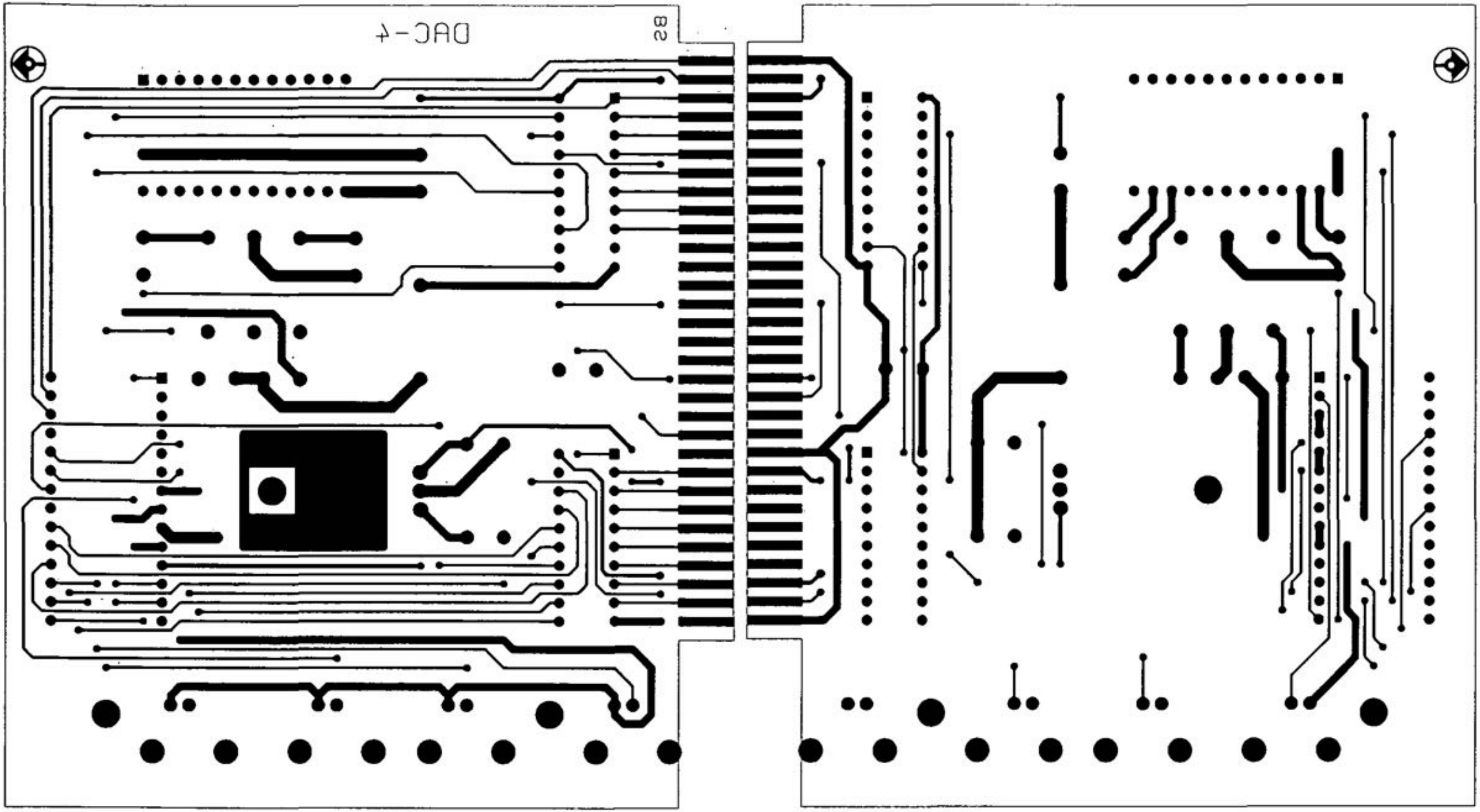
Table 2. PC-to-DAC interface logic.

FFFH	+ full scale
800H	zero
7FFH	zero – (1 LSB)
000H	– full scale

Since the DAC4815 comes laser-trimmed by the manufacturer, no external adjustments are required. ■

The printed circuit board for the four-fold DAC is available ready-made from

Kolter Electronic
Steinstrasse 22
D-5042 Erftstadt
GERMANY
Telephone (+49) 2235 76707
Fax: (+49) 2235 72048



VHF-LOW CONVERTER

Communication receivers for the VHF-low band (30-50 MHz) do not appear to exist commercially, and that is why we propose an up converter that allows you to use a VHF FM radio, or the VHF/UHF receiver described a few months ago, to listen to signals in this interesting band.

Design by J. Barendrecht and B. Romijn

THE VHF-low band extends roughly from 30 MHz to 50 MHz, and is reserved for mobile communication, cordless telephones and microphones, and radio control of toys and models. The upper part of the band, starting at 48 MHz, is no longer used for terrestrial television in the UK, which has been moved to the VHF-2 and UHF bands. On the European continent and in other parts of the world, however, channel E1 (48.25 MHz) is still widely used for television broadcasts (see the *DX Television* column featured every other month in *Elektor Electronics*).

Since most communication in the VHF-low band is in FM (frequency modulation), the proposed converter may be used in conjunction with a normal FM radio. If you want to listen to AM (amplitude modulation) signals as well, you need an AM demodulator, such as the one contained in the VHF/UHF receiver.

Principle of operation

The block diagram shown in Fig. 1 illustrates the basic operation of the converter. The antenna signals are amplified and subsequently mixed with a 58-MHz signal supplied by a crystal oscillator. The high mixing product (88-108 MHz) is extracted and fed to the FM receiver. The low mixing product (2-28 MHz) is suppressed. A signal at, say, 47 MHz appears on the radio at 47+58=105 MHz. The frequency up conversion principle is illustrated in Fig. 2.

If the VHF/UHF receiver is used with its digital frequency dial (Ref. 2), the counter may continue to be used in 'up' count mode, which leaves the relevant preset to be programmed only (more details further on).

Bandpass filter

The circuit diagram of the converter is given in Fig. 3. A low-noise transistor Type BFG65 (Philips) is used to amplify the antenna signal. Ahead of the RF amplifier sits a parallel tuned circuit, L_1-C_{13}-C_{14}. Actually, only L_1 and C_{13} form the tuned circuit, while C_{14} is a coupling device only that prevents the transistor's base voltage (approx. 0.6 V) from being short-circuited to ground by the inductor. Capacitors C_3, C_4 and C_6 elsewhere in the circuit have similar functions.

The frequency range selected by L_1-C_{13} covers about 30 to 50 MHz. After amplification by T_1, the signals are taken through a second filter, which is more complex. It consists of two parallel tuned circuits, L_2-C_5 and L_4-C_8, which are top-coupled by a series tuned circuit, L_3-C_7. The two parallel tuned circuits are designed to resonate at about 50 MHz, and the series tuned circuit at about 30 MHz. The combination of the three tuned circuits forms a bandpass filter for the frequency range 30-50 MHz. Figure 4 illustrates how the desired (ideal) bandpass characteristic is approximated with the aid of the four tuned circuits ahead of the mixer.

Sliding bias RF amp

Returning to the circuit diagram in Fig. 3, it is seen that transistor T_1 has no fixed base current. Rather, the base current is drawn from the collector circuit via resistor R_2. Consequently, the base current is determined by the collector voltage, which, in turn, depends on the emitter-collector current. If the latter fluctuates as a result of temperature variations, the base current varies accordingly, and causes the collector current to return to its normal value. This principle is called 'sliding bias'.

Resistors R_1 and R_3 form the collector resistor of T_1 as far as the d.c. bias setting is concerned. For RF currents, R_1 alone forms the collector resistance, since C_{15} decouples all RF voltages to ground. Network C_{15}-R_3-C_{10} forms a noise suppressor, and prevents RF signals from travelling to and from T_1 via the positive supply line.

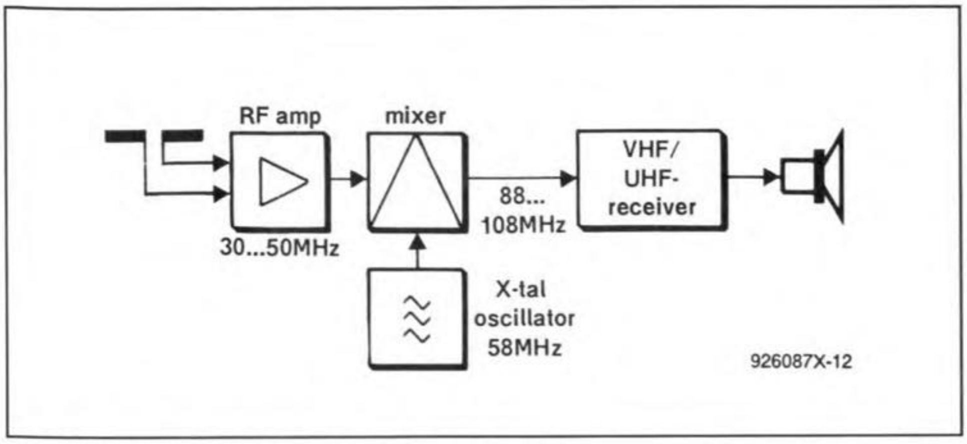


Fig. 1. Block diagram of the VHF-low converter.

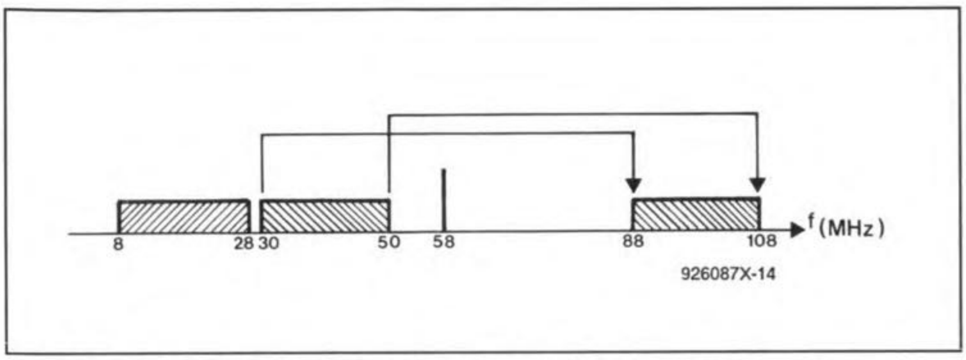


Fig. 2. Frequency up conversion by heterodyning (mixing) with a 58-MHz signal.

Mixer/oscillator

The mixer and local oscillator required for the frequency up conversion are contained in a single IC Type NE602 (or NE612). As illustrated by Fig. 5, the NE602 also contains a voltage regulator that affords a stable oscillator supply voltage.

The frequency determining external components required to complement the on-board oscillator are connected to pins 6 and 7 of the IC. Pins 1 and 2 form the (balanced) input, and pins 4 and 5, the (balanced) output. Returning to Fig. 3, the RF input signal (30-50 MHz) is fed to pin 1 of IC1. The second input of IC1, pin 2, is decoupled to ground via C9. This means that the input of IC1 is used 'single-ended'. Resistor R4 ensures equal voltages at pins 1 and 2, and at the same time provides some damping of L4-C8. Damping is required to prevent a too narrow filter characteristic of the two 50-MHz parallel tuned circuits (see Fig. 4b).

The quartz crystal, X1, is connected to the input of an amplifier contained in the NE602. The signal developed across the crystal is amplified in IC1, and output via pin 7. The signal is fed back to crystal via C1 to sustain the oscillation. Capacitor C2 and inductor L6 force the crystal to resonate at its third overtone (58 MHz), instead of its fundamental frequency, 19.33 MHz.

The 58-MHz local oscillator signal is fed to the mixer in the NE602 via an internal buffer. The sum and difference signals (88-108 MHz and 2-28 MHz respectively) that result from the mixing operation are available at pins 4 and 5 of IC1. The sum signal is selected by a parallel tuned circuit, L5-C11-C12, and the converter output signal is taken from the junction of the capacitive voltage divider, C11-C12.

Construction and adjustment

The converter is constructed on a compact, double-sided printed circuit board, of which the artwork is given in Fig. 6. This PCB is available ready-made through the Readers Services. To assist in the construction, Fig. 7 shows the locations of the component wires that have to be soldered at both sides of the PCB. Start the construction with these parts. Next, mount the remaining parts. Do not use an IC socket for IC1.

There is only one inductor that has to be home made: L1. It consists of four

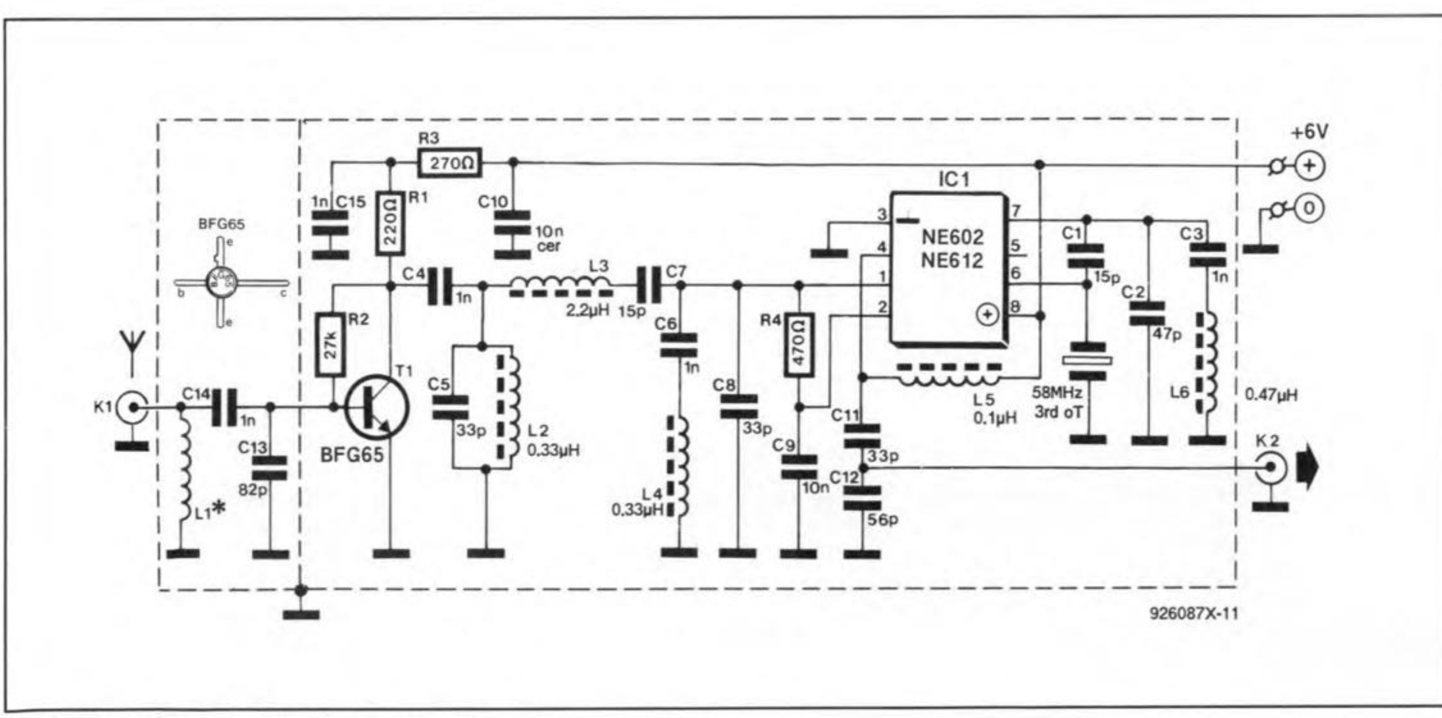


Fig. 3. Circuit diagram of the VHF-low converter.

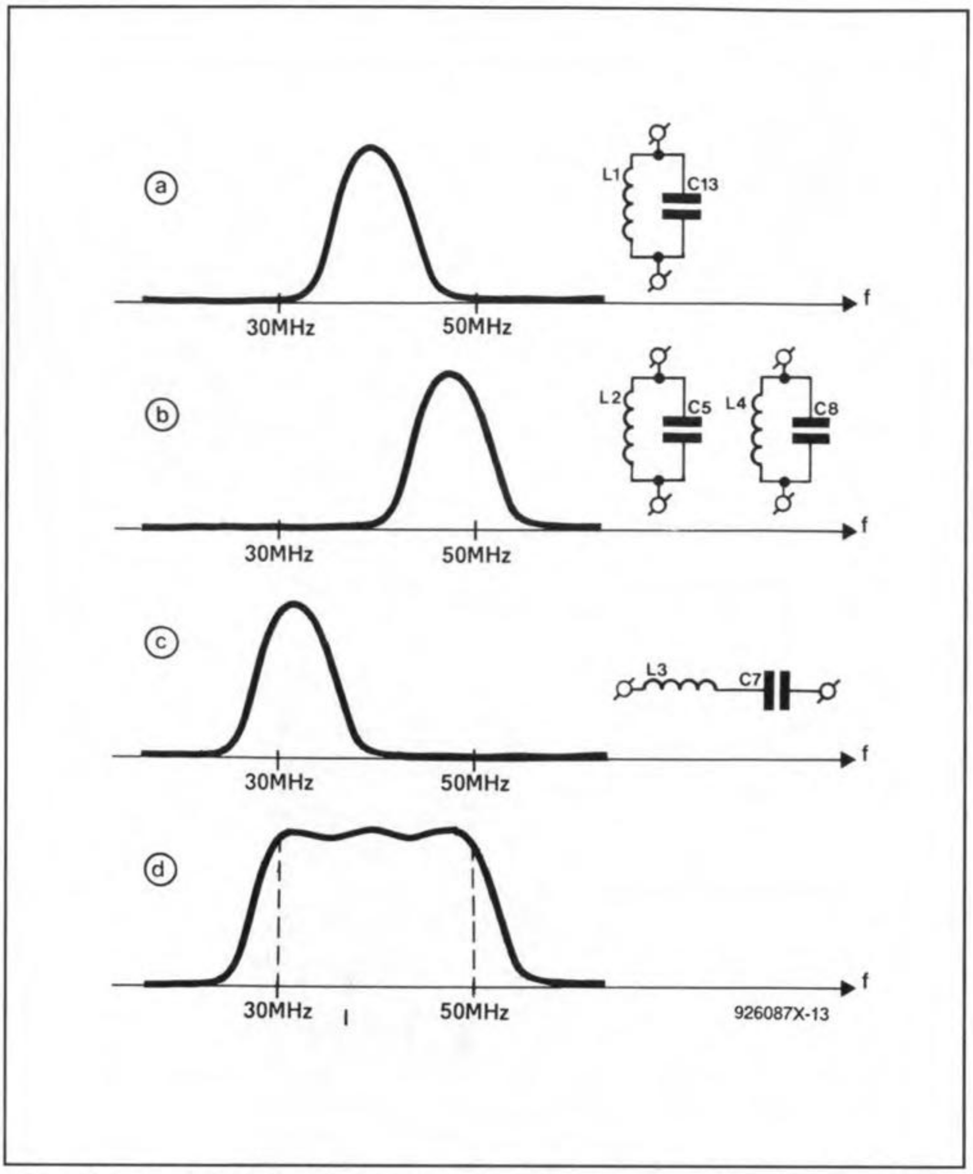


Fig. 4. The pass-band characteristic required for the converter is obtained with the aid of series and parallel tuned circuits.

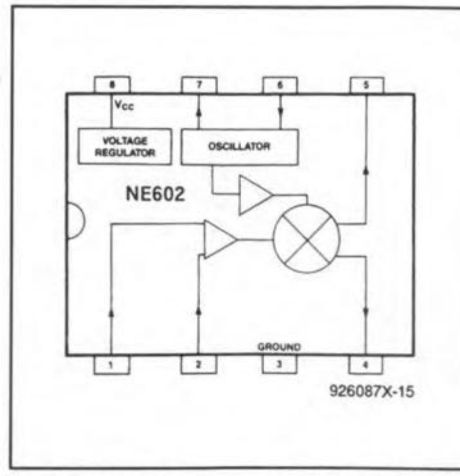


Fig. 5. Internal structure of the NE602/NE612 oscillator/mixer (Signetics).

turns of 1-mm dia. enamelled copper wire, and has an internal diameter of 5 mm. Stretch the inductor to a total length of about 15 mm, and then space the turns evenly. Next, compress the turns again until they have a spacing of about 1 mm. Cut the wire ends, and tin them. Next, mount the inductor on the board, a little above the board surface (see the photograph of an early prototype in Fig. 8). The BFG65 is seated in a hole, so that its connecting terminals (four) rest flush on the copper tracks. Note that the transistor is turned such that its type indication is legible from the solder side of the board.

When all the parts are on the board, fit a metal screen across T1. This serves to prevent oscillation caused by stray coupling between the input and output inductors in the circuit. Oscillation may also occur if the converter is not fitted in a metal case, or if the input and output coax cables are not properly connected or terminated. If oscillation persists in the face of these measures, try fitting a 10-100 kΩ damping resistor across L1.

Prepare the metal case by fitting the input and output socket, and the supply (d.c. adaptor) socket. Next, mount the completed PCB into the case, and connect it to the sockets.

Connect the converter output to the input of the VHF/UHF receiver, or an FM radio, via a 75-Ω coax cable. Connect the converter input to a wire dipole antenna or similar via a 75-Ω coax cable. A dipole antenna can easily and quickly be made of two straight wires with a length of 1.5-2.5 m. The core of the coax is connected to one wire, and the screen (braid) to the other.

When the converter is switched on, the receiver's noise output should increase. Possibly you hear signals straight away. If not, tune the VHF/UHF receiver below 88 MHz and switch to FM. This will probably bring in a couple of 27-MHz CB (citizen's band) stations (up-converted to about 85 MHz). Note that although 27 MHz is strictly speaking outside the pass-band of the receiver, these signals will still be converted, albeit at less conversion gain. The same goes for the 50 MHz (6-m amateur radio) band, hence it is also listed in Table 1.

Radio amateurs may want to have a go at modifying the converter's output tuned circuit to 3.55-8 MHz. This allows the 6-m band (50-54.45 MHz in most countries) to be received on an AM/SSB communications receiver tuned between 3.55 and 8 MHz. It should be noted, however, that the converted frequency spectrum is 'upside down', which means, among others, that USB transmissions should be received using LSB, and vice versa. As a matter of course, different quartz crystal frequencies may be used for further experiments.

In case you notice that stations around 50 MHz are received better than those around 30 MHz, you may have a problem with the converter's pass-band characteristic. This may be solved quite elegantly by compressing

26.965 - 27.405 MHz:	Citizens' band (CB) channels 1-40
28.0 - 29.7 MHz:	Radio amateurs (10m band)
35.01 - 35.20 MHz;	
40.665 - 40.695 MHz:	R/C models
37.07 - 38.00 MHz:	Cordless telephones
46.97 - 49.97 MHz:	Inexpensive cordless telephones
50 - 54.45 MHz:	Radio amateurs (6m band)

Table 1. Some frequency allocations in and around the VHF-low band.

COMPONENTS LIST

Resistors:

1	220Ω	R1
1	27kΩ	R2
1	270Ω	R3
1	470Ω	R4

Capacitors:

2	15pF	C1;C7
1	47pF	C2
5	1nF	C3;C4;C6; C14;C15
3	33pF	C5;C8;C11
2	10nF	C9;C10
1	56pF	C12
1	82pF	C13

Inductors:

1	see text	L1
2	0µH33	L2;L4
1	2µH2	L5
1	0µH47	L6

Semiconductors:

1	BFG65 (Philips Semiconductors)	T1
1	NE602 or NE612 (Signetics)	IC1

Miscellaneous:

1	58MHz quartz crystal (3rd overtone)	X1
2	BNC socket 75Ω	K1;K2
1	Printed circuit board 926087 (see page 70)	
1	Diecast case, e.g., Hammond 1590A	

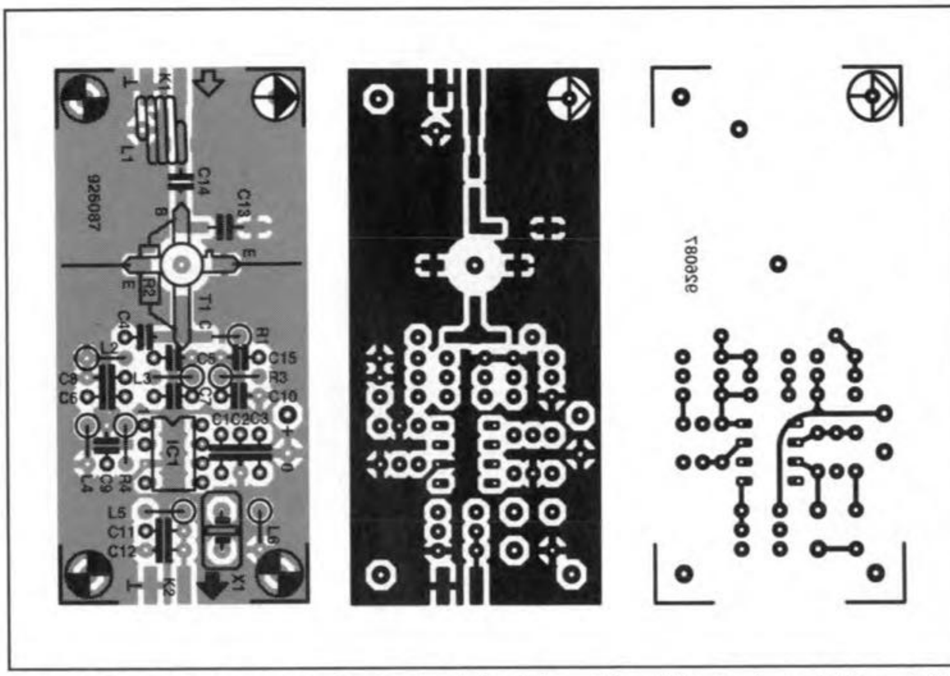


Fig. 6. Artwork for the double-sided printed circuit board (available ready-made through the Readers Services).

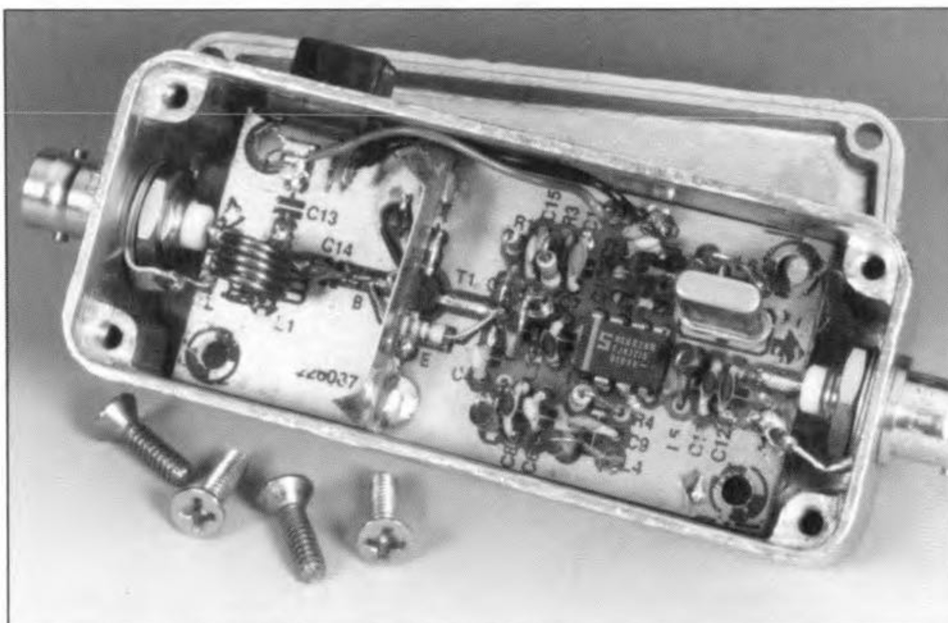


Fig. 8. An early prototype of the up converter.

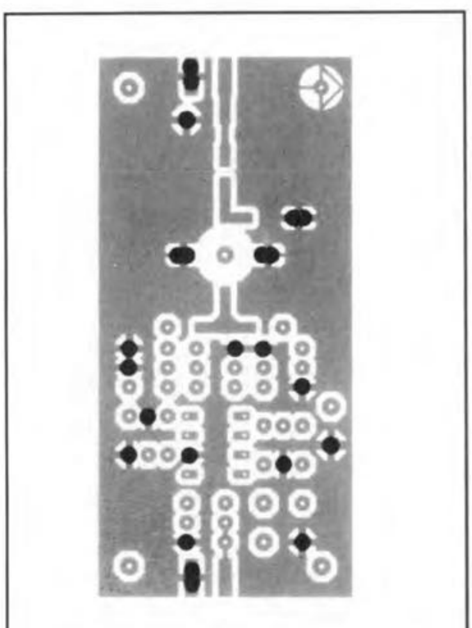


Fig. 7. The black spots show where to solder component wires at both sides of the printed circuit board.

the turns of L1 a little. This increases the self-inductance, so that L1-C13 resonates a little lower in the band. This, in turn, results in higher amplification of signals towards the low end of the band. By contrast, the gain around 50 MHz is boosted by stretching L1 a little.

If reception suffers from strong signals in the VHF FM broadcast band, reduce the length of the coax cable between the converter and the radio, so that it can not function as an antenna.

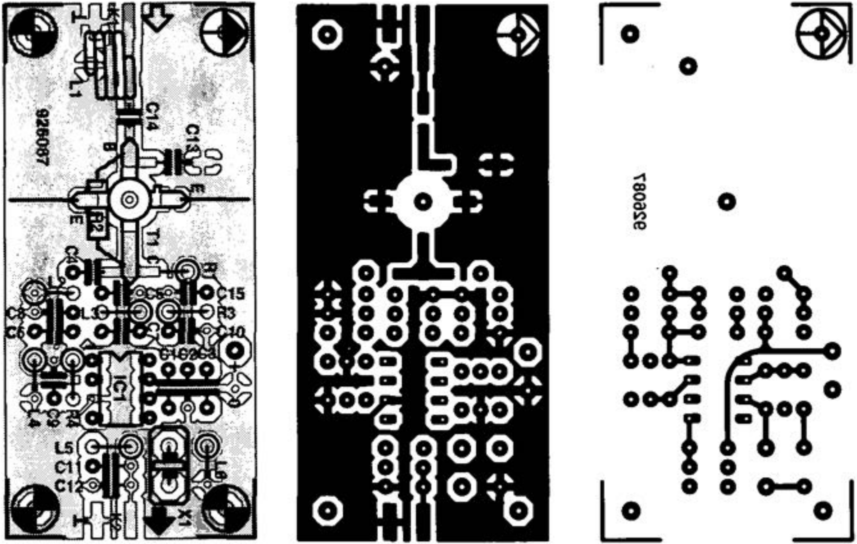
Digital frequency readout

Those of you who use the VHF/UHF receiver in combination with the digital frequency readout and the present converter will be pleased to learn that the counter (frequency readout) is readily programmed in accordance with the VHF-low band we have just added.

The new preset required is 10000–953=9047 (the original preset was 9627; see Ref. 2). This is achieved by fitting the jumpers as follows. First digit: jumpers D1-D4 ('1' and '8'; these are already fitted). Second digit: remove jumpers D6-D7 ('2' and '4'). Third digit: move jumper D11 ('2') to D10 ('4'). Fourth digit: no changes ('1', '2' and '4'). With a little dexterity, the presets for normal operation (9627) and 'VHF-low' (9047) may be swapped by relays or logic circuitry. ■

References:

1. VHF/UHF receiver, *Elektor Electronics* May 1993.
2. Digital frequency readout for VHF/UHF receiver, *Elektor Electronics* June 1993.



K1
L1
C14
C13
926087
B
E
R2
T1
C4
R1
C15
C8
L3
R3
C6
C5
C10
C7
IC1
C9
R4
L4
L5
C11
C12
L6
K2
X1

CORRECTIONS AND UPDATES

I^2C interface for PCs

February 1992

The PCD8584 iused in this project is no longer manufactured by Philips Semiconductors, and replaced by the PC**F**8584. This is a fully compatible IC and only improved as regards the 4-wire long-distance mode, which did not work correctly on the PCD8584.

Real-time clock for 80C32 computer

June 1993

Contrary to what is implied by the description of the parallel connection of the SmartWatch IC pins with the EPROM pins, pin 1 of the SmartWatch should be connected separately to +5V, for instance, to EPROM pin 28, via a short wire. This is necessary because pin 1 on the SmartWatch is 'reset', while on the EPROM it is address line A14, which may be made high by 'high' addressing or glitches, causing the clock to be reset.

VHF-low converter

June 1993

The parts list should be corrected to read:

1	2µH2	L3
1	0µH1	L5

The circuit diagram is correct.

The sub-1µH chokes used in this project are available from, among others, Cricklewood Electronics.

1.2 GHz multifunction frequency meter

December 1992

The recommended LCD module Type LTN211-F10 is no longer manufactured by Philips Components, and may be replaced by the compatible types LM016L from Hitachi, or the LM16A21 from Sharp.

FIGURING IT OUT

PART 7 – MORE DUALS

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas. We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

This month we examine the converse of several topics dealt with in previous parts. Although nothing fundamentally new is involved, since we are looking at familiar topics upside-down or back-to-front, a change of approach often turns an intractable problem into a solvable one.

Nodal analysis

In Part 4 we saw how to analyse a network by applying Kirchhoff's Voltage Law and calculating the currents circulating in the meshes of the network. This is **mesh current analysis**. Now we look at the dual of this technique. We apply Kirchhoff's Current Law, calculate the currents flowing into and out of the nodes of the network, and thus find the node voltages. This method, known as **node voltage analysis**, is more appropriate when a circuit is powered mainly by current sources rather than by voltage sources.

We begin with a very simple circuit (Fig. 55) that has only two nodes. Apply KCL to node A:

total current entering A = total current leaving A, or
$\frac{1}{3}+\frac{2}{3}=I$, so that
$I = 1$A.

Taking the potential at node B to be 0 V, the pd across the 2 Ω resistor is 1×2=2 V. Thus, the potential at node A is 2 V. This is the one of the circuits that we analysed in Part 4, but with the voltage sources replaced by current sources.

In Fig. 56 (another circuit modified from Part 4), there are three current sources and four nodes. Currents are determined either by the current sources or by the pds across resistors ($I=UR$). The unknown potentials at nodes A–D are denoted by U_A–U_D. Applying KCL to each node:

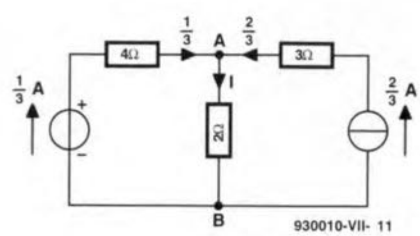


Fig. 55

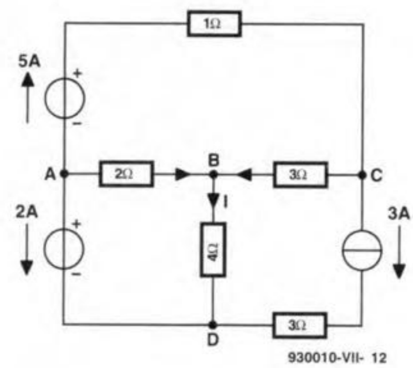


Fig. 56

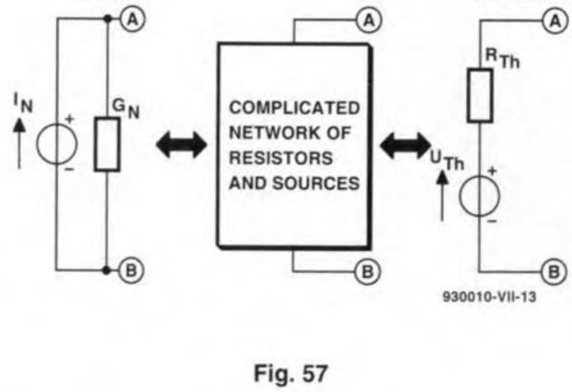


Fig. 57

Node A:

$(U_B-U_A)/2=5+2;$

$U_A-U_B=-14.$ [Eq. 37]

Node B:

$(U_C-U_D)/5+5=(U_B-U_A)/2;$

$5U_A-7U_B+2U_C=-50.$ [Eq. 38]

Node C:

$(U_C-U_B)/5-3;$

$U_B-U_C=-10.$ [Eq. 39]

Node D:

$(U_D-U_B)/4=2+3;$

$-U_B+U_D=20.$ [Eq. 40]

This provides a set of four simultaneous equations for calculating the four unknown voltages. In Part 4, we solved such equations by using determinants (and a computer program), but this technique does not work here. If you try, you obtain one or more determinants with zero value. The potentials are indeterminate under this technique. This makes sense, for we can take any one of the potentials to be 0 V (or any other voltage that we care to choose) and all other potentials are relative to this. For example, adding 100 V to **all** the potentials makes no difference to the currents flowing in the circuit. In particular, if one node is said to be at 0 V, one column of the determinant has all zero elements, and its value is zero.

Usually, we can evaluate the potentials by solving the equations one at a time, beginning with the simplest. Let us take U_D as the zero level. Eq. 40 gives:

$-U_B=20$, or, $U_B=-20$ V.

Putting this value into Eq. 37:

$-20-U_A=14$, so that

$U_A=-34$ V.

Similarly, with Eq. 39:

$-20-U_C=-10$, so that

$U_C=-10.$

As a check, we evaluate Eq. 38:

$$5\times-34-7\times-20+2\times-10=$$

$$-170+140-20=-50.$$

Summing up nodal analysis: on the circuit diagram, note currents provided by current sources; evaluate other currents by reference to known potential differences and resistances. At each node, equate the sums of the entering and leaving currents. Solve the set of equations obtained for the nodes. It is not always necessary to work at every node and solve every equation to obtain the information required.

Norton equivalent

The Norton equivalent is the dual of the Thévenin equivalent. In Part 3 of this series we showed how a 2-terminal network of voltage or current sources and resistors could be replaced by a single **voltage** source in **series** with a **resistance**. This is the Thévenin equivalent and is a very useful concept for simplifying the analysis of networks. When we drive the Norton equivalent of a network, we obtain a single **current** source in **parallel** with a **conductance**. Note the three sets of duals: voltage-current, series-parallel, resistance-conductance. Figure 57 contrasts the two equivalents.

If the Norton equivalent circuit is short-circuited, all of the current from the internal current generator flows through the external short circuit:

$$I_N=I_{sc}, \qquad \text{[Eq. 41]}$$

where 'sc' stands for short circuit. We can calculate I_N by finding out what current flows when terminals A and B are connected together.

When the Norton equivalent is an open circuit, the current from the generator flows through the Norton conductance and produces an open-circuited pd across it:

$$U_{oc}=I_N/G_N,$$

giving

$$G_N=I_N/U_{oc}.$$

Substituting from Eq. 37:

$$G_N=I_{sc}/U_{oc}. \qquad \text{[Eq. 42]}$$

To find the Norton equivalent

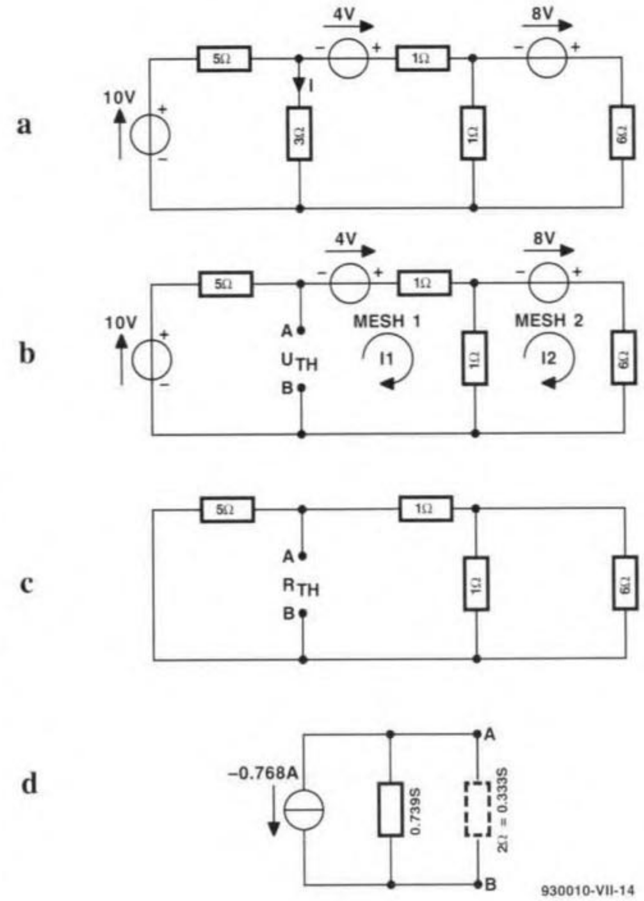


Fig. 58

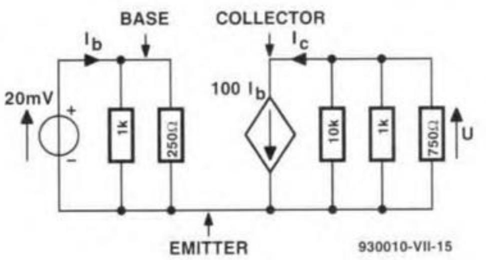


Fig. 59

conductance, divide the short-circuit current by the open-circuit pd.

Comparing Eq. 42 with the equations for the Thévenin equivalent quoted in Part 3, we see that:

$$G_N=1/R_{Th}. \qquad \text{[Eq. 43]}$$

Also,

$$I_N=I_{sc}=U_{Th}/R_{Th}. \qquad \text{[Eq. 44]}$$

The following example (see Fig. 58) shows how the Norton equivalent is used for solving network problems. Its use may be contrasted with the way we used the Thévenin equivalent in Part 3. We are required to find the current I flowing through the 3 Ω resistor. This problem could be tackled by mesh analysis and solving three simultaneous equations, but we will calculate the Norton equivalent instead.

Figure 58b shows the situation: we wish to obtain the equivalent of the whole network **except for** the branch that contains the 3 Ω resistor. The pd across AB is $U_{oc}=U_{Th}$. To find this, we need to know the current flowing through the 5 Ω resistor, and this is found by mesh analysis. Mesh 1 includes the 10 V and 4 V sources as well as the 5 Ω and two 1 Ω resistors, so:

$$7I_1-I_2=14. \qquad \text{[Eq. 45]}$$

In mesh 2:

$$-I_1+7I_2=8. \qquad \text{[Eq. 46]}$$

Solving equations 45 and 46 gives:

$$I_1=2.208 \text{ A and } I_2=1.458 \text{ A}.$$

The pd across the 5 Ω resistor is 5×2.208=11.04 V. Thus, the potential at A=10–11.04=–1.04 V. This tells us that U_{Th}=–1.04 V. Note that the 3 Ω resistor does not enter into the calculations since this branch is considered to be an open circuit and no current flows through that resistor.

Figure 58c, in which voltage sources are replaced by short circuits, is used for calculating R_{Th} by network reduction (see Part 1). Resistors of 6 Ω and 1 Ω in parallel result in 0.8571 Ω. In series with 1 Ω, this makes the righthand side of the figure equivalent to 1.8571 Ω. In parallel with 5 Ω, the network is equivalent to 1.354 Ω. In other words, R_{Th}=1.354 Ω.

We now have the quantities required for using equations 43 and 44 to calculate the Norton equivalent:

$$I_N=-1.04/1.354=-0.768 \text{ A}$$

$$G_N=1/1.354=0.739 \text{ S}.$$

Figure 58d shows the Norton equivalent. The 3 Ω resistor is drawn in dashed lines across AB as a load. Its admittance is 1/3=0.333 S. The current flowing through the 3 Ω resistor is proportional to its admittance:
$I=-0.768\times0.333/(0.739+0.333)=-0.239$ A.

An applicable example

The network of Fig. 59 is in two parts but, as they share only one common line, they can be treated separately. The problem is to calculate the value of current I and then pd U across the 750 Ω resistor.

The 1 kΩ and 250 Ω resistors in parallel are equivalent to 200 Ω, so the total current flowing from the source is $20\times10^{-3}/200=0.1$ mA. This is split between the resistors in inverse proportion to their resistance, so

$$I_b=0.1\times10^{-3}\times4/5=80\times10^{-6} \text{ A}.$$

Moving on to the other side of the diagram, the controlled current source produces a current of

$$100I=100\times80\times10^{-6}=8\times10^{-3} \text{ A}.$$

The three resistors in parallel are equivalent to 410.96 Ω. The pd across them is

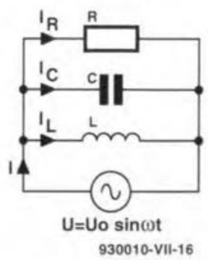


Fig. 60

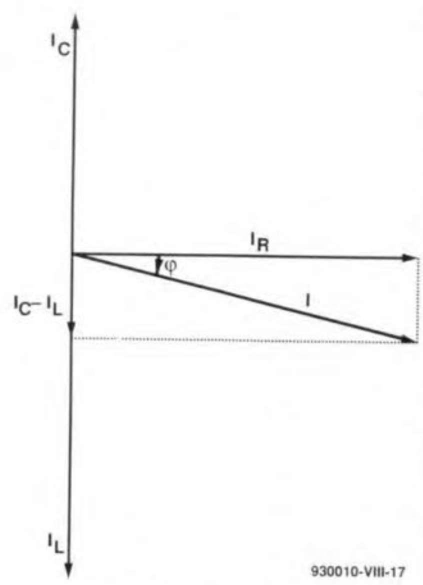


Fig. 61

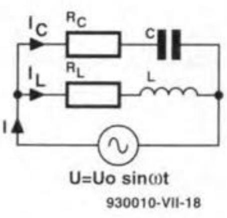


Fig. 62

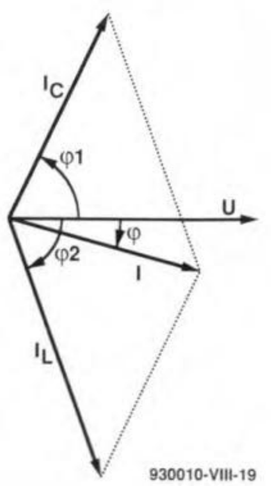


Fig. 63

$$U=8\times10^{-3}\times410.96=3.288 \text{ V}.$$

An examination of the stages of this calculation shows that, if we alter the output from the voltage source, U varies in proportion. The ratio between source voltage and U is:

$$A = 3.288/20\times10^{-3} = 164.$$

The circuit has a voltage gain of 164. This is not suprising, for the circuit of Fig. 59 is the equivalent of a common-emitter amplifier. The 20 mV source represents the input to the amplifier, perhaps from a microphone. The 1 kΩ resistor represents the biasing resistors in parallel. The 250 Ω resistor represents the input resistance, h_{ie}, at the base of the transistor. Since this resistance varies with base current, it would be better represented by a controlled voltage source. Current I_b is the base current, node B is the base terminal, and node E is the emitter terminal. On the other side, the controlled current source represents the amplifying action of the transistor with a small signal gain, h_{fe}, of 100. The 10 kΩ resistor represents the output admittance, h_{oe}, the 1 kΩ is the collector resistor, and the 750 Ω is the load.

This fairly simple example has been included to allay the suspicions of those who may believe that network analysis is concerned with resistances and sources connected in unlikely configurations. Figure 59 is a representation of an entirely practical circuit and, drawn in this form, it immediately becomes amenable to the techniques of network analysis described here and in previous issues.

LCR in parallel

In the sense that the parallel connection is the dual of the series connection, we now examine the dual of the *LCR* circuit described in Part 6, Fig. 45. Figure 60 shows the classic *LCR* circuit with a sinusoidal pd applied to it. At any instant, the same pd appears across all three branches. KCL applies to the instantaneous currents through the branches and I is the sum of these. To determine the currents, we use a technique similar to that employed in Part 6 for the *LCR* series circuit. Figure 61 shows the vector diagram, which is similar to Fig. 48 except that, showing currents instead of pds or impedances, I_C **leads** the applied pd and I_L **lags** it. The technique for analysing the currents is the same as for pds or impedances.

More often than not, we connect a capacitor and an inductor in parallel, but without a parallel resistor. Even so, resistance plays a part in the circuit, a small resistance being inherent in the construction of practical capacitors and inductors. Figure 62 shows the more practical situation. If we think of the branch that contains the capacitor as a **series** *RC* circuit with an applied pd of $U_0\sin\omega t$, we calculate the phase angle as we did last month for pds, except that this now has a lead instead of a lag (duals again!). Similarly, the inductor current lags the applied voltage.

For the capacitor:

$$\varphi_1=\tan^{-1}X_C/R_C;$$

and for the inductor

$$\varphi_2=\tan^{-1}X_L/R_L.$$

The vectors are shown in Fig. 63. The total current, $\boldsymbol{I}$, is the vector sum of $\boldsymbol{I_C}$ and $\boldsymbol{I_L}$. If there were a third branch in the network, with pure resistance, its vector would lie in the same direction as $\boldsymbol{U}$ and could be added to the other two currents to obtain the total current. If R_C or R_L is very small (ideal components), the phase angles are the inverse tangents of very large values and approach 90°. We are back to Fig. 61 again. In practical circuits, the working is not as simple. We can analyse the circuit graphically by making scale drawings like Fig. 63 and measuring I. This is not a precise technique. We need a non-graphic technique for adding vectors when they are not conveniently perpendicular to each other. This is to be the subject of next month's instalment.

Test yourself

1. By using nodal analysis, calculate the potential at each of the nodes in Fig. 64a, taking node C to be at 0 V. Calculate the current along each branch.
2. Calculate the Norton equivalent of the network in Fig. 64b, omitting the 4 Ω resistor on the right. Hence, calculate the current, I, through this resistor.
3. Figure 64c is the equivalent circuit of a common emitter amplifier in which bias is provided by a feedback resistor, R_F, connected between the collector and base of the n-p-n transistor. Given the values shown in the figure, calculate the voltage gain of the circuit for small signals.

Answers will be given in next month's instalment.

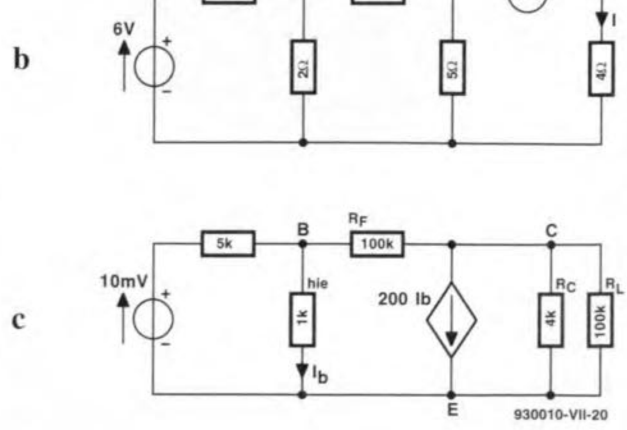


Fig. 64

Answer to problem set by Fig. 54. U_{out}/U_{in} is low at most frequencies, peaking sharply at 2 kHz (Fig. 65). This is a **band-pass** filter with a narrow pass band. Phase angle is a lag of 80° or more at frequencies below 1.5 kHz, zero at 2000 Hz, and leading by 80° or more at frequencies above 2.5 kHz.

Answers to Test yourself (Part 6)

1. 2199 rad s^{-1}
2. (a) 0.010 Ω; (b) 2073 Ω.
3. (a) 663 Ω; (b) 0.133 Ω.
4. See Fig. 66
5. 6.03 Hz.
6. 31.8 pF.

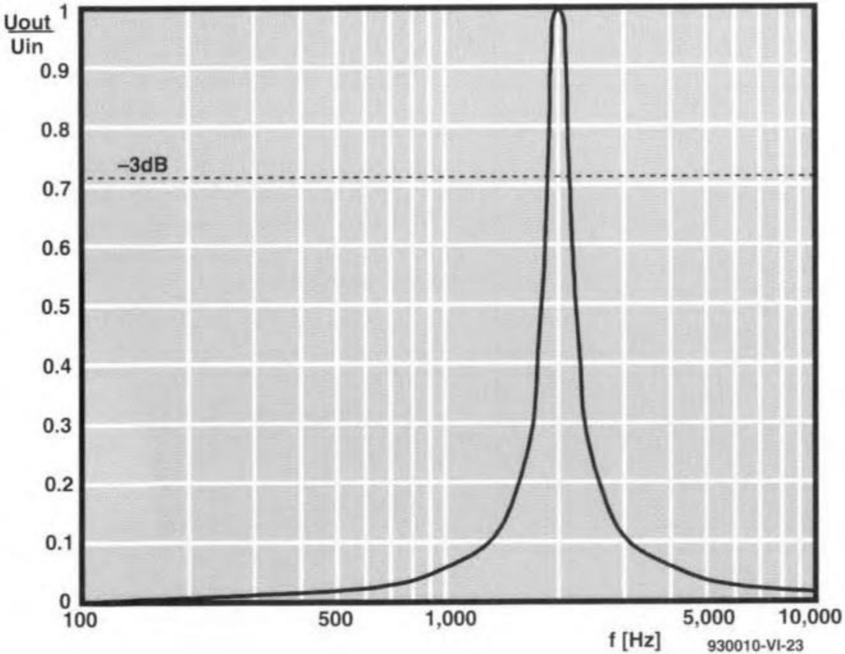


Fig. 65

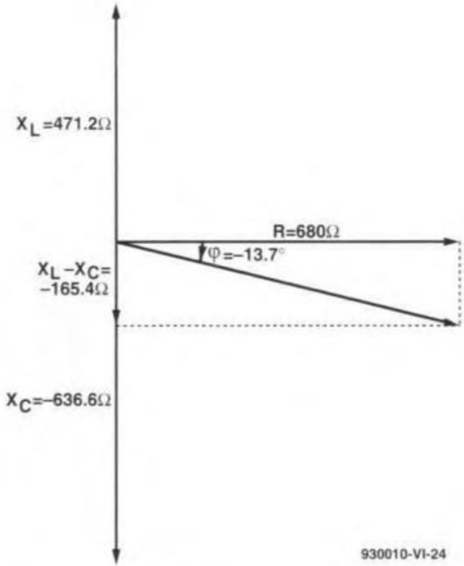
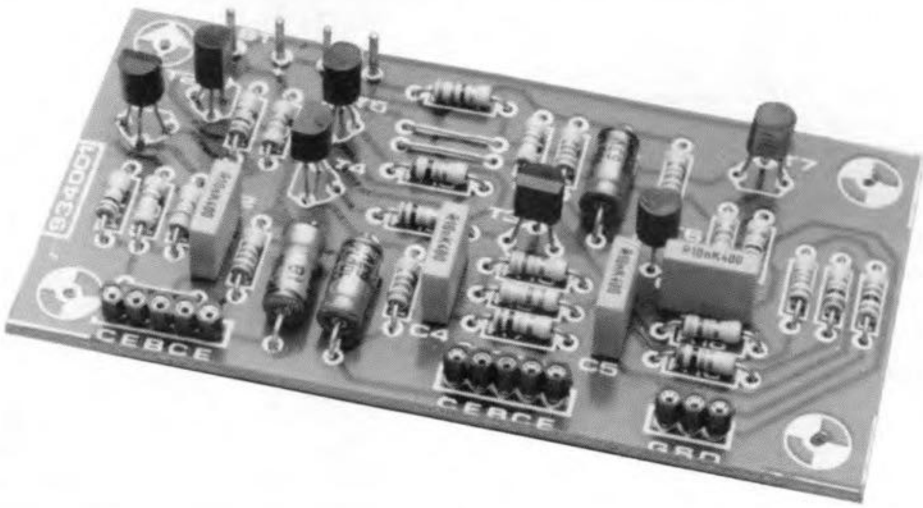


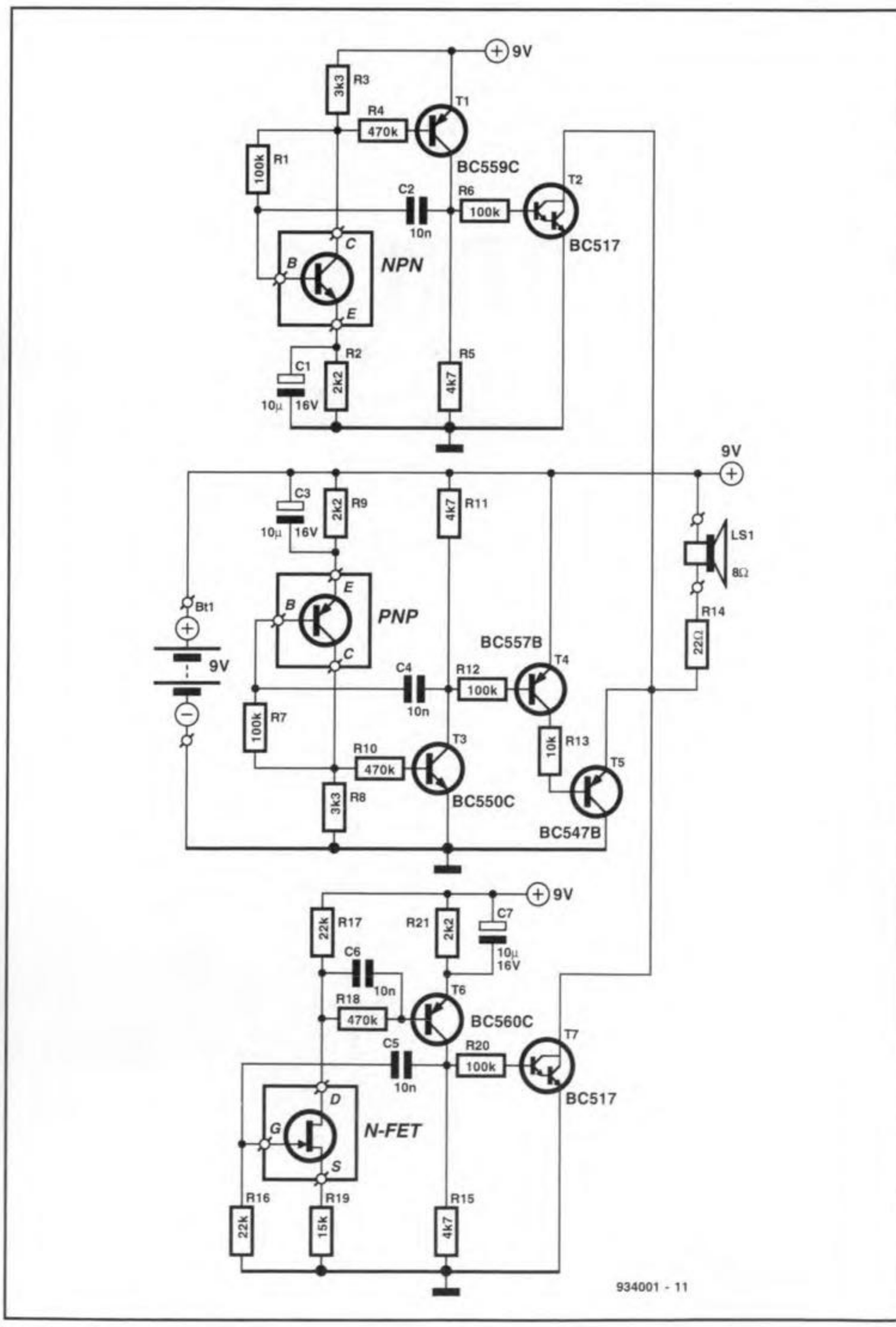
Fig. 66

AUDIBLE TRANSISTOR TESTER

The tester has no operating controls and draws current only when the loudspeaker sounds.

It comprises three oscillators that drive a common loudspeaker via buffering OR gates. Each oscillator consists of a permanently fitted transistor and the transistor on test (TOT). The oscillator frequency depends on the time constant of the appropriate *RC* network (R_5C_2, $R_{11}C_4$, or $R_{15}C_5$) and the current amplification of the TOT. Th n-p-n and p-n-p sections are virtually identical, but, of course, the supply lines are reversed. The buffer stages differ slightly to ensure that the loudspeaker remains connected correctly.





The design of the stage for testing n-channel FETs is based on that of the n-p-n stage. In operation, the gate must be negative with respect to the source. Transistors T_1, T_3 and T_6 are high-gain types to prevent TOTs with a low β being assessed as non-functional.

The design is such that unknown transistors can be tested without any danger of damage to them. The pinout of bipolar transistors may be EBC or BCE. Test leads with terminals EBCE or even CEBCE (both n-p-n and p-n-p) are, therefore, recommended. The transistor to be tested is connected with its pins to different terminals until the loudspeaker sounds. This does not only give the pinout, but also makes it clear whether the TOT is an n-p-n or a p-n-p type. The frequency of the emitted tone indicates whether the emitter and collector have been connected correctly. When they have not, the current gain is only about 1, so that the oscillator frequency is high. If the two connections are interchanged, the current gain is higher and the oscillator frequency lower.

Owing to the symmetry of n-channel FETs, it is not possible to distinguish between source and drain. When the loudspeaker sounds, it means only that the gate has been connected correctly.

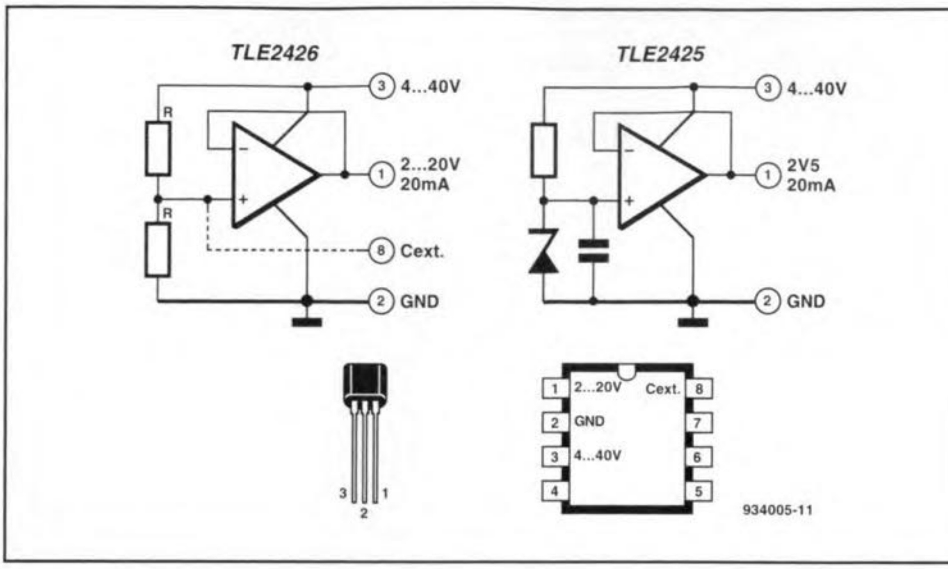
The test leads may be terminated into crocodile clips or quick-release, spring-loaded terminals.

[R. Radius - 934001]

VIRTUAL GROUND

Often only a single power supply is available for an opamp circuit. Since such circuits normally require a double supply, this means that in some way the available supply has to be converted. This is normally done by a potential divider and capacitor across the single supply, sometimes followed by an ancillary opamp.

Nowadays, a more sophisticated way is the use of the TLE2426 from Texas Instruments. This IC is not only small and easy to use, but, owing to the built-in opamp, also gives a better performance. This is particularly noticeable at low frequencies, where the impedance of the usual electrolytic capacitor across the potential divider may cause problems. In the TLE2426, a voltage follower takes the place of the capacitor, so that the impedance, even at low frequencie, remains low—typically 7.5 mΩ.



The quiescent current is only 170 μA, while the peak current that can be provided is 20 mA.

The device is available in two different enclosures: transistor-like, indicated by the suffix 'LP' in the type number, and 8-pin DIL, suffixed D, P or JG. The LP model is eminently suitable as a replacement of a conventional potential divider and capacitor (and possibly, opamp). The DIL model has the advantage that a capacitor may be connected to its pin 8, which in some cases is desirable since each of the two output voltages is half the supply voltage, which may contain much noise.

There is alo the Type TLE2425, which provides a fixed voltage of 2.5 V at a peak current of 20 mA. Pin 8 of this model is not connected.

(K. Walraven – 934005)

STATE OF BATTERY INDICATOR

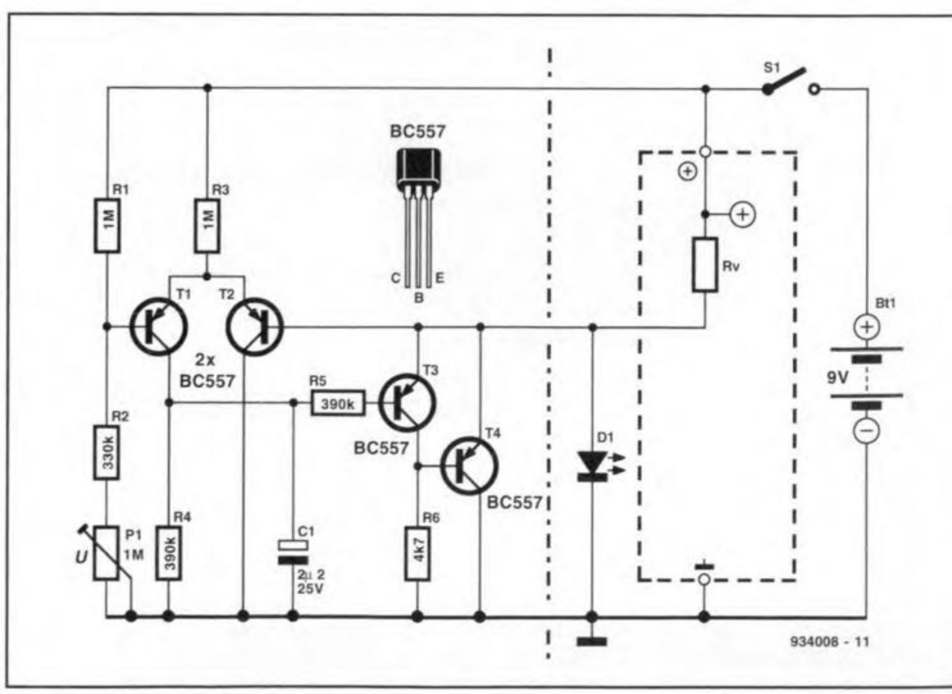
Many battery-operated appliances have an LED indicator that shows the state of the battery. If the battery is almost flat, the LED flickers. The circuit described enables this facility to be built into apparatus that has no indicator.

The drop across D_1 (U_{LED}) depends on the type of LED and is normally around 2 V. This voltage is applied to the input of a discrete differential amplifier, T_1-T_2. When the base voltage of T_1 is smaller than U_{LED}, the transistor switches on, whereupon C_1 gets charged via R_1. The rising potential on C_1 will cause T_3 to switch off, whereupon T_4 begins to conduct. Because of that, D_1 goes out and remains so until C_1 is discharged. Assuming that the base voltage of T_1 is still smaller than U_{LED}, the cycle starts afresh. In this way, the LED continues to flicker until the base voltage of T_1 becomes larger than U_{LED}. Only T_2 and T_3 are then on and T_4 is off, so that the LED lights continuously.

The setting of preset P_1 determines the point at which the LED begins to flicker. The frequency at which it flickers depends on the supply voltage and the value of C_1. The frequency may therefore be changed by altering the value of this capacitor.

The value of R_V determines the current through D_1 and thus the brightness at which this LED lights. In practice, the LED and R_V are often already present in the appliance.

[F. Rimatzki - 934008]



TWO-WAY AUDIO LEVEL INDICATOR

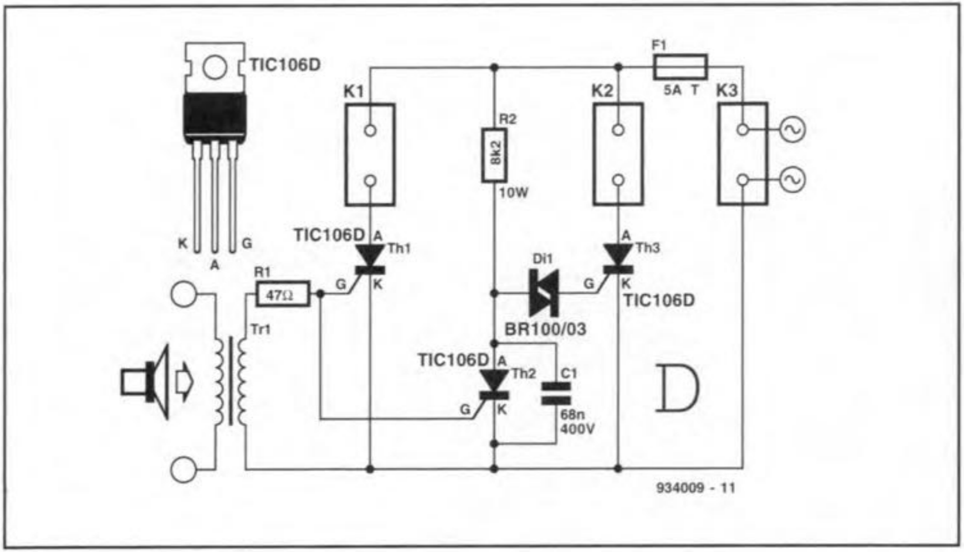
The audio signal is applied to the indicator circuit via a 1:1–1:4 microphone transformer. Its level determines which of the two lamps, L_1 and L_2, connected to K_1 and K_2 respectively, is switched on by the relevant silicon-controlled rectifier (SCR).

In the absence of an input signal, both SCRs are switched off. During the positive half period of the mains voltage, drive voltage is applied to the gate of Th_3 via R_2 and diac Di_1. The SCR is then on and switches on L_2. When the audio input reaches a certain level, Th_1 and Th_2 also conduct and switch on L_1. At the same time, Th_2 removes the drive to Th_3, so that this SCR switches off at the next zero crossing of the mains and lamp L_2 goes out.

Great care must be taken in the construction of the indicator, because potentially fatal mains voltages are present in it during operation. It must, therefore, be built into a well-insulated (man-made fibre) enclosure. The transformer **must be a type whose insulation is approved for operation in mains-carrying circuits.**

Resistor R_2 must be high-power type, and C_1 a 400-V type.

[Dr U. Kunz - 934009]



COPYING IN MS-DOS WITH ONE DRIVE

MS-DOS is a program developed for controlling a computer system, which it can do with a variety of procedures. One of these is the making of batch files in which a number of commands can be combined to form a new instruction. Batch file ACOPY copies files from drive A to drive A. A small alteration makes it possible for the instruction to to be used with other drives as well. The batch file contains several infrequently used procedures and is intended for systems that have only one diskette drive or have two incompatible drives.

The nature of the alteration can be seen in the listing. It has been assumed that the computer has a hard disk or a RAM disk with sufficient free space. This is because the batch file stores all files to be copied temporarily on the hard disk or RAM disk,

The file starts as usual with **echo off**., and then checks whether the temporary directory already exists and whether it contains any files. If so, the user is warned and asked for a decision. This is effected by the first section of the batch file from **if** to the lable **:endif**. The **if** command checks whether the directory is empty with **not** and **exist**. If that is the case, the commands up to **:endif** are ignored. If the directory is not empty, its content is shown on the monitor with **dir**, followed by a message directed to the screen by **echo**. The **pause** command gives ample time for a decision to be taken.

After the decision to continue, the content of directory c:\copy.tmp is erased automatically. The question 'are you sure?' is diverted to the zero-device (and is, therefore, not displayed) and replied to automatically by 'y(es)'. The zero device is a fictitious peripheral, which, as far as the computer is concerned, acts in the same way as a monitor or printer. It does nothing else and is eminently suitable for use as a waste basket. The automatic reply 'y(es)' is effected by the **echo** instruction, whose output (y) with the 'pipe' symbol as input is passed on to the **delete** instruction.

The next step is the creation of subdirectory c:\copy.tmp, which, if all is in order, is not present. Should this directory exist after all, it will be empty and be therefore usable.

Next, the diskette with the files to be copied is put into drive A. The files that must be copied are identified in the same way as with the MS-DOS **copy** instruction. No drive letter should be input, because that is already in the file.

The batch file cannot be used for $3^1/_2$ in **and** $5^1/_4$ in drives: each of these needs its own batch file. The only difference between these two files will be the drive letter in the **copy** instructions.

Once the files to be copied are stored in the temporary directory, the disks can be exchanged after which the copying can be completed.

Finally, the folders in c:\copy.tmp must be earased and the directory removed

The second **copy** instruction contains a full stop where a *.* might be expected. This full stop, in MS-DOS, means 'the present directory' and may be used in many cases instead of *.*. For instance, **del a:.** is the same as **del a:*.***.

Echo instructions followed by a full stop provide an blank line on the screen. This is so even with some other versions of MS-DOS, although the manual does not state this.

The syntax of the ACOPY command is

ACOPY pathname [pathname].

Bear in mind that with this method subdirectories must not be copied together with the files.

A last remark: the **pathname** must be given from the root directory.

[A.N. Other – 880191]

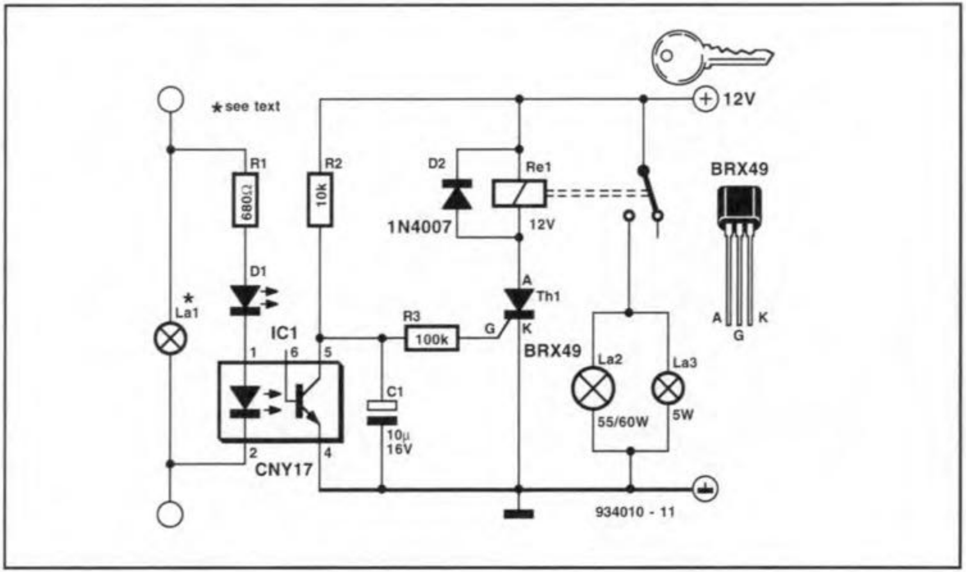
```
@echo off

if not exist c:\copy.tmp\*.* goto
endif
dir c:\copy.tmp /w
echo.
echo directory C:\COPY.TMP al-
ready exist
echo press ctrl-C to abort ACOPY
echo press any key to delete
C:\COPY.TMP\*.*
echo and continue ACOPY
pause > nul:
echo y | del c:\copy.tmp > nul:
:endif

ctty nul:
mkdir c:\copy.tmp
ctty con:
echo. echo Insert SOURCE
diskette in drive a:
pause
copy a:%1 c:\copy.tmp

echo.
echo Insert TARGET
diskette in drive a:
pause
copy c:\copy.tmp\. a:%2 >
nul:
echo y | del c:\copy.tmp >
nul:
rmdir c:\copy.tmp
```

LIGHT SWITCH FOR MOTORCYCLES

Many motorcycles have the facility that as soon as the ignition is switched on, their dipped headlights (or day running lights) come on. Although on most modern machines these lights go off briefly while the engine is started (so as to prevent a heavy load on the battery), there are some on which this is not so. This shortcoming may be remedied by the circuit described here.

The circuit is based on optoisolator IC_1, which detects whether the 'N(eutral)' light, La_1, on the dashboard is on. As long as this light is on, the machine is not in gear. When the engine is started (with the gear box still in 'N'), La_1 remains on, so that the LED in the optoisolator lights. This switches on the transistor in IC_1. The gate of the thyristor is then at earth potential and the relay remains deenergized. Headlights LA_2 and La_3 are off and the engine can be started without these lights coming on. When a gear is selected, La_1 goes out and the transistor in IC_1 switches off. The thyristor then gets a gate current via R_2 and R_3, whereupon Re_1 is energized. The headlights then come on and the machine can be ridden away safely. The thyristor remains on as long as the battery is being charged, so that when 'N' is selected, for instance, at traffic lights, the lights remain on.

[M. Weber - 934010]



VOICE OPERATED RECORDING

VOICE operated recording (VOR) is a feature found on most handheld memo-recorders. The function of the VOR control is to start the recorder's tape on detection of a speech signal. This relieves the user from pressing a 'record' button. Typical memo-recorder users include the too energetic manager and the detective on TV, unravelling a murder plot by talking to himself behind the wheel of his fast sportscar. The circuit shown here adds a VOR function to any standard cassette recorder, provided this has a 'remote' control input. Alternatively, it could be used as a VOX (voice operated switch) to control a transmitter, or a sensitive sound detector in an alarm system.

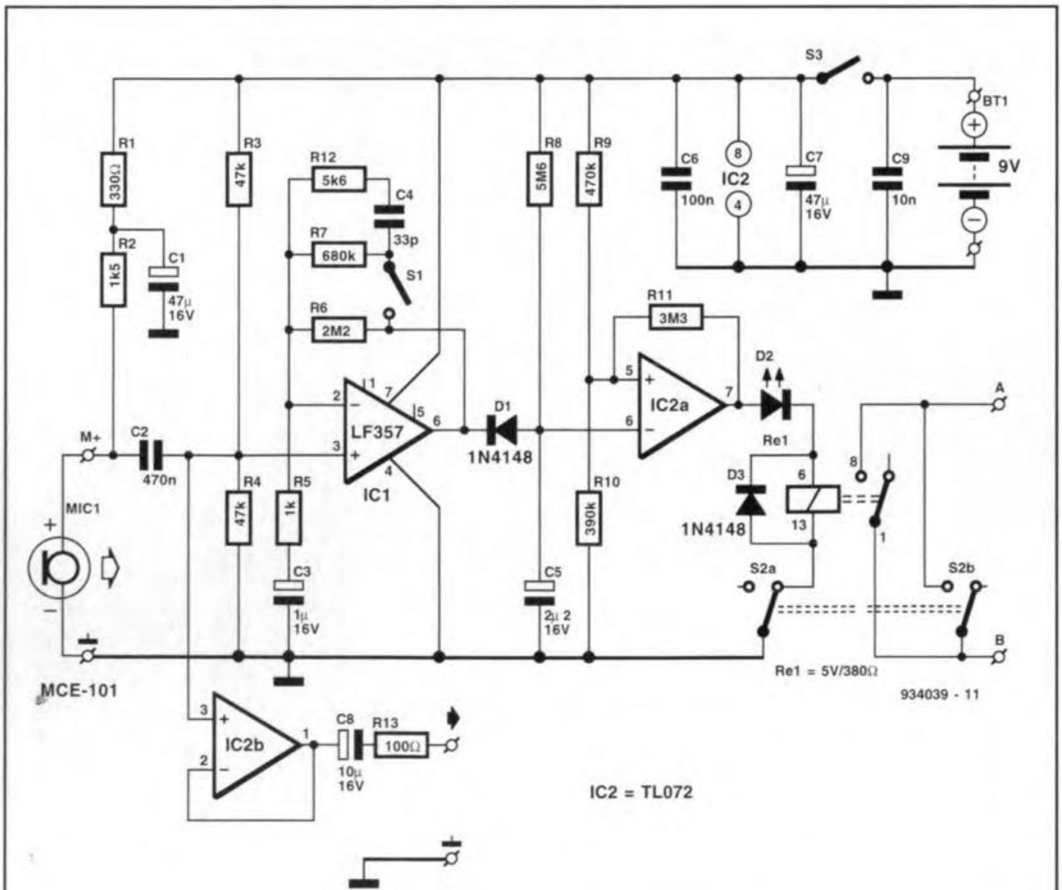
Depending on the position of switch S_1, the output signal from the electret microphone (approx. 5 mV) is amplified ×520 or ×2,200 by opamp IC_1, which also functions as an active filter with a (speech) bandwidth from 160 Hz to about 9 kHz. Components D_1, C_5 and R_8 turn the amplified audio signal into a direct voltage, which is applied to the inverting input of comparator IC_{2a}. If the sound picked up by the microphone is sufficiently loud, the voltage at the inverting input falls below that at the non-inverting input, and the comparator output changes state. The resulting logic high at pin 7 of IC_{2a} causes relay Re_1 to be energized, so that its contact closes. If the contact is connected to the remote control input on the recorder, the tape will be started, and recording begins. This condition is indicated by the lighting of D_2.

The supply voltage to the circuit and the microphone is switched on and off with S_3. Switch S_2 allows the VOR function only to be switched off, while S_1 serves to select between 'low' and 'very high' sensitivity.

Since the gain-bandwidth product of the LF357 is about 20 MHz, the opamp will have a bandwidth of about 9 kHz if its amplification is ×2,200. With S_1 closed, the gain is reduced to ×520 times, which results in an effective bandwidth of 38.5 kHz. Since this is not practical or desirable for speech signals, capacitor C_4 is included in the feedback loop to give a bandwidth of 9 kHz at the lower gain also. Resistor R_{12} prevents the opamp oscillating by maintaining a minimum amplification of of ×5.

When the relay is energized, the circuit draws about 18 mA from the 9 V battery; if not, the current is roughly halved. The minimum battery voltage for reliable operation is about 6.8 V.

The VOR is an extremely sensitive circuit, and to ensure that it can not be upset by strong RF signals, it is best built on the printed circuit board shown here. The dimensions of the board allow it to be fitted into a ready-made plastic enclosure with an integral battery compartment. A suggestion for a nice looking front panel is also given. Notice that all inputs and outputs of the circuit are located at one side of the PCB — this is



PARTS LIST

Resistors:

1	330Ω	R1
1	1kΩ5	R2
2	47kΩ	R3;R4
1	1kΩ	R5
1	2MΩ2	R6
1	680kΩ	R7
1	5MΩ6	R8
1	470kΩ	R9
1	390kΩ	R10
1	3MΩ3	R11
1	5kΩ6	R12
1	100Ω	R13

Capacitors:

2	47μF 16V	C1;C7
1	470nF	C2
1	1μF 16V	C3
1	33pF	C4
1	2μF2 16V	C5
1	100nF	C6
1	10μF 16V	C8
1	10nF	C9

Semiconductors:

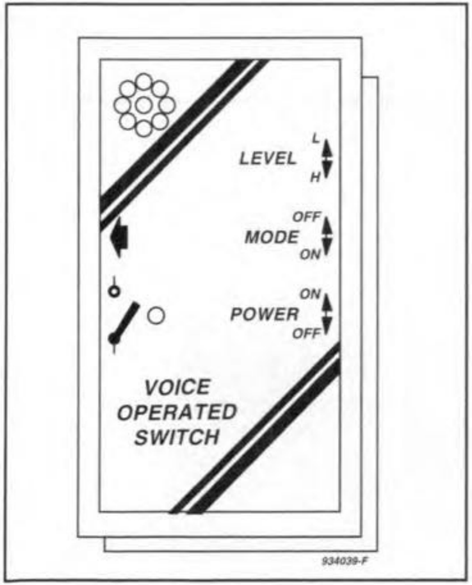
2	1N4148	D1;D3
1	LED, red, 3mm	D2
1	LF357	IC1
1	TL072	IC2

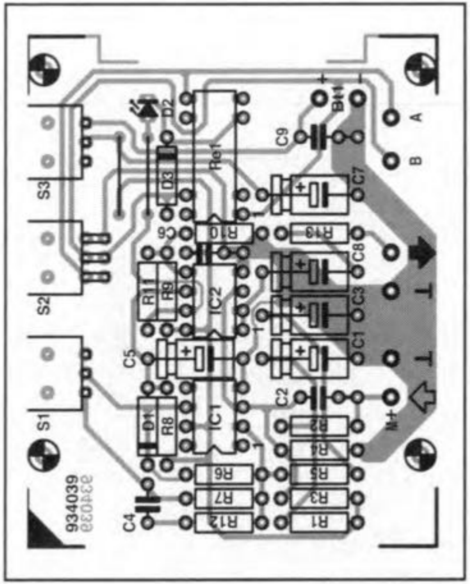
Miscellaneous:

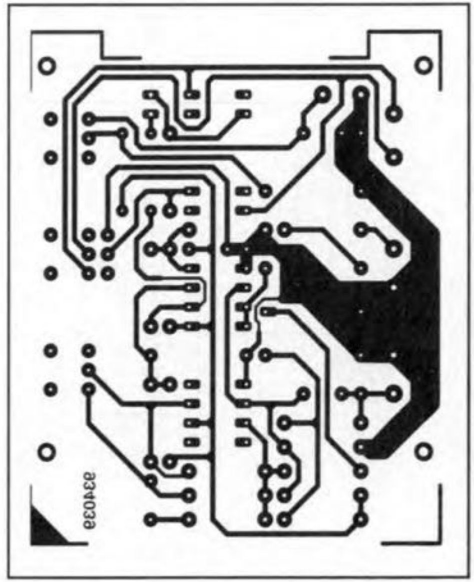
- 2 AS1D-5M single-pole changeover slide switch, angled wires, PCB mount (Fujisoko; Van Reijsen Electronica) S1;S3
- 1 AS2D-5M double-pole changeover slide switch, angled wires, PCB mount (Fujisoko; Van Reijsen Electronica) S2
- 1 MCE-101 electret microphone (Monacor/Monarch) X1
- 1 V23100-V4005-A001 relay (5V/380Ω; Siemens) Re1
- 1 Plastic case; dim. 61×97×26mm (PacTec Type HM-kit)
- 1 9-V battery clip
- 1 Printed circuit board 934039 (see page 112)

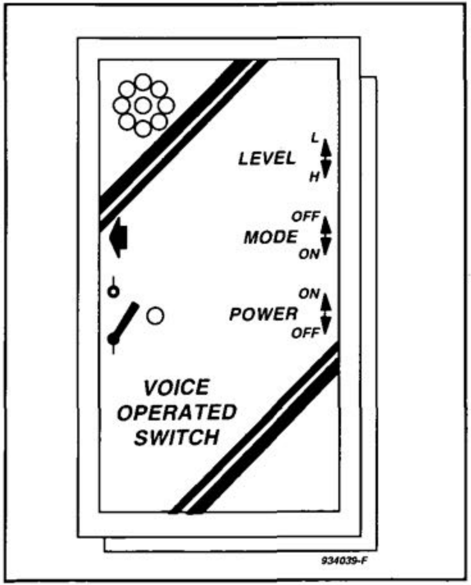
done to ensure that common-mode signals induced in the cables are not 'forced' to cross the PCB, where they could affect the operation of the electronics. All connections to ground are soldered to a wide copper pad that forms a low impedance. An exception has been made in the case of capacitor C_5: to prevent the discharge currents interfering with the microphone signal, they should not be allowed to travel through the ground pad.

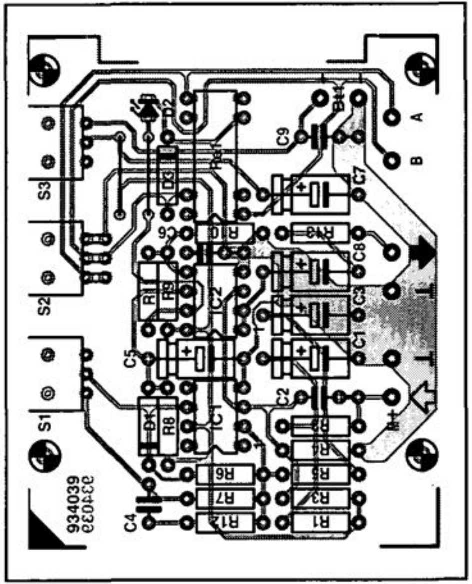
[J. Ruiters - 934039]

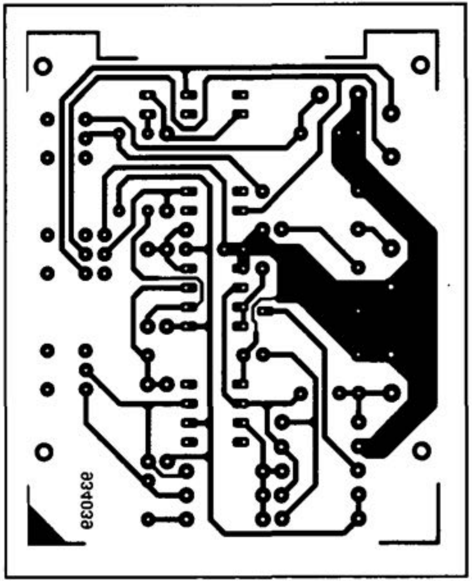







LEVEL
L
H
MODE
OFF
ON
POWER
ON
OFF
VOICE
OPERATED
SWITCH
934039-F


S3
S2
S1
A
B
934039



TELEPHONE-OPERATED NIGHT LIGHT

In most telephone networks, a direct voltage of 48–60 V exists at terminals *a* and *b*. When the call signal comes in, a 50–60 V alternating current at a frequency of 25–27 Hz is superimposed on the direct voltage. This open-circuit voltage is more than halved when the receiver is taken from the hook (which closes the current loop). The current in the loop is 20–100 mA.

The present circuit starts a timer when it detects the call signal.The timer energizes relay Re_1 and this switches on lamp La_1. After the last ring, the timer remains on for a short time, but then returns to its quiescent state, whereupon the lamp goes out. When, however, the receiver is taken from the hook, the timer is triggered permanently, so that the lamp remains on during the telephone conversation and for some time afterwards. Furthermore, the lamp may be switched on and off manually by S_1 being pressed.

One part of the circuit is connected in parallel, and the other in series, with telephone terminals *a* and *b*. The parallel-connected part, which detects the call signal, is isolated from the direct voltage by C_3. Because of this, no current flows through the LED of optoisolator IC_2, so that the phototransistor is off. The second optoisolator, which is in series with the *a* and *b* lines, also carries no current since the circuit is broken by the cradle switch.

The call signal, however, is applied to IC_2 via C_3. Consequently, a current flows through the LED and the transistor is on. The current is limited by R_1, while diodes D_2 and D_3 chop the peaks off both the half waves to prevent any damage to the optoisolator.

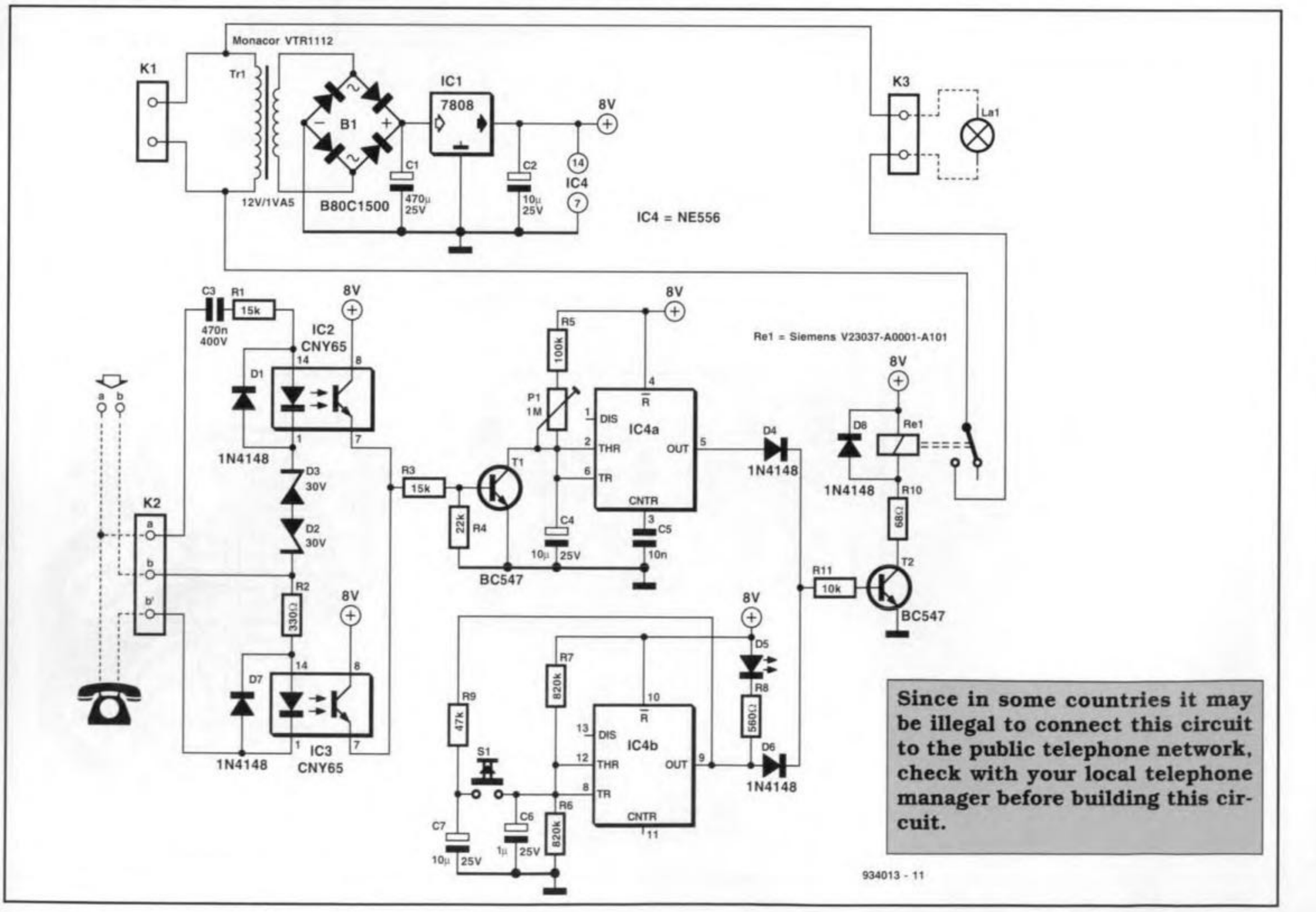
The phototransistor provides sufficient base voltage to T_1 to switch on this transistor. Capacitor C_4 discharges, whereupon the potential at pins 2 and 6 of timer IC_{4a} drops to earth, which triggers the timer. Its output changes state (goes high), whereupon T_2 is switched on so that the relay is energized. The relay contact connects one of the mains lines (in the UK, the live) to lamp La_1. The time constant of the timer has been arranged so that even the longest intervals in the call signal can be bridged. If the receiver is not lifted, and the last call signal has passed, the lamp remains on for the duration of the time constant, and then the circuit returns to the quiescent state.

When the receiver is lifted off the hook, a current flows through the LED in IC_3, whereupon the phototransistor in the optoisolator, and consequently T_1, is switched on. The timer is triggered as long as the telephone conversation lasts. When the receiver is replaced into the cradle, the lamp stays on for a short while determined by the setting of P_1.

The second timer in the NE556, IC_{4b}, is used for manual operation of the lamp. Until this timer is triggered, half the supply voltage exists at pins 8 and 12. The output (pin 9) is low, so that C_7 cannot be charged. When S_1 is pressed, the trigger voltage briefly drops below the threshold, whereupon the output goes high. Since, owing to the time constant R_9C_7, the potential across C_7 rises only gradually, the level at the trigger input (pin 8) remains below the upper threshold. When the switch is operated again, the full supply voltage is across C_7, so that the level at the trigger input exceeds the upper threshold. The timer output then goes low again and the lamp goes out.

The power supply is a standard design, but could be replaced by a suitable 12 V mains adaptor. The lamp could then be a 12 V car type. From a safety point of view, that is not a bad alternative.

[F. Hueber - 934013]



Since in some countries it may be illegal to connect this circuit to the public telephone network, check with your local telephone manager before building this circuit.

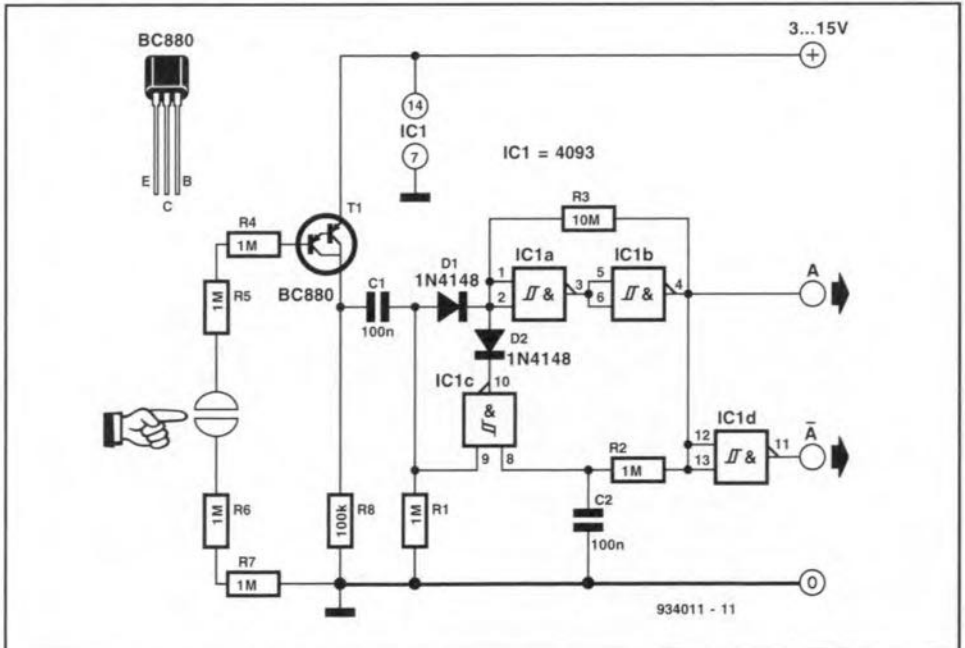
SEQUENTIAL TOUCH SWITCH

The touch switch is based on a Schmitt trigger with hold contact, which is built from IC_{1a}, IC_{1b} and R_3. The output of IC_{1b} is fed back to IC_{1a} via R_3, resulting in both gates maintaining their stable status. As long as the circuit is quiescent, that is, the anode of D_1 is low and the cathode of D_2 is high, this situation continues.

Consider that output A is low. When the touch contact is shorted with a finger, the potential acorss R_8 rises. This rise is translated into a short pulse by R_1-C_1. This pulse renders the input of IC_{1a} high via D_1. The output of IC_{1b}, and output A, will then also go high and, as explained before, this status is maintained.

At the same time, the potential across C_2 will rise gradually. When this capacitor is fully (or nearly so) charged, the circuit is ready to change state at the next pulse across R_1. This pulse will make the output of IC_{1a} low, whereupon the output of IC_{1b} becomes low and the potential across C_2 drops. Output A is then low again and the circuit is back to where it started.

[P. Sicherman - 934011]



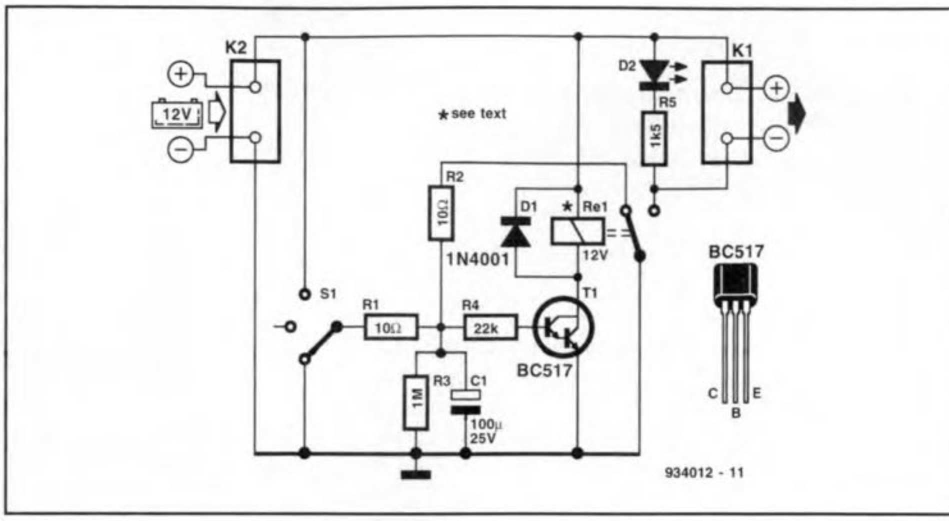
AUTOMATIC POWER OFF

The power off circuit ensures that the load is automatically removed from a battery. It is particularly useful in cars, so that if the driver forgets to switch off the headlights, he does not return to a car with a flat battery a couple of hours later*.

The only operating control is a spring-loaded change-over switch with a test position. When the switch is pressed, the +ve supply line is connected to the base of T_1 via R_1 and R_4, so that the transistor switches on. The relay is then energized and its contact links the –ve terminal of the load to ground, so that the load is powered.

At the same time, C_1 is charged via R_1. This capacitor and R_3 determine the time constant of the circuit. When S_1 is set to its centre position, C_1 discharges slowly via R_3. When the base potential of T_1 drops below 1.2 V, the transistor switches off, whereupon the relay contact connects C_1 to ground via R_2. The capacitor then rapidly discharges completely. The current drawn by the circuit in this situation is nil.

When S_1 is in its lowest position, that is, connected to ground, the discharging process is halted immediately.



Diode D_2 shows the status of the circuit. Resistor R_5 should not have too low a value, otherwise the LED may not show up properly.

The value of C_1 may be increased to 4700 µF: the power-off time is then increased from 10 min to about 30 min. This capacitor should be a type with very low leakage current.

The relay may be 6 V or 9 V type to prevent it clattering when the car engine is started or a large load is switched on. Do not omit protection diode D_1.

Since the circuit has no polarity protection, take great care that correct polarity is observed.

[W. Zeiller - 934012]

* In many modern cars, this switch off facility is not needed since the ignition switch removes all loads from the battery. [Editor]

AUDIBLE TESTER

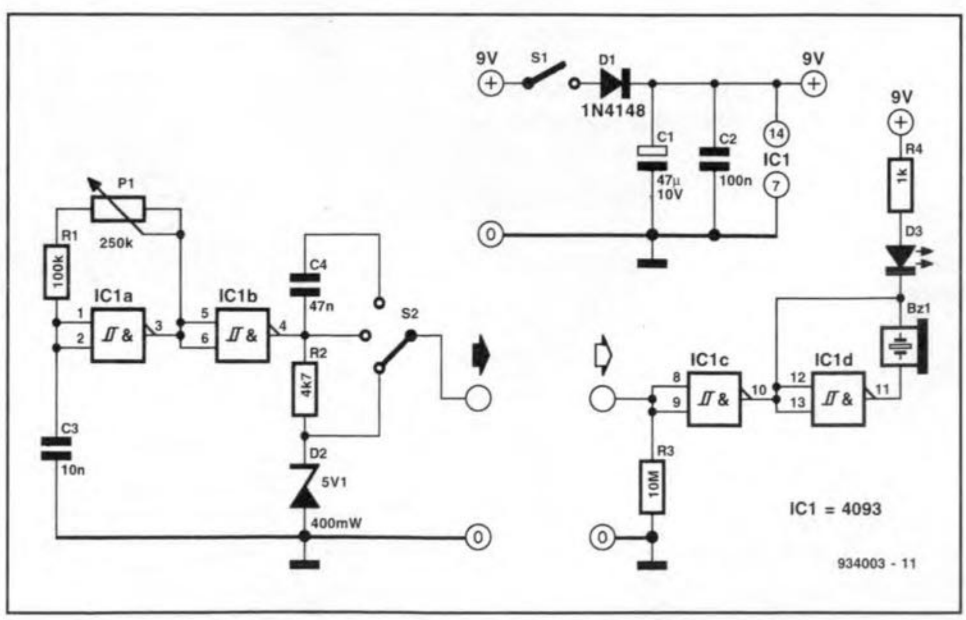
A simple unit for testing an audio, TTL or CMOS circuit is always useful. Here, an oscillator, based on IC_{1a}, generates a frequency that can be preset between about 500 Hz and 1.5 kHz with P_1. Before the signal is ready for use, it is buffered by IC_{1b}.

Switch S_2 serves to match the signal to the equipment on test. With the switch in the upper position, the tester can be used with audio circuits; in the centre position, with CMOS circuits, and in the lowest position, with TTL circuits (only HC or HCT types, not standard TTL logic or LS2 types). If testing of LS2 type circuits is required, the value of R_2 must be reduced to 1 kΩ. The 4093 cannot provide sufficient current for testing standard TTL logic circuits.

The output of the probe that is used to check various points in the equipment on test is applied to the input of IC_{1c}. If there is a signal present, the LED lights and the buzzer sounds.

The tester may be powered by a 9-V battery or a 9-V mains adaptor. It may also be possible to obtain the power from the equipment on test. The tester draws a current of about 11 mA.

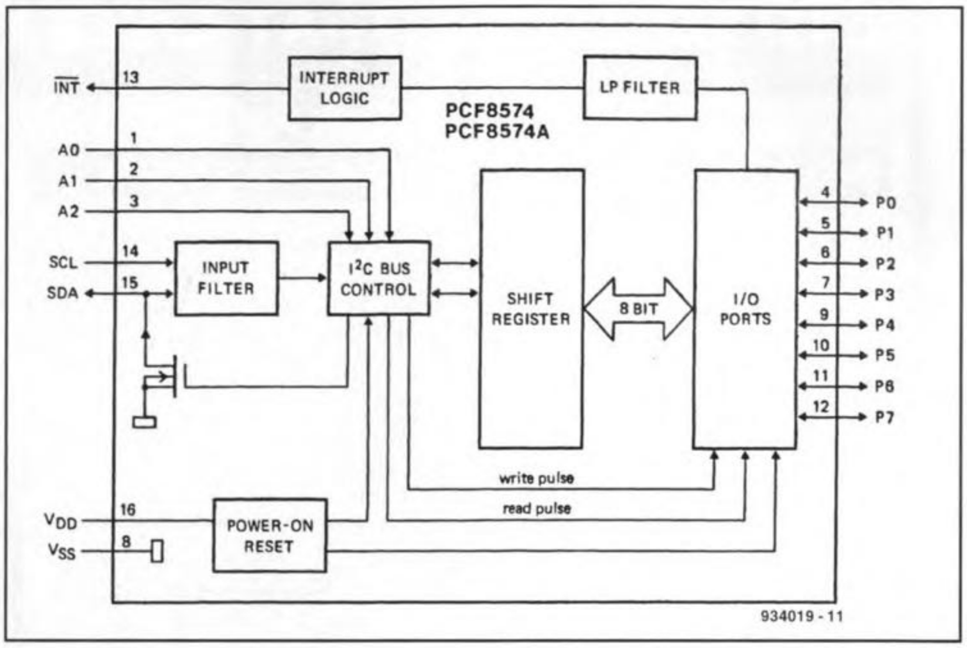
[Amrit Bir Tiwana - 934003]

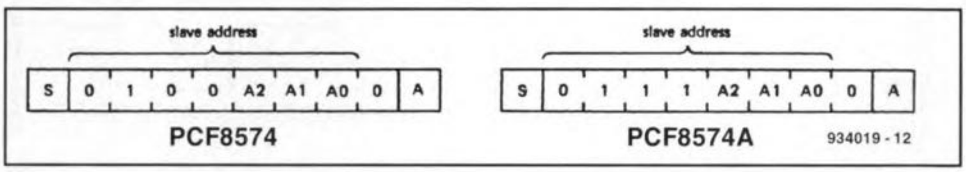


I²C TIP

The 8-bit I/O chip Type PCF8574 is very popular: it is used, among others, on the AD-DA I/O module, the opto relay card and the LC display – all published in this magazine within the past 12 months. Since it is often desirable to connect a number of 8574 modules to the I²C bus, it is possible to give each IC a different address via inputs A0–A2. As therefore only eight addresses can be set, it might be expected that this is also the maximum number of ICs that can be connected to the bus. It is, however, possible to connect 16 PCF8574s, because there are two variants of the IC, each with a different address. The standard PCF8574 has, as is well known, the basic address 40_{HEX}. The variant, the PCF8574A, has the basic address 70_{HEX}. Otherwise the two types are identical and, therefore, completely interchangeable. Thus, eight standard **and** eight variant types of the IC can be connected to one bus.

[J. Ruiters – 934019]





SINGLE-CHIP TEMPERATURE CONTROL

The UAA2016 from Motorola lends itself admirably to making a proportional mains-operated temperature control. A resistor with negative temperature coefficient (NTC), R_3, which must have a value of about 100 kΩ at 25 °C, serves as sensor. In the diagram, the heating element is a 100 W light bulb, which generates enough heat for a greenhouse or an environmental chamber (for testing electronic equipment).

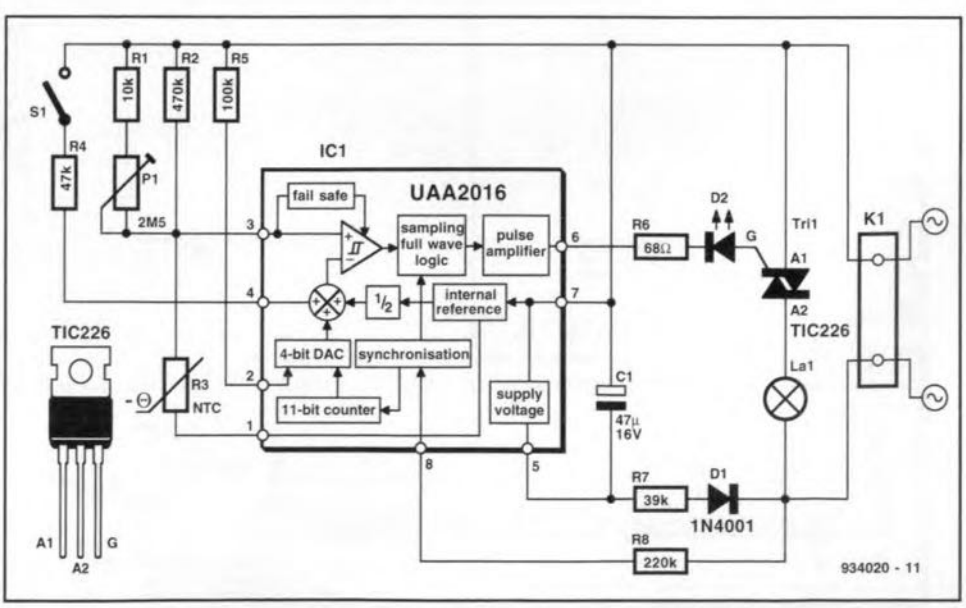
The triac can switch up to 3 A if it is fitted on a heat sink.

The properties of the regulator are determined by a the resistors at pins 1–4. Resistors R_1 and R_2 establish over which range the temperature may be set with P_1. With values shown, the range is 0–70 °C. Closing S_1 causes the temperature to drop by about 2 °C. The degree to which the temperature drops is determined by R_4: a value of 100 kΩ causes a drop of about 1.5 °C, while 10 kΩ results in a reduction of about 5 °C.

Resistor R_5 determines the hysteresis of the regulator: here, it is about 150 mV. How much that is in °C depends on the type of NTC and the temperature: an NTC has a non-linear temperature characteristic.

Do bear in mind that the entire circuit is connected to the mains and that, therefore, the regulator must be constructed with due care to safety regulations. It should be installed in a well-insulated, earthed (if metal) case.

[A. Rietjens - 934020]



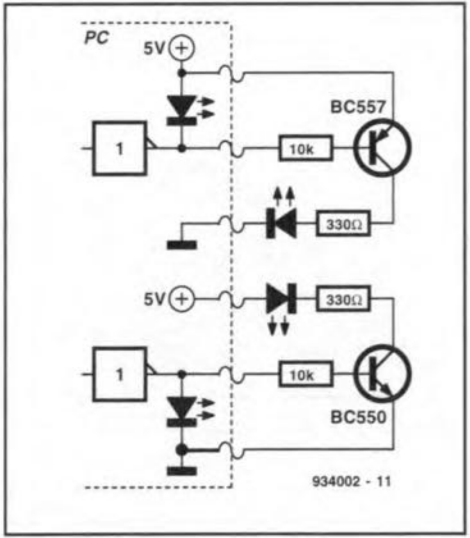
EXTENDED PC DRIVE LED

When the disk drive indicator of a computer cannot be read (since the computer is under or beside the table) it may be useful, if difficulties arise, to have an extended indicator alongside the monitor.

The circuit described here is just that. It consists of a p-n-p transistor, an LED, and two resistors, or an n-p-n transistor, an LED, and two resistors, Which is to be used depends on the LED drive of the floppy disk or hard disk.

The LED in the computer must be branched off to a D connector to be fitted at the rear of the computer.

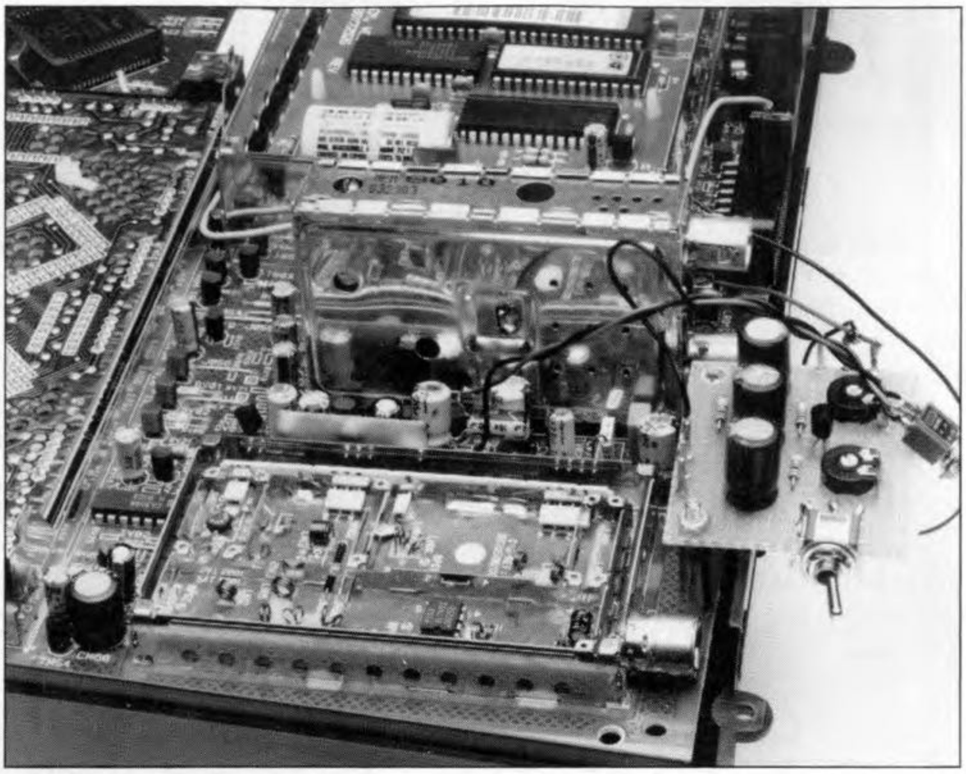
The small circuit is easily constructed on a small piece of prototyping board and built into a small case. The whole is then connected via a cable terminated into a D connector to the computer. The small case can be located wherever desired.



[L. Lemmens - 934002]

FERGUSON BSB RECEIVER AS D2MAC DECODER

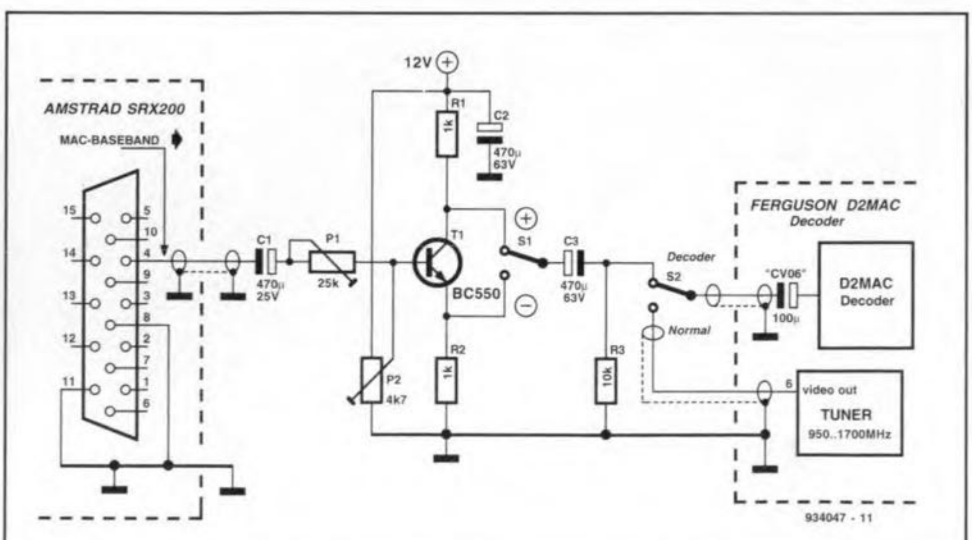
After the demise of BSB (British Satellite Broadcasting) in early 1991, the electronics surplus market was flooded with tens of thousands of brand new, but otherwise unusable, satellite TV indoor units plus 55-cm dish antennas or squarials. Initially, unmodified ex-BSB sets went at give-away prices of £20 and less, which is a fraction of the original price. Unfortunately, these receivers are designed to handle DMAC only, a TV transmission standard used by BSB and a couple of Scandinavian broadcasters only. Not surprisingly, the 'scrapped' BSB sets were soon modified to handle the more widely used D2MAC format, and then sold as 'upgraded' receivers. One of the companies that has been actively involved in this is Trac Satellite Systems (address below). The modification from DMAC to D2MAC essentially consists of fitting a new EPROM, reprogramming the receiver's built-in EEPROM, and disabling the access control module (ACM). The result is a complete D2MAC satellite receiving system that costs less than a D2MAC decoder alone. The pictures and sound produced by the modified sets are, in a word, superb.



In addition to complete, modified BSB receiver sets, there are also many 'scrapped' receiver boards around. This article shows you how to use such a board, in this case from a Ferguson SRB1 receiver, as a D2MAC decoder for use with an Amstrad SRX200 Astra receiver. The conditions for being able to do so are: (1) your Ferguson board is in working order, and (2) it has been modified for D2MAC.

One transistor does the job! The interface shown here consists of an inverter, which is required to ensure the correct polarity of the baseband signal fed into the D2MAC decoder. The baseband signal is taken from pin 4 on the 15-way sub-D socket on the rear panel of the SRX200, and fed to a simple transistor inverter. Switch S_2 is added to enable the tuner inside the Ferguson receiver to be used as a second input to the D2MAC decoder. This allows you to keep using the Ferguson set for reception of the D2MAC channels on the German and French high-power DBS channels at 19° West. A second switch, S_1, selects the polarity of the video signal (which can give quite interesting effects on some D2MAC broadcasts). Finally, the interface should be connected with short coax cables. If you cannot avoid the use of a relatively long cable (for instance, a standard length SCART cable), fit a 10 pF speed-up trimmer across the fixed terminals of P_1.

[B. Romijn - 934047]



Converted SRB-1 receiver systems are available from Trac Satellite Systems, Commerce Way, Skippers Lane, Middlesbrough, Cleveland TS6 6UR. Telephone: (0642) 468145. Fax: (0642) 440927.

GENERAL PURPOSE DIGITAL TESTER

The tester cannot only detect rectangular-wave signals, it can also generate them. The test signals may be generated continuously or for a preset period as required.

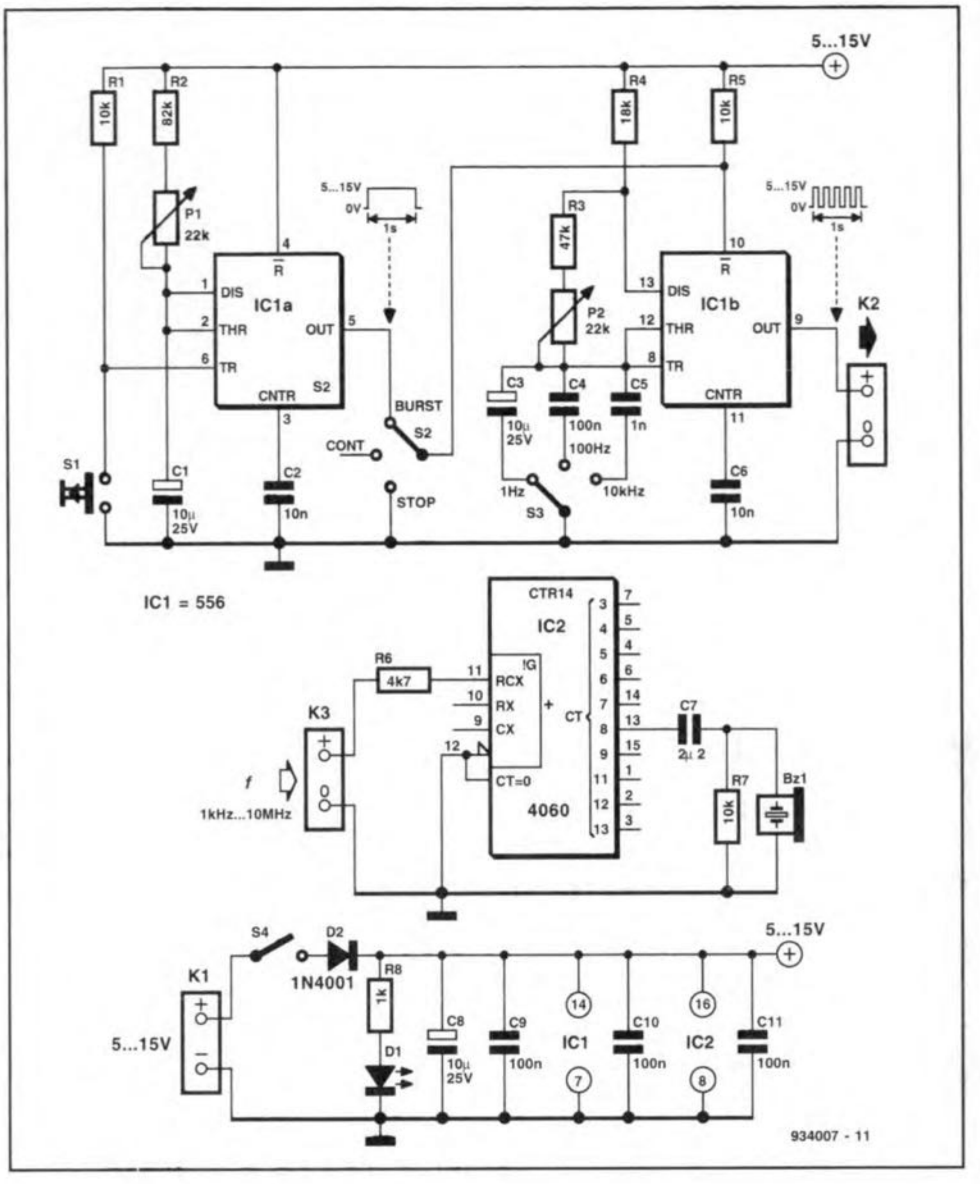
The detector part is based on IC_2. A signal input to K_3 is applied to IC_2, where it is divided by 1024 and then fed to a buzzer. The buzzer will sound at a frequency between 1 Hz and 10 kHz if a signal is present at a frequency between 1 kHz and 10 MHz respectively.

The generator is formed by monostable IC_{1a} and variable oscillator IC_{1b}. The pulse width of the monostable may be set to 1s maximum with P_1. The monostable is started when push-button switch S_1 is pressed. Rotary switch S_2 serves to select a continuous or a burst signal; in its third position there is no output.

The oscillator operates over three ranges selected by S_3: 1 Hz, 100 Hz, 10 kHz. The wanted frequency is selected with P_2 The required signal is available at K_2.

Although the tester may be powered by a 9-V battery, provision has been made for an external 5–15 V source. Diode D_1 serves as power on indicator. Switch S_4 is the on-off switch. The tester draws a current of about 50 mA.

[Amrit Bir Tiwana - 934007]

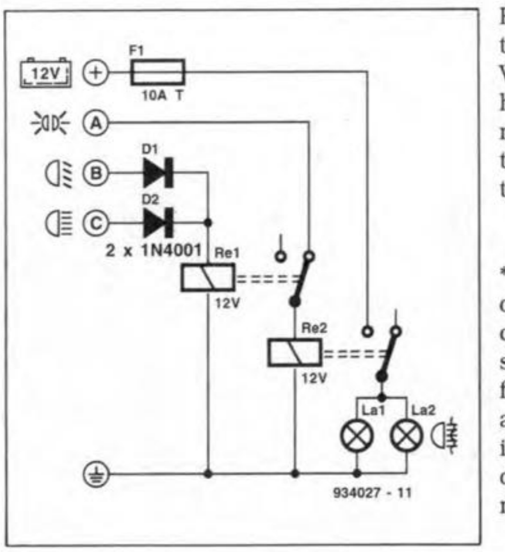


FOG LIGHT SWITCHING

Many cars are not fitted with fog lights and, what is more, there is often no provision for them. However, since in most European and North American countries it is illegal to drive on sidelights and many cars still have a switch for these, the fog lights may be operated with this. The only trouble is that the fog lights are then on also when the headlights are on. This difficulty* can, however, be remedied with two relays and two diodes.

When the sidelights are switched on, relay Re_2 is energized and the fog lights are on. When the headlights are switched on, relay Re_1 is energized via one of the diodes. Relay Re_2 is then no longer actuated and the fog lights go out.

Relay Re_1 may be any 12-V type with a break or change-over contact, but for

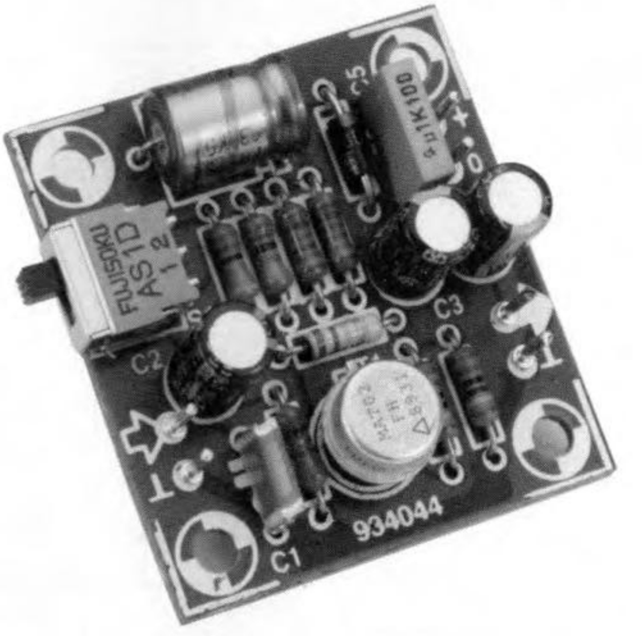


Re_2 a vehicle type must be used owing to the heavy current through the fog lights. Vehicle type relays can normally easily handle 10 A direct current. They are not normally available from electronics retailers, but most car accessory shops stock them.

[J. Bosman - 934027]

* There is a fundamental difference of opinion between the designer and most car manufacturers, who make it impossible—rightly in our opinion— for the fog lights to be on when the headlights are not. In fact, in many countries it is illegal to drive with fog lights on only; even during daylight fog, headlights (dipped) must be on also. [Editor]

LOW-NOISE MICROPHONE AMPLIFIER



WITH the arrival of matched transistors in a single package it seems that the quest for the zero-noise microphone preamplifier has received a new impetus. The present amplifier is based on PMI's MAT02 transistor pair. The performance of this amplifier is such that even the most demanding audio enthusiast will be impressed by its very low noise contribution. The amplifier is simple to match to a wide range of microphone impedances, and offers a voltage gain of 20 dB or 23.5 dB (amplification of ×10 or ×15 respectively), selectable with a switch.

The preamplifier is a two-stage direct coupled type with feedback and an input stage that runs on a very low collector current to minimise the noise contribution. The output impedance is about 70 Ω at a gain of 23.5 dB. The preamplifier is built on a compact printed-circuit board as illustrated. Its low current drain (approx. 2.5 mA) and 9-V battery supply make it ideal for portable use in conjunction with a high-quality dynamic microphone.

Since resistor R_3 is responsible for the input impedance and some other important characteristics of the preamplifier, we will have a look at some measured data with different values of R_3.

With R_3 = 220 kΩ, the input impedance is about 30 kΩ. A THD+N (total harmonic distortion plus noise) of 0.045% was measured; the noise level was –65 dB, measured in a bandwidth of 22 kHz, at an output voltage of 15 mV. Source impedance and overall gain were found to have a negligible effect on these figures.

Changing the value of R_3 to 6.8 kΩ resulted in a THD+N figure of 0.042% at an input voltage of 1 mV, and a noise level of –65 dB. At this resistor value the input impedance dropped to about 1 kΩ. At a source impedance of about 600 Ω, the THD is smaller than –95 dB (noise floor). On the down side, however, the lower input impedance results in a reduction in overall gain of about 4 dB. An even lower source impedance of 25 Ω was found to result in a THD of –94 dB.

(W. Zeiller — 934044)

COMPONENTS LIST

Resistors:
1% resistors: metal film from E96 series

1	56Ω2 1%	R1
1	22kΩ1 1%	R2
1	220kΩ	R3
1	562Ω 1%	R4
1	866Ω 1%	R5
2	280Ω 1%	R6;R7
1	2kΩ21 1%	R8

Capacitors:

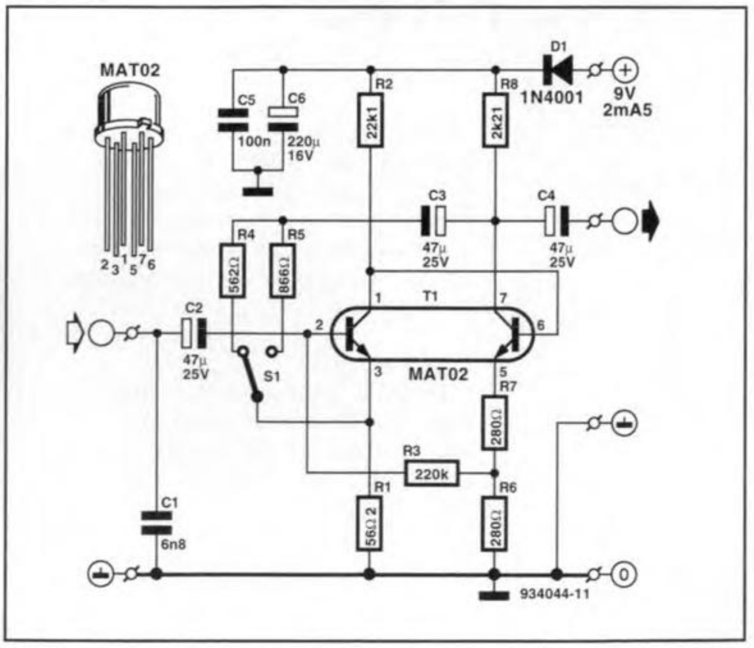
1	6nF8	C1
3	47µF 25V radial	C2;C3;C4
1	100nF	C5
1	220µF 16V	C6

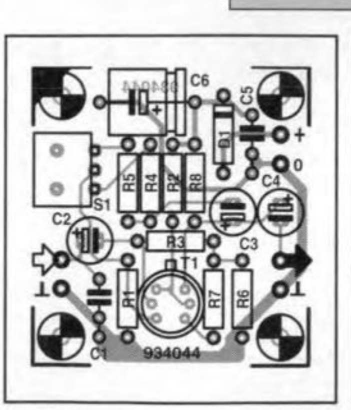
Semiconductors:

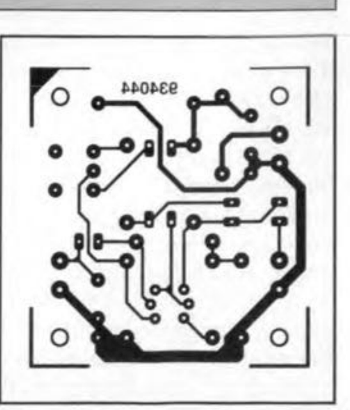
1	1N4001	D1
1	MAT02 (Precision Monolithics Inc., Analog Devices)	T1

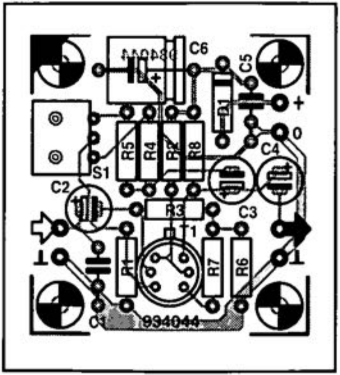
Miscellaneous:

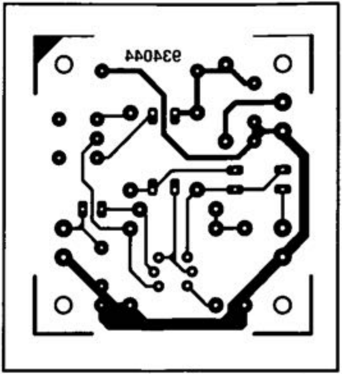
1	PCB mount change-over switch, Fujisoku AS1D-5M	S1








C6
C5
+
0
D1
R5
R4
R2
R8
C4
S1
C2
R3
C3
T1
R1
R7
R6
C1
934044



SPEED CHESS CLOCK

In speed chess a player has to make his move in 5–10 seconds. If he/she does not, he/she loses the game (depending on the agreed conventions). The 'clock' described here indicates whose turn it is and who has defaulted. It may, of course, also be used for other speed games.

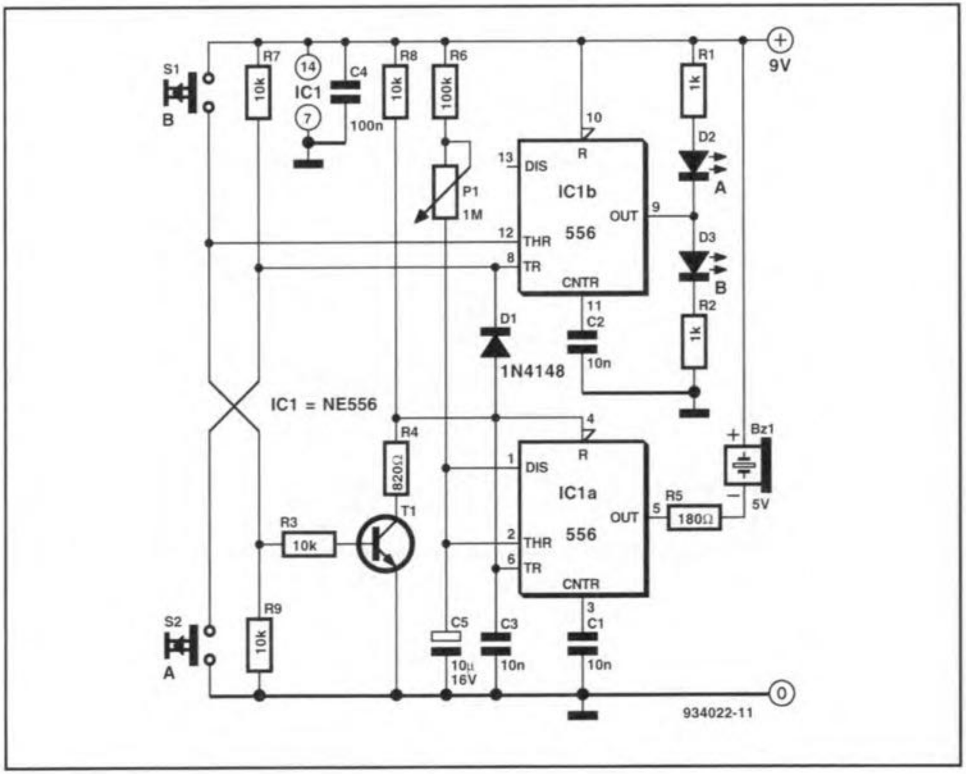
When player A presses S_2, D_3 lights indicating that it is player B's turn. This player returns the game to A by pressing S_1, whereupon D_2 lights. If a player fails to return the game to his/her opponent in the time set with P_1, the buzzer will sound.

The circuit is built around a 556, the dual version of the well-known 555. One half, IC_{1b}, functions as a bistable (flip-flop) that 'remembers' which switch was pressed last: this is indicated by the associated LED lighting. The other half, IC_{1a}, operates as a monostable. It is kept in its triggered state (buzzer inactive) as long as it is retriggered regularly, and timely, by switches S_1 and S_2.

The trigger input is made low by S_1 with the aid of T_1, and by S_2 with the aid of D_1. Every time IC_{1a} is retriggered, C_5 is discharged via pin 1 (DIS) of IC_{1a}. The time set with P_1 then starts afresh. If neither of the switches is pressed in time, the voltage across C_5 becomes so high that IC_{1a} is reset. Its output is then low and the buzzer sounds.

The circuit draws a current of about 20 mA (with one of the LEDs lighting). When the buzzer sounds, the current rises to around 40 mA.

[C.R. Suthikshn Kumar - 934022]



DC-DC CONVERTER

The converter raises a direct voltage to almost twice its level and is, therefore, particularly suited to increase the output of solar cells to the level required for charging lead-acid or NiCd batteries. It can deliver a current of up to 3 A. Measurements with a load current of 2 A are given in the table. The open-circuit output voltage is about 1–1.5 V higher.

In the following description it is assumed that the input voltage to the circuit is 12 V and the output voltage 22 V. IC_{1a}, R_2 and C_5 form a rectangular-voltage generator. This signal is also available in inverted form at the output of IC_{1d}. Network $R_2.C_6$ delays the output of IC_{1a}, so that the output of NAND gate IC_{1b} has a duty factor >0.5 (the negative half is shorter than the positive half). The same is true of the output of NAND gate IC_{1c}. (The input signal to this gate is delayed by R_5-C_7).

The output of IC_{1c} is inverted and buffered three times: in IC_{3f}, IC_{3a}, and the four parallelled gates IC_{3b}–IC_{3e}. It is then used to drive power FET T_3.

The output of IC_{1b} drives small-signal transistor T_1. When this transistor is on, junction R_6-R_7 is pulled to 2 V without diode D_1. However, IC_{2a} needs an input signal of 11–22 V, since the supply voltage for this inverter (and, of course, also for inverters IC_{2b}–IC_{2e}), as well as the collector voltage of T_1 is already derived from the doubled output voltage. The negative supply voltage for this IC is therefore derived from the positive input voltage. Diode D_1 ensures that the potential at the input of IC_{2a} does not drop below 10.5 V.

Transistors T_2 and T_3 conduct alternately. When T_2 is on, C_{10} is charged to the level of the input signal via T_3 and D_3. When T_2 is off and T_3 is on, C_9 is charged similarly. Capacitor C_{10} retains its charge, since D_3 prevents its discharging. Since the two capacitors are in series, the output voltage is twice the input potential.

Owing to the multiple inversion of both signals following the delay networks, it is impossible for T_2 and T_3 to be switched on simultaneously.

Input and output voltages with a load current of 2 A.

U_{in}	U_{out}
10 V	18 V
12 V	22 V
14 V	26.4 V
15 V	28.3 V
16 V	30 V

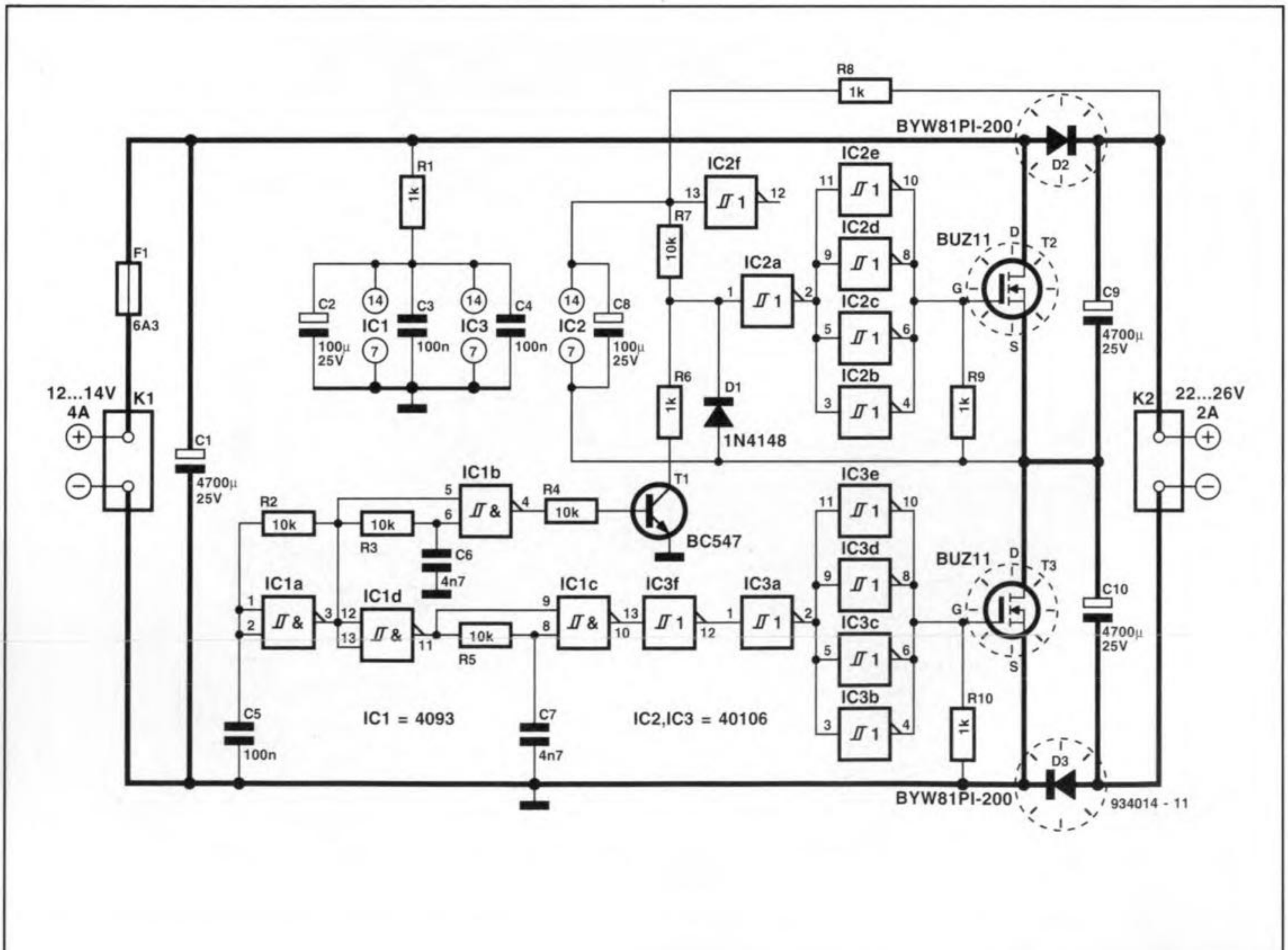
Capacitor C_1 buffers the input signal, so that its loading is constant in spite of the varying current drawn by the circuit.

It is essential that D_2, D_3, T_2 and T_3 are well cooled. It is best to mount these components on a common heat sink.

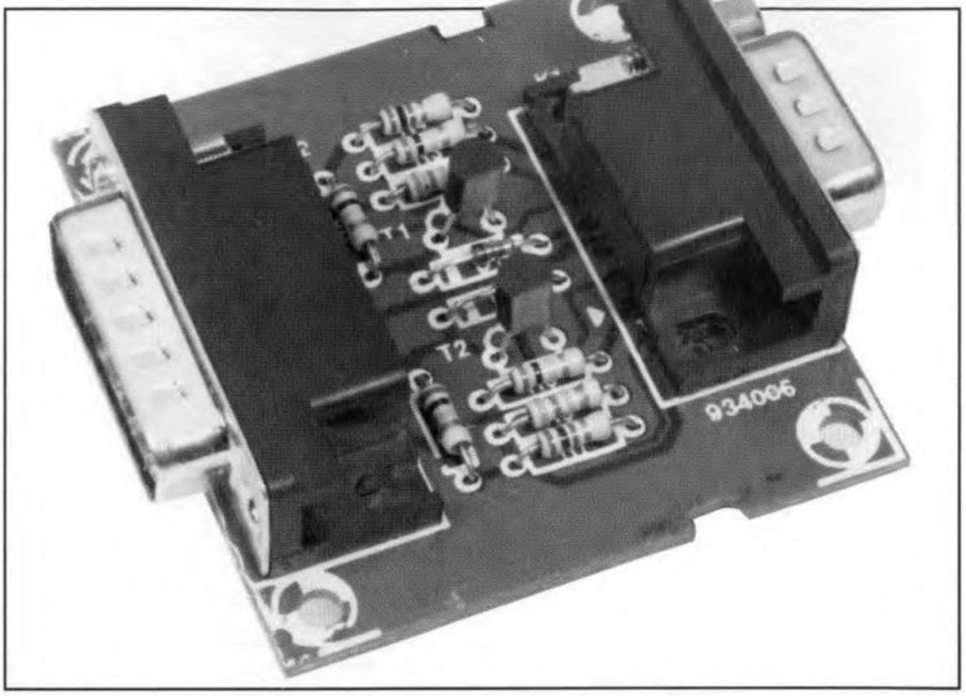
The bold lines in the circuit diagram represent heavy-duty wires that should be as short as possible, since they carry a current of 6 A.

A carefully constructed convertor should give an efficiency of 94% (at 22.2 V and 1.8 A).

[U. Kardel – 934014]



JOYSTICK CONVERTER FOR PCs



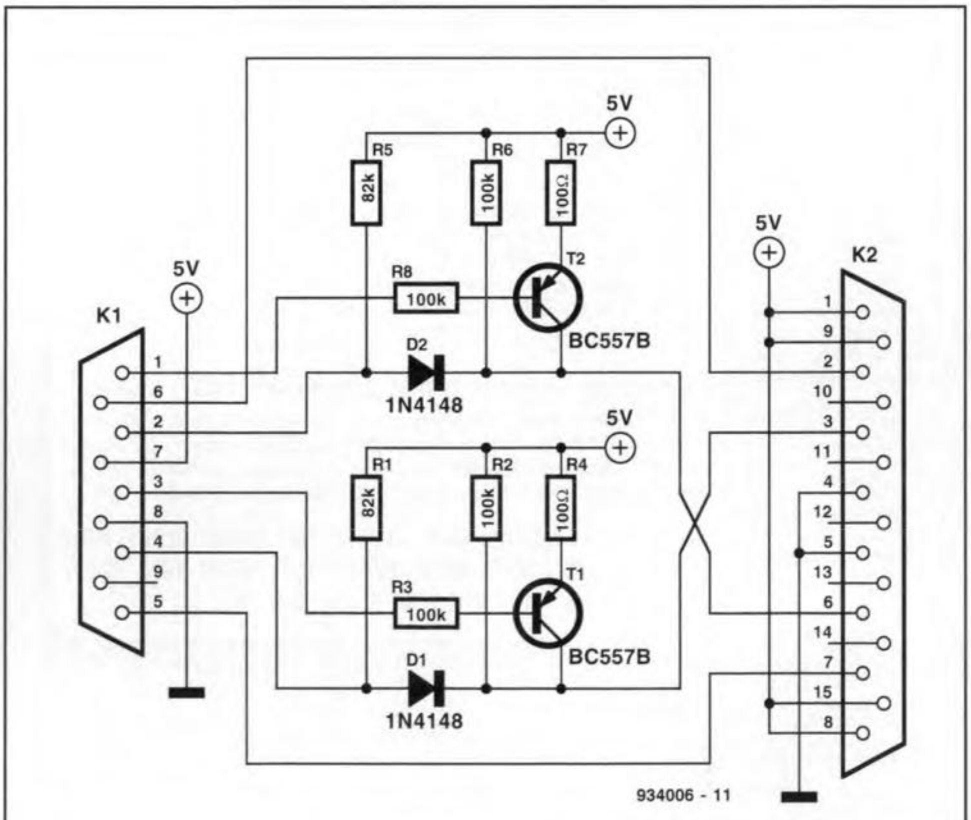
ALTHOUGH frowned upon by many as addictive and devoid of educational value, computer games can be great fun to play. This little circuit allows you to connect a digital (switch-based) joystick to the analogue inputs on the Game Control Adapter (GCA) which is available on most IBM PC/ATs and compatibles.

A normal joystick interface consists of two inputs, each of which is connected to the two potentiometers in the joystick. Traditionally, the resistance formed by each of these potentiometers is used to charge a capacitor. When the capacitor voltage has reached a certain level, a bistable toggles. This means that the charging time is a function of the potentiometer position. The output state of the four bistables and the four digital inputs is a bit pattern found at address 201H. A software counter then tells the system how long it takes before a bistable bit toggles.

This interface works fine in a closed, unchanging system. Unfortunately, the immense increase in processor speeds that has come about in recent years has caused compatibility problems with regard to the use of the analogue joystick inputs. The reason is simple: the same joystick read routine gives different results on PCs with different clocks speeds. The effect of clock speed differences has been reduced to some extent by making the timing variable with the aid of a switch. In this way, the program still reads the desired value (well, roughly ...).

Most analogue joysticks have a centre position adjustment, which, unfortunately, can be turned by accident in the excitement of a game. Consequently, the program (game) will not respond properly to your signals any more, so that you will be unable to reach the highest score. Fortunately, many programs have a fairly wide margin around the centre position, so that a small misadjustment need not be fatal. This type of software does not follow the above concept of reading the joystick position and turning it into a proportional value. Instead, it just reads left, right, up and down, and no proportional value at all. This circumstance is exploited by the present design.

Arguably, if proportional movement translation is not used, there is no rea-



COMPONENTS LIST

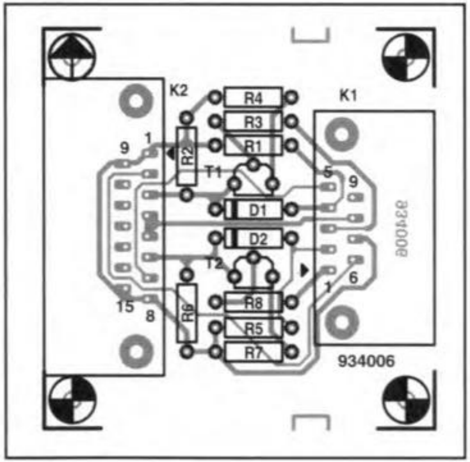
Resistors:
2 82kΩ R1;R5
4 100kΩ R2;R3;R6;R8
2 100Ω R4;R7

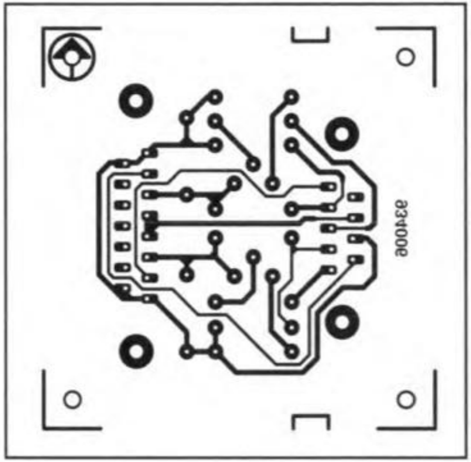
Semiconductors:
2 1N4148 D1;D2
2 BC557B T1;T2

Miscellaneous:
1 9-way male sub-D connector, PCB mount, angled pins K1
1 15-way male sub-D connector, PCB mount, angled pins K2

son why the potentiometers in the joystick could not be dropped altogether and replaced by simple switches, while still using the analogue input on the PC. Fortunately, switch-based joysticks exist — they also go under name 'digital joystick'.

With reference to the circuit diagram, the joystick is in the centre position if it is not operated, and none of the four switches is closed. This means that the PC 'sees' the resistance formed by $R_2//(R_1+D_1)$. The value of R_1 is calculated such that the GCA card 'sees' a joystick resistance of about 50 kΩ, in spite of the diode. If the lever is moved to the right, pin 4 on K_1 is pulled to ground, so that D_1 blocks and the computer 'sees' R_2 (100 kΩ) only. If the lever is moved to the left, pin 3 on K_1 will go low. This causes T_1 to be switched on via R_3, and, consequently, the GCA card to 'see' a joystick resistance of about 100 Ω. In this way, the input signals are converted into resistance values 100 Ω, 50 kΩ or 100 kΩ. The same applies to the up/down section built around T_2.

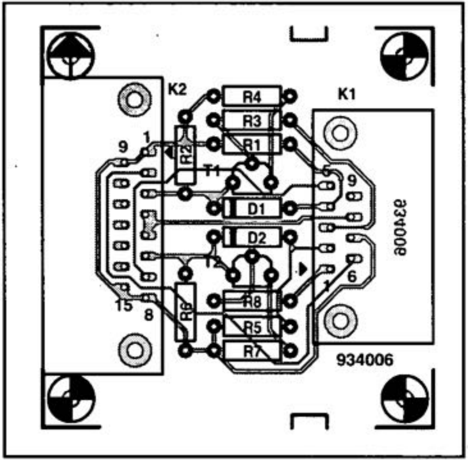


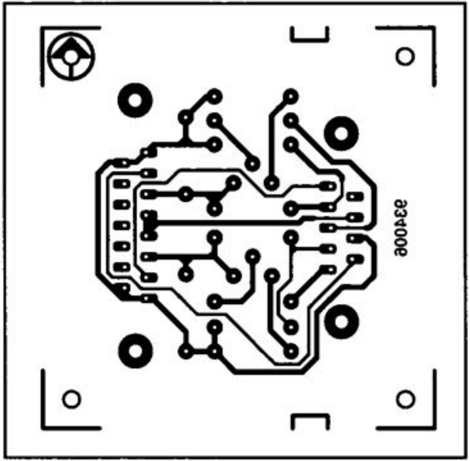


The 'fire' button is read directly via pin 6 on K_1. If your joystick has a second button, this is read also via pin 5 on K_1. Also note the presence of the +5 V supply voltage on pins 8 and 15 of K_2 (computer's side) — this may be used for inductive joysticks and/or types with an 'auto-fire' button.

Finally, R_1 (R_5) may be replaced by a series combination of a 68-kΩ resistor and a 50-kΩ preset. This allows you to adjust the joystick's centre position, and play 'older' games that lack an up-to-date initialization routine.

(R. Zandbergen — 934006)


K2
K1
R4
R3
R1
R2
T1
D1
D2
T2
R8
R5
R7
R6
9
1
15
8
5
9
6
1
934006

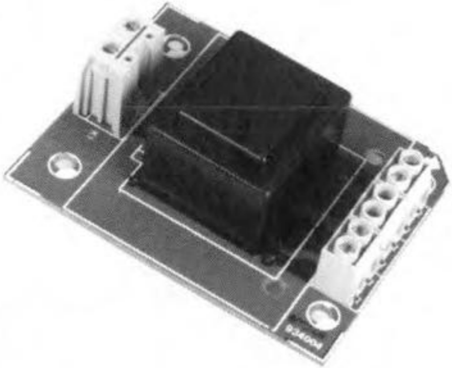


GENERAL TRANSFORMER PCB

The board may be used either as shown in Fig. 1 or as in Fig. 2 It may be used with transformers rated from 1.5 VA to 12 VA. When used with a very small transformer, it is possible to cut off the part marked K2–K4. Place the terminal strips in the same position along the new edge as they were originally.

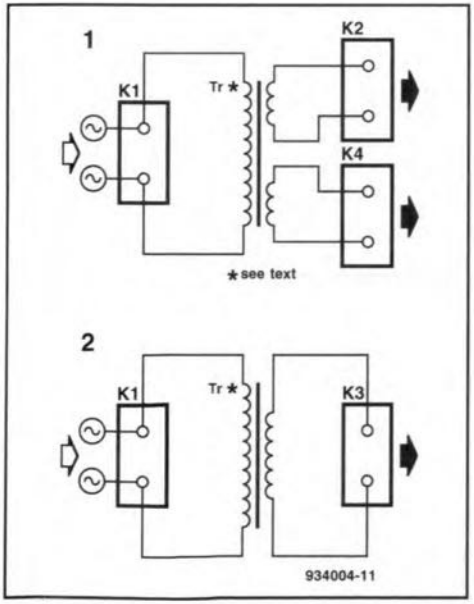
When non-short-circuit-proof transformers are used, a fuse in the mains entry plug is essential.

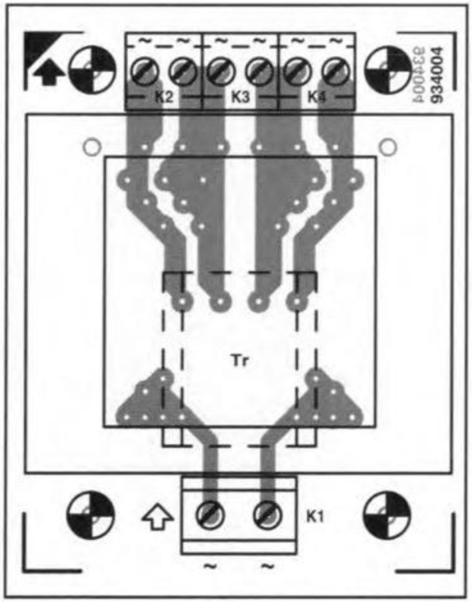
[L. Pijpers - 934004]

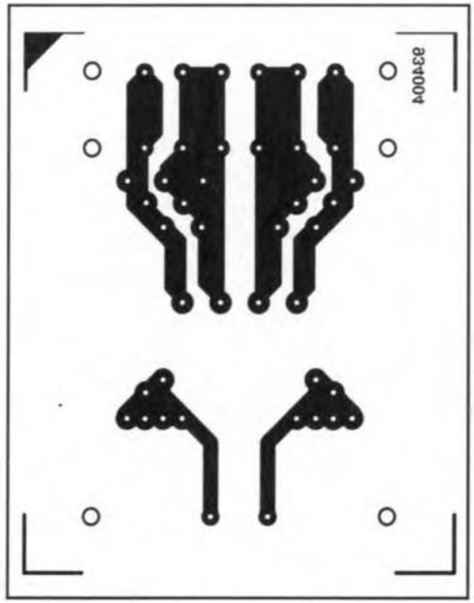


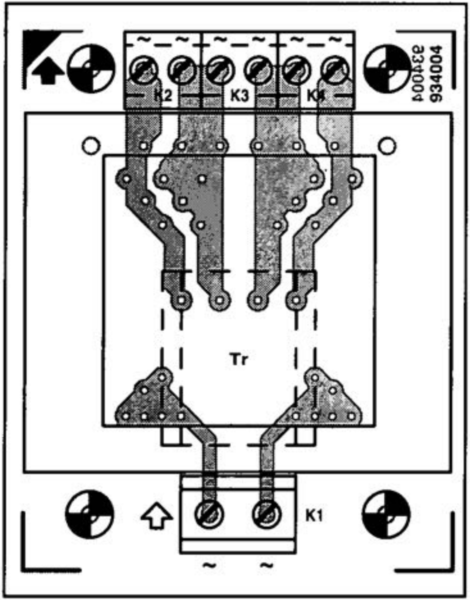
PARTS LIST

K1=two-way terminal block, pitch 7.5 mm
K2, K3, K4=two-way terminal block, pitch 5 mm
Tr=transformer for PCB mounting

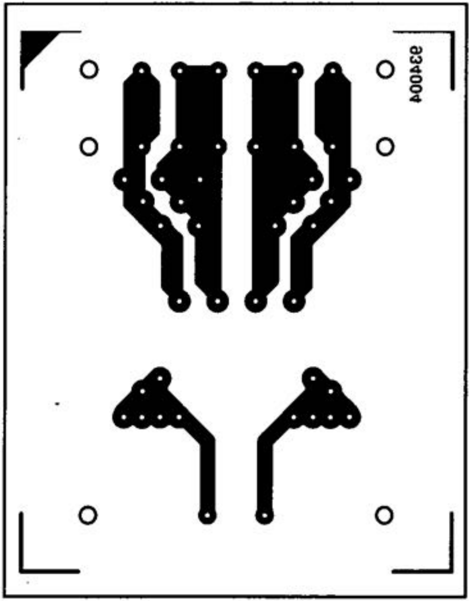








K2
K3
K4
934004
Tr
K1



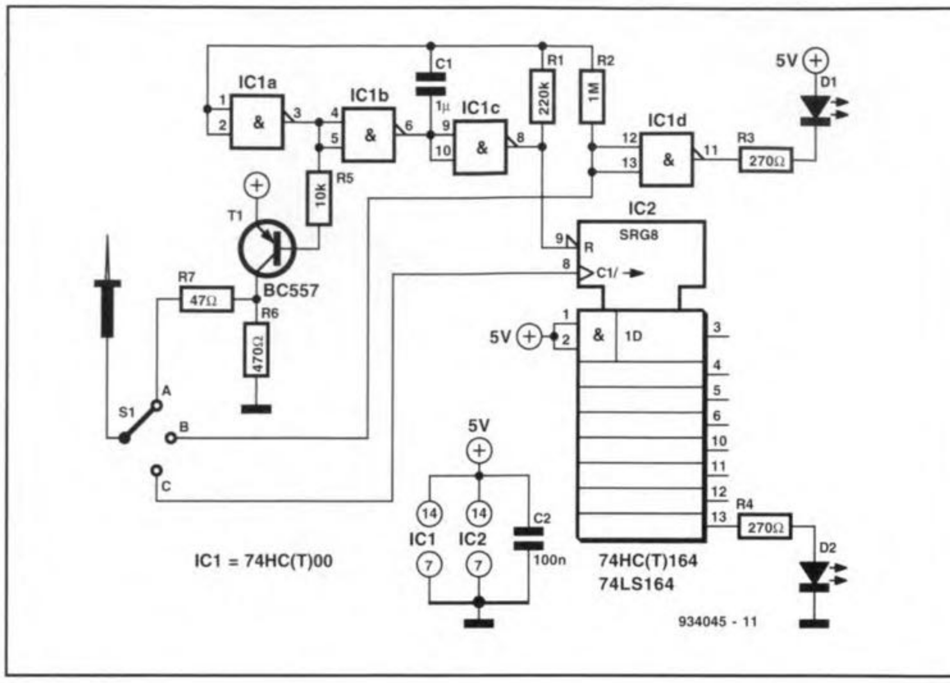
MULTI-FUNCTION TEST PROBE

Based on only two ICs and a transistor, this circuit can perform three test functions selected by switch S_1.

The circuit is based on a clock generator, IC_{1a-1c}, that provides a rectangular signal at a frequency of 3 Hz.

With S_1 in position A, T_1 applies the clock signal to the probe. This buffer stage can deliver a current of up to 100 mA via limiting resistor R_7. In practice, this means that the probe can inject the 3 Hz signal into any digital circuit, thereby overruling all logic outputs. Only short circuits can cause the signal to disappear and the probe can, therefore, quickly detect them. This is, however, a rough and ready test, because it means that the tester short-circuits all outputs connected to the track that is being examined, whereas manufacturers allow this to happen to only one output at a time.

With S_1 in position B, the tester can detect logic levels. When the probe is not connected to anything, the clock is present at the input of IC_{1d}, so that D_1 flickers in a 3 Hz rhythm. When a logic level is applied to the probe, it suppresses the clock; if the level is low, the LED goes out; if the level is high, the LED lights continuously.



The switching thresholds of IC_1 lie at about $^1/_3$ and $^2/_3$ of the supply voltage, which is suitable for the CMOS 4000 series and the HCT family. Standard TTL, LS, and HCT ICs may be used, but the thresholds are then not optimum.

The tester can work with input signals at frequencies of up to 30–40 MHz.

With S_1 in position C, the input signal is applied to the clock input of shift register IC_2. After eight clock pulses at the probe, the high level that is constantly present at inputs A and B of IC_2 will have arrived at the Q_H output, so that D_2 lights. Since IC_2 is constantly reset by the 3 Hz signal, the LED will flicker as long as there is a clock at the probe, provided that its frequency is higher than 25 MHz.

The circuit draws a current of 10 mA when both LEDs flicker. To ensure that the pulse generator works satisfactorily, the power supply should be able to provide a current of some 100 mA.

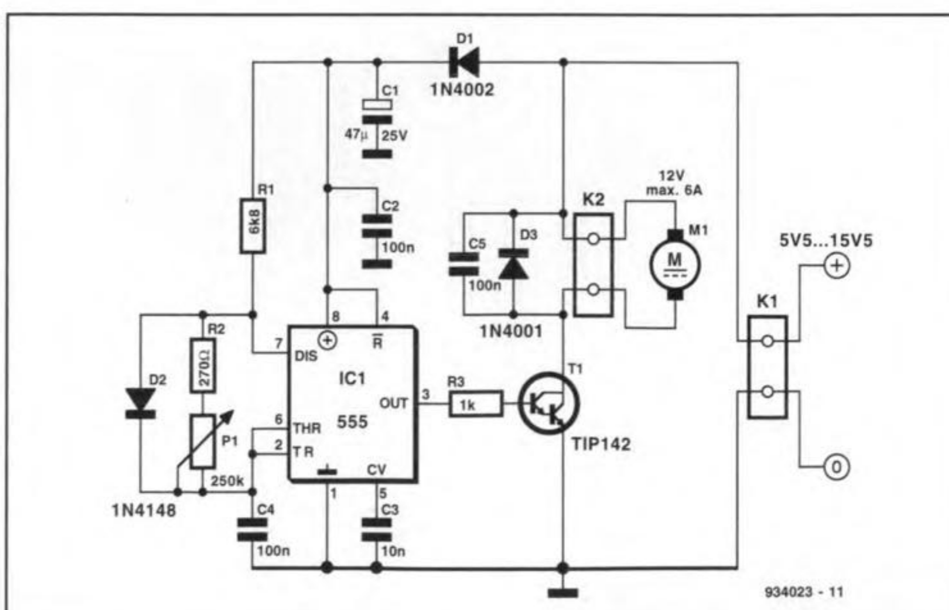
[S. Mitra - 934045]

DC-MOTOR REGULATOR

The regulator shown in the diagram enables the speed of a DC motor to be set over a wide range. Although the most frequently used method for achieving this is pulse-width modulation, in the present design pulse position modulation (also called pulse time modulation) is used. That is, rectangular modulator IC_1 always switches the motor on for 0.5 ms ($=R_1C_4$). For most motors that is sufficient to start. The speed at which it will finally rotate is influenced by the pulse rate. The pulse spacing, which is the discharge time of C_4, may be set between 1 µs and 14 ms with R_2 and P_1.

T_1 can switch up to 6 A, provided it is fitted on a heat sink of not less than 20 K W^{-1}.

[L. Pijpers - 934023]



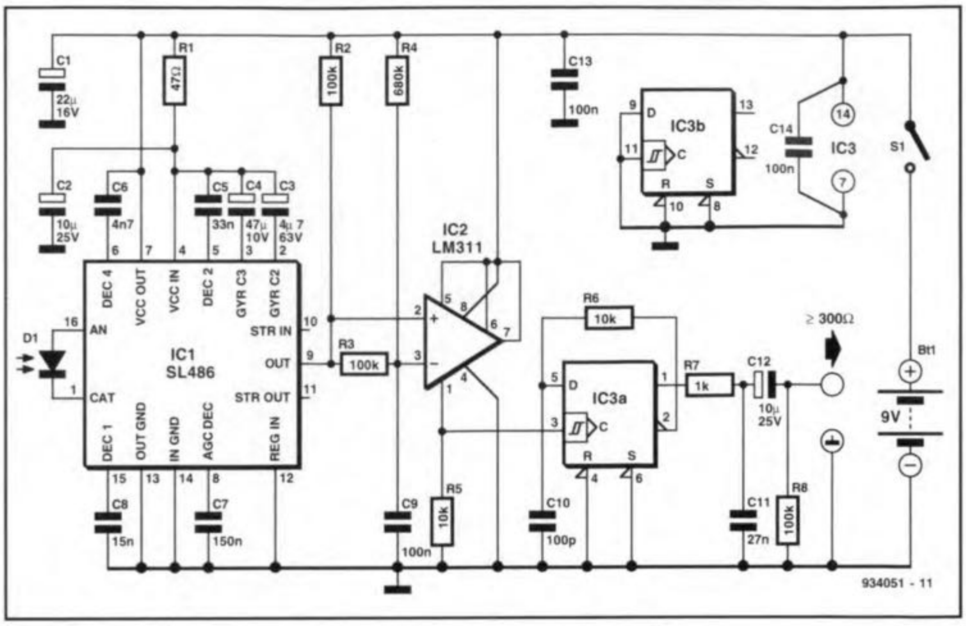
INFRA-RED A.F. RECEIVER

This audio-frequency receiver is the counterpart of the infra-red transmitter. The output from the infra-red diode D_1 is amplified by IC_1. The diode is a Type BPW41N which works well with the transmitter and is fairly fast (≈200 ns). The output pulses of IC_1 are taken to comparator IC_2, which separates them from the ambient light. To that end the d.c. component of the output of IC_1 is stored in C_9 and used as reference for IC_2.

The received signal is decoded by a simple binary scaler, IC_{3a}. Each pulse results in a change of level at the output of this stage, so that the output resembles the original pulse-duration-modulated (PDM) signal. The output is filtered by R_7-C_{11}, decoupled by C_{12}, resulting in an audio signal that is suitable for driving high-impedance headphones.

The binary scaler stage contains a delay network, R_6-C_{10}, which prevents the scaler being clocked by ambient noise. Owing to this network, the previous level is retained at the D input for a microsecond during which the state of the clock input is immaterial.

The output signal may be enhanced by an active 3rd or higher order low-pass (10–12 kHz) filter, which may also serve as a buffer



In combination with the infra-red transmitter, and ignoring the residual high-frequency components, the audio signal is virtually undistorted. With the receiver at a distance of 1 m (3.3 ft) from the transmitter and 30% modulation, the signal-to-noise ratio is about –50 dB. The maximum distance the prototypes worked without interference was about 5 m (16 ft).

The circuit draws a current of some 15 mA, which makes it advisable to use a rechargeable (NiCd) battery.

[T. Giesberts - 934051]

INFRA-RED A.F. TRANSMITTER

It often happens that someone wants to watch television without the sound disturbing other people in the room.. To avoid having a long cable trailing over the floor, the present transmitter (and accompanying receiver) make it possible to listen to the TV sound 'wirelessly'.

The transmitter uses pulse-duration (=pulse-width) modulation. The modulated signal is produced in the traditional manner of having the audio signal set against a pure high-frequency triangular signal in a comparator. A suitable triangular signal generator can be found on page 55. If another generator is used, make sure that its off-set is equal to half the supply voltage of 5 V and its peak value is 2.5 v_{pp}.

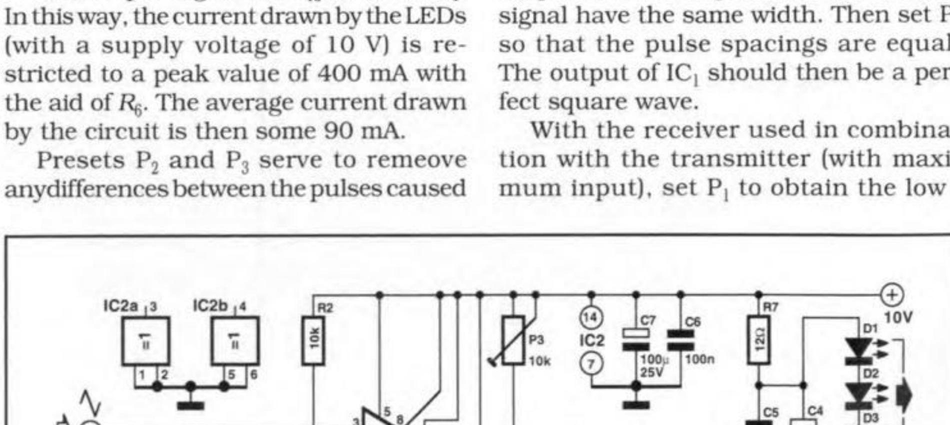
For a good range, the current through the LEDs must be large. Since, however, LEDs should not draw large continuous currents, the pulses must be short, and this is why PDM is used. (In this type of modulation, the time of occurrence of the first and last transition, that is, the leading and the trailing edge, is varied from its unmodulated position).

The pulses are generated by XOR gate IC_{2d}, which compares the original PDM signal with one delayed by R_5-C_3-IC_2. This results, for every level change, that is transition, into a short pulse whose width is equal to time constant R_5-C_3.

The output signal of IC_{1d} switches T_1. In this way, the current drawn by the LEDs (with a supply voltage of 10 V) is restricted to a peak value of 400 mA with the aid of R_6. The average current drawn by the circuit is then some 90 mA.

Presets P_2 and P_3 serve to remeove anydifferences between the pulses caused by a first transition (leading edge) and those resulting from a last transition (trailing edge). In the absence of an input signal, set P_3, with the aid of an oscilloscope, so that all pulses in the output signal have the same width. Then set P_2 so that the pulse spacings are equal. The output of IC_1 should then be a perfect square wave.

With the receiver used in combination with the transmitter (with maximum input), set P_1 to obtain the low



est possible interference on the received signal

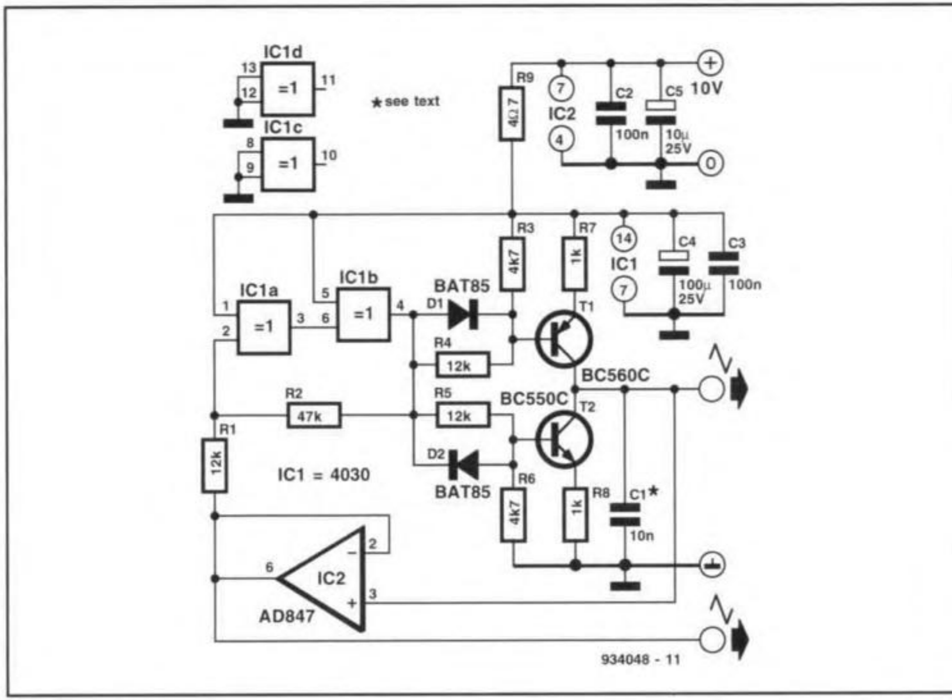
Finally, note that IC_{2a} and IC_{2b} may be used for the triangular wave generator.

[T. Giesberts - 934052]

TRIANGULAR-SIGNAL GENERATOR

Pulse duration modulators require a perfect triangular signal, that is, one that is absolutely symmetrical with straight sides. Any deviation from this results in distortion.

The design is the usual rectangular-triangular oscillator. A Schmitt trigger converts a triangular wave into a rectangular one, which is used to continuously charge and discharge a capacitor via current sources. Positive current source T_1 arranges the charging and negative current source T_2 the discharging.



The Schmitt trigger is formed by EXOR gates IC_{1a} and IC_{1b}, which are connected as inverters. The remaining gates in the IC are not used here, but are available, for example, for a pulse duration modulator. Resistors R_1 and R_2 determine the hysteresis of the convertor.

The output of the trigger switches on T_1 and T_2 alternately. If the output of IC_{1b} is high, T_2 conducts; if it is low, T_1 is switched on. Resistors R_3 and R_4 ensure that the base voltage of T_1, and thus the potential across emitter resistor R_7, is constant if the output of IC_{1b} is low. This ensures that C_1 is charged via T_1 with a constant current. Diode D_1 arranges a rapid switching off of T_1. Transistor T_2 operates similarly, but when the output of IC_{1b} is high.

The amplitude of the triangular signal must not be greater than the base voltage of T_1 and T_2. Assuming a supply voltage of 10 V, this means that the amplitude of the triangular signal is 2.5 V_{pp}.

Opamp IC_2 is a good-quality, fast type (200 V μs^{-1}) that functions as a buffer between the output and the input of the Schmitt trigger. Since it is a good-quality type, it has virtually no effect on the shape and quality of the triangular signal. If required, the symmetry of the triangular signal may be corrected to a small degree by connecting a low-value (1 kΩ) preset in series with R_3 or R_6. The value of the chosen resistor must be reduced by half the value of the preset.

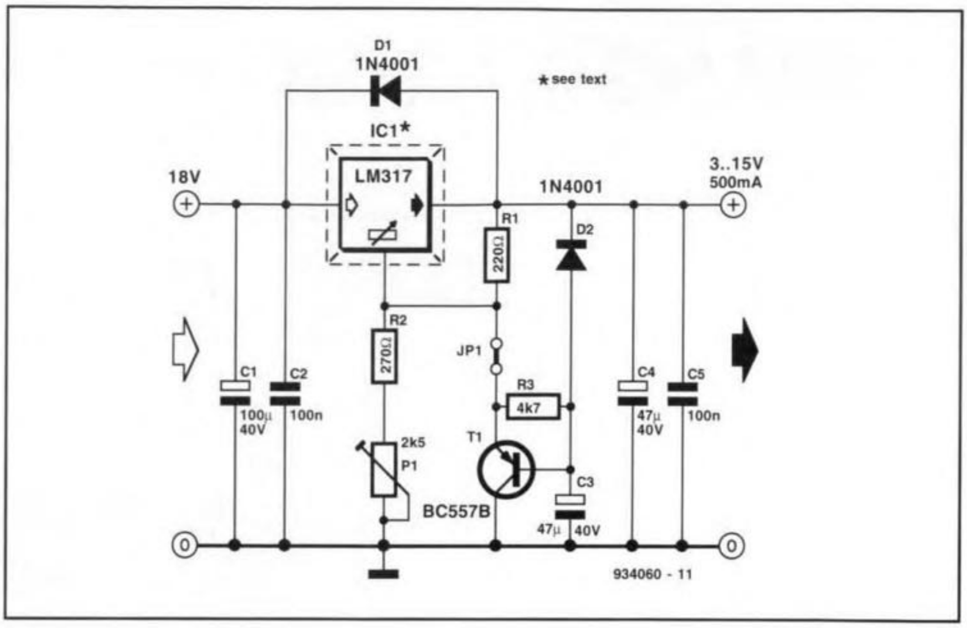
The circuit can generate signals at frequencies up to 300 kHz. With values as shown, the output frequency is 38 kHz, which is required by the pulse duration modulator elsehwere in this issue. The frequency may be altered by changing the value of C_1.

The circuit draws a current of about 8 mA, of which 5 mA are on account of IC_2.

[T. Giesberts - 934048]

REGULATOR OUTPUT DELAY

In this circuit, a Type 317 variable voltage regulator is provided with a facility whereby, after the input voltage has been switched on, its output voltage rises only slowly to the value set with P_1. For instance, if P_1 was set for an output of 15 V, that value is reached only after 5 seconds. If the output voltage was set at 7.5 V, that value would have been reached after 2.5 second. In other words, the delay time is directly proportional to the set output voltage. The delay may be lengthened by increasing the value of R_3 and/or C_3. The delay may be disabled by removing the link at JP_1 (which is useful when P_1 is being adjusted).



Diode D_2 protects the base-emitter junction of T_1 against too high a reverse bias. Without the diode, if the output of the regulator were short-circuited, charged capacitor C_3 could cause a U_{BE} that exceeds the permissible maximum 6 V.

Diode D_1 protects the regulator if the input voltage were (accidentally) to drop faster than the output voltage.

The circuit draws a current of only a few mA, although the peak current may amount to 1.5 A. To prevent the temperature protection of the LM317, mounted on a 14 K W^{-1} heat sink, coming into operation, the constant output current should not exceed 500 mA.

[J. Ruiters - 934060]

FAST HIGH-POWER 'ZENER DIODE'

The aim of the design of this 'zener diode' is to obtain a fast voltage regulator that can provide high currents. The circuit is based on a Type BUZ10 SIPMOS transistor, which has a high dissipation and can cope with high currents.

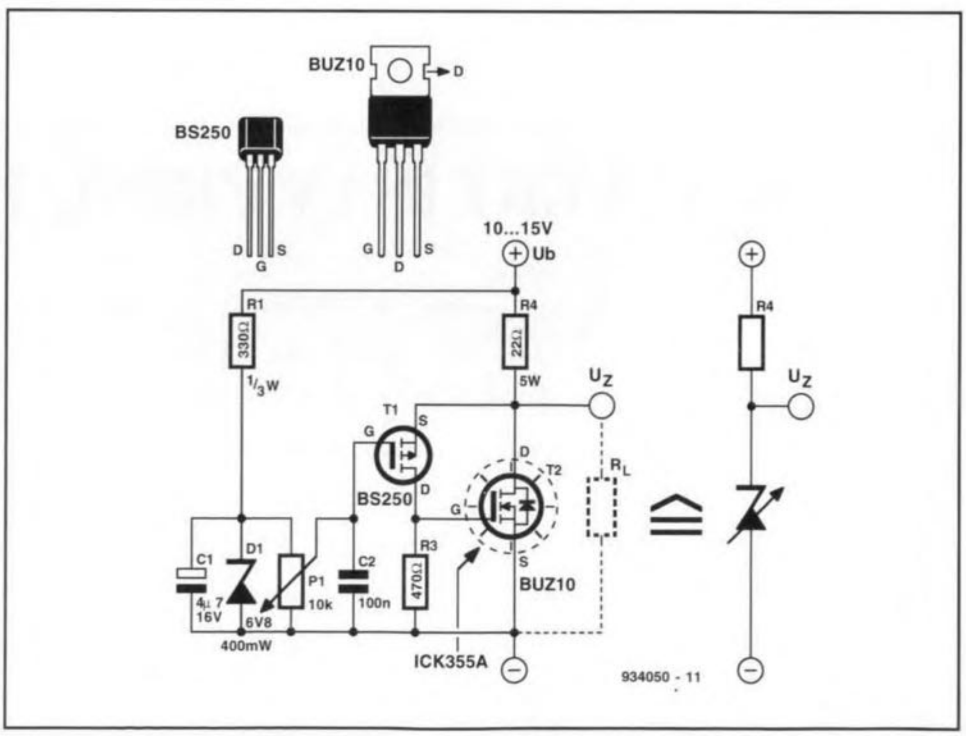
From an external power source, whose voltage must be about 2 V higher than the wanted stabilized potential, P_1-C_1-D_1 derive a stable voltage in the range 0–6.8 V. This voltage switches on T_1, which in turn switches on T_2. If U_z drops slightly, the potential at the gate of T_1 also drops, whereupon T_2 raises its on-resistance, so that the output voltage rises. The control speed is determined by R_4.

The circuit as drawn delivers output voltages in the range 3.9–9.6 V at a supply voltage of 15 V.

In the prototype an open-circuit voltage of 6 V was set (the current drawn by T_2 was 490 mA), which dropped to 5.94 V when a load of 15 Ω (minimum value) was connected across T_2 as shown in dashed lines. The drain current was then 13 mA.

The dynamic behaviour is excellent: with a rectangular input signal at a frequency of 10 kHz, and a load of 15 Ω, the recovery time was smaller than 1 µs.

[B.C. Zschocke - 934050]



MATRIX KEYPAD ENCODER

Matrix keypads may be readily purchased by the home constructor from a large number of sources, and they are usually competitively priced. However, unlike the generally more expensive non-encoded types, they require a decoder IC, which may work out almost as expensive as the keypad itself. A low-cost alternative is presented here.

By C.J. Bearman

HAVING recently purchased a miniature 16-way keypad for a computer project I was designing, I had the option of either obtaining a decoder chip (such as the MC14419) or building my own out of general-purpose integrated circuits. One of the problems I immediately noticed while perusing the specification of the Motorola IC was the fact that it seemed to deliver a strobe pulse only on the ten numeric keys, whilst my circuit was going to require a strobe from all sixteen.

The circuit I have come up with uses five readily available ICs and a small handful of other components. Although designed to work into a microprocessor data bus, it may readily be adapted in a number of ways to suit the use of any experimenter. Likewise, it will easily adapt for use with a twelve or nine key pad, by simply omitting the appropriate decoder lines.

Circuit description

The matrix connections for a typical keypad are evident from the circuit diagram, Fig. 1. A free running 2.5-kHz oscillator is made from a gate in IC1, and feeds into the count (CNT) input of IC2. The first two counter outputs are connected to decoder IC3a to select the column signals on pins 4 to 7. The row connections are taken to the four AND gates in IC4 via pull-down resistors. The second decoder, IC3b, splits the other two outputs from the counter into four lines to feed the remaining inputs of IC4.

When a key is pressed, the corresponding line from decoder IC3a is routed to one of the gates in IC4. Within a few milliseconds the other input of that gate will be addressed by the scan of IC3b, causing its output to go high. The four diodes at the IC4 gate outputs create a wired-OR gate which will supply a leading edge to the latch input of IC5 via two gates in IC1. As the inputs from the latch are being fed with all four outputs of the counter, the leading edge on pin 11 will load the binary number associated with the particular key being pressed. As soon as the decoders pass on to the next key, the output from the wired-OR will, of course, go low. However, because of the charge stored in capacitor C2, the voltage on input pins 12 and 13 of IC1 will remain high, so avoiding any further toggling of the data available line until the key is released.

The four output pins from IC5 will present a binary code between '0' (when the top left-hand key is pressed) and '15' (when the bottom right-hand one is pressed). If it is desired to decode these into sixteen lines, all that has to be added to the circuit is a 16-line decoder using the appropriate technology; i.e., a CMOS 4515, a TTL 74154, or similar.

Current consumption in the quiescent state was measured on the prototype at 44 µA with a +5-V supply, rising to almost double this value when a key was pressed. ■

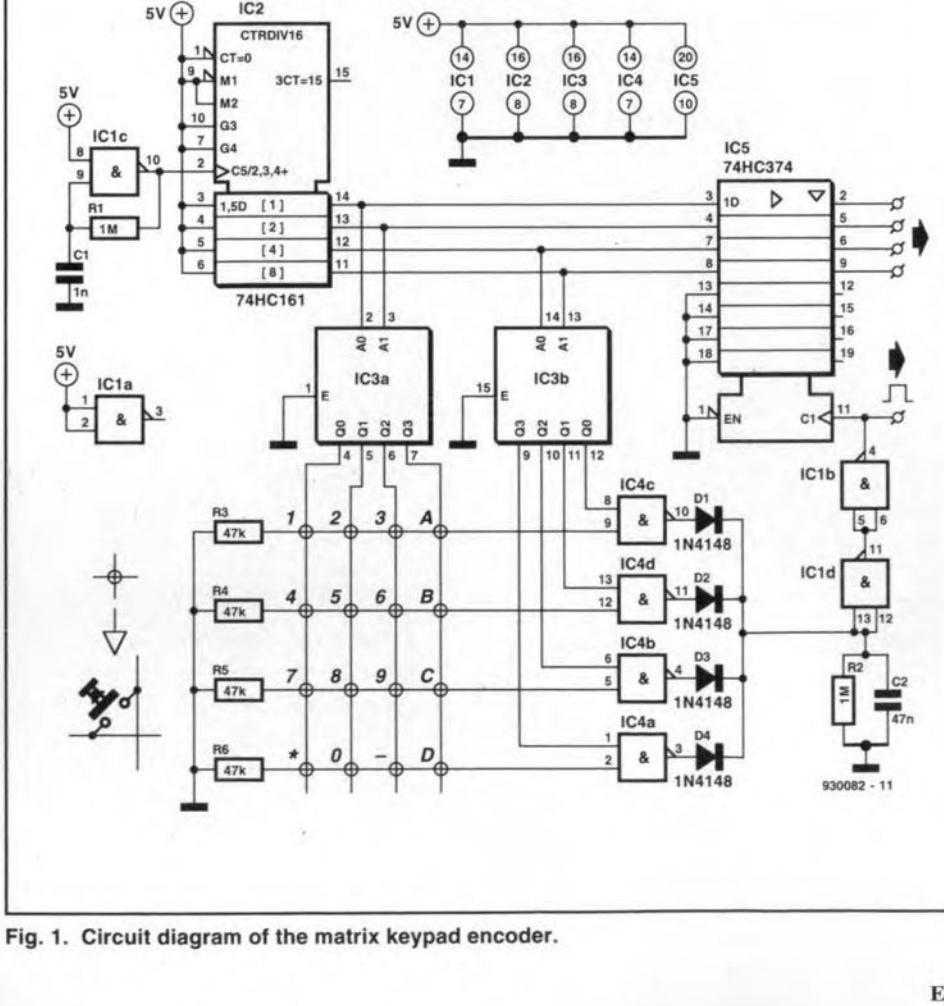


Fig. 1. Circuit diagram of the matrix keypad encoder.

CLASS A PREAMPLIFIER

Nowadays, the design of preamplifiers is almost invariably based on integrated circuits. However, there are still many constructors who like a discrete design and they may like the one described here.

The design of the amplifier is symmetrical. In the differential input stages dual transistors T_1 and T_2 are used. The differences in current amplification between these make bias correction at the input stage essential and this is provided by T_{12}. The required reference voltage source is provided by D_2. Preset P_2 sets the output voltage to 0 V. The only drawback of the correction stage is that an input coupling capacitor, C_1, becomes essential. However, with the value of this capacitor as shown, the cut-off frequency is only 1.5 Hz, so in practice this no real problem.

The input stages drive a push-pull amplifier that consists of T_5 and T_6. To increase the stability of this amplifier, its gain is limited by R_{18}-R_{19}.

The output stage is a conventional emitter follower, T_9 and T_{10}. A current source, consisting of T_7 and T_8, ensures constancy of the quiescent current through the emitter follower. For best performance, the plane sides of T_7 and T_9, and of T_8 and T_{10}, should be joined together after some heat conducting paste has been applied to them. A brass clamp to make the mechanical coupling secure is advisable, but not absolutely essential.

The quiescent current through T_9 and T_{10} is set to 20 mA with P_1. The value of the current may be measured on the basis of the voltage drop across R_{22} and R_{23}.

Capacitor C_2 serves to improve the response of the stage and to suppress overshoot. The effective bandwidth is thus reduced to 2.4 MHz, which is more than ample for audio applications.

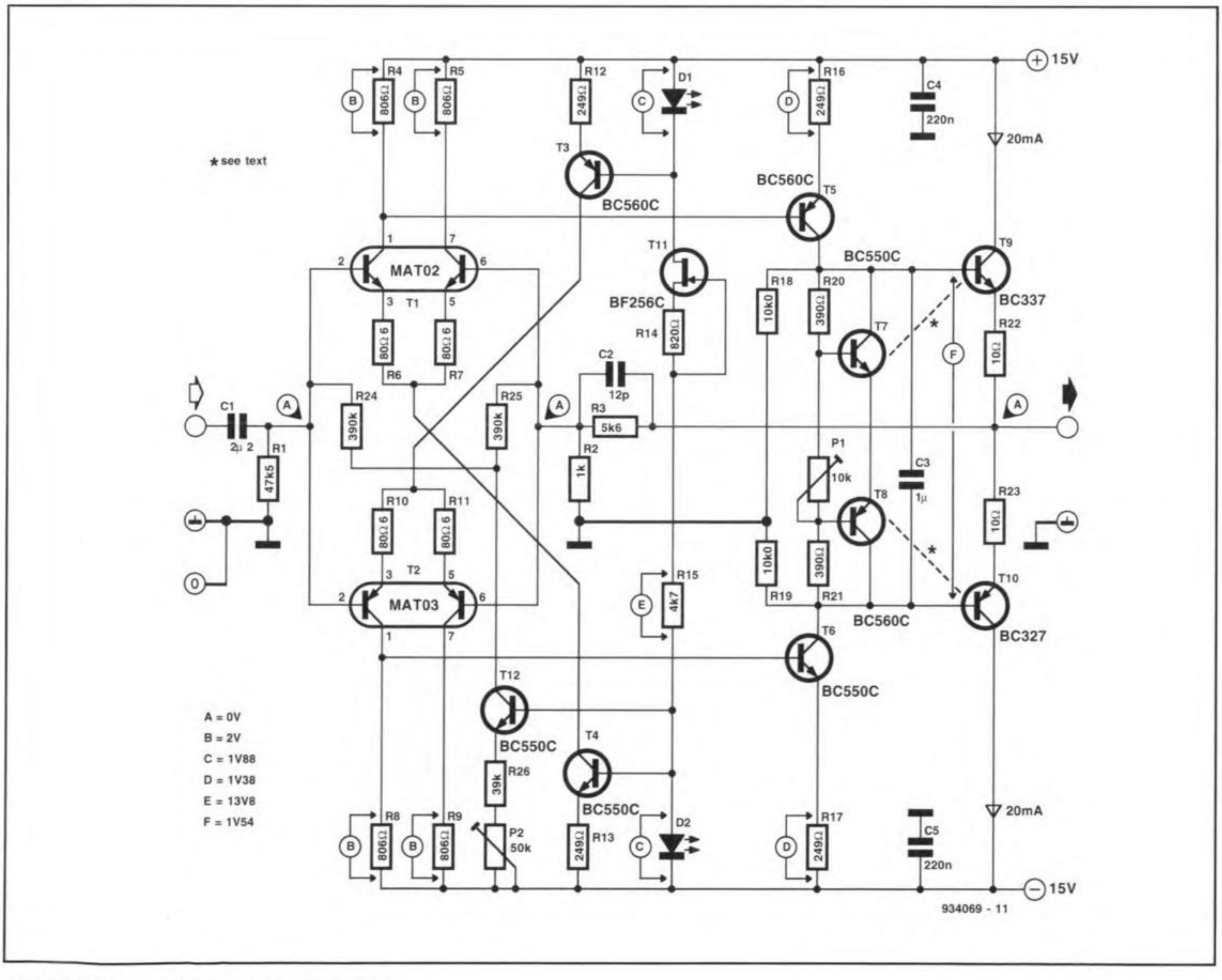
For optimum performance, all transistors operate in Class A, which results in the relatively large quiescent current of 40 mA.

To ensure correct feedback, the earths of input and output, R_1, R_2, R_{18}, and R_{19}, as well as the negative supply line must be taken to one earthing point only.

Some measuring results (with U_b=±15 V, 1 V r.m.s. output and a 1 Ω load):

Gain: 16 dB
Input sensitivity 150 mV)
Slew rate 200 V μs^{-1}
Signal-to-noise ratio: 100 dB (unweighted)
Third harmonic distortion <0.00025%
(20 Hz–20kHz).

[T. Giesberts - 934069]



I²C BUS FUSE

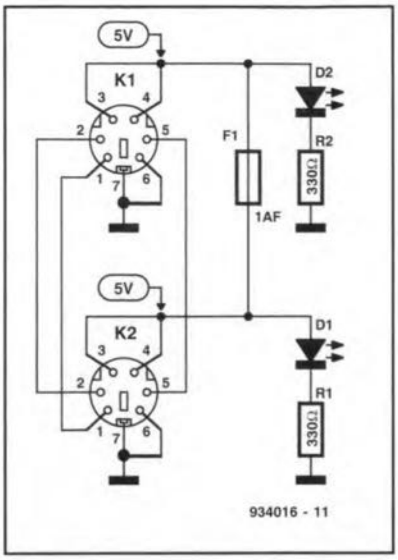
ALL I²C cards published so far in this magazine (see References) are normally powered by the 5 V supply in the PC via the 6-way DIN socket on the I²C PC insertion card. A drawback of this arrangement is that the 5 V supply in the PC is at risk of being short-circuited by a fault on any of the I²C cards connected. Since most PC power supplies are capable of supplying very high currents at the 5-V output (20+ A is typical!), short-circuits can cause a lot of smoke and annoyance.

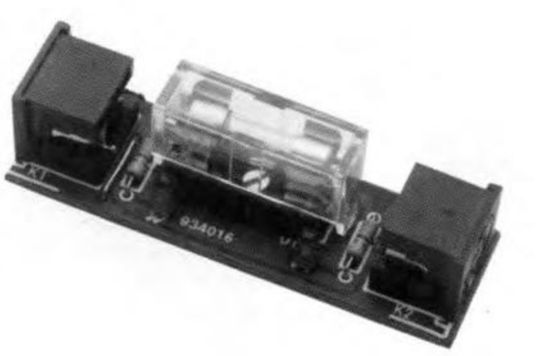
The circuit shown here protects your costly hardware by inserting a fast 1 A glass fuse in the 5 V supply line between the I²C interface card in the PC and the first I²C card in the chain. Alternatively, for even better security, a fuse card may be inserted into every cable that connects two I²C boards.

Two LEDs are fitted on the fuse card to signal the presence of the 5 V supply voltage ahead of and behind the fuse. The LEDs cause an additional current consumption of about 20 mA.

The printed circuit board shown allows five fuse cards to be built.

(J. Ruiters — 934016)



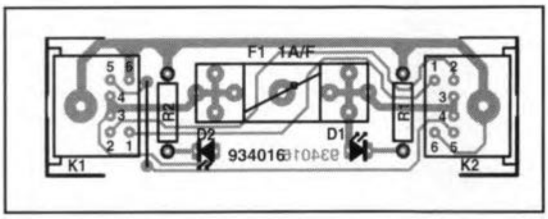


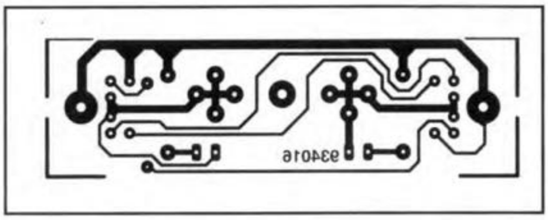
PARTS LIST

Resistors:
R1, R2 = 330 Ω

Semiconductors:
D1, D2 = LED, 3mm, red

Miscellaneous:
K1, K2 = 6-way mini-DIN socket for PCB mounting
F1 = fuse, 1 A (fast), with holder for PCB mounting
PCB Type 934016 (see p. 112)





References:

1. I²C interface for PCs. *Elektor Electronics* February 1992.
2. ADC/DAC and I/O for I²C bus. *Elektor Electronics* March 1992
3. I²C LED display. *Elektor Electronics* June 1992.
4. I²C opto/relay card. *Elektor Electronics* February 1993.

ZENER BRIDGE

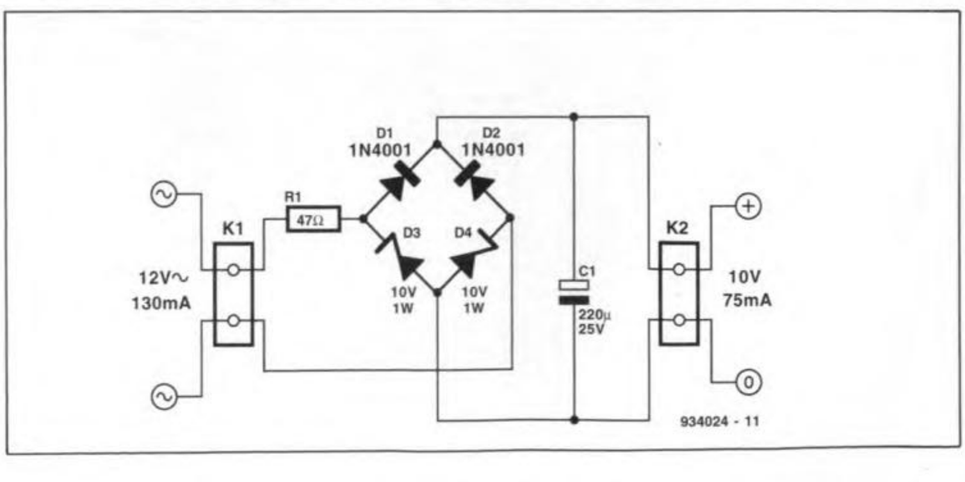
The supply described here uses two zener diodes and two rectifier diodes instead of the usual four rectifier diodes and one zener.

Assuming that the potential at junction D_1-D_3 is positive with respect to that at D_2-D_4 and is greater than the zener voltage (+0.6 V), D_3 will act as a zener and limit the voltage to 10 V. The current flowing from D_1 through C_1 and the load will return to the transformer via D_4 (the voltage across C_1 is then equal to the zener voltage of +0.6 V).

The potential across D_3 (**not** plus 0.6 V) will be applied to C_1, which (in theory) can be charged to 0.6 V below the zener voltage (in practice, without a load, it is slightly higher).

If the voltage at junction D_2-D_4 is positive with respect to that at D_1-D_3, the same process happens, but now D_4 will act as the zener, while D_2 and D_3 are forward biased.

The zener current is determined by R_1. Bear in mind, however, that the diodes act as zeners only half of the time. In other words, the average current through the zener diodes is only



half the peak current. The same applies to the power. With the value of R_1=47 Ω (as drawn) and an average zener current of 90 mA, the diodes dissipate about 450 mW, but the peak dissipation is nearly 1 W. It is, therefore, not possible to lower the value of R_1.

Since the current through the zeners is not constant (owing to the alternating voltage at the input), the maximum output current of the circuit is rather lower than twice the average zener current. Here, it is around 75 mA. At this current, the output voltage is about 9.4 V; without a load, it is around 10.4 V.

Just as in a traditional circuit, the output voltage of the present circuit rises to the peak value of the input potential (that is, about 18 V) when one of the zener diodes fails.

Also, bear in mind that the circuit limits only the peak value of the output voltage. As soon as the diodes no longer act as zener diodes, the output voltage depends entirely on the charge of C_1. That charge will, of course, drop and so, therefore, will the output voltage. The present circuit is consequently intended primarily for use with circuits that tolerate ripple or that draw only a small current. With C_1=220 μF (as drawn), and the load current 75 mA, the ripple amounts to 7 V. To lower that value appreciably, the value of C_1 must be raised considerably or the load current must be limited to about 1 mA.

[L. Pijpers – 934024]

SCSI ACTIVE TERMINATION

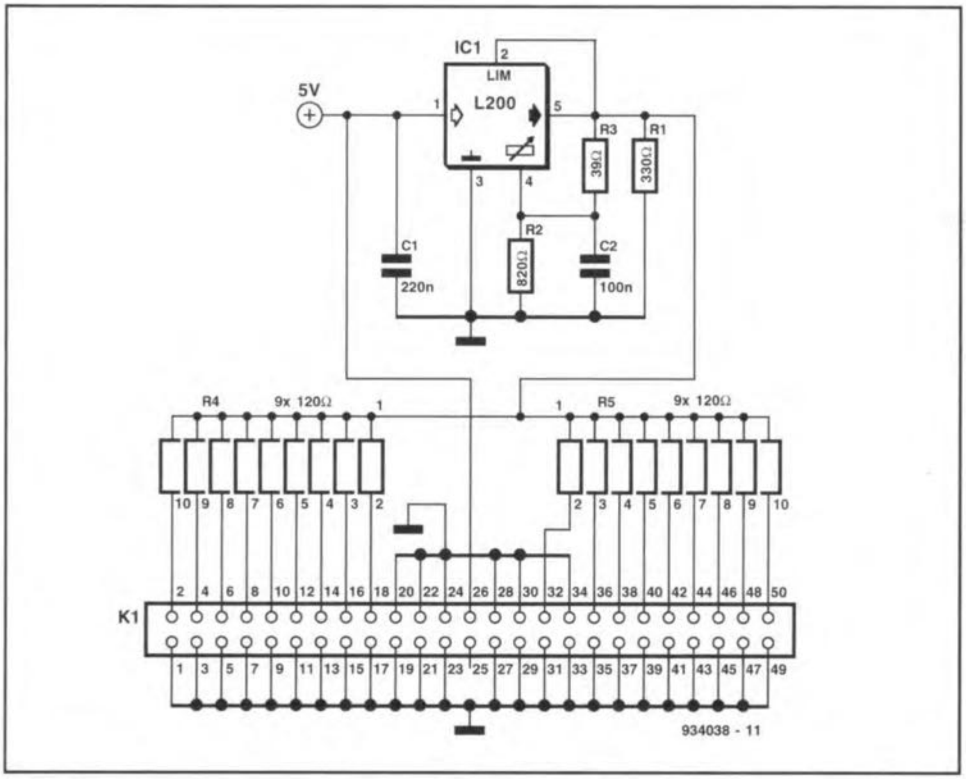
THE SCSI bus (small computer systems interface) is a high-speed bidirectional parallel bus used mainly for connecting 'intelligent' peripheral devices to a CPU. The bus is the *de facto* standard in the Apple computer environment, where it is used to connect devices like CD-ROM readers, scanners, and internal as well as hard disks. The SCSI bus has also found wide acceptance in the IBM PC world, where the main application is with large hard disks. SCSI hard disks feature a 'software' type identifier, and require an appropriate hard disk controller and associated software.

An SCSI bus network should have only one bus termination device. These devices are usually supplied with the PC, or the SCSI peripheral, and consist of resistors only, connected between the TERMPWR (+5-V) line, the transmission (signal) lines, and ground (SCSI drivers are open-collector types). The termination resistance 'seen' on every signal line is critical to ensure the highest possible data speed on the network.

The circuit shown here is an active termination device, which eliminates the 220 Ω and 330 Ω resistors required for each signal line with a passive termination scheme, reducing significantly the continuous system power drain. When placed in series with 120 Ω resistors, the L200 voltage regulator matches the impedance level of the transmission line, and eliminates reflections. Here, the L200 supplies a fixed voltage of about 2.75 V across R_1.

The active termination device should replace the single, passive, termination device already fitted in the SCSI system. Construction is easy on the PCB shown. The voltage regulator, IC_1, may be bolted direct to the PCB without a heat-sink.

(T. Scherer — 934038)



PARTS LIST

Resistors:

1	330Ω	R1
1	820Ω	R2
1	39Ω	R3
2	9-way 120Ω SIL array R4;R5	

Capacitors:

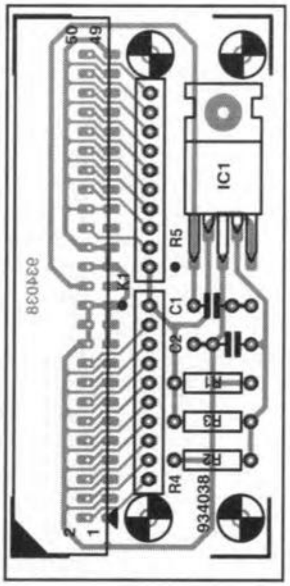
1	220nF	C1
1	100nF	C2

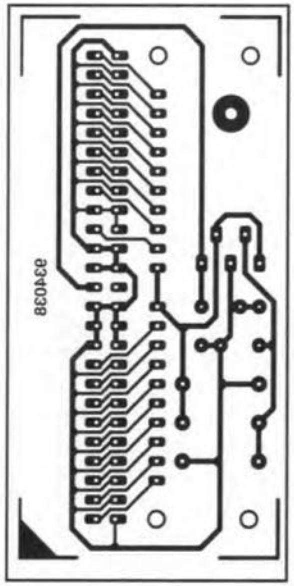
Semiconductors:

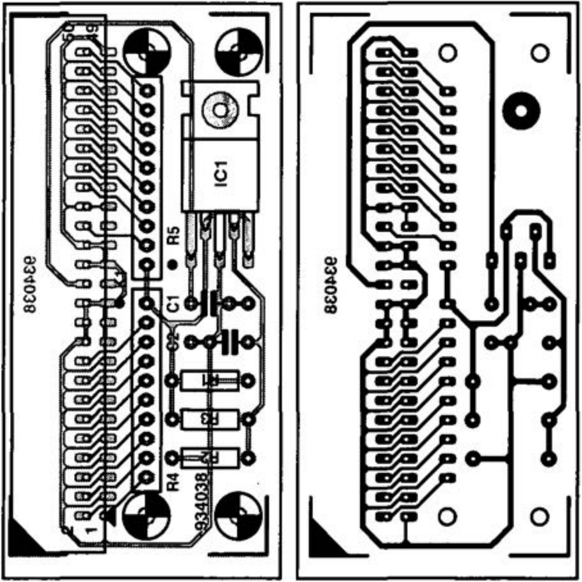
1	L200	IC1

Miscellaneous:

1	50-way box header, PCB mount, angled solder pins	K1
1	50-way IDC socket	







IC1
R5
C1
C2
R1
R3
R2
R4
934038
1

HALOGEN LIGHT SWITCH

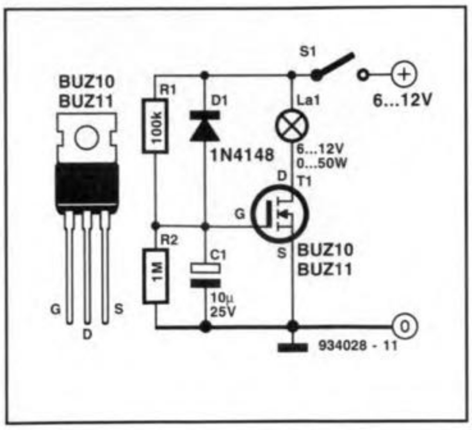
Halogen lamps are rightly popular because they give a good light and have an excellent efficiency. Unfortunately, they tend to be costly. Moreover, they often give up the ghost when they are switched on, because they then draw a very high current (some ten times normal, equivalent to roughly a hundred times their normal power ating). A simple passive circuit can remedy this, but only for DC operated lamps. If they are AC operated, a rectifier could be added, but that gives relative high losses at 6 V or 12 V.

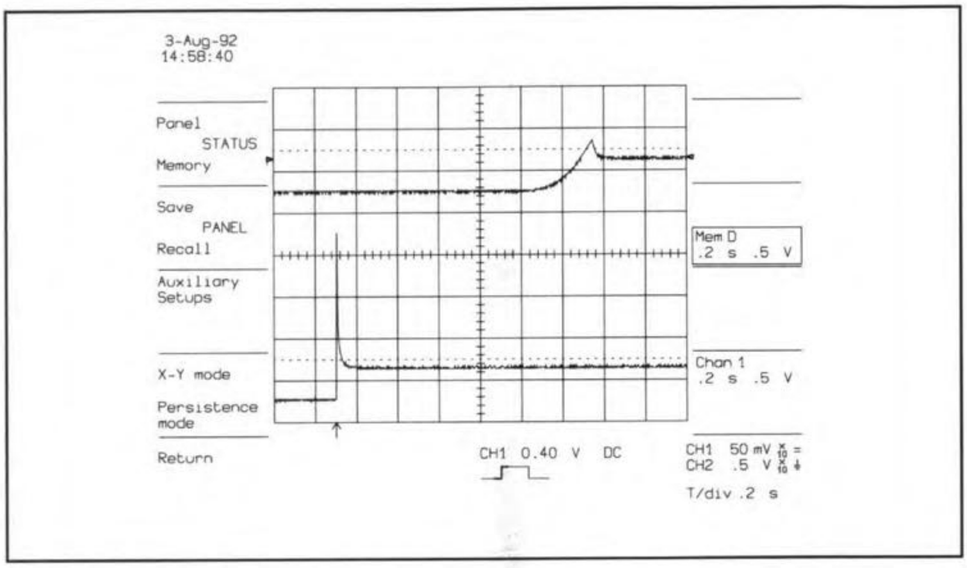
The present circuit takes a different tack: it is based on the fact that a power FET passes a current that is dependent on its gate voltage. If the gate voltage rises gradually, the current will increase gradually also. Here, the gate voltage is determined by the potential across C_1, which is charged slowly via R_1. Slowly means some tens of milliseconds, which is long enough for the filament to warm up.

Note that the FETs specified need at least 6 V gate voltage to switch on. The maximum gate voltage is 12 V, so that the circuit is suitable for 12 V lamps also. The value of R_1 for 6 V lamps should be about 100 kΩ, and for 12 V lamps, about 470 kΩ. Figure 2 shows the effect of the circuit. The lower curve represents the current without limiting: its peak is about 4.5 times the nominal current through the lamp. With the limiting circuit, the lamp current no longer assumes such high values, as shown by the upper curve.

The MOSFET may be any suitable type. The BUZ10 can handle some 20 A, so that it can switch 12 V, 20 W lamps without any problems. In practice, even 50 W lamps may be switched, because the large currents last only a very short time. A Type BUZ11 can handle up to 30 A. Losses are small: a BUZ10 has an on resistance of 0.08 Ω, which at 1.67 A gives a loss of 230 mW. This heats the transistor in free space by about 17 °C. A heat sink is, therefore, not needed.

[K. Walraven - 934028]



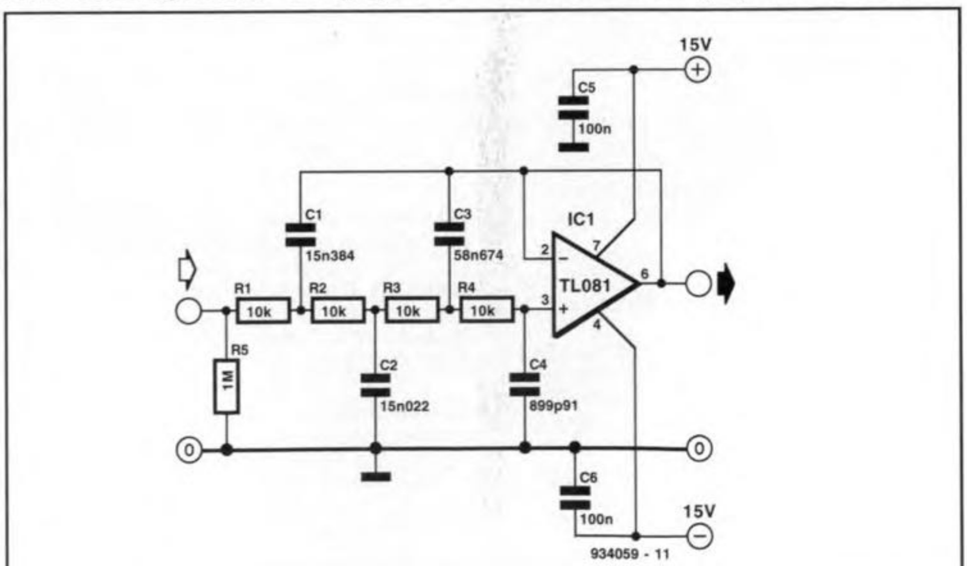


FOURTH ORDER SINGLE CHIP FILTER

High-order filters are normally designed by using two or more 2nd-order sections in series. This means that a 4th-order filter needs at least two opamps. The present filter, however, uses only one opamp, which results in smaller distortion, intermodulation, and so on. Also, there is no internal resonance rise, which typifies combinations of 2nd-order sections. Because of this, the peak input signal may be equal to the peak output of the opamp (but consider the common-mode input range).

Drawbacks of the circuit are the rather high ratio C_3: C_4 and the minimum value of the resistors. The resistor values are determined by the load on the output of the opamp (of which the resistors form a part). The maximum load (with large signals) of a TL081 is 2 kΩ. Resistors R_1–R_4 constitute an impedance of 2.5 kΩ, so that the external load must not be smaller than 10 kΩ. If an opamp is used that can handle a load of 600 Ω, it is advisable to give R_1–R_4 a minimum value of 2.5 kΩ. This will lower the noise emanating from the filter, which is generated chiefly by the resistors.

The characteristic of the filter is a 4th-order Bessel polynomial. A Butter-



worth characteristic is difficult to obtain with this type of filter, since, owing to the unity amplification of the opamp, the ratio $C_3 : C_4$ becomes very high. With component values as shown, the –3 dB point is 1 kHz. Other cut-off frequencies can be obtained by recalculating the component values: these are directly proportional to the frequency. The 4th-order Bessel and Butterworth polynomials and the transfer function of the circuit are given below.

Bessel: $1 + s + {}^{3}/_{7}s^2 + {}^{2}/_{21}s^3 + {}^{1}/_{105}s^4$.
Butterworth: $1 + 2.6131s + 3.4142s^2 + 2.6131s^3 + s^4$.

$$
\begin{aligned}
A\mathrm{j}\omega = 1 &+ \mathrm{j}\omega 2C_2C_4 + \\
&(\mathrm{j}\omega)^2[C_1(C_2+3)C_1(C_4+4)C_2(C_4+3)C_3C_4] + \\
&(\mathrm{j}\omega)^3[2C_1C_2(C_4+2)C_1C_3(C_4+2)C_2C_3C_4] + \\
&(\mathrm{j}\omega)^4\, C_1C_2C_3C_4.
\end{aligned}
$$

[T. Giesberts - 934059]

GENERAL PURPOSE SCHMITT TRIGGER

The thresholds voltages, and thus the hysteresis, of the Schmitt trigger described can be set independently.

The circuit is based on opamp IC_1, which has been connected as an inverting Schmitt trigger. The two voltages at which the output of the circuit changes level are preset with P_1 (high level = U_H) and P_2 (low level = U_L). Provided that U_H is greater than U_L;

$$U_i<U_r->U_o=U_H->U_r=U_H;$$

$$U_i>U_r->U_o=U_L->U_r=U_L.$$

These two expressions are shown as a graph in Fig. 2. From this it is clear that when the output is high, it will change ot low only if the input has risen above U_H, and will then change back to high again only if the input voltage has dropped below U_L.

The changing state of the output results in the changing of the reference voltage at the non-inverting input of IC_1. In the first instance, this is caused by the weak, but fast, positive feedback via R_6 and R_7. The hysteresis caused by this network amounts to only 20 or 30 mV, but that is more than enough to give the opamp good switching performance.

Once the opamp has changed state, U_r changes for the second time, because IC_{2a}-IC_{2d} alter the level at junction R_6-C_7 from U_H to U_L or vice versa.

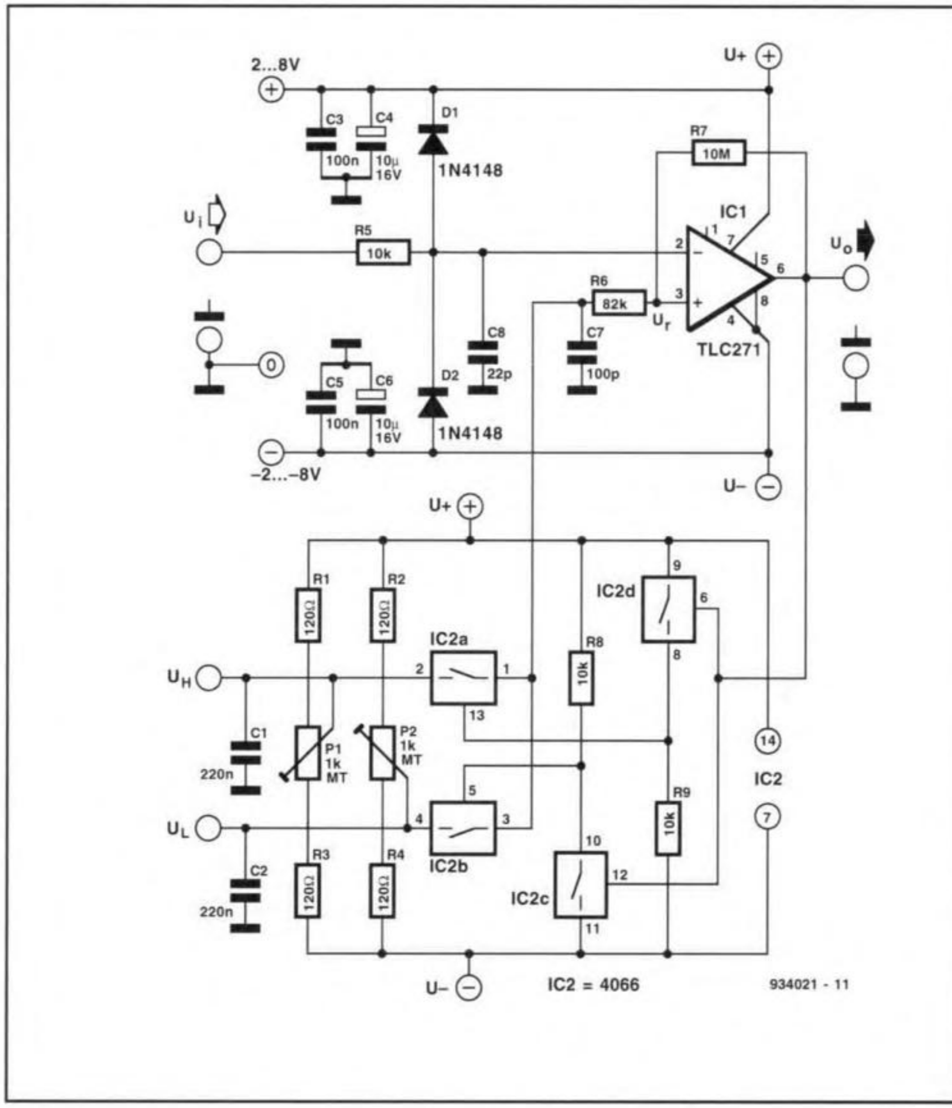
Capacitor C_7, in conjunction with the on resistance of the analogue switches, slows the transfer process: the reference voltage changes smoothly from one value to the other in about 250 ns. This assumes that the speed at which the input signal changes is not too high. To enable signals with steep transitions (edges) to be processed, there is a low-pass filter at the input (R_5–C_8).

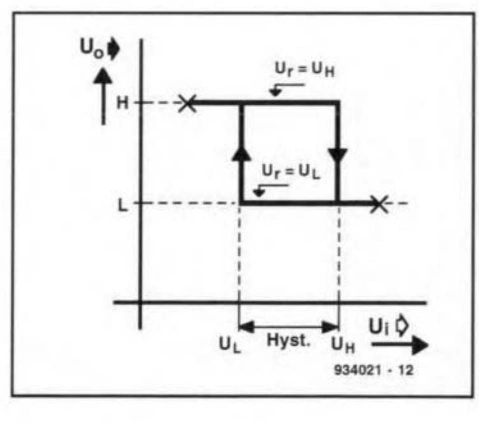
The threshold voltages are set with the aid of a millivoltmeter connected successively to U_H and U_L. Presets P_1 and P_2 should preferably be multiturn types to enable the levels to be set accurately to within a millivolt.

The trigger may be used with unipolar and bipolar input signals. If only positive inputs need to be processed, the power supply for the trigger may be asymmetrical. To that effect, the negative supply line is linked to '0' and the supply (max. 16 V) connected to + and 0. With a supply voltage of 16 V, the trigger draws a current of ≤30 mA.

The switching performance is tested with a triangular voltage (0–5 V) at the input. With that input and the thresholds set to 2 V and 4 V, the trigger switched accurately at frequencies up to 5 kHz.

[J. Ruiters - 934021]

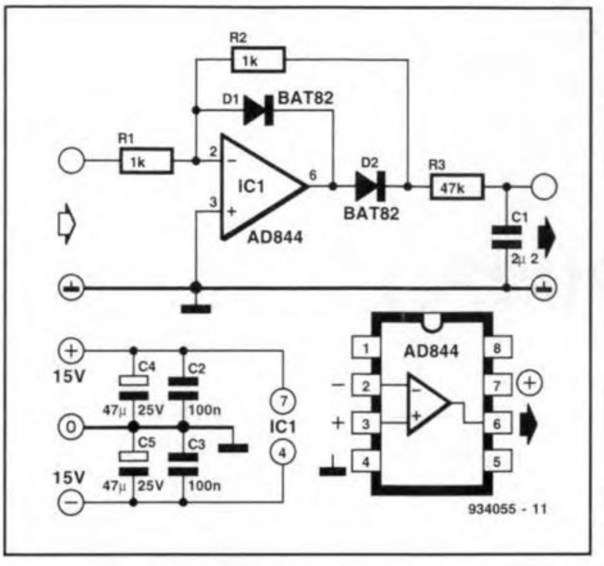




FAST ACTIVE RECTIFIER

The rectifier is intended to improve the bandwidth of most digital multimeters. It is based on a half-wave rectifier in a traditional set-up.

Diode D_2 is in series with the outputof opamp IC_1. Since feedback resistors R_1 and R_2 are connected after D_2, the negative halves of the signal at the input of the circuit will appear with the same value at the cathode of D_2. During the positive halves at the input, the output of the opamp is kept at almost 0 V by D_1. Network R_3-C_1 at the output integrates the half-wave rectified voltage, so that the direct voltage across C_1 represents the average value of the negative halves of the input signal. If the original signal was sinusoidal, the direct voltage will be $1/\pi$ times the peak value of the input signal.



The opamp used is a current-feedback type. Its bandwidth is determined by the values of the resistors in the feedback loop. For a large bandwidth, the value of these resistors should be as low as feasible. To retain an acceptable input impedance (1 kΩ in the prototype), both R_1 and R_2 have a value of 1 kΩ. The bandwidth is then 30 MHz. When the value of both resistors is halved, the bandwidth becomes 60 MHz.

The accuracy of the rectifier is determined not only by the bandwidth, but also by the slew rate. That of the AD844 is not less than 1200 V µs^{-1} (typically 2000 V µs^{-1}). Furthermore, the capacitance and forward bias of D_1 and D_2 play a role. The BAT82 has a capacitance of 1.6 pF,and a forward bias of about 0.5 V at a forward current of 4 mA.

In the prototype circuit, the frequencies were measured when the output voltage was dropped by 1% (about 0.1 dB). At 1 V output, the –1% point was at 1.7 MHz; at 100 mV at 400 kHz, and at 10 mV at 45 kHz.

The circuit draws a current of 6.5 mA with a supply voltage of ±15 V. The supply to the AD844 may lie between ±4.5 V and ±18 V.

[T. Giesberts - 934055]

INTERIOR LIGHT SWITCH FOR CARS

Most cars are provided nowadays with interior lighting that comes on automatically when one of the doors (or the tailgate of a hatchback) is opened. This can have a drawback if one of the doors of the car is not closed properly and the car is left in the garage for a few days.* This may effectively flatten the battery. There is, however, a remedy for this.†

For the proposed circuit, the usual lights switch takes over the control of the interior lights from the door switches. The circuit is based on a 555 (7555, 555C) timer and acts as a monostable that switches on the lights for four minutes as soon as it is switched on. This time may be altered by changing the values of R_2 and C_2. Network R_1-C_1 ensures that the IC is enabled as soon as voltage is applied.

The output of IC_1 controls relay Re_1 via T_1. The only requirement of the relay is that it should be actuated by a voltage of 12 V and a current of less than 200 mA. As soon as the timer switches the lights off, the current drops to 6 mA if a 555 is used and to 0.5 mA if a 555C is used.

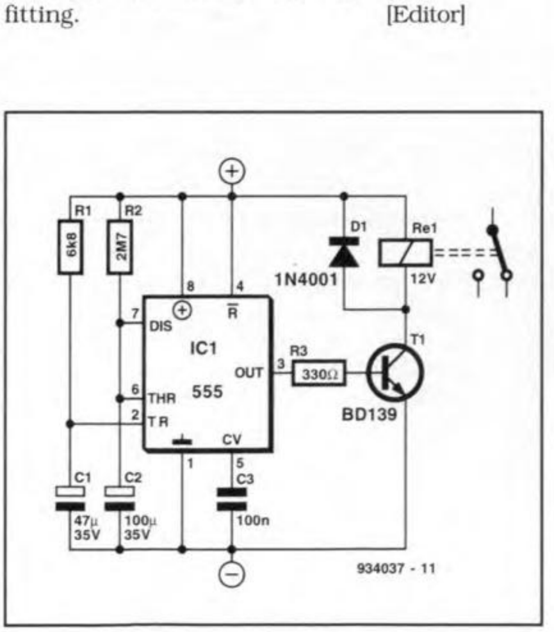
Before the circuit is built into the car, check how the door switches are connected. If they are in the +ve battery line, use variant *a* (top); if in the –ve (earth) line, use variant *b*.

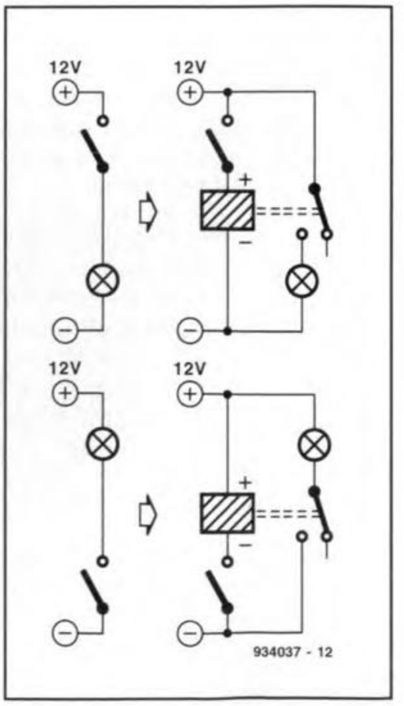
Finally, connect the circuit to the junction of the wires from the door switches, otherwise it will react to only one door.

[B. Klein - 934037]

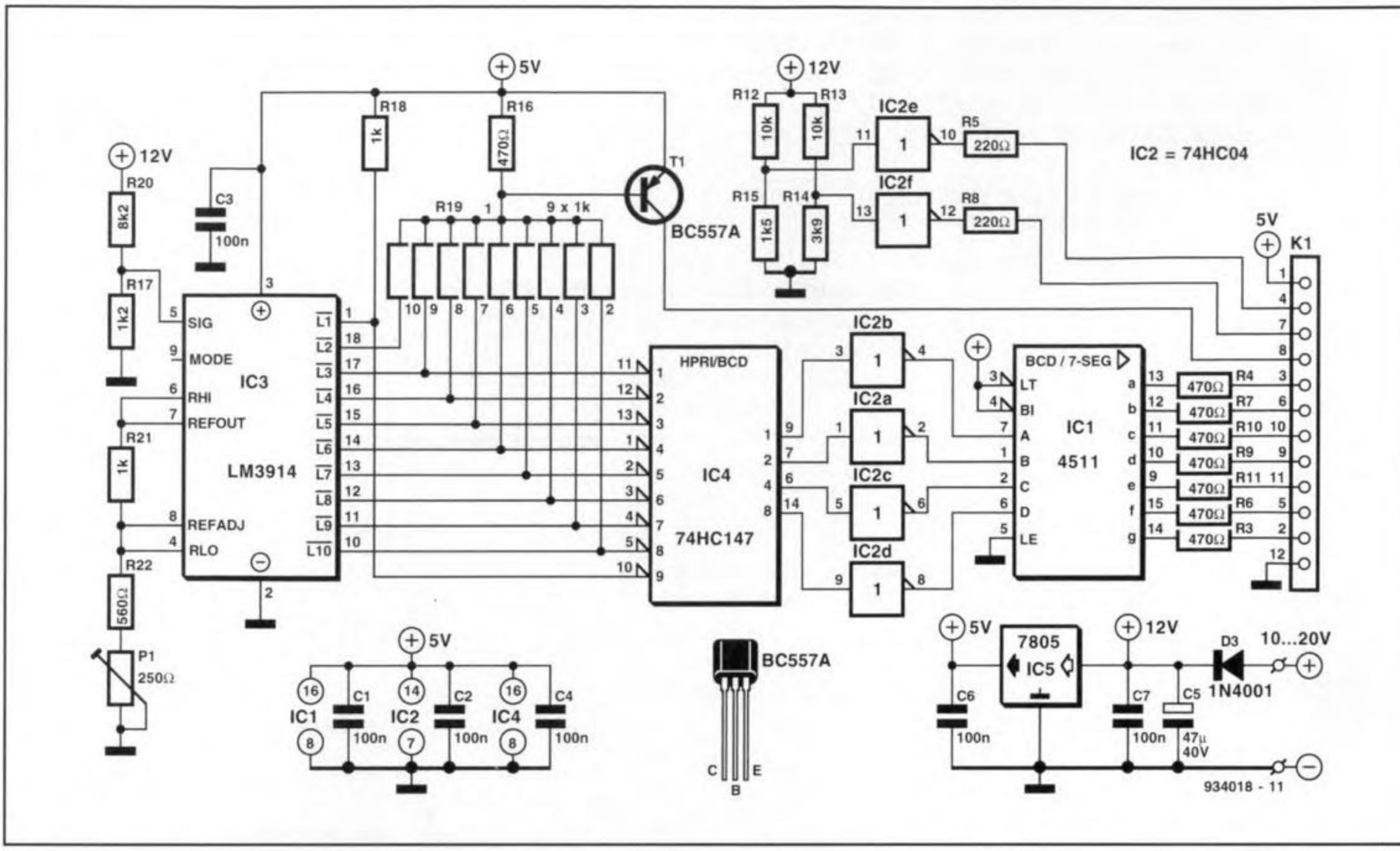
* These days, even in the garage, it is best for security's sake to properly shut and lock all doors and, if fitted, set the alarm. [Editor]

† In quite a few modern cars, the circuit described (or one like it) is a standard fitting. [Editor]





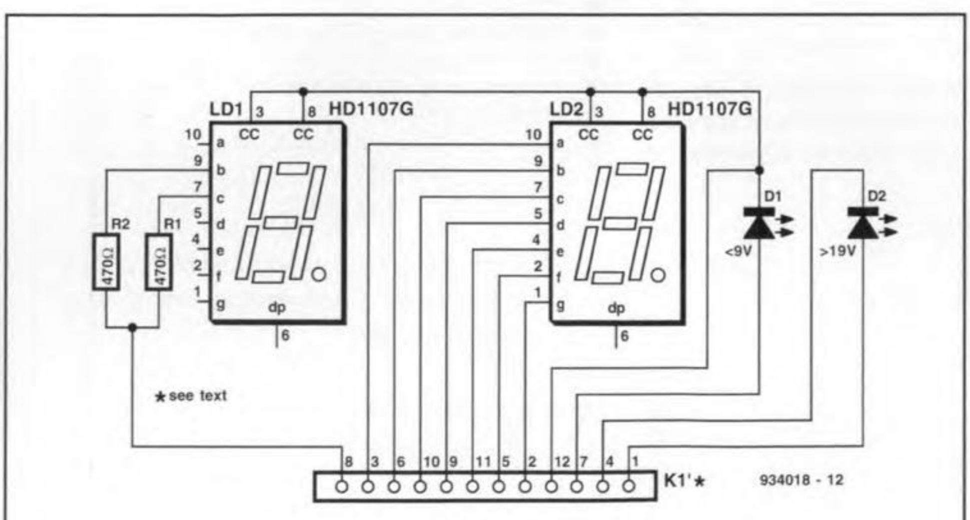
CAR BATTERY VOLTAGE MONITOR



THE monitor uses a two-digit 7-segment LED display to indicate the battery voltage. In addition to this, there are 'low' and 'high' voltage warning indicators. The display range is 9–18 V.

The battery voltage is scaled down by a factor 7.83 by resistors R_{20} and R_{17}. This means that the voltage range at the SIG input of the LM3914 (IC_3) is 1.14–2.29 V for battery voltages between 9 V and 18 V. The LM3914 has an internal voltage reference that maintains a voltage difference of 1.25 V between pins 7 and 8. Each of the ten resistors in the ladder network thus drops a voltage of 0.125 V. The voltage difference for each step between 9 V and 18 V is (2.29-1.14)/9, or 0.1278 V, which is close enough to 0.125 V for the resolution required by the present application. Here, the LM3914 is operated in 'dot' mode.

The open-collector outputs of the LM3914, L1\-L10\, are connected to 1 kΩ pull-up resistors, nine of which are contained in SIL array R_{19}. The resistors also cause a '1' to appear on the most significant (MS) display for all input voltages greater than 9 V. Only if 9 V is measured will the base of T_1 remain high, so that the MS display is quenched. If the input voltage is higher, the base of T_1 is pulled low, and the B and C segments of LD_1 light, showing a '1'. Output L_2 of the LM3914 is not connected because the priority encoder, IC_1, automatically encodes a '0' if none of its inputs is pulled low. Together with the '1' on the MS display this produces the reading '10'. Inverters IC_{2a}–IC_{2d} ensure the correct input levels for the inputs of the 4511 display driver, IC_1.



Inverters IC_{2e} and IC_{2f} function as undervoltage (<8.7 V) and overvoltage (>18.7 V) monitors and indicator drivers. The actual switching thresholds will depend on the IC used, so some experimentation with the value of R_{15} (ovevoltage) and R_{14} (undervoltage) may be called for. The undervoltage LED should light at input voltages between 5 V and about 8.5 V. Lower voltages are not possible because the LED light would not be seen any more. The overvoltage LED should light at an input voltage of about 19 V.

The circuit is simple to adjust with an accurate digital voltmeter and an

COMPONENTS LIST

Resistors:

10	470Ω	R1-R4;R6;R7; R9;R10;R11; R16
2	220Ω	R5;R8
2	10kΩ	R12;R13
1	3kΩ9	R14
1	1kΩ5	R15
1	1kΩ2	R17
2	1kΩ	R18;R21
1	9-way 10kΩ SIL array	R19
1	8kΩ2	R20
1	560Ω	R22
1	250Ω preset H	R22

Capacitors:

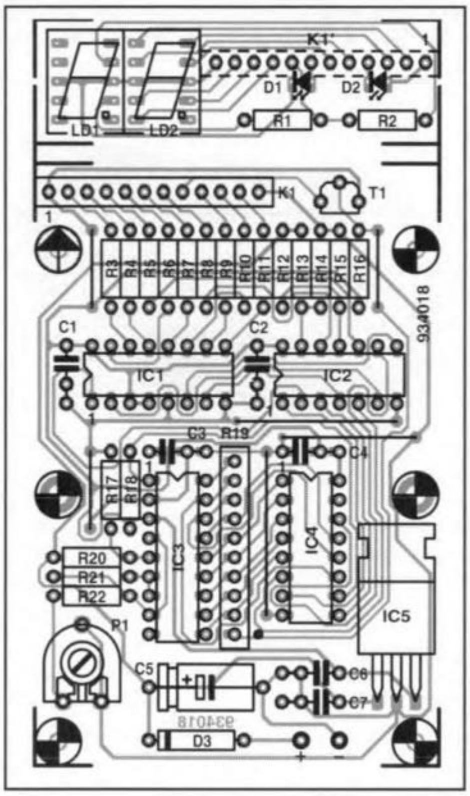
6	100nF	C1-C4;C6;C7
1	47μF 40V	C5

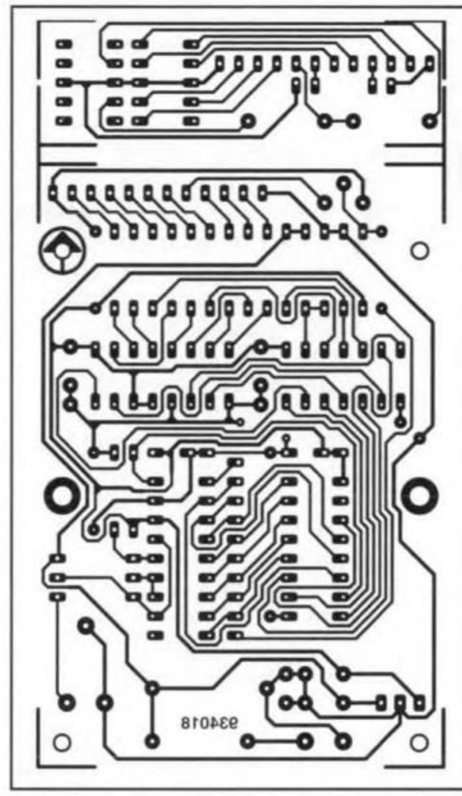
Semiconductors:

1	LED, yellow, 3mm	D1
1	LED, red, 3mm	D2
1	1N4001	D3
1	BC557A	T1
1	4511	IC1
1	74HC04	IC2
1	LM3914	IC3
1	74HC147	IC4
1	7805	IC5
2	HD1107G (Siemens)	LD1;LD2

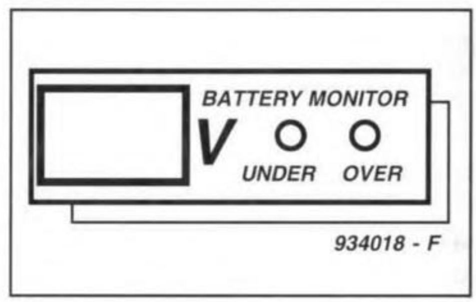
Miscellaneous:

1	12-way angled SIL header	K1
1	12-way SIL socket	K1'
1	Plastic enclosure HM-Kit (Pactec), 61×97×23mm (w×l×h); order code 6600-902.	
1	Printed circuit board 934018 (see page 112)	





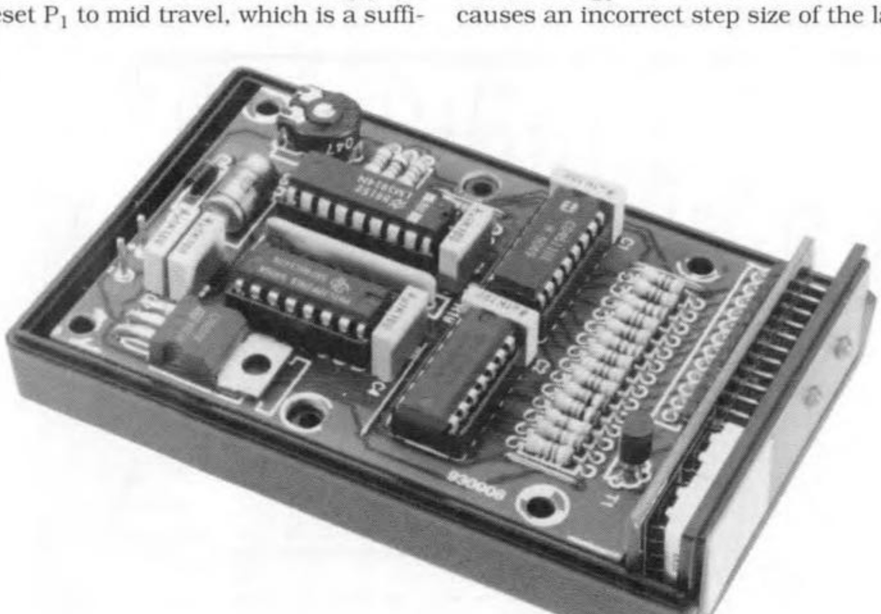
adjustable power supply. If these instruments are not available, simply set preset P_1 to mid travel, which is a sufficiently accurate setting in most cases. If precision is needed, apply 12.0 V to the input of the circuit. Adjust P_1 until the display reads '12'. This value should remain on the display if the input voltage is varied between 11.5 V and 12.5 V. If not, adjust P_1. Next, reduce the input voltage to 8.0 V. The undervoltage LED should light, and the display read '0'. Slowly increase the voltage to 9 V. At a value of 8.5 V the display should change to '9'. The exact voltage at which this happens is not terribly important. Next, apply 17 V, which should cause the display to indicate '17'. Slowly increase to 18 V, and check that the display changes to '18' between 17.2 V and 17.8 V. If this does not work, R_{21} is out of tolerance, which causes an incorrect step size of the ladder network. The cure is simple: replace R_{21} (and, if necessary, R_{17} and R_{20} also) by a 1% tolerance type.

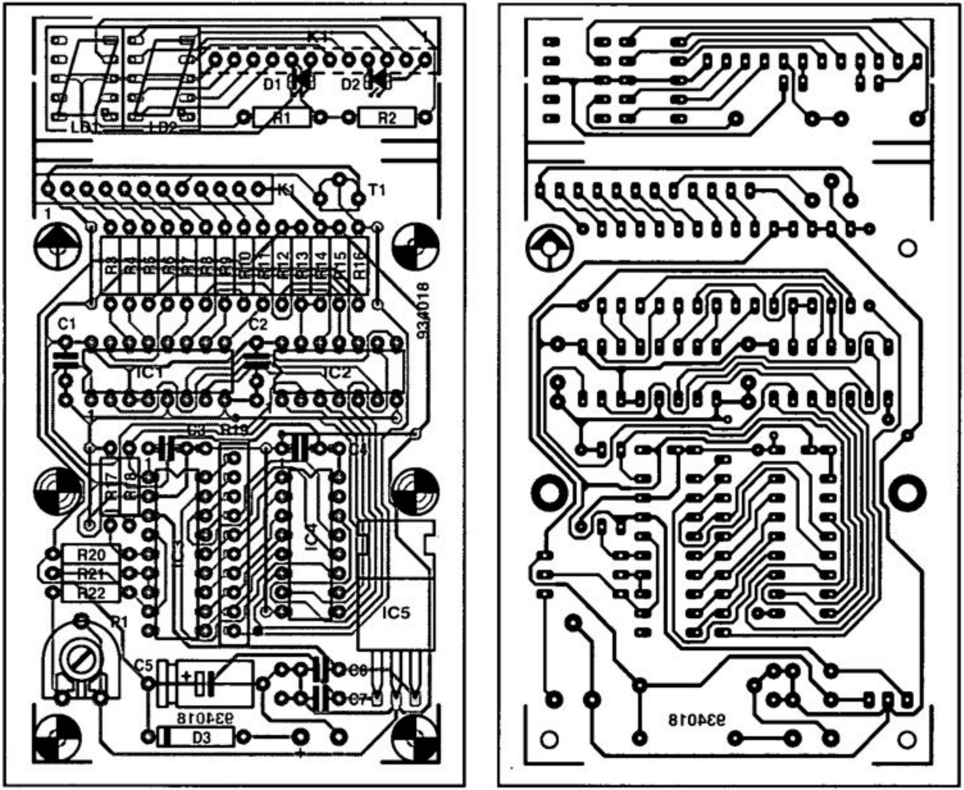


The circuit draws a fairly high current, which causes voltage regulator IC_5 to run hot if it is not fitted on a small heat sink.

The printed circuit board consists of two sections (display and voltage monitor proper), which must be separated by cutting before they are populated. The connection between the two units so obtained is formed by a 12-way pin header, K_1, on the main board, and a mating 12-way SIL socket, K_1', on the display board. The display unit is mounted at a suitable position on the car's dashboard, with the displays covered by a bezel. Alternatively, the driver circuit and the readout may be fitted into a small plastic enclosure, for which a suggested front panel layout is shown.

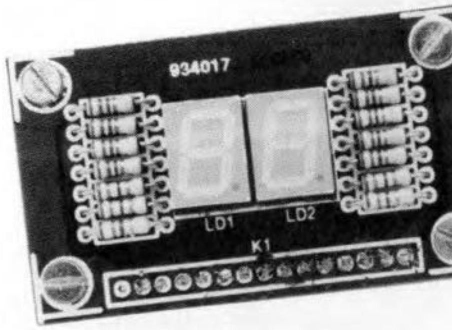
[K.M. Reddy - 934018]




LD1
LD2
K1
D1
D2
R1
R2
K1
T1
R3
R4
R5
R6
R7
R8
R9
R10
R11
R12
R13
R14
R15
R16
934018
C1
C2
IC1
IC2
C3
R19
C4
R17
R18
IC4
R20
R21
R22
P1
IC5
C5
C6
C7
D3
934018

SPEED INDICATOR FOR PCs

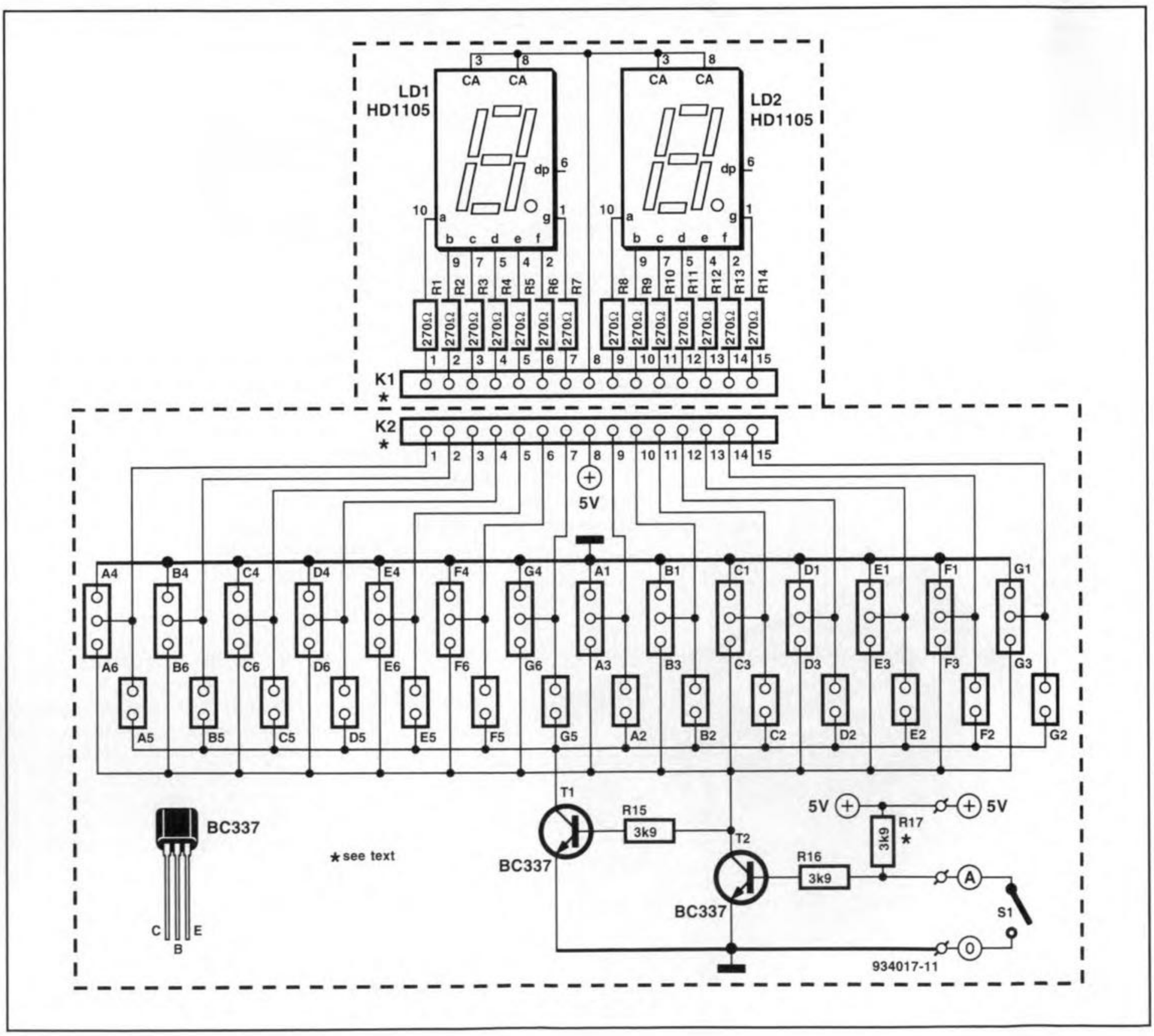
FOR some reason or other, the principal feature of a PC seems to be its CPU clock speed, and that is probably why many PC users have developed a habit of judging a PC by what appears on the 2-digit speed readout. Machines with a display reading below '33' (for 33 MHz), or which lack a display altogether, are not worth talking about. In other words, if your PC is not fast enough, you are 'out' before you know it.

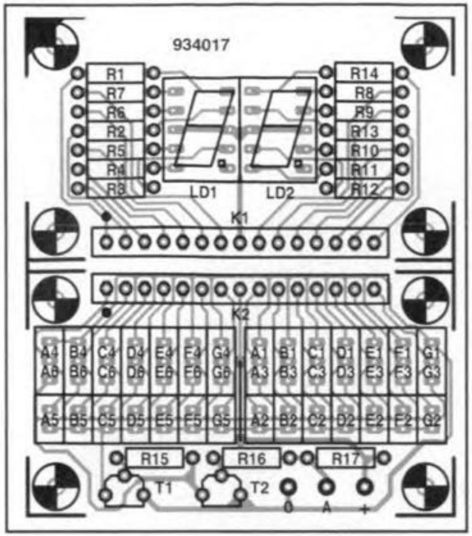


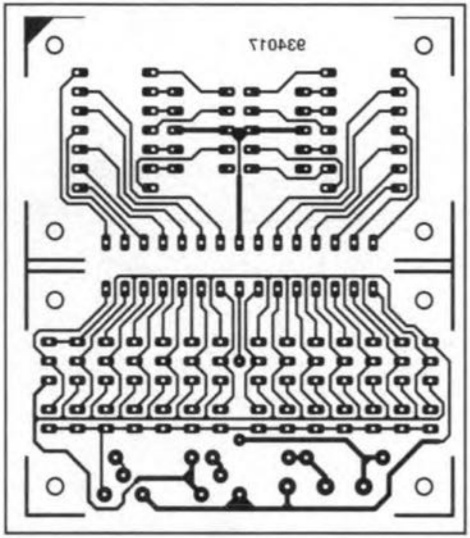
To fool these sufferers of 'PC speed phobia', and to bring you up to par again, we propose a circuit that will feign a much higher CPU speed than actually used in the machine. In this way, you will be able to accredit, say, an 'old' 12-MHz AT with a 'state-of-the-art' CPU turbo speed of 66 MHz, or even 99 MHz! After all, only you will know for sure what's really ticking inside the grey desktop case.

Joking apart, resistors R_1–R_{14} limit the LED segment currents, and their value may be changed as required (within limits, of course) to reduce or increase the readout brightness. The 'turbo' speed indication is switched on via transistor T_2. Applying +5 V to the base of T_2 causes T_1 to be switched off. The reverse situation, T_2 off and T_1 on, applies when the turbo switch is set to the 'normal' position.

Unfortunately there exist differences between PCs as regards the logic levels used to switch between 'normal' and 'turbo' CPU speed. In most cases, turbo speed is selected when the switch contact is a 'make' type that closes to ground. The table shown applies to that configuration. The table is also valid if the 'normal' position of the switch makes contact to +5 V; only R_{17} must then be removed. In case turbo speed is







PARTS LIST

Resistors:

14	270Ω	R1-R14
3	3kΩ9	R15;R16;R17

Semiconductors:

2	BC337	T1;T2
2	HD1105O	LD1;LD2

Miscellaneous:

14	2-way pin header	
14	3-way pin header	
1	15-way SIL socket	
1	15-way SIL header	
1	Changeover switch (turbo switch in PC)	S1

Normal	Turbo	Always fitted	Normal	Turbo
8	12	A1;B1;D1;E1;G1	C3;F3	B5;C5
8	16	A1;C1;D1;E1;F1;G1	B3	B5;C5
8	25	A1;C1;D1;F1;G1	B3;E3	A5;B5;D5;E5;G5
8	33	A1;B1;C1;D1;G1	E3;F3	A5;B5;C5;D5;G5
8	66	A1;C1;D1;E1;F1;G1	B3	A5;C5;D5;E5;F5;G5
8	99	A1;B1;C1;D1;F1;G1	E3	A5;B5;C5;D5;F5;G5

selected when the switch contact closes to +5 V, the table needs to be modified, and R_{17} omitted. These modifications are also required, but R_{17} must be left in place, if 'normal' speed is selected with the switch contact closed to ground.

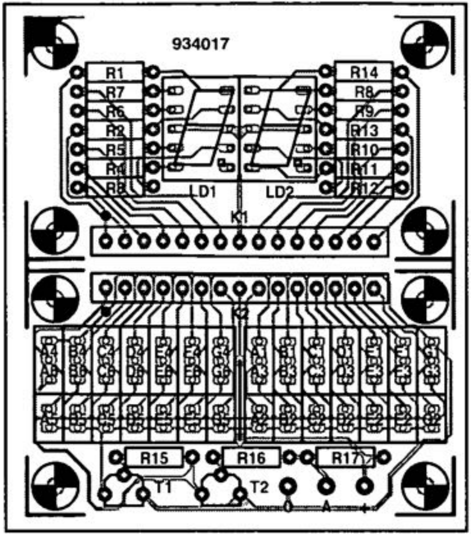
The modifications to the table are as follows:

(1) in the fourth ('Normal') column, change all entries *x*3 into *x*2

(2) in the fifth ('Turbo') column, change all 5s into 6s.

Construction of the readout is simple on the printed circuit board shown (not available ready made). The board is cut into two to separate the jumper section from the display section. The two sections are interconnected by a 15-way SIL socket and a 15-way SIL header. Finally, note the wire link that runs between the speed setting jumpers — if you forget to fit it, the displays will remain dark forever in the absence of their supply voltage.

(L. Pijpers, 934017)


934017
R1
R7
R6
R2
R5
R4
R3
R14
R8
R9
R13
R10
R11
R12
LD1
LD2
K1
K2
A4
B4
C4
D4
E4
F4
G4
A1
B1
C1
D1
E1
F1
G1
A3
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D3
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F3
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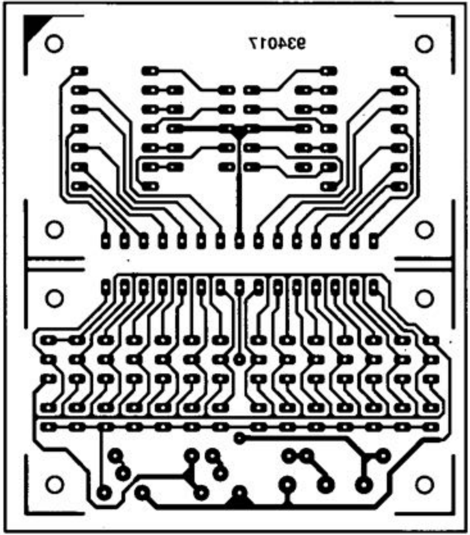
T1
T2
0
A
+

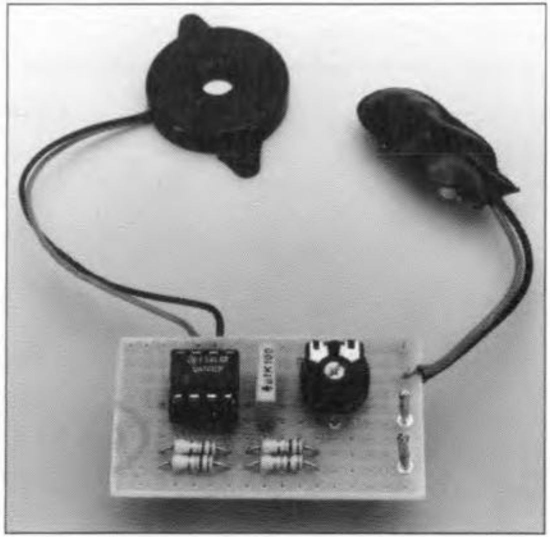

934017

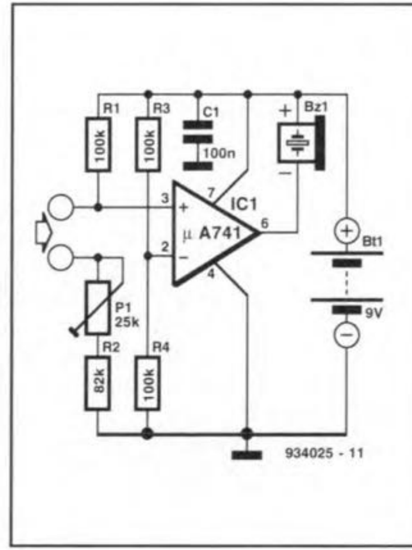
CONTINUITY TESTER

The continuity tester is based on the well-tried µA741, which is used as a comparator that operates a buzzer at its output. The inverting (–) input is at half the supply voltage via R_3 and R_4. When the potential at the non-inverting (+) input is lower than that at the –ve input, which is the case when the resistance between the two terminals is small enough, the buzzer will be actuated.

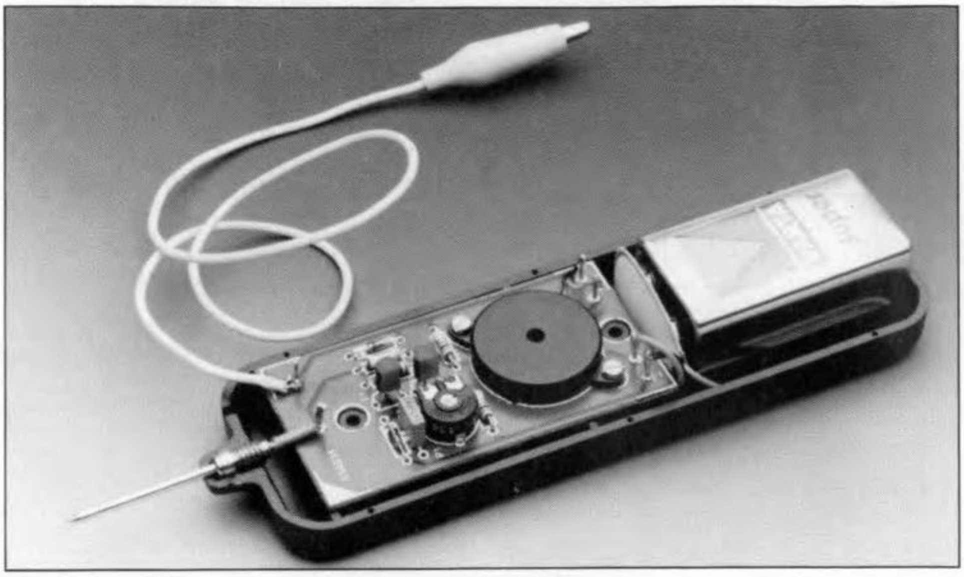
With values of R_1, R_2 and P_1 as shown, the sensitivity of the tester is about 1 kΩ. The circuit is set by connecting a 1 kΩ resistor between the input terminals and adjusting P_1 so that the buzzer just does not sound. The sensitivity may be increased to 100 Ω by making the values of R_1, R_2 and P_1 ten times smaller. The circuit then draws a slightly larger current (2.5 mA instead of about 2 mA).

[A. Rietjens - 934025]





AUDIBLE CONTINUITY TESTER



WHATEVER the latest trend in digital multimeters, no electronics workshop is complete without a continuity tester for fast and reliable testing of connections, components and PCB tracks. The little instrument discussed here is an audible continuity tester whose output frequency is determined by the resistance measured. Handheld and extremely simple to use, the tester allows you to concentrate on the connections you wish to examine rather than on a LED, a display or a moving-coil meter. So, using the continuity tester is a matter of listening as you 'probe through' the circuit under investigation.

The circuit diagram shows a two-transistor oscillator whose output frequency is a function of the resistance measured between the input terminals, A and B. The feedback element in the oscillator is formed by capacitor C_1. Diode D_1 and the base-emitter diode in T_1 ensure smooth charge reversal in C_1, without creating a potential divider with base resistor R_1. The collector resistance of T_1 is formed by a diode, D_2, which ensures stable amplification in spite of battery voltage reduction. The distortion caused by this arrangement is of little consequence since the output is not a clean sine wave anyway. The transistors draw the greater part of their supply current through a passive piezo buzzer, and so generate an audible tone.

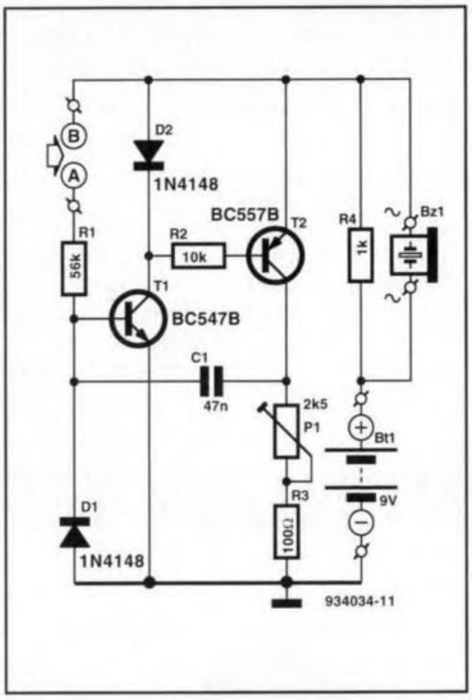
A combination of one n-p-n and one p-n-p transistor is used to make sure that no battery power is consumed when the test terminals are not connected. This obviates an on/off switch. Depending on the setting of preset P_1, the current drain of the actuated circuit is a modest 3–5 mA.

The frequency of the tone produced by the tester drops with increasing resistance measured. It is practically impossible to damage components or connections with the tester, since the voltage difference between the input terminals is about 8 V, and the maximum current about 50 μA.

Preset P_1 is adjusted such that an agreeable tone is produced with the input terminals short circuited. Next, connect a resistor of 22 kΩ, whereupon the tone frequency should drop considerably.

The completed printed circuit board (not available ready made) is fitted into a probe-type enclosure, together with the 9-V (PP3) battery. Input 'A' is connected to a solid pin (the actual probe tip) that protrudes from the front side of the enclosure, while 'B' is connected to a flexible, red wire fitted with an insulated croc clip.

[F. Hueber — 934034]



COMPONENTS LIST

Resistors:

1	56kΩ	R1
1	10kΩ	R2
1	100Ω	R3
1	1kΩ	R4
1	2kΩ5 preset H	P1

Capacitors:

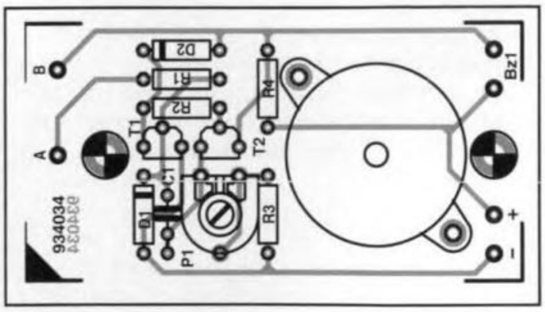
1	47nF	C1

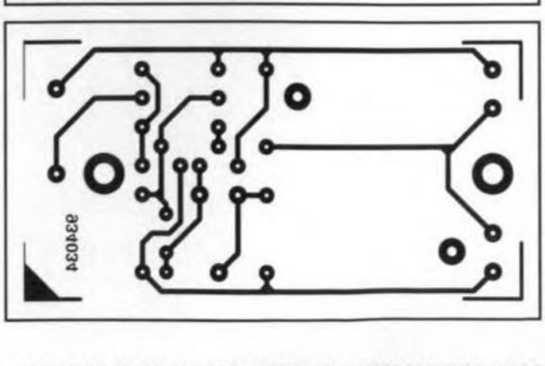
Semiconductors:

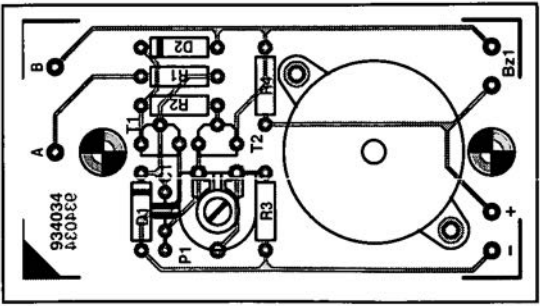
2	1N4148	D1;D2
1	BC547B	T1
1	BC557B	T2

Miscellaneous:

1	Passive (AC) piezo-ceramic buzzer	Bz1
1	9-V (PP3) battery with clip-on leads	Bt1
1	Probe enclosure, e.g., Type 52.68.86-66 (Conrad)	







B
A
934034
D2
R1
R2
T1
C1
D1
P1
R3
R4
T2
Bz1
+
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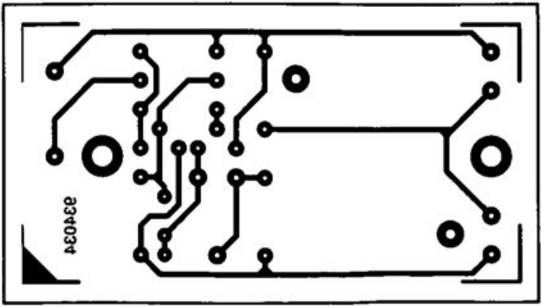
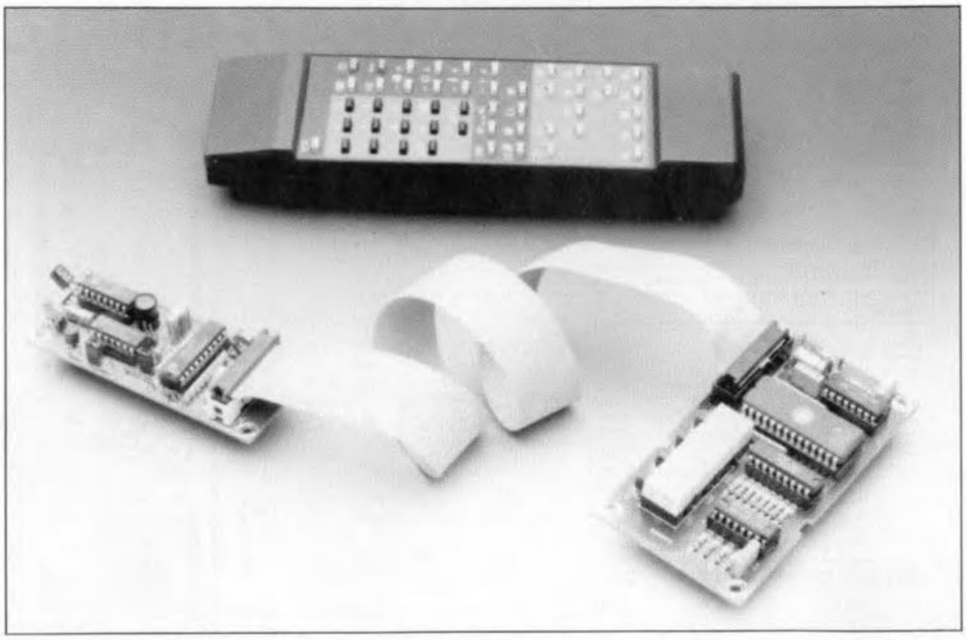


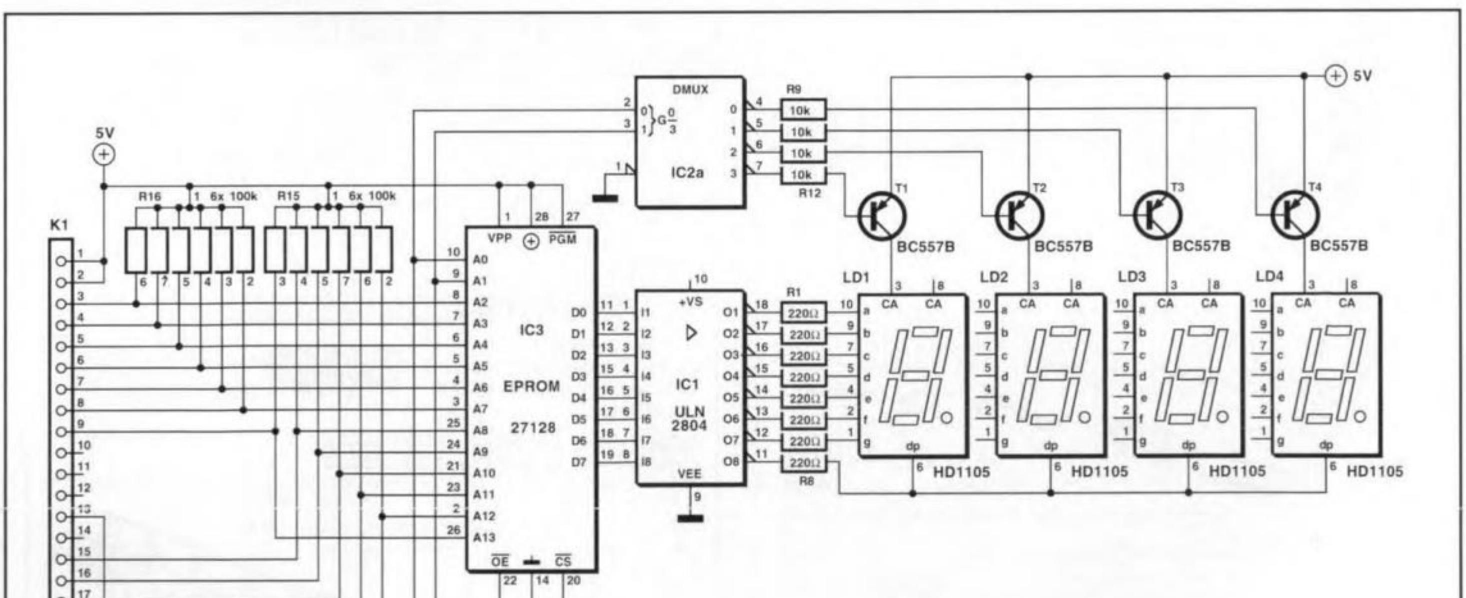
934034

MULTI-PURPOSE DISPLAY DECODER

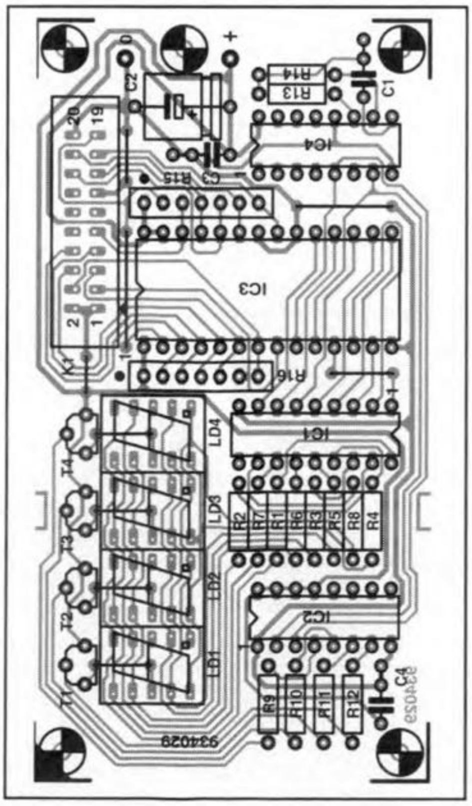
DESIGNING a suitable LED display for a particular application can be time consuming. In particular, the decoder that sits between the driver circuit and the display often needs to be made to measure. The main advantage of the circuit presented here is that the same hardware can be used to implement many different types of display decoding. This is achieved by using an EPROM, which most readers will be able to program or have programmed.
The circuit translates a 12-bit input code into suitable drive signals for a 4-digit readout built from 7-segment common-anode LED displays. Four digits (displays) require four bytes in the EPROM. If these bytes are stored at four successive addresses, they can be read by applying binary code 00 through 11 to the two least significant address lines, A0 and A1. Since the four displays are multiplexed, driving the A0 and A1 lines in this way enables codes to be fetched from the EPROM and indicated on the readout. This leaves the remaining address bits of the EPROM free to apply the code to be converted.
As an example, an EPROM is programmed to function as a code translator for an RC5 infra-red remote control

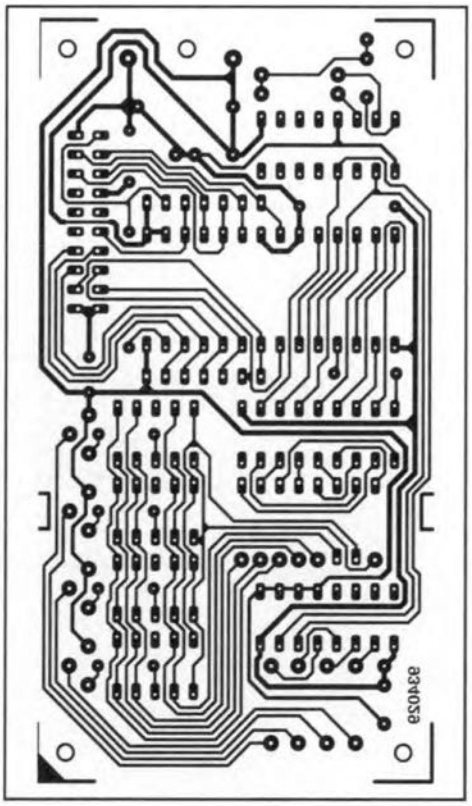


transmitter. The RC5 receiver described in Ref. 1 may be connected directly to the decoder input connector, K_1, via a short flatcable. An RC5 code consists of five address bits and a 6-bit key code. In this example, the two most significant digits are used to show the address, and the other two to show the key code. The six data bits, the five address bits and the toggle bit are connected to the EPROM address lines. The toggle bit is used to drive the decimal point on the digit at the extreme right, which provides an indication that a key has been pressed on the remote control unit.
The contents of the EPROM are generated with the aid of the Pascal program



```
program IR_display_decoder;

uses dos,crt;

var
count : word;
i, j, k, l, m : integer;
g : file of byte;
displayandtoggle : array[0..9] of byte;

const
{The display array contains the databyte for each display value}
display : array[0..9] of byte =($3f,$06,$5b,$4f,$66,$6d,$7d,$7,$7f,$6f);

begin
 assign (g,'display.dat');                              {Open the desired filename}
 rewrite (g);
 for i:=0 to 9 do displayandtoggle[i]:=display[i]+$80;          {Include toggle bit}

                                                                 {Calculate readout}
for m:=1 downto 0 do                                 {Toggle led on and off addresses}

                                                               {Maximum count is 31 63 }
 for i:=3 downto 0 do                            {i and j define the received IR address}
   for j:=9 downto 0 do
    begin
      if (i=3) and (j=9) then j:=1;
      for k:=6 downto 0 do                      {k and l define the received IR key code}
        for l:=9 downto 0 do
                              {Calculate the four display bytes including the toggle bit}
           begin
             if (k=6) and (l=9) then l:=3;
             if m=1 then  write (g,displayandtoggle[l])
                   else  write (g,display[l]);
             write (g,display[k]);
             write (g,displayandtoggle[j]);
             write (g,display[i]);
           end;
  end;
close (g);                                                {Close the destination file}
end.
```



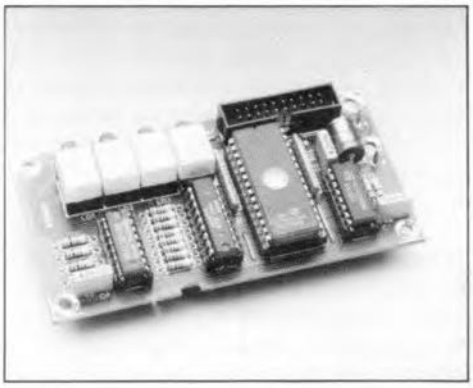


shown. When analysing the program, bear in mind that the outputs of the IR receiver are active low. This means that counting has to be done backwards, that is, from 31 to 0 for the addresses, and 63 to 0 for the data, rather than in rising order, 0 to 31 and 0 to 63, respectively. In this way, many decoder variants may be implemented simply by adapting the EPROM contents as required. The EPROM to implement the RC5 code readout is available ready-programmed through the Readers Services Reference 6261 (page 112).

(A. Rietjens — 934029)

Reference:

1. Universal RC5 code infra-red receiver. *Elektor Electronics* January 1992.



COMPONENTS LIST

Resistors:

8	220Ω	R1-R8
5	10kΩ	R9-R12;R14
1	100kΩ	R13
2	6-way 100kΩ SIL	R15;R16

Capacitors:

1	10nF	C1
1	220µF 16V	C2
2	100nF	C3;C4

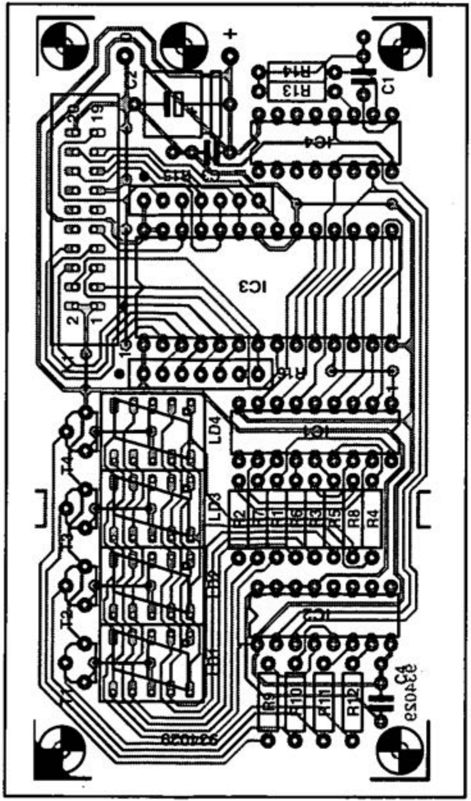
Semiconductors:

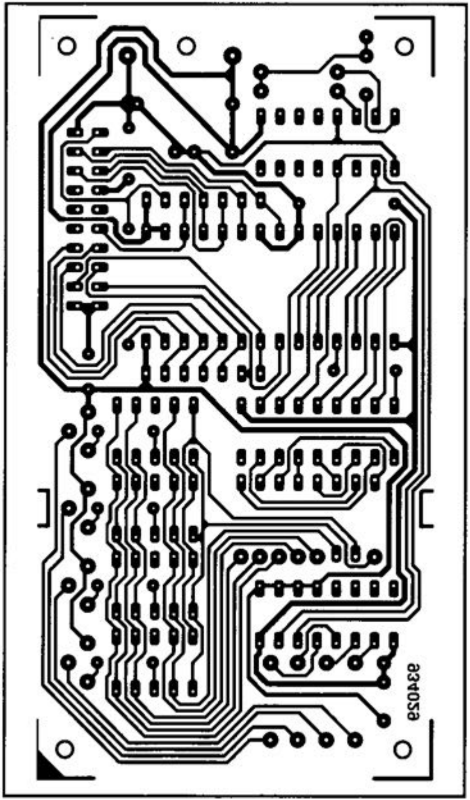
4	BC557B	T1-T4
1	ULN2804	IC4
1	74HCT139	IC2
1	27128 EPROM (order code 6261; see page 110)	IC3
1	4060	IC4

Miscellaneous:

1	20-way straight box header	K1
4	HD1105O *	LD1-LD4
1	Enclusure Heddic 222	
1	Printed circuit board 934029 (see page 110)	

* Siemens product. ElectroValue, 28 St Jude's Road, Englefield Green, Egham, Surrey TW20 0HB. Telephone: (0784) 442253. Fax: (0784) 460320.


IC3
LD4
LD3
R13
R14
C1
C2
R2
R7
R1
R6
R3
R5
R8
R4
R9
R10
R11
R12
C4
T1
T2
T3
T4



MULTI-PHASE RECTANGULAR WAVE GENERATOR

The circuit consists essentially of two distinct parts. A rectangular-wave oscillator, based on IC_1, generates two signals which may be phase-shifted between 0° and 360° with respect to one another. The frequency of the signal at a supply voltage of 5 V is around 1 kHz and at a supply voltage of 12 V about 1.4 kHz.

The stage based on by-eight-counter/divider IC_2 provides eight signals, each of which is 45° out of phase with the preceding and the following signal.

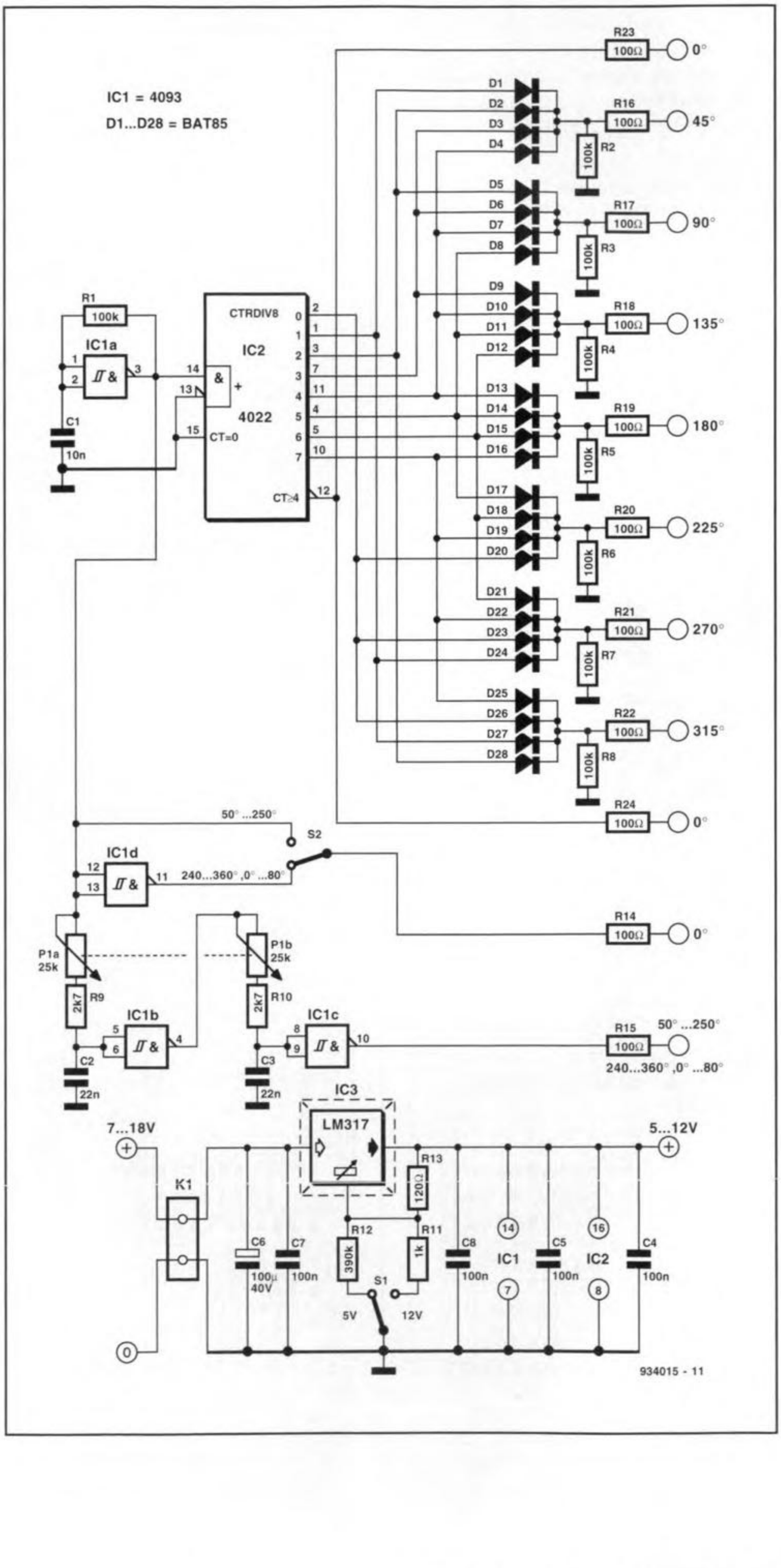
The output of IC_{1a} may be shifted between 50° and 250° with networks P_{1a}-R_9-C_2 and P_{1b}-R_{10}-C_3. Since the 0° signals may be inverted (which appears like a 180° shift) with IC_{1d} and S_2, the range 240°-80° is also covered with P_1 (via 360°=0°). The 0° output at R_{14} must not be confused with those at R_{23} and R_{24}, because (since IC_2 is a by-eight divider), the signals at the R_{14} output and the other two differ in frequency by a factor eight.

Each counter state of IC_2 is reached when the associated output (of eight) goes high. If the counter cycle is thought of as a circle (360°), each counter state corresponds to a phase angle of 45°. The eight output signals are obtained by connecting four of the IC_2 outputs to the circuit outputs via an OR gate for every subsequent step in the count cycle. In this way, the circuit output is a square-wave (duty factor 1:1). The OR gates are formed by D_1–D_{28}. No OR gate is required for producing a 0° signals, since that is already arranged by the carry output of IC_2 (CT≥4) which is high only when the counter state is ≥4.

The use of the circuit is enhanced by S_1 which allows the supply voltage to be switched between 5 V and 12 V.

The current drawn by the generator is negligible.

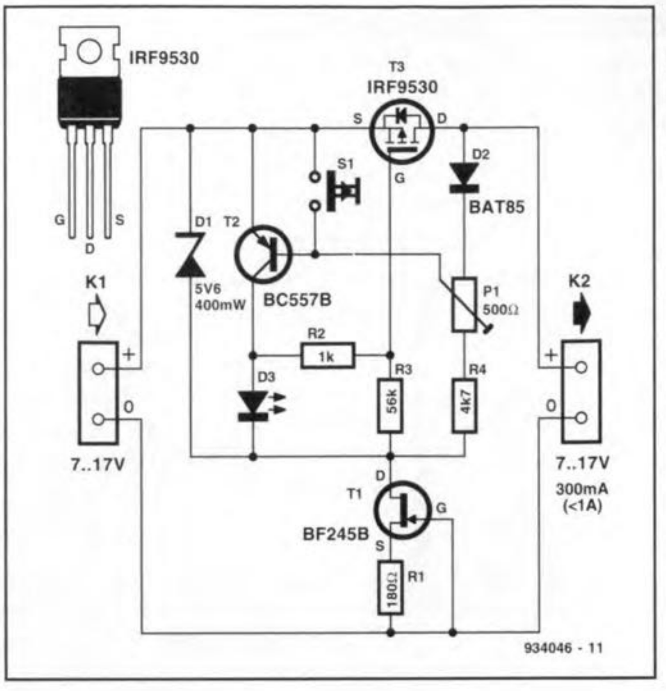
[L. Pijpers - 934015]



ELECTRONIC FUSE

Most modern voltage regulators have an integral current limiter. These are, unfortunately, normally provided with a thermal 'fuse', which in the first instance does not control the current but the dissipation. Since this type of fuse is very slow, in many applications a standard glass fuse is added to protect a number of components against over-current. Also, many small power supplies and mains adaptors have no short-circuit protection. During the design and testing of new equipment, this may have catastrophic consequences.

In all these cases, the fast electronic fuse described here may be found of great value. The fuse is connected between the power supply output and the load. After it has 'blown', and the current has returned to a normal value, it can be reset with S_1.



Transistor T_1 and resistor R_1 form a 6 mA current source that, together with zener diode D_1, provides a constant voltage of 5.6 V with respect to the +ve supply line. This gives the gate of T_3 a sufficiently high negative voltage to keep this p-channel MOSFET conducting. (and thus providing the link between power supply and load).

When the load current rises to about 1 A the potential drop across T_3, D_2, and part of P_1 reaches a level at which T_2 switches on, so that the gate-source junction of T_3 is short-circuited. Diode D_3 then lights to indicate that the fuse has 'blown'. In practice, the LED will already light dimly when the critical current is approached. The circuit is reset by pressing S_1.

With values as shown in the diagram, the maximum load current is around 300 mA. The drop across T_3 is only 100 mV. Higher load currents may be accommodated by taking a higher rated p-channel MOSFET for T_3 or connecting two or more FETs Type IRF9530 in parallel. A heat sink is not needed, since the maximum dissipation does not exceed 40 mW.

The LED should be a low-current type, as a current of only 6 mA is available.

The circuit is calibrated by providing a current into K_1 at the level at which the fuse should operate and then adjusting P_1 until the fuse 'blows'.

As already stated, other types of p-channel MOSFETs may be used for T_3. Note, however, that the lower the on-resistance of this transistor, the less precise the value of the load current at which the fuse will act.

[J. Schäper - 934046]

LOGIC LEVEL TESTER

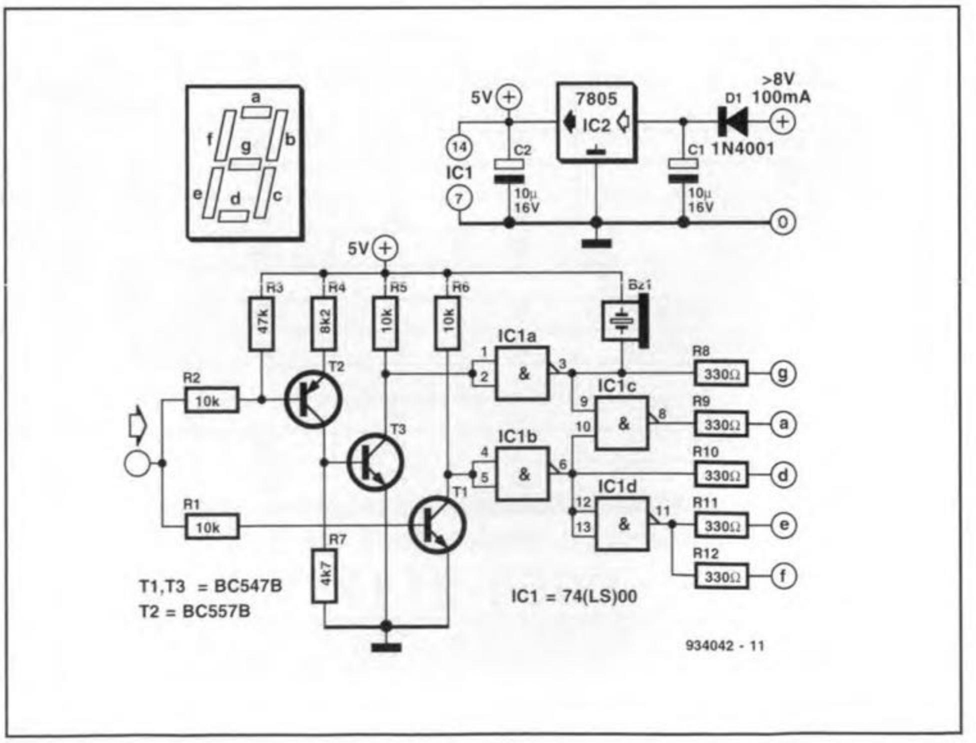
The tester shows on a seven-segment, common-cathode display whether its input is logic 1 (H on the display) or logic 0 (L on the display). An undefined level is indicated as 'n'.

When the input is low, T_1 is switched off and T_2 and T_3 conduct. This results in a high output at IC_{1a} and a low one at IC_{1b}, whereupon segment 'd' lights. Together with permanently actuated segments 'e' and 'f', this causes an 'L' to be displayed.

When the input is high, T_1 is on and T_2 and T_3 are off. The output of IC_{1a} is then low and that of IC_{1b} is high. Apart from segments 'e' and 'f', 'b', 'c', and 'g' also light, which causes an H to be displayed.

When the input is undefined, or open, all transistors are on (because of R_1, R_2 and R_3). The outputs of IC_{1a} and IC_{1b} are then both high, which causes segments 'a', 'b' and 'c' to light together with 'e' and 'f'; an 'n' (no level) is then displayed.

The buzzer is not essential, merely useful; it may thus be omitted. If it is used, it should be a type with integral oscillator.



The type of the components used is not critical. The display may be any seven-segment, common-cathode type. IC_1 may be a standard 7400 or an LS version of this or the open-collector version 7401; HC and HCT types are less suitable because these cannot provide the required output current.

The switching thresholds of the tester are at 1 V and 3 V, which is not standard, but in practice they do well. The 3 V threshold may be lowered slightly by increasing the value of R_4.

The input impedance of the tester is 5 kΩ, which has no effect on the circuit on test.

The current drawn by the tester is determined by the type of display; in the prototype, it is about 60 mA.

[Iyer Mahesh Nagajaran - 934042]

ACTIVE BASS CORRECTION

Active loudspeakers offer the only way of obtaining good bass reproduction from inexpensive or small enclosures. The design described does not make use, therefore, of large, heavy enclosures to obtaion a good result, but of acoustic feedback. A microphone placed in close proximity of the bass drive unit unfailingly registers every movement of the loudspeaker. It is, of course, important that proper attention is paid to the maximum movement of the speaker.

The microphone output is coupled into the negative feedback loop of the output amplifier. In this way, the input signal to the amplifier is compared with the acoustic signal produced by the speaker. In practice, this arrangement appears to work well only with low-frequency signals. Experiments have shown that if the microphone is placed about 10 mm from the cone of the woofer, signals at frequencies of up to 500 Hz are fed back faithfully. To make absolutely certain of correct operation, in the present circuit the upper frequency has been set to 300 Hz; above it, the correcting action gradually ceases. Note, however, that the phase behaviour of the loudspeaker is corrected also for signals above 300 Hz. If the change-over frequency of the cross-over filter of the loudspeaker lies at 300 Hz, it is advisable to make the cut-off frequency of the present circuit, determined by R_6-C_8, lower than 300 Hz.

The gain of IC_2 over the operating range of the circuit is 20 dB, which reduces to 0 dB for frequencies above 300 Hz. This amplifier, which provides the correction up to the cut-off point, also serves as buffer for the microphone signal.

Preset P_1 serves to set the signal level on the basis of the power rating of the outpu amplifier and the efficiency of the microphone. If this control is set too high, corection is also applied to frequencies above the cut-off point; if it is set too low, little correction will be applied and signals between 20 Hz and 300 Hz will increase along a standard 1st order characteristic.

The choice of microphone is a matter of some experimentation, particularly with high-power amplifiers. That used in the prototype proved to work well with low-power systems with a relatively low efficiency. If another type is used, make sure that the potential across the microphone is about half the supply voltage. This is arranged by R_8 and R_9. Also, make sure that the cut-off point set by P_1-C_9 remains well below 20 Hz (no signal at P_1 results in an increase of the final amplification).

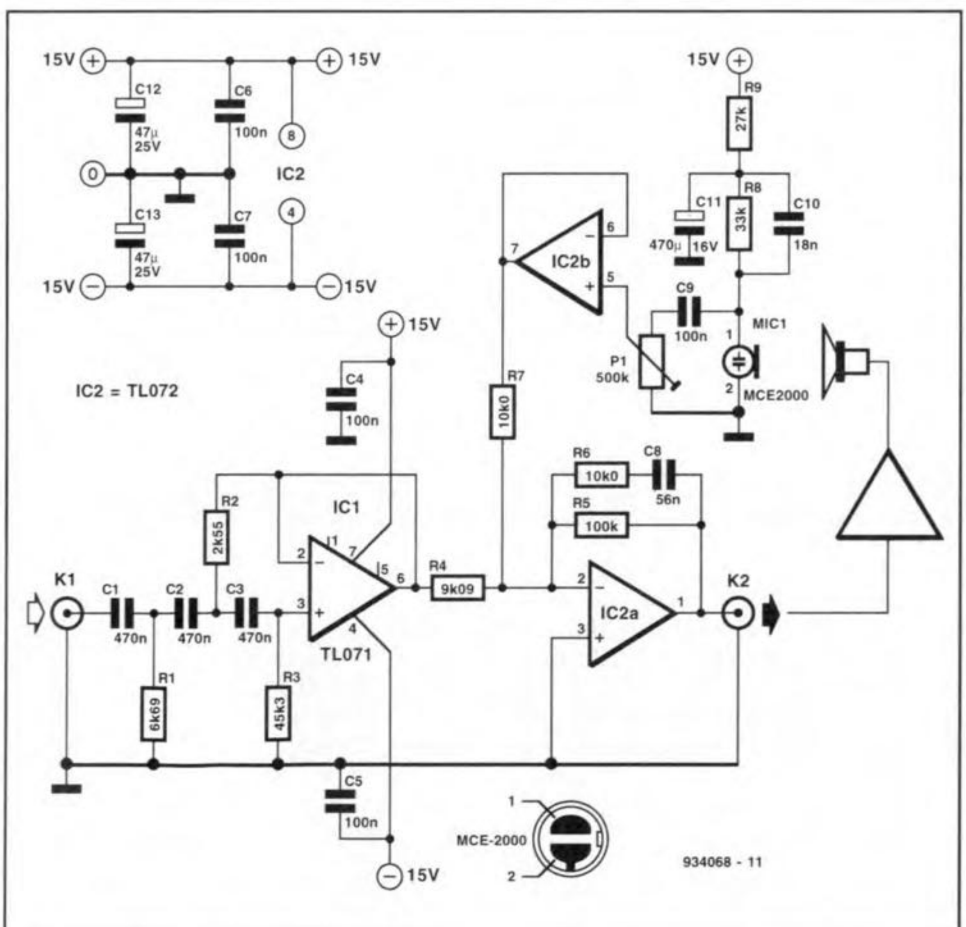
The frequency up to which the microphone signal is compensated is determined by R_8-P_1-C_{10}. This time-constant must be equal to R_6-C_8.

The present circuit can magnify frequencies down to 20 Hz by roughly 20 dB. Since most loudspeakers cannot cope with that frequency, the circuit includes a 3rd-order Butterworth section with a cut-off point of 37 Hz. This frequency may be altered by changing the values of C_1, C_2, and C_3. This filter prevents the loudspeaker being loaded with signals which it cannot reproduce.

The correction circuit is of particular use with active loudspeaker systems. Make sure that the loudspeaker phase is shifted by 180° to prevent positive feedback. This may be done by adding an inverter-buffer before K_2.

The circuit draws about ±6 mA, of which only 0.25 mA is drawn by the microphone.

[T. Giesberts - 934068]



UHF REMOTE CONTROL TRANSMITTER

This low-power UHF FM transmitter is intended for remote control applications such as garage doors and wireless alarm systems. It is a single transistor design that operates at a frequency reserved for low-power wireless signalling. The operating frequency is determined by a resonator, Fl_1. The following types may be used:

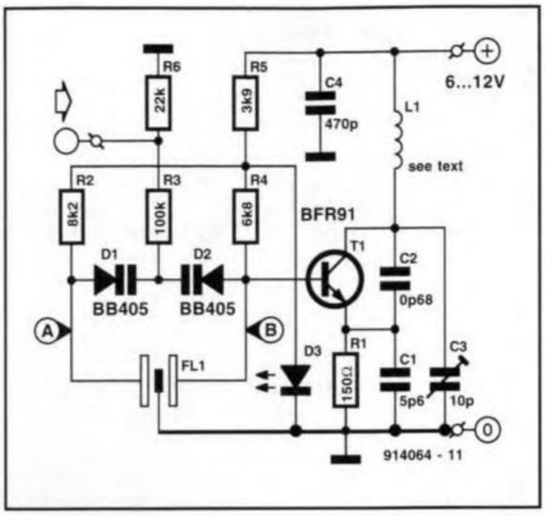
UK: Type R2528 for 418 MHz;
USA: Type R1530 for 315 MHz;
Holland, Belgium, Germany: Type R2554 for 433.92 MHz;
France: Type R2523 for 224.5 MHz.

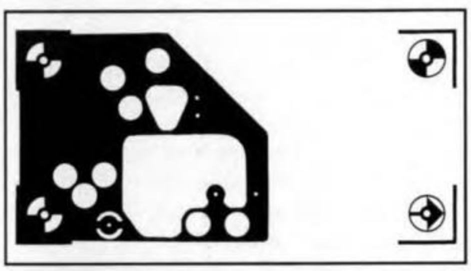
The transmitter is frequency-modulated by an audio (or digital) signal applied to the junction of a varactors D_1, D_2 via R_3. The varactors effectively change the shunt capacitance of the resonator as a function of the modulation signal, which results in FM. The design may be changed to produce AM (amplitude-modulation) by omitting components D_1, D_2, R_2, R_3 and R_6, and interconnecting points 'A' and 'B'.

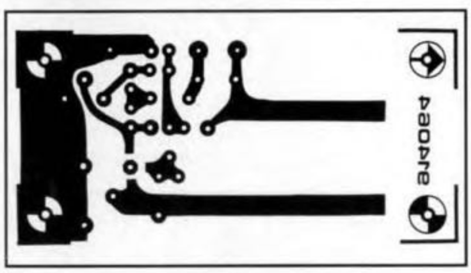
Transistor T_1 is fitted at the solder side of the printed circuit board, as indicated by the dashed outline on the component overlay. Also on this side of the PCB is a short wire link (use silver-plated wire) that connects strip line inductor L_1 to the positive supply track that runs in parallel with it. The position of the wire link depends on the transmit frequency. The lower the frequency, the more inductance is required, that is, the further the link has to be 'moved' towards the edge of the PCB. Some experimenting may be necessary to find the best position. Start by setting the trimmer, C_3, about mid-way, and fit the wire link about 'half way' on the strip line. Monitor the received signal, and adjust C_3 until a maximum is found. If you can not find a maximum, try moving the wire link either towards the transistor (smaller inductance), or towards the PCB edge (greater inductance). The best position is that at which C_3 'peaks' when set about mid-way.

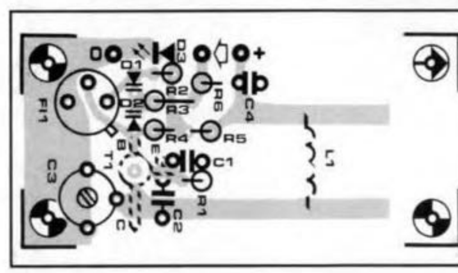
Since we are dealing with a UHF circuit, it goes without saying that all component terminals must be kept as short as possible. The transmitter must be housed in a plastic enclosure to enable it to radiate.

[J. Barendrecht - 914064]









PARTS LIST

Resistors:

1	150Ω	R1
1	8kΩ2	R2
1	100kΩ	R3
1	6kΩ8	R4
1	3kΩ9	R5
1	22kΩ	R6

Capacitors:

1	5pF6 ceramic	C1
1	0pF68 ceramic	C2
1	10pF trimmer	C3
1	470pF ceramic	C4

Semiconductors:

1	BFR91	T1
1	BB405	D1;D2
1	LED 5mm red	D3

Miscellaneous:

1 Ceramic resonator (see text).Fl1
Silver-plated wire 0.8 mm dia.

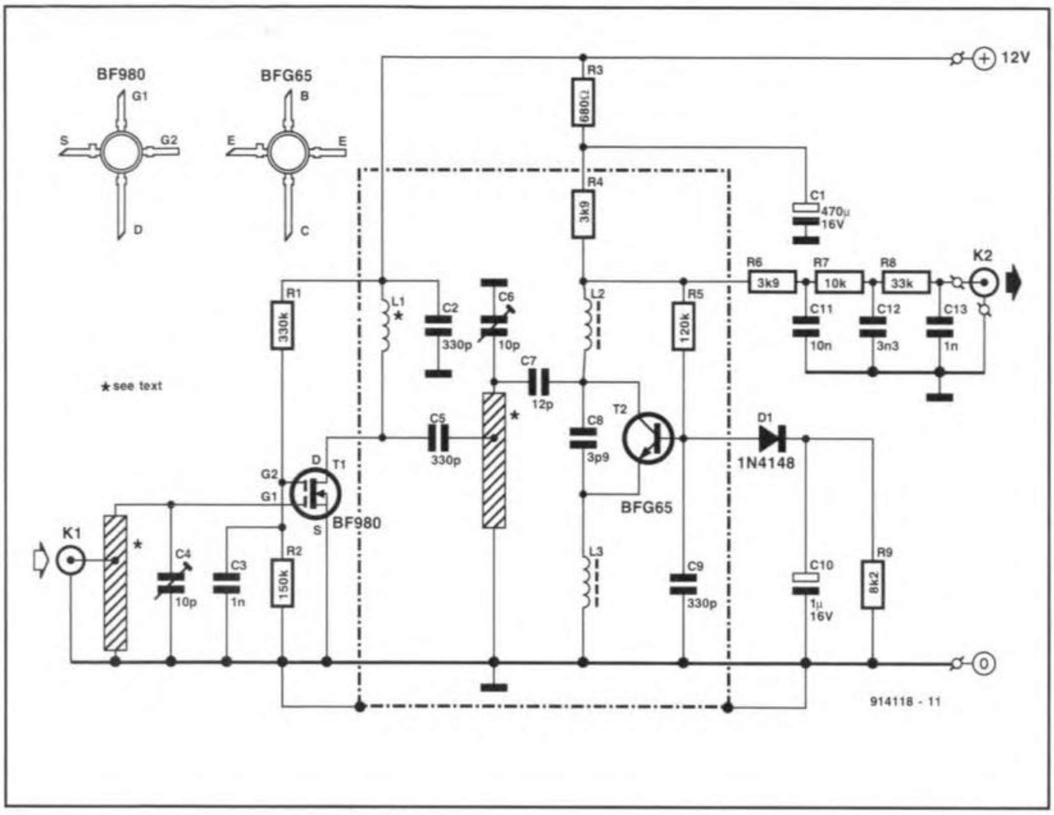
Note that radio regulations apply to the possession and use of this transmitter. In the United Kingdom, a licence must be obtained from the Department of Trade and Industry. [Editor]

UHF REMOTE CONTROL RECEIVER

The receiver is intended primarily for use with the remote control UHF transmitter described in the preceding article. It is a super-regenerative type with an active RF amplifier, T_1. The antenna signal is applied to the input inductor via a BNC socket, K_1. The input circuit is tuned by trimmer C_4.

The amplified RF signal is applied to the input of the super-regenerative stage based on transistor T_2. Although the oscillator is, strictly speaking, not tuned, it will lock on to the amplified RF signal applied via coupling capacitor C_7. The low-frequency modulation component is extracted from the oscillator signal with the aid of low-pass filter, R_6-C_{11}-R_7-C_{12}-R_8-C_{13}. The signal level at the demodulator output is 50 to 800 mV_{pp}, so that further amplification is required before the signal can be applied to a digital input.

The inductors in the RF amplifier input and output are made from 1 mm dia. silver-plated wire. The length of the pieces of wire is indicated by the component overlay. The wires run at a height of about 3 mm above the board surface. Note that the stator terminal of C_4 is bent upwards and soldered direct to the input inductor. The same goes for junction C_6-C_7, which is soldered 'in the air', directly on to the hot end of the in-



ductor wire.

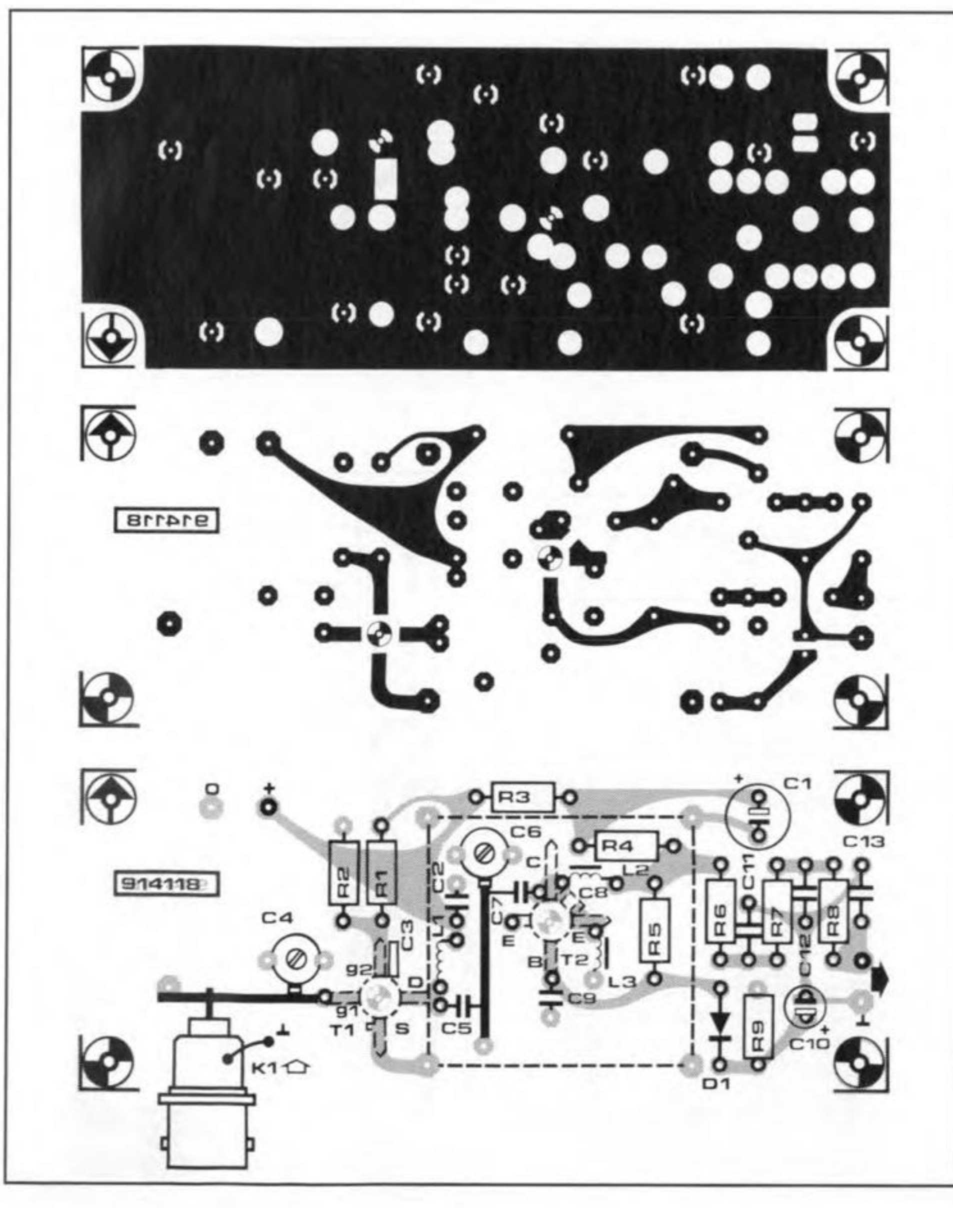
Inductor L_1 consists of 12 turns of 0.6-mm dia. enamelled copper wire. Its internal diameter is 3 mm.

Each of chokes L_2 and L_3 consists of 4 turns of 0.2-mm dia. enamelled copper wire through a 3 mm long ferrite bead.

Capacitor C_8 is a surface-mount technology (SMT) type which is fitted at the solder side of the board, as are the BFG65 and the BF980. The type indication printed on the transistors is legible from the component side of the board.

As indicated by the dashed lines on the component overlay, the super-regenerative section of the circuit must be screened from the rest. To do this, it is best to solder a 20 mm high tin plate box on to the PCB as indicated.

[J. Barendrecht - 934118]



PARTS LIST

Resistors:

1	330kΩ	R1
1	150kΩ	R2
1	680Ω	R3
2	3kΩ9	R4;R6
1	120kΩ	R5
1	10kΩ	R7
1	33kΩ	R8
1	8kΩ2	R9

Capacitors:

1	470µF 16V radial	C1
3	330pF ceramic	C2;C5;C9
1	1nF coffin-style leadless ceramic	C3
2	10pF trimmer	C4;C6
1	12pF ceramic	C7
1	3pF9 SMA	C8
1	1µF 16V radial	C10
1	10nF	C11
1	3nF3	C12
1	1nF	C13

Semiconductors:

1	BF980 or BF966	T1
1	BFG65	T2
1	1N4148	D1

Inductors:

Construction details are given in the text.

2 3-mm ferrite bead

0.6-mm dia. enamelled copper wire

0.2-mm dia. enamelled copper wire

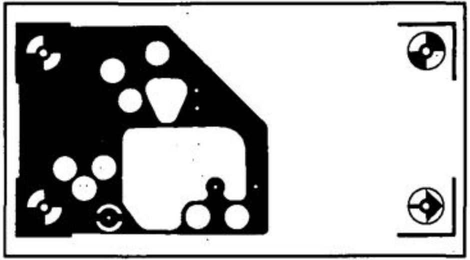
1-mm dia silver-plated copper wire

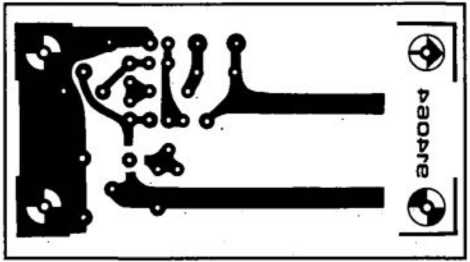
Miscellaneous:

2	BNC socket	K1;K2

the junction of a varactors D_1, D_2 via R_3. The varactors effectively change the shunt capacitance of the resonator as a function of the modulation signal, which results in FM. The design may be changed to produce AM (amplitude-modulation) by omitting components D_1, D_2, R_2, R_3 and R_6, and interconnecting points 'A' and 'B'.

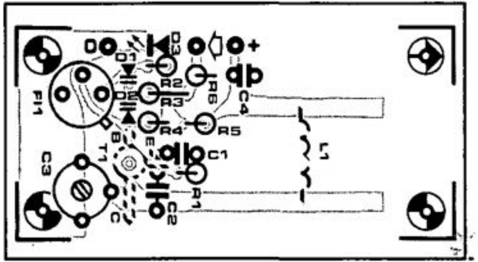
Transistor T_1 is fitted at the solder side of the printed circuit board, as indicated by the dashed outline on the component overlay. Also on this side of the PCB is a short wire link (use silver-plated wire) that connects strip line inductor L_1 to the positive supply track that runs in parallel with it. The position of the wire link depends on the transmit frequency. The lower the fre-

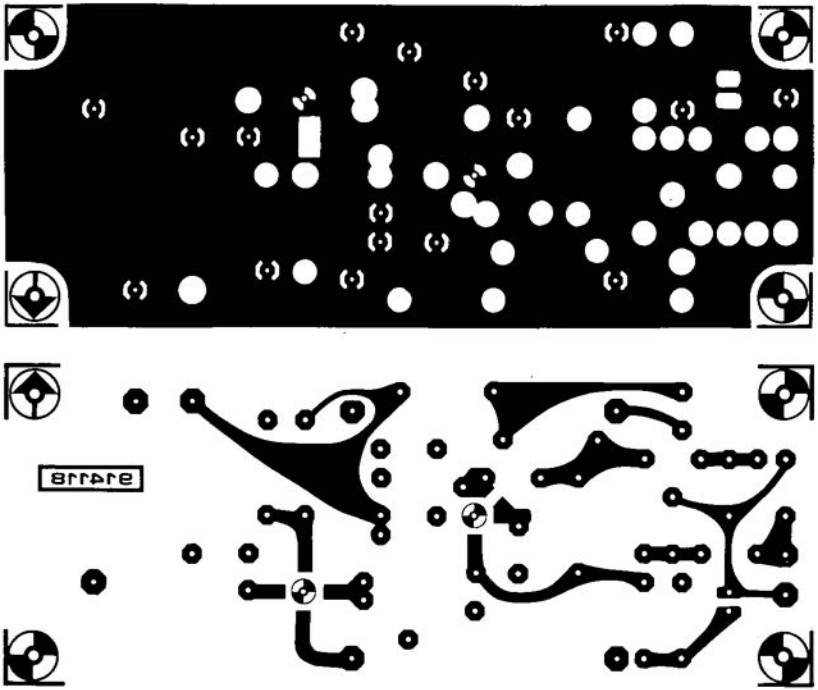




1	BFR91	T1
1	BB405	D1;D2
1	LED 5mm red	D3

Miscellaneous:
1 Ceramic resonator (see text).FI1
Silver-plated wire 0.8 mm dia.





PARTS LIST

Resistors:

1	330kΩ	R1
1	150kΩ	R2
1	680Ω	R3
2	3kΩ9	R4;R6
1	120kΩ	R5
1	10kΩ	R7
1	33kΩ	R8
1	8kΩ2	R9

Capacitors:

1	470µF 16V radial	C1
3	330pF ceramic	C2;C5;C9
1	1nF coffin-style leadless ceramic	C3
2	10pF trimmer	C4;C6
1	12pF ceramic	C7
1	3pF9 SMA	C8
1	1µF 16V radial	C10
1	10nF	C11
1	3nF3	C12
1	1nF	C13

LEAD-ACID BATTERY CHARGER

To make sure that a lead-acid battery (whether standard or sealed) is always fully charged, a constant voltage should be applied to it. Chargers that check whether the battery is still charged and only commence charging when it is (nearly) flat, do not guarantee that it is fully charged when it is required in an emergency.

The charger monitors the battery voltage constantly and tops up the charge if and when required. This has advantages and drawbacks. The battery is always fully charged, but it becomes sluggish, that is, it does not take the charge well. This effect disappears after normal use (a few duty cycles). If the voltage is not too high (≤13.8 V), the battery remains charged and will have a long life. It is not advisable to constantly apply a high voltage of, say, 14.4 V across the battery (this is in any case not necessary, because at 13.8 V it remains fully charged). The only reason for applying a high voltage (temporarily) is that this charges a flat battery quicker.

Always use a commercially available charger: this meets safety requirements and is normally relatively cheap. A partly discharged battery may be connected to such a charger without any precautions.

When the battery is fully charged, do not leave it connected to the charger, because that provides quite a high voltage. It is then that the charger comes in handy. The voltage at which charging should cease is preset with P_1. Turn the preset fully anti- clockwise and connect a voltmeter across the battery. The LED will not light. Turn P_1 slowly clockwise till the LED just lights: the battery is then being charged. When the battery voltage has reached the required value, turn P_1 slightly anti-clockwise until the LED goes out. This happens fairly gradually, so it is best to do this in semi-darkness if possible.

The 723 is powered by the battery via D_2 (current about 10 mA). The internal reference voltage is reduced to 2.2 V by R_1 and R_2. This voltage is compared with the divided battery voltage at the wiper of P_1. If it is lower, the output (pin 10) goes high, whereupon D_1 lights and the optoisolator switches on the thyristor. This results in the charger output being applied to the battery: the level of the current is determined by the charger. Since the output of the charger is an unfiltered, rectified alternating current, the thyristor will switch off at every zero crossing, but immediately switch on again if the battery voltage is not high enough.

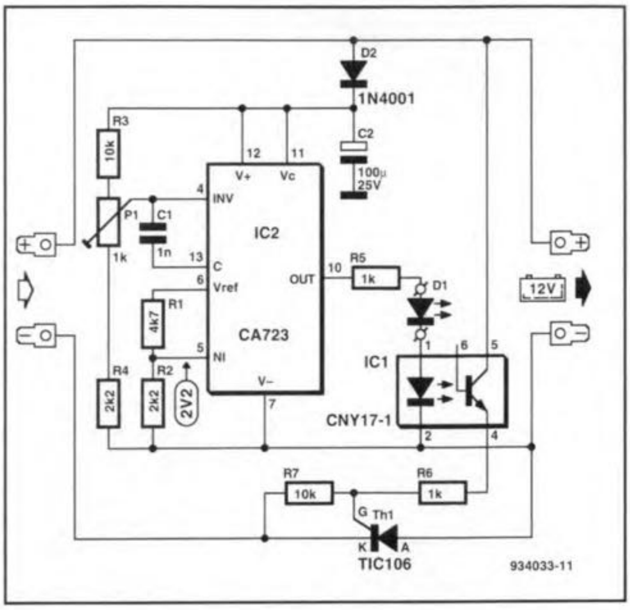
It is best to leave the monitor between the charger and the battery. Charging will then take a little longer because there is a loss of about 1 V across the thyristor.

The thyristor must be mounted on a heat sink if the current exceeds 1 A; the maximum current must not exceed 5 A.

The circuit may be built on the PCB illustrated. A large part of the board is occupied by the heat sink for the thyristor. For the four terminals that carry fairly high currents, use flat connectors that can be screwed on to the board

The board may be fitted neatly in a small case for which a front panel foil is available—see p.110.

[K. Walraven - 934033]



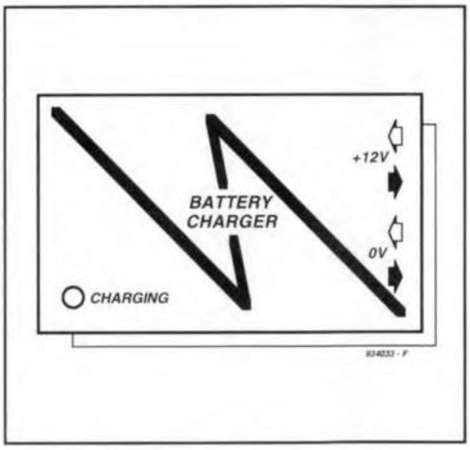
PARTS LIST

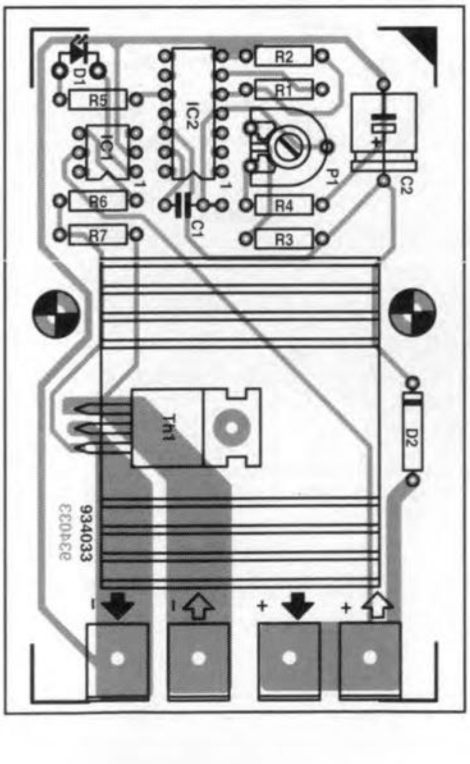
Resistors:
R1=4.7 kΩ
R2, R4=2.2 kΩ
R3, R7=10 kΩ
R5, R6=1 kΩ
P1=1 kΩ preset

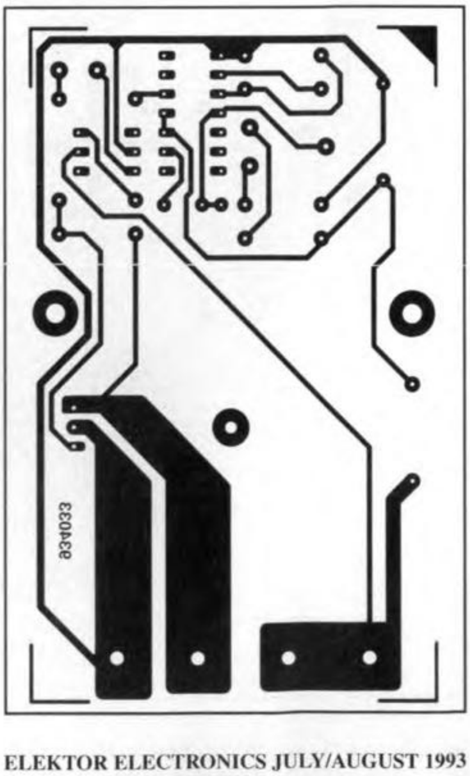
Capacitors:
C1=1 nF
C2=100 μF, 25 V

Semiconductors:
D1=LED, 3 mm, red
D2=1N4001
Th1=TIC106
IC1=CNY17-1
IC2=CA723

Miscellaneous:
Heat sink for Th1,e.g., SK59
4 flat connectors with screw fitting
1 case 95×60×24 mm ($3^3/_8$×$2^3/_8$×1 in)
PCB Type 934033
Front panel foil Type 934033-F



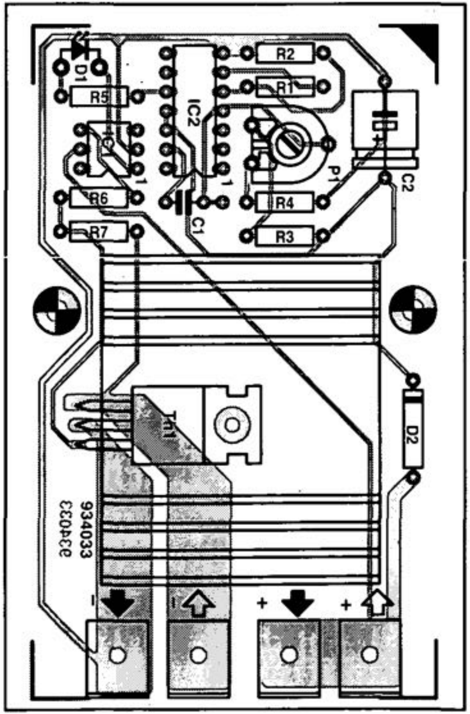


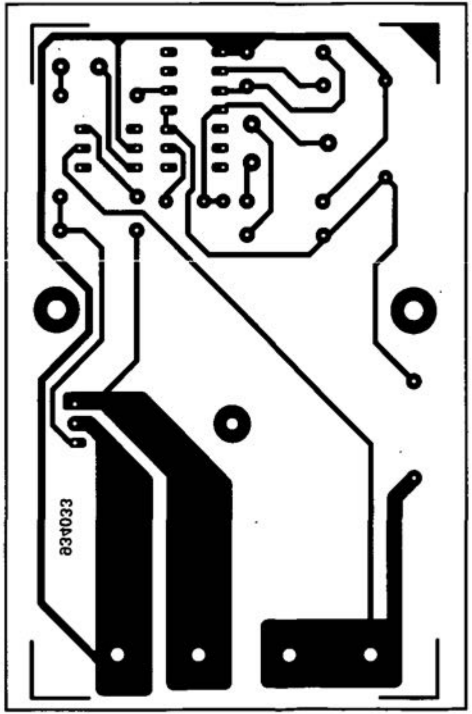


till the LED just lights: the battery is then being charged. When the battery voltage has reached the required value, turn P_1 slightly anti-clockwise until the LED goes out. This happens fairly gradually, so it is best to do this in semi-darkness if possible.

The 723 is powered by the battery via D_2 (current about 10 mA). The internal reference voltage is reduced to 2.2 V by R_1 and R_2. This voltage is compared with the divided battery voltage at the wiper of P_1. If it is lower, the output (pin 10) goes high, whereupon D_1 lights and the optoisolator switches on the thyristor. This results in the charger output being applied to the battery: the level of the current is determined by the charger. Since the output of the charger is an unfiltered, rectified alternating current, the thyristor will switch off at every zero crossing, but immediately switch on again if the battery voltage is not high enough.

It is best to leave the monitor between the charger and the battery. Charging will





SCIENCE & TECHNOLOGY

EMC TESTING OF PMR EQUIPMENT

By Brian P. McArdle

(1) Introduction

The European Community (EC) Directive 89/336 on electromagnetic compatibility (EMC) applies to radiocommunications equipment. It was believed that the directive would assist in the removal of barriers to trade by having common standards throughout the EC. This remains to be seen, but the directive is coming into force and its requirements must be followed. The purpose of this article is to review some of the main tests which, in due course, will become mandatory. The comments are purely personal and the reader is referred to the directive and associated specifications for the exact position.

Before introduction of the directive, type approval tests on equipment for radiocommunications consisted mainly of performance tests. For example, Performance Specification MPT 1326 refers to the testing for voice communications of analogue equipment using angle modulation. The various tests are listed in Appendix 1. There are six transmitter and eight receiver tests which are intended to maintain a basic operating standard for fixed, mobile and handportable units. The individual tests refer to specific operating conditions and parameters are designed to assess the general quality of a unit. In order to obtain type approval, a unit must pass every test. In the case of EMC testing, it is not the quality of a radio as a radio which is in question, but its overall compatibility with other services – either other radio, telecommunications or ISM equipment. For example, in the future, a handportable unit must be tested to ETSI Specification 300-086 (Appendix 2) covering the various performance tests for the type approval of radiocommunications equipment. However, in order to comply with the directive, it must also be tested to ETSI Specification ETS RES 0901, which covers the EMC tests. This is a new draft standard and the tests are discussed in Sections 4 and 5. The principal difference between the two standards is that EMC refers to compatibility requirements between different services. If a unit passes the EMC tests, the official CE mark in accordance with Article 10 of 89/336 may be affixed. If it satisfies both specifications, it meets the requirements for full type approval.

To some extent, the directive cuts across traditional methods of type approval. Originally, radiocommunications equipment had to be type approved separately in every country of the Europen Community. Although the specifications in each state could be very similar, they were still separate standards in their own right with their own documents. In some countries the official government publications describing the tests were actually legal documents. ETSI standards have changed this tradition because the new spcifications are harmonized. The same specification will apply throughout the EC. If a unit is type tested and approved in a member state, then it is approved for all countries of the EC.

The various EMC tests required by ETSI Specification ETS RES 0901 are explained in Sections 2–5. The reader should consult the specification for the exact position and the comments in this article are purely personal.

(2) Performance criteria

A unit must operate in the frequency range 30 MHz to 1 GHz. The tests are divided into two categories: emission and immunity. A unit on test must be in the operating mode. In emission (Fig. 1), the signals from the unit are measured. These are non-functional signals in the sense that they are not intended to be specific emissions due to the fact that the unit is radio equipment. In immunity (Fig. 2), the unit is placed in an environment where it would experience large external signals which could disrupt or block its operation. Again, the external signals need not be functional for radiocommunications but simply large electromagnetic fields. It should be noted that emission and immunity for EMC are not the exact equivalent of transmitter and receiver tests for performance. A receiver must be tested for performance as well as immunity. Similarly, a transmitter must undergo immunity tests. For both categories, the assessment for compatibility examines the quality of the detected audio signals and requires that a communication link between a unit and test equipment should be maintained during each test. There are four main criteria for the full range of tests as follows.

(a) Continuous phenomena applied to analogue transmitters: the detected audio signals should not be distorted by more than 25%. Controls should be serviceable and stored data should not be affected. If a transmitter has a standby mode, transmission should not be actuated.
(b) Transient phenomena applied to analogue transmitter: after each transient, a unit should return to normal operation with no permanent degradation. Controls should be serviceable and stored data should not be affected. If a transmitter has a standby mode, transmission should not be actuated.
(c) Continuous phenomena applied to analogue receivers: the detected audio signals should not be distorted by more than 25%. A unit should continue to operate without loss of controls or stored data. If a unit is a transceiver, transmission should not be actuated.
(d) Transient phenomena applied to ana-

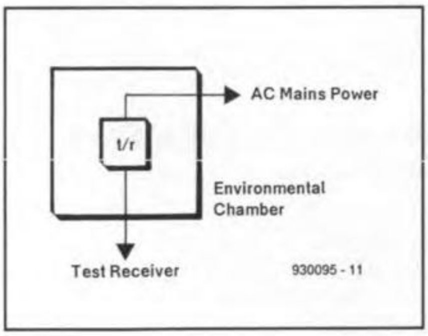


Fig. 1. *Emission tests.* Emissions may be by conduction via cables or by radiation. The tests are carried out in an anechoic chamber to reduce external effects. The measuring apparatus is normally outside the chamber. If the cable from the output to the receiver is replaced by 2 antennas, the method is for radiation.

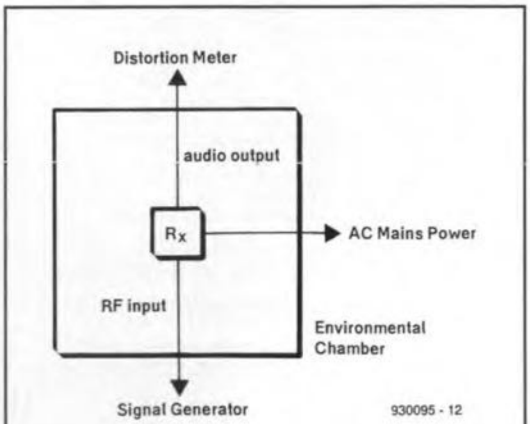


Fig. 2. *Immunity tests.* This arrangement is to test wideband spurious responses in a receiver by conduction via an r.f. cable to the input terminal. The signal source and test instrument are outside the chamber.

logue receiver: after each transient a unit should continue to operate normally with no permanent degradation. Controls and stored data should not be affected or altered. If a unit is a transceiver, transmission should not be actuated.

It should be emphasized that these criteria are not a restatement of performance conditions for PMR equipment. For example, in the case of transmitters, the test for spurious emissions outside the designated channels is not an EMC test (see Appendix 5).

(3) Test methods

All tests must be carried out in the operation mode. In certain cases, tests are repeated in the standby mode. If a unit forms part of a system (for instance, installed in a host unit), all tests must be undertaken on the system with an adequate number of sub-systems installed – at least to the extent to ensure operation of the ports. If a handportable or mobile is powered by a.c. mains, the tests and limits for a base station apply. If a portable unit is powered by a battery in a vehicle, the tests and limits for vehicular equipment are used. These points are discussed further in Sections 4 and 5.

If a unit has a large number of terminals, an adequate number should be used for the various tests to ensure that operational conditions are properly simulated. Input or output ports for radio frequency signals must be terminated to avoid radiation (for instance, by direct connection to dummy non-radiating loads). For equipment of a highly specialized application, the specification can be modified provided that the criteria of Section 2 are followed. However, the new version must not result in a lower level of immunity and the manufacturer must declare the amount of degradation as required by the standard. All tests are carried out within the normal operating anechoic range and at the rated supply voltage. The following points are of special importance.

(a) The term 'normal test modulation' means that the modulating signal is a 1 kHz audio tone at a level to produce 67% of maximum frequency deviation (2.5 kHz for PMR equipment). In certain situations (for instance, voice communication between ATC and aircraft) where AM is still in use, a modulation depth of 60% is specified.

(b) For immunity tests: a communication link between the unit and test equipment must be established at the start of every test and maintained throughout the duration. A transmitter is operated at the maximum output power. For a receiver, the input signal is modulated as in (a) above. For a transceiver on duplex operation (Appendix 2 on additional performance tests), the input signal for the receiver is modulated as in (a) and the transmitter operated at maximum power with modulation applied.

(c) For transmitter tests: an exclusion band of 25 kHz extends around the nominal carrier frequency. The audio signal can be generated internally or externally provided that the modulation is as per (a). The output is connected directly by r.f. cable to test equipment located outside the chamber. If the unit has an integral antenna an no external connector, a receiving antenna installed within the chamber replaces the direct connection.

(d) For transceiver and receiver tests: an exclusion band is determined by the switching range in accordance with the manufacturer's specification as follows. The Lower Limit is the lower limit of the switching range minus 5% of centre frequency or 10 MHz to give the minimum value. The Upper Limit is the upper limit of the switching range plus 5% of centre frequency or 10 MHz to give the maximum value.

(e) For receiver tests: the signal source is located outside the chamber and connected directly by r.f. cable to the receiver input. The level must be 60 dB above the level of minimum sensitivity. For example, if this threshold is 6 dB μV emf, the level should be 66 dB μV emf. If a unit has an integral antenna with no external connector, a transmitting antenna installed in the chamber replaces the direct connection. The audio output is connected via a non-metallic acoustic tube to a distortion meter outside the chamber. Other arrangements may be used, but must not reduce the accuracy of the test.

(f) In the case of receivers and transceivers on duplex operation, the unit may have a response to an unwanted signal which results in distortion at the output. This is partly covered by the performance tests for Spurious Response Rejection (see Appendix 6). However, it could be due to wideband phenomena, which would come under EMC, and it is necessary to distinguish between these two different effects. If a response occurs to an unwanted signal, its frequency is increased by twice the bandwidth of the intermediate frequency stage before the demodulator. The frequency is reduced by the same amount and the test repeated. If the distortion is removed in both situations, the response is considered to be of the narrowband variety. If it continues, the frequency is increased or decreased as appropriate by 2.5 times the bandwidth. If it still persists, the response is deemed to be caused by wideband phenomena and the EMC test standards are applied. Narrowband responses come under performance and not EMC testing.

(4) Emission tests

There are three tests as follows. In each test, the measuring instrument, which is a receiver, must have a certain level of performance to ensure accuracy. CISPR Publication 16, Section 1, gives a detailed breakdown of the receiver characteristics.

(a) **Enclosure** This refers to the emission, by either conduction or radiation, of signals in the range 30 MHz to 1 GHz generated within the unit. The test is not required where the performance specification (Appendices 1 and 2) has tests for spurious emissions (Appendix 5) from the transmitter and from the receiver. Consequently, it does not apply to PMR equipment. It would apply to associated apparatus such as encoders which are not installed within the transceivers, but connected externally as separate units. The electric field strength, generated from inside the unit, is measured at a distance of 10 metres and must be less than 30 dB μV m^{-1} for the range 30 MHz to 230 MHz and 37 dB μV m^{-1} between 230 MHz and 1 GHz.

(b) **DC ports** This test refers to input or output ports provided that the voltage is direct. Each port on a unit is tested in turn with the particular port on test, connected as in Fig. 4 and the other ports terminated into a 50 Ω dummy load. Measurements are made with two methods of detection (Appendix 3) with the limits

Frequency range	Quasi peak	Average
150 kHz–500 kHz	79 dB μV	66 dB μV
500 kHz – 30 MHz	73 dB μV	60 dB μV

Note that the averages are smaller than the quasi peak measurements.

(c) **AC mains power ports** This test is for base and fixed mobiles. It examines the presence of internally generated signals from the unit on the a.c. power input terminals. Measurements are made with the two methods of detection and the limits are:

Frequency range	Quasi-peak	Average
150 kHz–500 Khz	61±5 dB μV	51±5 dB μV
500 kHz–5 MHz	56 dB μV	46 dB μV
5 MHz–30 MHz	60 dB μV	50 dB μV

(5) Immunity tests

There is a total of eight tests as follows. The first two apply to all PMR equipment whether fixed or mobile. In a short article it is not possible to analyse the individual tests to any extent. However, it suffices to state that the criteria of Section 2 have been followed in the design of the tests and interpretation of the results.

(a) **Electrostatic discharge** This assesses the capability of a unit to continue to operate in the presence of an electrostatic discharge applied to exposed surfaces. The only part of a surface which is exempt from the test is the centre pin of an r.f. connector. Test levels are 4 kV for a contact discharge and 8 kV for an air discharge.

(b) **EM Field 80 MHz to 1 GHz** This asses the capability of a unit to continue to operate in the presence of a large external electromagnetic field. The field strength is 3 V m^{-1} and the test is performed over the full range with the exception of the exclusion band (see Section 3). The interfering field is a functional signal for radiocommunications in that it is modulated by a 400 Hz tone to a depth of 80%.

(c) **Radiofrequency Common Mode** This test applies to base and mobiles, but not to handportable stations. It is performed on signal and d.c. power ports where the attached cables are longer than 2 metres and on a.c. power ports. It is different from (b) in that the interfering signal, modulated by a 400 Hz tone to a depth of 80% as in (b), is present at the ports. The frequency range is 150 kHz to 80 MHz with the exception of the exclusion bands (see Section 3).

(d) **Fast transients Common Mode** This test applies to base and fixed mobiles. It is performed on signal and d.c. power ports where the attached cables are longer than 3 m and on a.c. power ports. It examines the ability of a unit to operate in the presence of fast transients at the ports. For signal and a.c. ports, the transient is a 1 kV open-circuit voltage. For an a.c. mains power, it is 2 kV. For the power ports, the transients are applied in parallel to all wires of a cable with reference to the cabinet ground.

(e) **Voltage Dips and Interruptions** This applies only to base and fixed mobile stations where a.c. mains power is in use. It examines the ability of a unit to continue to operate following dips and interruptions at the a.c. input ports. The voltage level is reduced by 50% for 100 milliseconds and by 95% for 500 milliseconds. There is provision for a situation, albeit unlikely, where a mobile or handportable is powered solely by a.c. power with no battery back-up. The test is carried out, but the communication link, as per the criteria of Section 2, need not be maintained.

(f) **Transients Common and Differential Mode** This only applies to base and fixed mobile stations. A unit must continue to operate following the presence of transients on the a.c. input ports. The transients are a 1 kV and 0.5 kV open-circuit voltage for common and differential modes respectively and are applied in parallel to all wires in a cable with reference to the cabinet ground.

(g) **Transients and Surges in Vehicles**. This applies to mobiles and handportables intended for installation or operation from within a vehicle. Transients and surges on the d.c. power input ports must not disrupt operation of a unit. The specification has two tables depending on whether a unit is connected directly to a vehicle's battery.

(h) **Voltage Variation**. This test is not required if an equivalent test is specified by the performance standard. If required, the test is carried out on d.c. input ports which are not connected to dedicated uninterruptable power supplies. For PMR equipment, it is not usually needed. For example, ETSI Specification 300-086 (Appendix 2) requires tests to be undertaken at normal and extreme conditions where the supply voltage is varied by ±10%.

(6) Summary

This article examines the EMC requirements for PMR equipment. Further standards for specific areas will be introduced in due course. It should be noted that the article refers to radiocommunications and not to telecommunications equipment intended for connection to the public telephone network. EMC tests for telecommunications equipment are required by the directive, but are covered by separate specifications. It has been agreed that in cases of overlap the more stringent of the standards will apply.

Although the directive refers to type approval, there is provision in Article 6 for the protection of communications equipment which is experiencing interference owing to electromagnetic incompatibility in an operational situation. Indvidual administrations, in order to overcome interference to an emergency service at a specific site, may impose additional requirements. This is the only case where the directive permits additional protective measures which are not part of the standards of Article 4. An interesting point is that since the directive and, therefore, Article 6 refer only to compatibility, interference, which is caused by an inadequate performance specification (for instance, levels of spurious emissions), is not automatically covered. However, this is unlikely to cause serious problems.

The foregoing comments are purely personal and readers are referred to the directive and specifications. Not all relevant points are included as the article is intended as an overview.

(7) References

1. CISPR Publication 16: 'CISPR Publication for Radio Interference Measuring Apparatus and Measurement Methods'.
2. IEC 801 Series Publications.
3. Anritsu Corporation: 'Practical Advice for EMI Measurements' (2nd Edition, 1991).
4. Anritsu Corporation: 'Personal EMI Measurement System' (1991).
5. Rohde & Schwarz Application Note: 'Electromagnetic Susceptibility Measurements – New Trends for EMS Standards'. *Rohde & Schwarz News*, Number 138, Volume 32 (1992).
6. Rohde & Schwarz: 'EMC Test Technology to Meet the 90s'.

Appendix 1

The tests under normal conditions in Performance Specification 1326 are as follows.

Transmitter

1. Frequency error: ±1 kHz for a base or fixed station;
   ±1.5 kHz for a mobile station.
2. Carrier power: ±1.5 dB of rated value.
3. Frequency deviation: ±2.5 kHz.
4. Adjacent channel power: 60 dB or more below carrier power.
5. Spurious emissions: 1 μW (max) in operating mode for a base or fixed station;
   2.5 μW (max) for a mobile station;
   20 nW in standby mode for a base or mobile station.
6. Intermodulation attenuation: 15 dB or more for any intermodulation component.

Receiver

7. Sensitivity: SINAD of 20 dB or higher for an input of 6 dB μV e.m.f.
8. Amplitude response: variation of 3 dB or lower in the audio level for in-

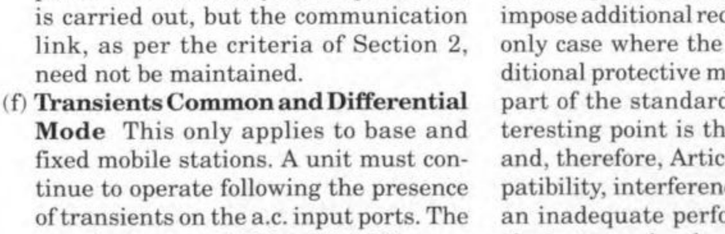


Fig. 3. Quasi-peak detector. The circuit performs a weighted peak measurement.

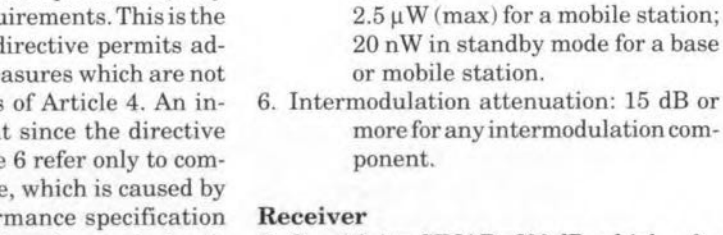


Fig. 4. Line Impedance Stabilizing Network (LISN).

puts at 6 and 100 dB μV e.m.f.

9. Co-channel rejection: –12 dB (min).
10. Adjacent channel selectivity ratio: 60 dB (min).
11. Intermodulation response rejection ratio: 70 dB (min) for a base or fixed station;
65x dB (min) for a mobile station.
12. Spurious response rejection ration: 70 dB (min).
13. Blocking: 90 dB μV e.m.f. (min).
14. Spurious emissions: 2 nW (max).

Appendix 2
ETSI Specification 300-086

This specification is for analogue speech equipment very similar to MPT 1326. Recently, a number of standards have been introduced for digital modulation where the speech signals are represented in digital form (for instance, speech codec for GSM and DSRR), but 300-086 is for angle modulation. There are three additional tests as follows.

1. **Transmitter Transient Behaviour** This is the variation in frequency from the nominal value with time when transmission is actuated and deactuated. The standard has a table for the difference for three time intervals.
2. **Receiver Desensitization in Duplex Operation** This is similar to blocking, but in duplex operation a unit can be transmitting and receiving simultaneously. The sensitivity test is repeated while the unit is also transmitting with modulation applied.
3. **Receiver Spurious Response Rejection in Duplex Operation** The test for spurious response rejection is repeated while the unit is also transmitting an unmodulated carrier. The ratio should be at least 67 dB.

Appendix 3
Quasi-peak and average detection

EMC tests are made in accordance with CISPR recommendations and require quasi-peak and average measurements of signal levels.

Figure 3 illustrates the principle for quasi-peak. The operational amplifier provides the necessary gain. The purpose of the diode is to ensure that the voltage acrosss capacitor C does not fall as the input moves down from a maximum. However, although the capacitor charges up to a maximum through R_1, it does not remain at this value owing to the effect of R_2. These are charging and discharge resistors respectively with the charging time (e.g., 1 ms) very small in comparison with the discharge time (e.g., 550 ms). Hence, the term quasi-peak rather than just peak. The indicator stage for the actual measurements is on the right in parallel with R_2. The circuit performs a weighted measurement of interference signals at the input.

For an average detector, the input is applied direct to R_1 with the operational amplifier and diode removed. In addition, the time constants are equal. The reader should consult Ref. 1 for a detailed analysis.

In addition to the above methods, the measuring apparatus must be protected against unwanted signals, such as those introduced from a power supply. The protective circuit of Fig. 4 is for the frequency range 150 kHz to 30 MHz. CISPR Publication 16 gives a generalized circuit with a table of values for components for different frequency ranges. These points need not be considered further in this type of paper.

Appendix 4
CISPR measurements

In Fig. 4, the measuring receiver on the right is represented by 50 Ω, which is the required input impedance, with an indicator in parallel. It must have certain technical characteristics, which are sometimes referred to as the CISPR standards for interference measurements in manufacturers' catalogues and manuals. The principal characteristics are as follows.

6 dB IF bandwidth: 9 kHz for the range 150 kHz to 30 MHz;
120 kHz for the range 30 MHz to 1 GHz.

Charging time: 1 ms for the full range of 150 kHz to 1 GHz.

Discharge time: 160 ms for the range 150 kHz to 30 MHz;
550 ms for the range 30 MHz to 1 GHz.

Amplitude response: CISPR standard requires a response to certain pulses and repetition rates equal to the response to a sinusoidal unmodulated r.f. carrier at a level of 66 dB μV e.m.f. at the nominal receive frequency.

Selectivity: CISPR standard has a graph of response vs frequency (e.g., response should be at least 20 dB down at 10 kHz from the nominmal receive frequency).

Intermodulation response rejection: 40 dB (min).

Image rejection: 40 dB (min).

Spurious response rejection: 40 dB (min).

Accuracy: error of 2 dB (max) in field strength or signal level measurements.

Screening: field strength or signal level measurements should fall by at least 60 dB on removal of the antenna from the input terminal. (This requirement allows for the arrangement in Fig. 1 where the receiver is outside the anechoic chamber.)

Thus, a measuring receiver does not necessarily have a performance specification equivalent to MPT 1326 (Appendix 1). If a receiver is required for the complete range of type testing, this is a different matter but, in this case, it is EMC testing which is being considered. The reader is referred to CISPR Publication 16 for a detailed explanation.

Appendix 5
Spurious emissions

Refer to Test 5 in Appendix 1. A transmitter may have emissions at frequencies other than the assigned frequency. These are known as spurious emissions and can cause severe interference. The test is carried out with modulation applied, but at 20 dB above the level for normal test modulation. The specific details need not be examined in this article, but the output spectrum is monitored for out-of-channel emissions. Consider a handportable with an output power of 1 W. A spurious emission must not be greater than 2.5 μW or at least [60–10log2.5] dB below the level of the designated emission. For example, if the level of the wanted signal at the correct frequency is 60 dBμV, the level of the spurious signal must not be greater than 3 dBμV.

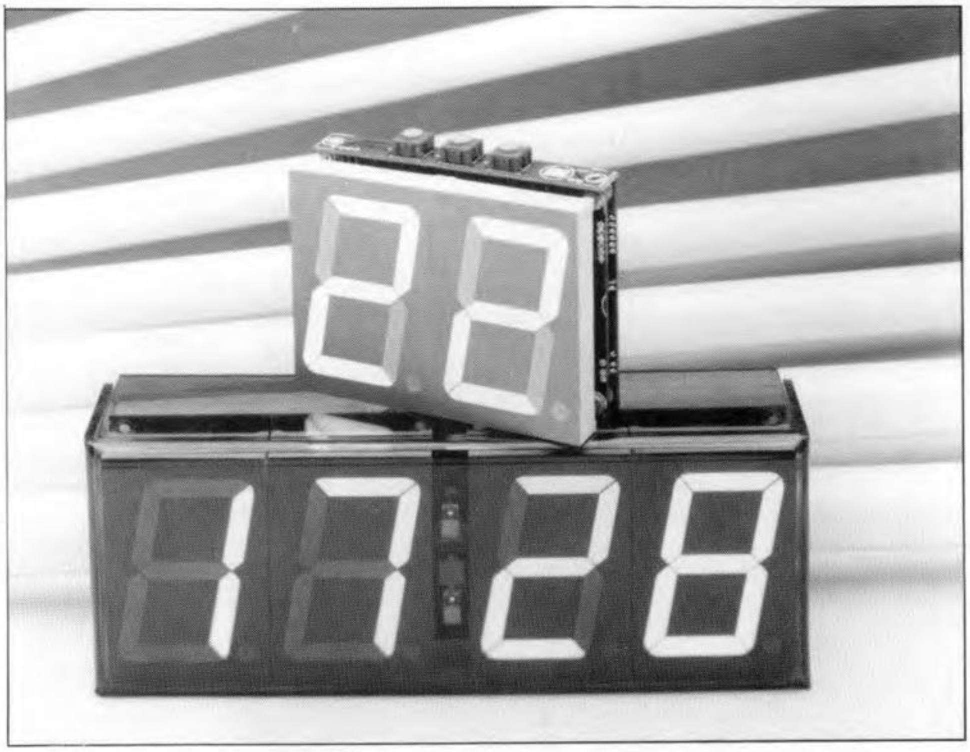
For the receiver, there is no designated emission, but specific signals (e.g., local oscillators) are generated internally. The test method is the same with the exception that the unit is in the receive mode. The limit of 2 nW (Test 14) is smaller by a factor of over 1000.

These two tests cover all emissions irrespective of communications and, consequently, the ENCLOSURE test in Section 4 is not required for en EMC test.

Appendix 6
Spurious response rejection

This is a measure of the capability of a receiver to distinguish between a modulated signal at the receive frequency and an unwanted signal at another frequency. Refer to Test 11 in Appendix 1. The test methods need not be considered, but the level of an unwanted signal in the range 100 kHz to 1 GHz must be at least 70 dB above the level of the wanted input. However, the difference between this performance test and EMC requirements is that the unwanted signal is also a functional signal intended for communications purposes. Intermodulation response and blocking are in the same category. Thus, in order to comply with EMC standards, the tests in Section 5 on immunity are required. ■

MAXI MICRO CLOCK



Digital clocks and timers are around us in a multitude of appearances, so it makes little sense these days to build one yourself. However, clocks with a really HUGE readout are few and far between, although they have many useful applications.

Design by A. Rietjens

THE maxi micro clock described in this article is based on the ST62T10 miniature microcontroller from SGS-Thomson. The remarkable thing about this microcontroller is that it couples the full functionality of a microprocessor to a small (20-pin) enclosure. The program that implements the clock function contains those functions that were considered useful for a clock of 'larger than normal' size. Unfortunately the capacity of the EPROM contained in the ST62T10 is only 2 KByte, so that three different programs, i.e., differently programmed microcontrollers, are required for the three applications of the clock. These applications are: a clock with alarm function; a darkroom timer/long-period timer; and a cooking timer.

Although the clock is controlled by a miniature microcontroller, the size of the displays allows it to be read from quite a distance. An automatic display intensity control is provided that reduces the display brightness if the ambient light intensity drops.

The clock may be set with the aid of three keys, whose function is supported by four LED indicators. Finally, audible signalling for the timer functions is provided by a buzzer. If desired, a relay may be connected to the switching output.

In a future issue of *Elektor Electronics* a modified version of the clock will be described using small displays. This clock will be as large as one display of the present clock, and will be called 'mini micro clock'.

ST62T10 microcontroller

This is the first *Elektor Electronics* application of a miniature microcontroller from ST (SGS/Thomson). The microcontroller used is one from a family consisting of four devices: the ST6110, ST6220, ST6215 and ST6225. The structure of these controllers is open and versatile, which makes them eminently suited to compact control systems.

The internal structure of the ST6 series of microcontrollers is shown in Fig. 1. The processors feature a bus width of eight bits, and contain an 8-bit counter, a 7-bit divider, an 8-bit A-D converter, and a watchdog timer with a power-up reset function. The ST6210 and ST6220 in addition feature eight

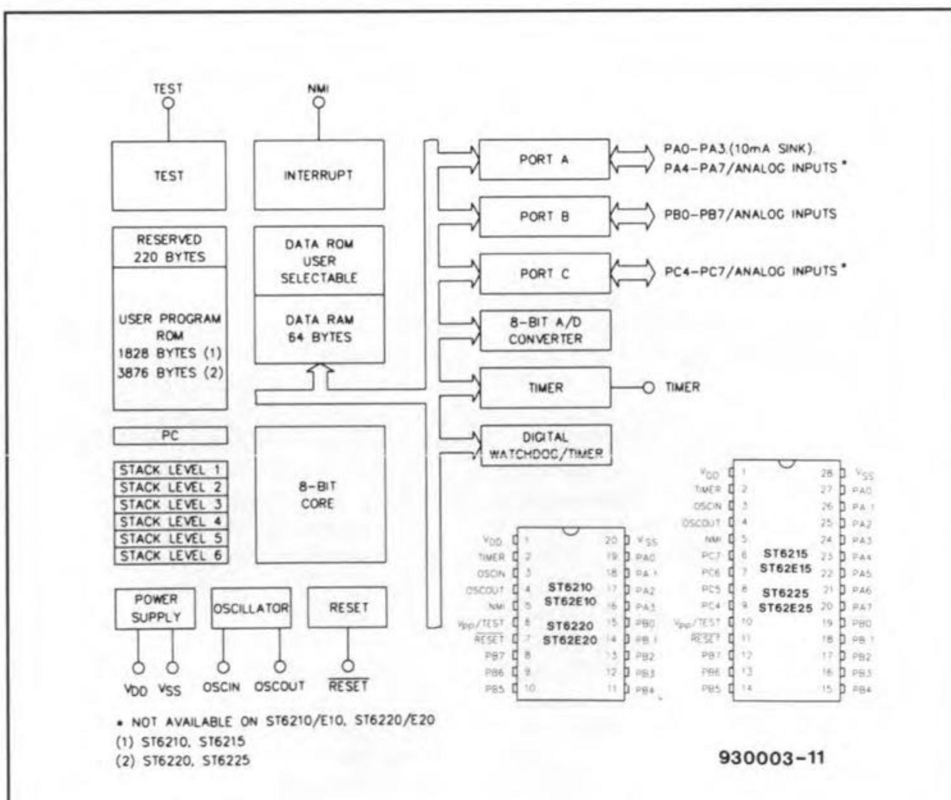


Fig. 1. Internal architecture and pinning of the devices in the ST62 family of microcontrollers (courtesy SGS-Thomson).

analogue inputs, while the ST6215 and ST6225 have 16 of these inputs. Interestingly, the analogue inputs may be programmed as digital inputs or outputs. The difference between the ST6210/6215 on the one hand, and the ST6220/6225 on the other, is that the first two types have an on-chip 2-KByte ROM, whereas the 6220/6225 have a 4-KByte ROM.

All four microcontrollers in the ST62 series are also available in -S and -T versions. The -E versions have a glass window and an EPROM that can be erased by ultraviolet light, and reprogrammed. The -T versions are one-time programmable devices (OTPs). The -T versions are three to four times cheaper than the -E versions, whence the controllers supplied through our Readers Services are -T devices.

Of the 2 KBytes of internal memory, 1,828 locations (bytes) are available to the user, while 220 bytes are reserved for other functions. Furthermore, an internal RAM space of 64 bytes is available. The processor's internal stack has 6 levels, and the chip works from a supply voltage between 3 V and 6.25 V. The clock frequency of these processors is 8 MHz, which is easily generated with the aid of the internal clock oscillator and an external quartz crystal.

Circuit description

The circuit diagram of the maxi micro clock is given in Fig. 2. Apart from the ST microcontroller there are five integrated circuits. The remaining electronics are required to make the digital signals suitable for driving the displays (LD1-LD4), the buzzer and the relay output. Finally, there is a simple circuit to detect action on the press-keys. The on-board voltage regulator, IC6, enables the clock to be powered by an ordinary mains adaptor with 12 VDC/500 mA output. The displays and the LEDs are multiplexed, which allows the four displays to be driven by the limited number of lines available on the I/O port.

Because of the internal structure of the displays (three LEDs in series for each segment, and two for the decimal point), these have to be connected to a supply voltage that is 5 V higher than that of the microcontroller. The display supply voltage of about 12 V calls for open-collector drivers, contained in

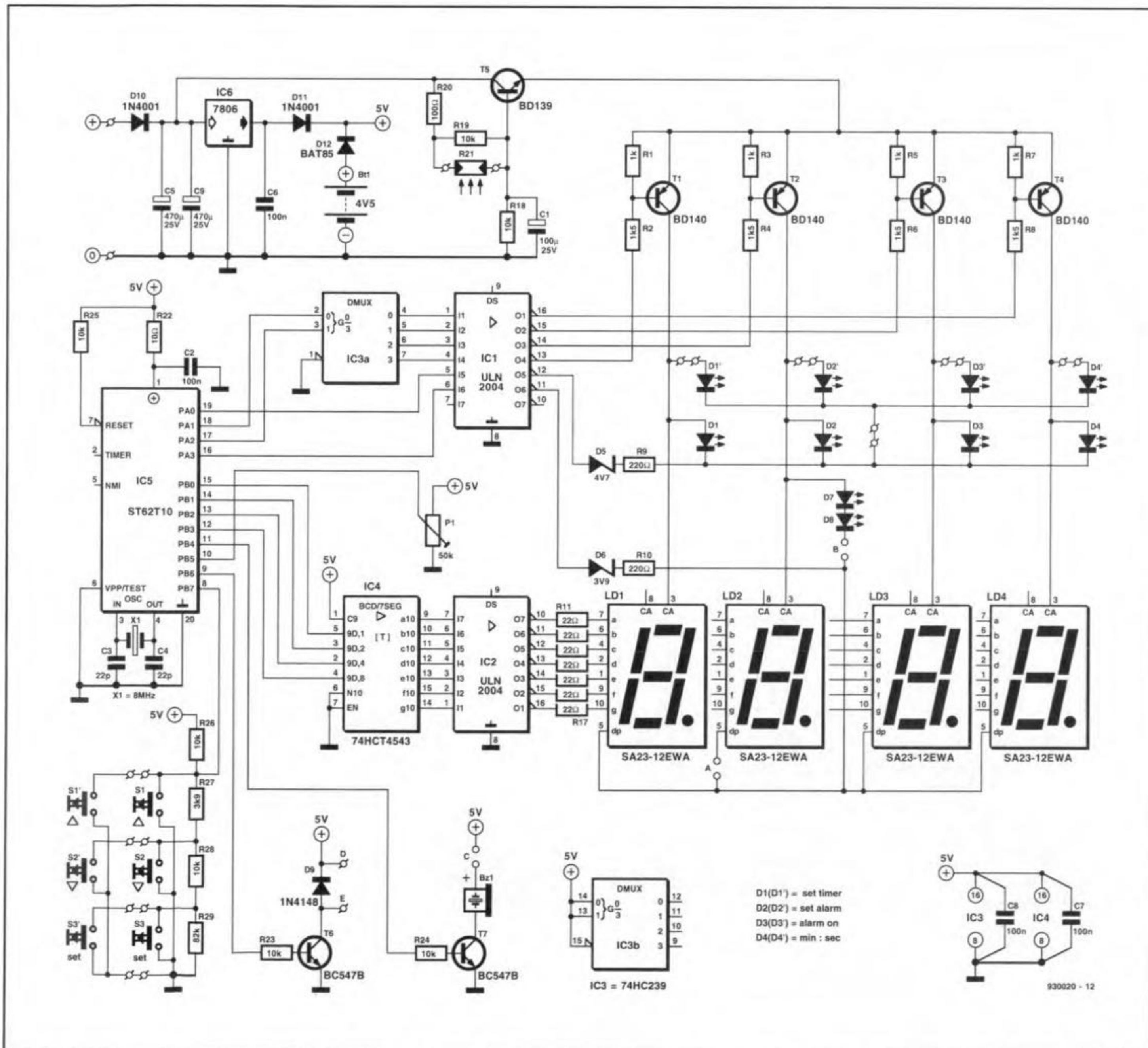


Fig. 2. Circuit diagram of the switching clock. The heart of the circuit is formed by a microcontroller Type ST62T10, which comes ready-programmed through the Readers Services.

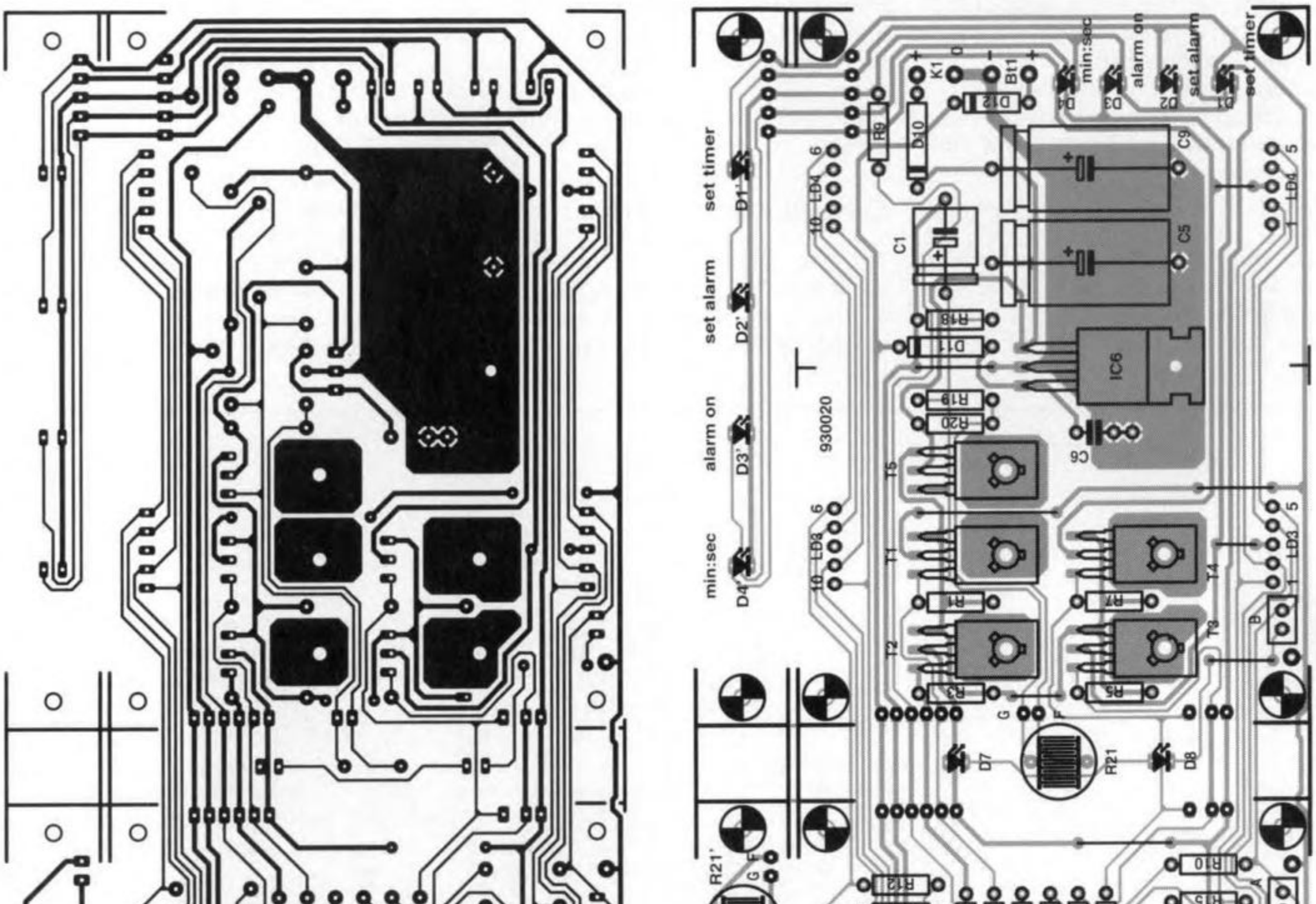


Fig. 3. Track layout and component mounting plan of the printed circuit board designed for the clock. Depending on your application and construction, the PCB may be cut into three or four pieces.

COMPONENTS LIST

Resistors:

4	1kΩ	R1;R3;R5;R7
4	1kΩ5	R2;R4;R6;R8
2	220Ω	R9;R10
7	22Ω	R11-R17
7	10kΩ	R18;R19; R23-R26;R28
1	100Ω	R20
1	LDR	R21
1	10Ω	R22
1	3kΩ9	R27
1	82kΩ	R29
1	50kΩ preset H	P1

Capacitors:

1	100µF 25V	C1
4	100nF	C2;C6;C7;C8
2	22pF	C3;C4
2	470µF 25V	C5;C9

Semiconductors:

4	LED red 3mm	D1-D4
1	zener 4V7 400mW	D5
1	zener 3V9 400 mW	D6
2	LED red 5mm square face	D7;D8
1	1N4148	D9
1	1N4001	D10;D11
1	BAT85	D12
4	BD140	T1-T4
1	BD139	T5
2	BC547B	T6;T7
2	ULN2004	IC1;IC2
1	74HC239	IC3
1	74HCT4543	IC4
1	ST62T10 (see below and page 110)	IC5
1	7806	IC6
4	SA23-12EWA* (Kingbright)	LD1-LD4

Miscellaneous:

1	4.5V battery	Bt1
1	buzzer	Bz1
1	adaptor socket	K1
3	press-key CTL3** (MEC)	S1;S2;S3
1	8MHz crystal	X1
1	Printed circuit board 930020 (see page 110)	

The microcontroller, IC5, is supplied ready-programmed through the Readers Services. Three different versions are available:
Order code 7081: Maxi clock;
Order code 7091: Maxi darkroom clock;
Order code 7101: Maxi cooking timer.

* Kingbright UK distributors: Hero Electronics (0525) 405015; Rapid Electronics (0206) 751166. International: C-I Electronics, Holland, fax (+31) 45 241877.

** C-I Electronics, Holland. Fax: (+31) 45 241877

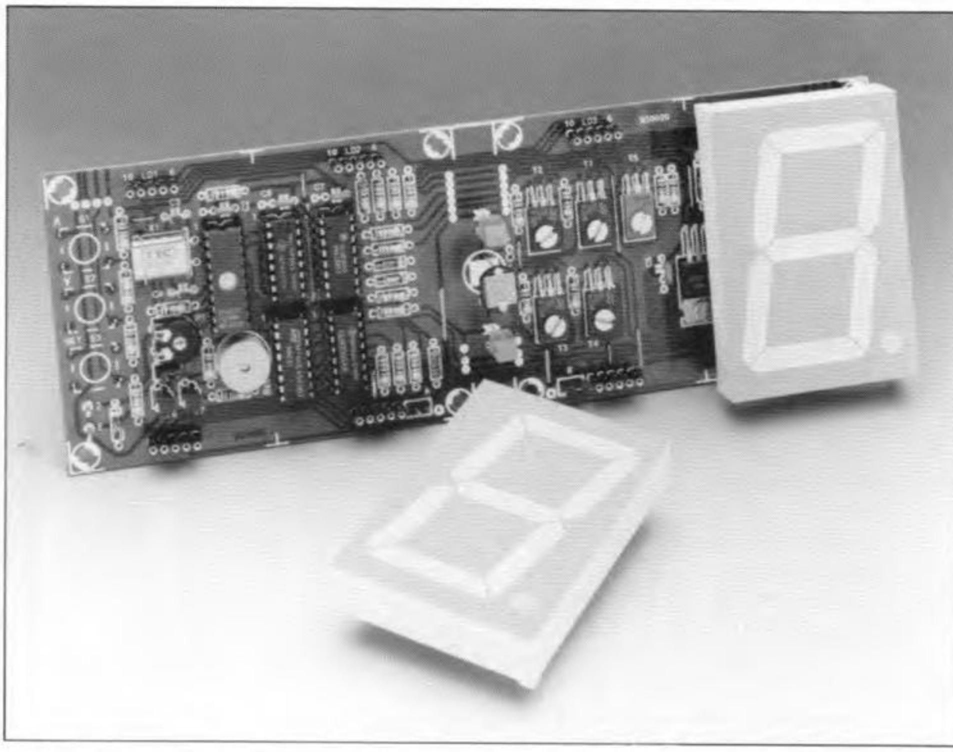


Fig. 4. The 'king-size' displays are fitted above the components on the printed circuit board.

two ULN2004's (IC1 and IC2). These ICs contain seven drivers with an open-collector output, and can easily accept a collector voltage of 12 V.

The multiplexing is arranged by two microcontroller outputs, which select one of four transistor switches T1-T4 via a 2-to-4 decoder. The binary value generated by the controller, and subsequently copied to the driver via IC4, determines which segments in the selected display are turned on. The translation from binary code to seven-segment code is performed by IC4, a 74HCT4543.

The controller also has an output line for the decimal points and the indication LEDs. The decimal point on LD2 has been made switchable, which allows you to select either one flashing decimal point, or two flashing LEDs (D7 and D8) to act as a separation between the two display pairs.

Two further drive lines are provided: pin 11 drives a buzzer that may be silenced using jumper 'C', while pin 9 drives a transistor that makes it possible to control an electrical apparatus via an externally connected relay. The maximum relay coil current that can be supplied is about 100 mA.

The clock control panel is rudimentary, but functional — three press-keys enable all clock functions to be controlled. The press-keys (S1, S2 and S3) are read in an unconventional manner by the microcontroller, because the input used to read the key code is analogue, and gives access to the on-chip A-D converter. The controller software detects key action if the voltage at the analogue input (pin 8) is between the two extremes defined by voltage divider R26-R29. When a key is pressed, the ratio between the resistors in the divider changes, which creates a total of three analogue voltage levels. The circuit diagram also indicates that three further switches, S1', S2' and S3' may be connected. These switches are fitted on a separate printed circuit board, and are used in case the switches on the main clock board are difficult to control because of the enclosure used. The same goes for the LEDs, D1-D4, which may not always be clearly visible. In this case, too, the extra PCB provides the solution.

The clock is calibrated with the aid of preset P1, whose wiper supplies a voltage that is fed to an analogue input of the microcontroller. The voltage read in this way is used to make small adjustments to the speed of the clock. This function is implemented in software, and does not affect the 8-MHz crystal clock oscillator in any way. The speed variation is ±0.01%, which is sufficient for an accurate setting (approx. 1 minute per week). If the clock option is used, an (optional) 4.5-V battery may be connected to ensure that the clock keeps running (with the displays quenched) in the event of a mains voltage failure of up to 48 hours.

Cutting and soldering

Since the final appearance of the clock must be neat and attractive, the printed circuit board (Fig. 3) is designed such that you have many options as regards the practical construction. A quick look at the PCB tells you that this can be cut into three or four parts. The narrow part cut off lengthwise is available to build the additional keyboard (if required).

Since the standard press-keys S1, S2 and S3 are fitted at the track side of the PCB, they are difficult to operate if the clock is built into an enclosure. In this case, the separate PCB provides the solution. The LED indicators are also accommodated on this board, so that they remain clearly visible. If the operation of the LDR is affected by the transparent plastic material fitted in front of the displays, the device may also be moved to the narrow PCB. The LDR controls the display intensity as a function of the ambient light intensity. If this function is not required, replace the LDR by a 1-kΩ resistor.

The PCB may also be cut into two over its width. This option is useful when constructing the cooking timer version which requires only two displays, and is extra compact (Fig. 5).

If the press-keys and the indication LEDs are fitted on to the main board, it is suggested to mount the LEDs at right angles so that they are clearly visible from the side of the PCB.

In case the display intensity is too low at low ambient light levels, resistor R19 may be reduced to 10 kΩ

Finally, fit either jumper 'A' or 'B'. Fitting jumper 'A' causes the decimal point on LD2 to act as a seconds indicator, while jumper 'B' causes LEDs LD7 and LD8 to perform this function.

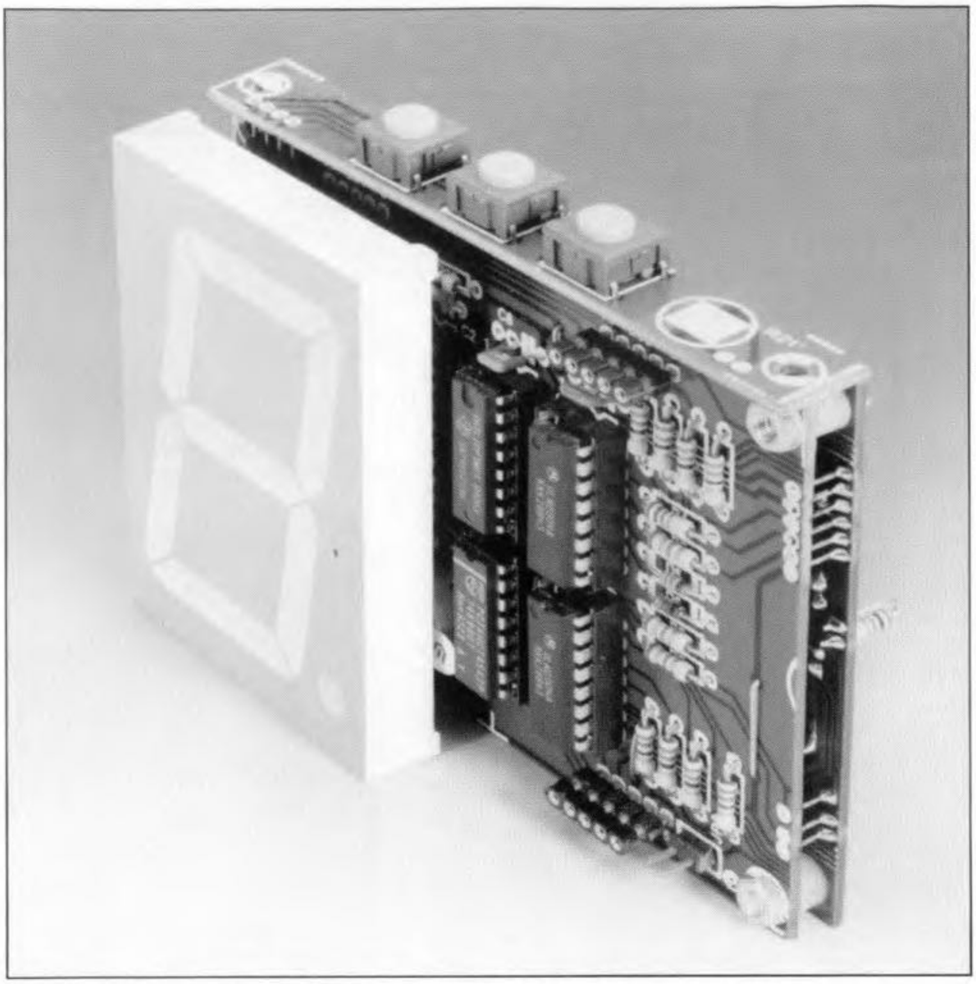


Fig. 5. The cooking timer is a compact unit thanks to the 'sandwich' construction.

Use as a clock

The SET press-key selects between clock setting and alarm time setting. Depending on the selection, LED D1 or D2 lights. The desired time is set by pressing the UP and DOWN keys. If neither of the two functions is selected (both LEDs off) the UP and DOWN keys enable you to select between an 'hours:minutes' or a 'minutes:seconds' display, both with the alarm enabled or disabled.

The dot (or the two LEDs) between the hours and minutes (or minutes and seconds) flashes at a rate of 1 Hz. One of the four options remains selected until the user selects another setting. Consequently, the clock will not automatically return to the preferred settings.

If the alarm is enabled, and the time equals the alarm time, the buzzer will produce beeps with pauses of 0.5 second. The alarm may be turned off by pressing S1 or S2. If no key is pressed, the alarm turns off automatically after one hour. Transistor T7 is controlled together with the buzzer. The PCB terminals provided may be connected to an external relay which, in turn, controls an electrical apparatus (such as a coffee machine). Jumper 'C' (or a switch) is used to switch off the buzzer.

Darkroom clock

The darkroom clock or long-period timer function allows a programmed timing interval to be signalled. Both repetitive and single signalling is possible. In darkroom clock mode, the 'min:sec' display function is not used, and replaced by a 'repetitive alarm' function. The maximum time between two short alarm signals is then 99 minutes and 59 seconds (99:59). The darkroom clock is selected using the DOWN key, which also serves to select between alarm (buzzer) on and off. The SET key allows the alarm time only to be programmed — there is no point in setting the actual time. The counter is reset if the UP key is pressed after setting the desired alarm time. Depending on your selection, the buzzer will sound once, or every time after a period has elapsed. The latter function is particularly useful to time the film developer flip rate.

Transistor T6 and the optional relay are actuated during the first period (after resetting). This allows an electrical apparatus, for instance, an enlarger, to be switched on for a maximum period of 99 minutes and 59 seconds.

Cooking timer

The UP and DOWN keys are used to set the desired time, which is counted down to zero. Three short beeps are produced when the programmed time has elapsed.

The cooking timer does not use displays LD3 and LD4, so that only the minutes and tens of seconds readouts are visible. Before programming the cooking time, the SET key must be used to select between cooking times longer or shorter than 10 minutes. Next, the desired time is set using the UP and DOWN keys. When the cooking time is shorter than 10 minutes, the first display shows the remaining minutes, and the second display the remaining tens of seconds. For instance, if the displays read '8.3', the remaining time is 8 minutes and 30 seconds. With cooking times longer than 10 minutes, the display shows the number of minutes only.

By virtue of the sandwich construction, the cooking timer remains a very compact unit (Fig. 5). The dot between the two displays flashes at a rate of 1 Hz. ■

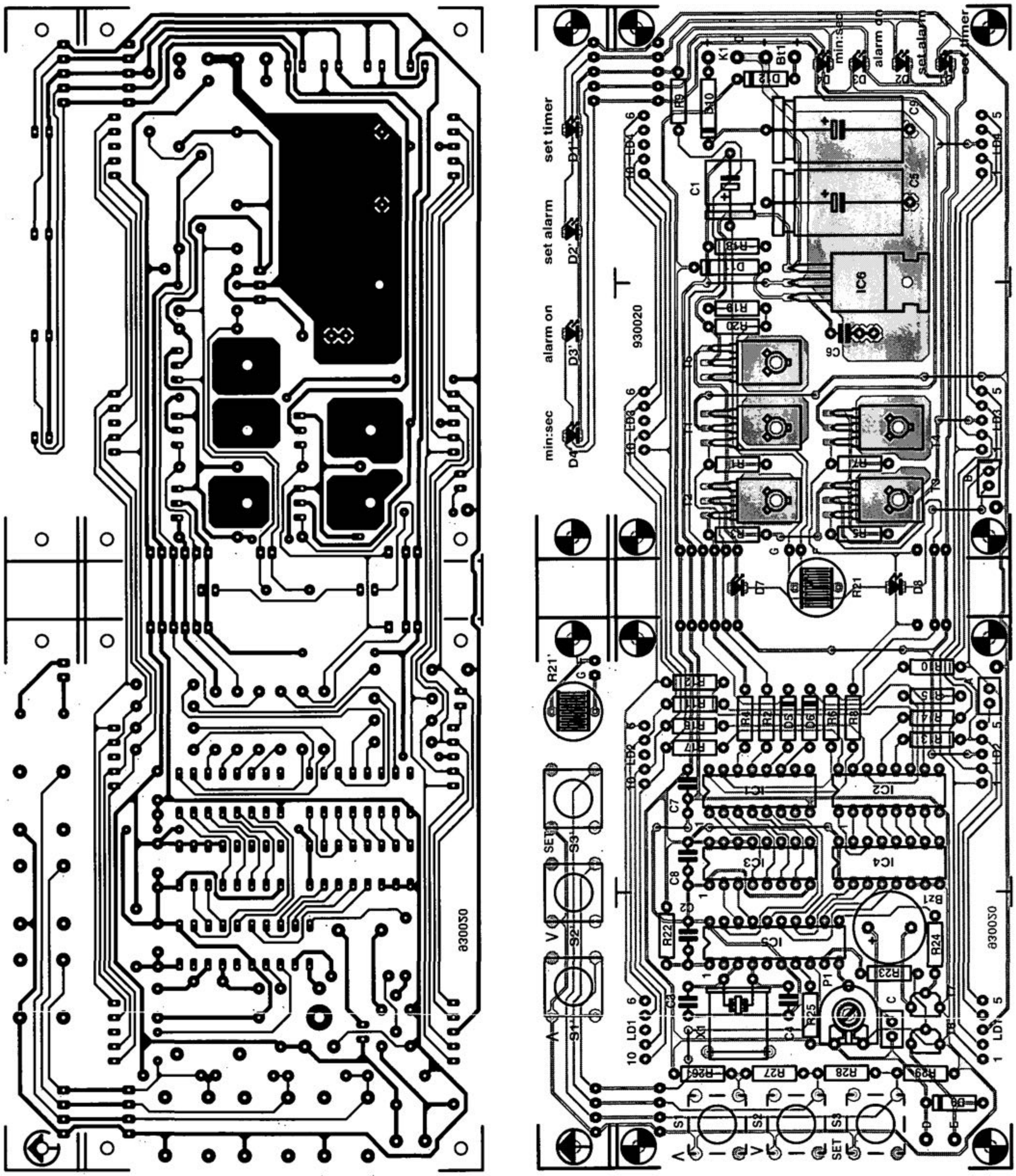


Fig. 3. Track layout and component mounting plan of the printed circuit board designed for the clock. Depending on your application and construction, the PCB may be cut into three or four pieces.

CAMCORDER AUDIO MIXER

939504

An ELV design (© 1993)

The mixer has been designed especially for use with a camcorder. During recording (or later copying), it allows the output of an external microphone to be mixed with up to two other audio signals. Also, the mixer has a headphone output, whose level may be adjusted independently.

A good video film should go together with good sound. It is the sound, however, that gives film makers problems. The microphone in many camcorders is not exactly of the best quality and has, moreover, a spherical polar pattern, so that a multitude of unwanted sounds are also recorded.

Many camcorder recordings are transferred to (S)VHS tape. Unfortunately, few VHS recorders afford access to the sound channel, so that during the copying no other audio signals can be added.

The mixer is connected to the external microphone input available on most camcorders. The external microphone, which is connected to the input of the mixer, may be arranged so that no spurious signals are recorded. It should, of course, have the polar pattern best suited to the individual application.

The (mixed) output signal may be monitored via a headphone, whose volume may be adjusted as required.

The unit may be used not only for mixing audio signals during recording, but also used during the copying of the camcorder film to the VHS recorder (or during the copying from one VHS recorder to another). It is thus possible to add a commentary or background music to the recording. When this is done, the headphone output (instead of the camcorder output) of the mixer is connected to the audio input of the video recorder. The recording level can then be set as required with the headphone volume control.

The mixer is fairly small: it is contained in a man-made fibre box measuring only 120×65×22 mm ($4^{11}/_{16}$×$2^{9}/_{16}$×$^{7}/_{8}$ in). It is powered by a 9 V battery.

The circuit

The line inputs are taken via BU2 and BU3 to buffer T_1 via networks C_5-R_6 and C_6-R_7 respectively, and summing resistors R_{10} and R_{11}—see Fig. 2.

The microphone signal is applied to BU1 and amplified in IC_{1a}. Since the mixer is powered by a single supply, the + input of the opamp is at half the supply voltage via potential divider R_1-R_2. Resistor R_4 ensures that the input impedance remains 47 kΩ.

The amplification of the opamp is determined by R_5-R_3 in the feedback loop. Since R_5 is variable, the amplification may be varied from ×1 to ×11. Capacitor C_7 suppresses any tendency to oscillation.

The amplified signal is taken to the junction R_{10}-R_{11} via R_8.

Owing to the d.c. setting of the opamp, its output is also at about half the supply voltage level. This voltage is used also as the d.c. setting for emitter follower T_1. The composite signal is then available at the low-impedance output of T_1.

The signal at junction R_8-R_{10}-R_{11}-R_{17} is also applied to the non-inverting input of headphone amplifier IC_{1b}. Capacitor C_8 prevents any direct voltage reaching the amplifier.

The d.c. setting of IC_{1b} is arranged via R_9 and potential divider R_1-R_2.

The feedback loop is similar to that of

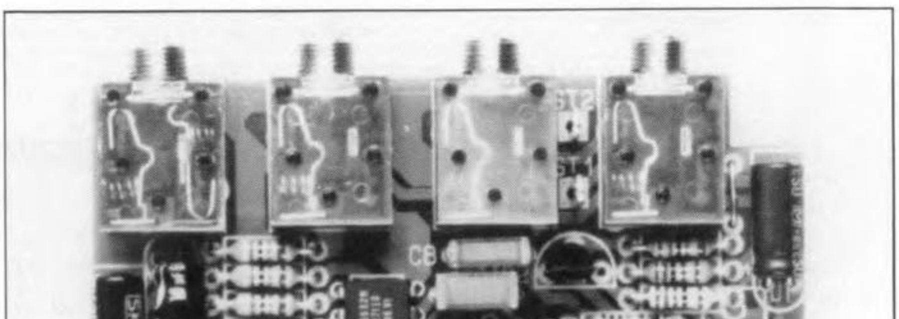


Fig. 1. The completed printed-circuit board of the mixer.

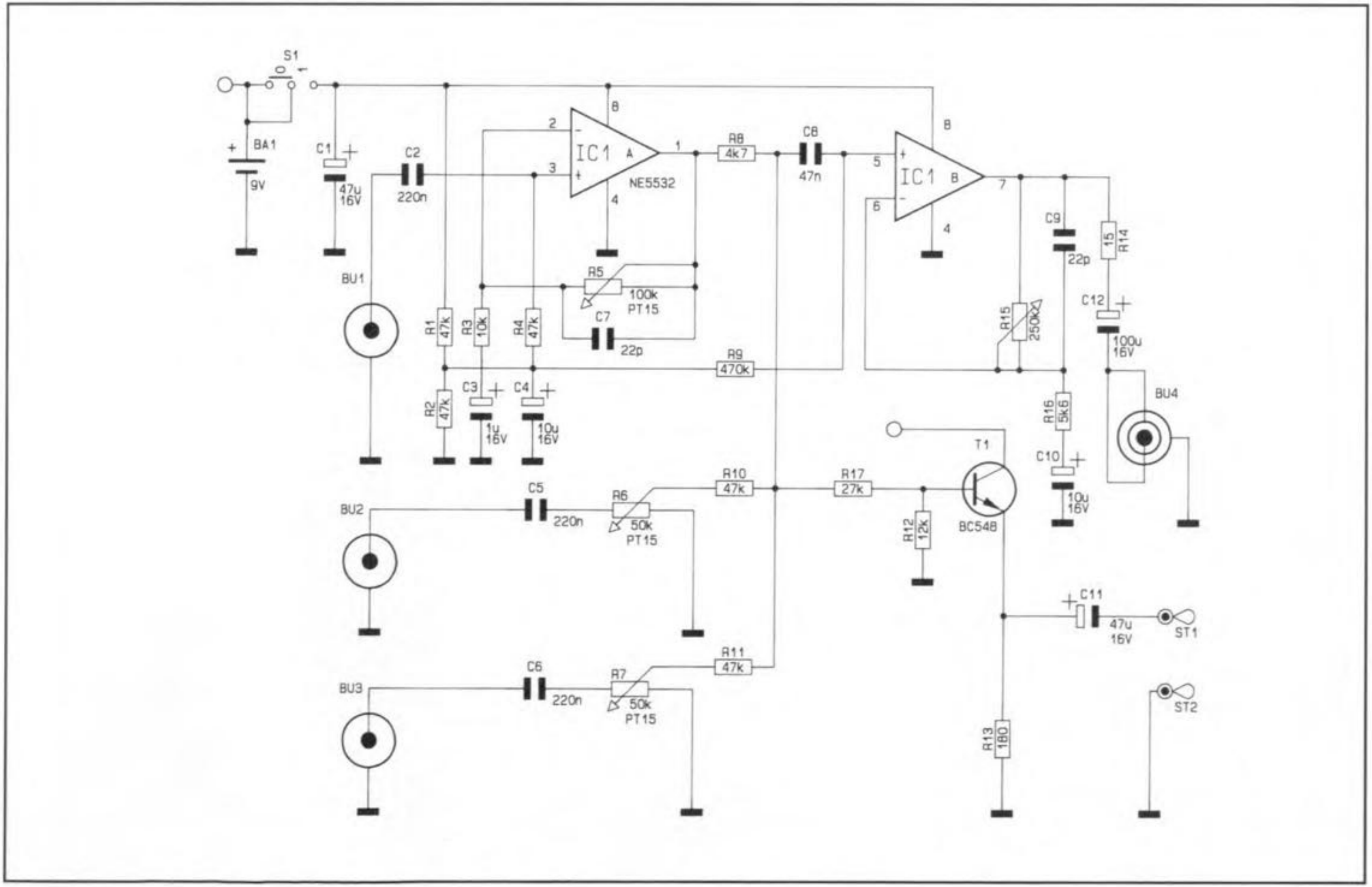


Fig. 2. Circuit diagram of the camcorder audio mixer.

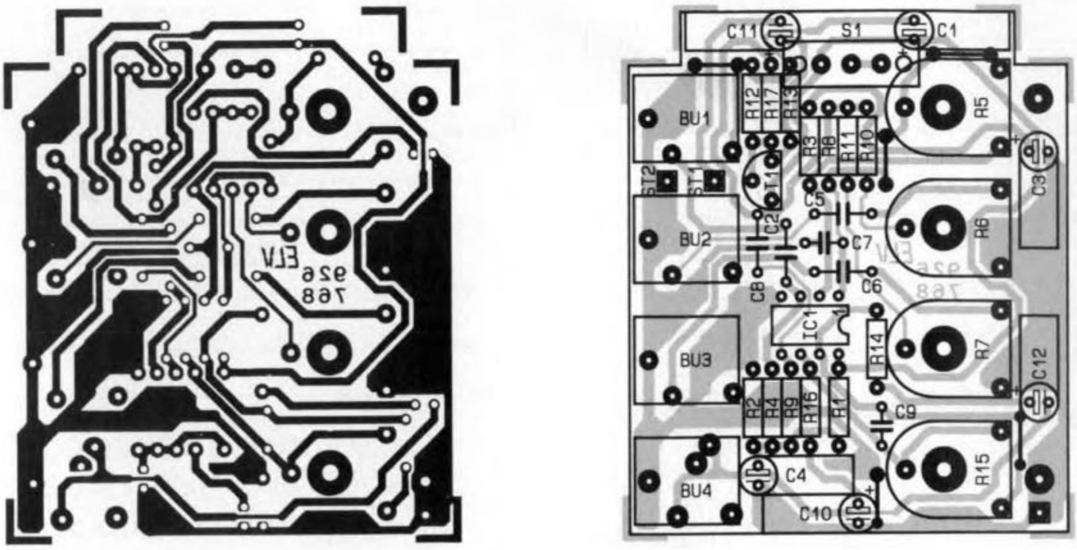


Fig. 3. Printed-circuit board for the camcorder audio mixer.

the microphone amplifier: the ratio R_{15}:R_{16} determines the amplification of IC_{1b}, which may be set between ×1 and ×45 with R_{15}.

The output of IC_{1b} is applied to the headphone via current limiting resistor R_{14}, coupling capacitor C_{12} and socket BU4.

Construction

The mixer is intended to be constructed on the printed-circuit board in Fig. 3. Populating is best commenced with the resistors and wire bridges, followed by the capacitors. Observe polarity of the electrolytic capacitors and note that these are all fitted horizontally on the board. Finally, mount the transistors, IC, audio sockets and preset potentiometers.

The mixer is linked to the camcorder by a 1 m (3 ft) long screened audio cable that is terminated at one end into a 3.5 mm stereo jack plug (interconnect the two channels in the plug). The other end of the cable is passed through a hole in the enclosure and soldered to ST1 and ST2 (screen). When the jack plug is inserted into the relevant socket on the camcorder, the internal microphone is switched off automatically.

The photo of the completed board shows that preset potentiometers are used into which a mand-made fibre spindle is inserted. These spindles are included in the ELV construction kit. After the board has been fitted into the enclosure, the spindles are pressed into the presets via the holes in the front panel.

Subsequently, fit the on/off switch into the enclosure and solder one terminal of it to the S_1 pin adjacent to C_{11} on the board (the other two pins are non-connected pads). Solder the other terminal of the switch to the red wire of the battery clip. Solder the black wire of this clip to the pad in the corner of the board next to R_{15}.

Finally, fit a strain relief sleeve on to the output cable, connect the battery to the clip and shut the enclosure. The mixer is then ready for use. ■

The ELV Construction Kit is available from
ELV
Customer Service
Postfach 1000
D-2950 Leer
Germany
Telephone +49 491 600 888
Fax +49 491 7016

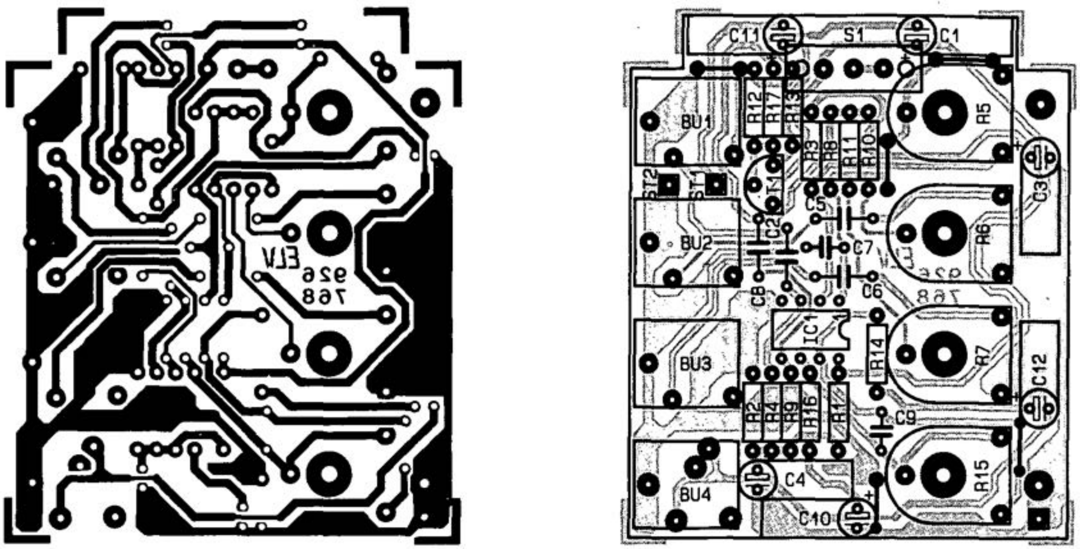
PARTS LIST

Resistors:
R1, R2, R4, R10, R11 = 47 kΩ
R3 = 10 kΩ
R5 = 100 kΩ preset, horizontal
R6, R7 = 50 kΩ preset, horizontal
R8 = 4.7 kΩ
R9 = 470 kΩ
R12 = 12 kΩ
R13 = 180 Ω
R14 = 15 Ω
R15 = 250 kΩ preset, horizontal
R16 = 5.6 kΩ
R17 = 27 kΩ

Capacitors:
C1, C11 = 47 μF, 16 V
C2, C5, C6 = 220 nF
C3 = 1 μF, 16 V
C4, C10 = 10 μF, 16 V
C7, C9 = 22 pF
C8 = 47 nF
C12 = 100 μF, 16 V

Semiconductors:
T1 = BC548
IC1 = NE5532

Miscellaneous:
S1 = miniature change-over slide switch
BU1–BU3 = 3.5 mm mono jack socket for PCB mounting
BU4 = 3.5 mm stereo jack socket for PCB mounting
4 man-made fibre spindles, 16.5 mm long
Connecting clip for 9 V battery
Enclosure 120×65×22 mm ($4^{11}/_{16}$×$2^{9}/_{16}$×$^{7}/_{8}$ in)



C11
S1
C1
BU1
R12
R17
R13
R3
R8
R11
R10
R5
ST2
ST1
T1
C3
BU2
C5
C2
C7
R6
C8
C6
IC1
R14
R7
BU3
C12
R2
R4
R9
R16
R1
C9
R15
BU4
C4
C10

MAKE YOUR OWN PCBs FROM DIRECT-READING ARTWORK

By our PCB design staff

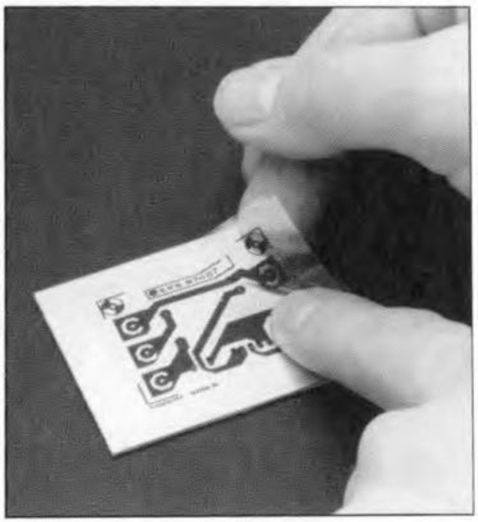
THIS notice is aimed at those of you who, occasionally or frequently, produce their own printed circuit boards from the artwork printed in *Elektor Electronics*. As of the September 1993 issue of *Elektor Electronics*, printed circuit board artwork (i.e., copper track layouts for single and double-sided PCBs) will be printed **DIRECT READING** instead of mirrored (reflected or wrong-reading) as we did for the past two years. This change can be made thanks to the availability of a new transparent film type, called Letracopy, that can be put into a photocopier machine.

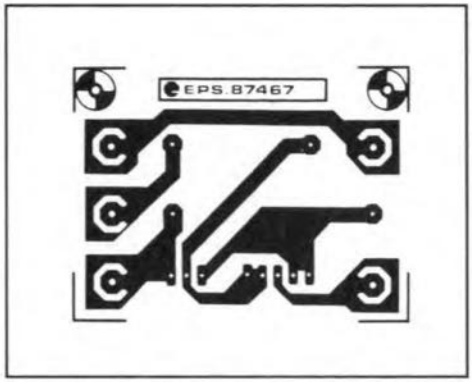
You can see that PCB artwork is direct reading by looking at the printed circuit board number, which is legible. By contrast, a printed circuit board number is not legible if the artwork is printed as a mirror image.

'Letracopy' film is a Letraset product, and should be available from larger stationery and draughting material stores. If you are unable to find a supplier in your area, call Letraset UK in London on (071) 620 0297. The film has a smooth and a rough (matt) side. The matt side has text printed on it, to inform you that the copy has to be made on the reverse (smooth) side. To make sure that you insert the Letracopy sheet correctly into the 'single sheet' tray of the photocopier, practice with a normal sheet of paper on to which you pencil a cross. If you are at a photocopy shop, call in the help of the shop assistant. Also be sure to turn the copier to the darkest possible copy setting. This is necessary to make sure that black stays black. Normally, the copy can be made at 100%, i.e., no reduction or 'blowing up' will be required. In some cases, however, artwork is printed at reduced size in *Elektor Electronics*, and you may have to calculate the enlargement factor to obtain the actual size of the board. Once the layout has been copied on to film, this can be used to expose the photosensitive printed circuit board material to an UV light source. Cut the Letracopy foil such that 1-2 cm is left free around the edges of the artwork. Next, stick the foil on the circuit board, making sure that there are no creases or bubbles. It is assumed that the rest of the production steps to make a printed circuit board are known.

(932003-8)

Using Letraset 'Letracopy' self-adhesive transfer film to prepare a piece of photosensitive circuit board for UV light exposure.





Example of positive, direct-reading, PCB artwork. Note that the production number, 87467, is legible.

THE ANALOGUE SUBSYSTEM (PART 1)

HOMING IN ON THE NON-DIGITAL SIDE OF COMPUTER BASED ELECTRONIC INSTRUMENTS

The digital computer revolutionized electronic design to such an extent that today even relatively minor devices contain microprocessors. Computers, from single chip models to personal computer 'desktop' units, to rack mounted mini-computers (if that term is still useful), to megacomputers, are used at the heart of electronic control, instrumentation and data acquisition systems. One need not look very far to find computer based scientific or engineering instruments, household devices, and assorted control systems. Hobbyists get in on the act by building everything from microprocessor based instruments to small microcomputers. The advertisements in this magazine often contain several single-board computers designed specifically for use as the heart of an instrument or control system. Personal computers that allow plug-in boards make the contriving of significant, complex devices particularly easy to accomplish.

By Joseph J. Carr, M.Sc.

ALTHOUGH interfacing and programming get a lot of coverage in the technical press, the analogue subsystem — on both input and output sides of the computer — is rarely covered. Yet the analogue subsystem is terribly important to the overall success of the project; so much so that it is dangerous to ignore it.

Another reason to examine the analogue subsystem is that an analogue solution is sometimes the **best** solution. Heresy! Heresy! But before you convene a witch hunting court, let me explain what I mean by the 'Digital Myth'. We have become enamored of the digital solution to all instrumentation problems, even though it is not always the best solution. Analogue chip makers are not exactly going out of business, and indeed are prospering quite well.

Several examples can easily be found where the selection of digital vs. analogue solutions are in question. For example, consider an absorbency ratio measurement in which two phototransistors are used to find the ratio between the light passing through a specimen vs. the light passing through a clear path, or one calibrated to light colour by a filter. An analogue logarithmic amplifier can provide high speed processing of the ratio with no clock or digital circuitry needed. The circuit is simple, as opposed to any computer implementation (which would be quite complex). If cost is a factor, then the analogue solution may actually be the preferred solution. In some applications, the fact that the analogue log-amp provides a continuous time domain output signal, as opposed to the inherently discrete time domain of a digitized signal, becomes important.

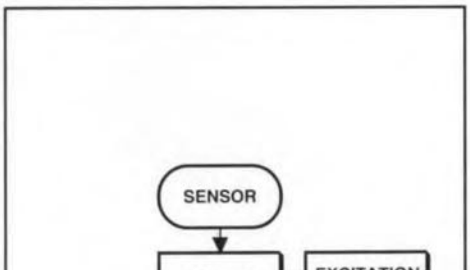


Fig. 1. Typical analogue instrument system.

Another application where the analogue solution is found when a minute analogue signal rides on top of a large DC offset signal. Two examples spring to mind. First, the human electrocardiogram (ECG) signal is an analogue signal with peak amplitudes in the 1-mV range, and has a Fourier frequency spectrum of the order of 0.05 to 100 Hz, or more. Unfortunately, the silver-silver chloride (Ag-AgCl) electrodes typically used on the skin surface to acquire the signal produce a half-cell potential of about 500 mV. Therefore, the desired signal is of the order of 500 ±1 mV. Medical device designers prefer 12-bit A-D converters for the 1-mV segment, so will require many more bits to have sufficient range to directly convert the 500 ± 1 mV signal to permit software to strip off the 500 mV offset component. The 12-bit converter is relatively low-cost, and has a 72 dB dynamic range; the 20-bit version is a lot costlier, and has a 120-dB dynamic range. While some very high resolution Δ-Σ A-D converters can do the trick (at high cost), a simple analogue DC restoration or DC clamp circuit will also do the trick at a cost that can be described as 'decimal dust'.

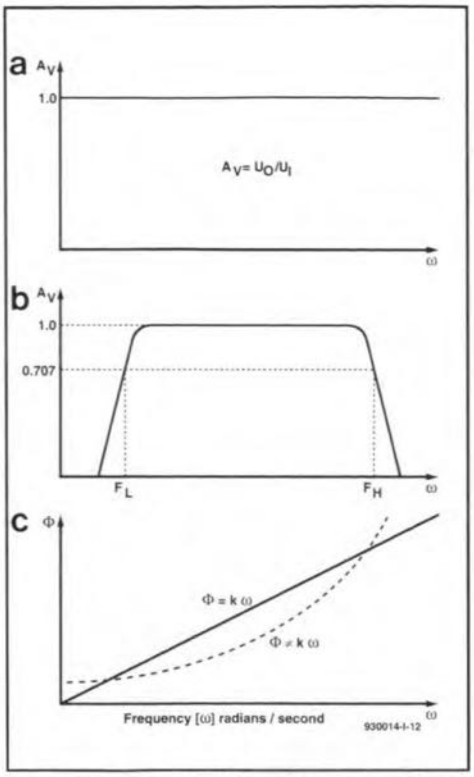


Fig. 2. a) perfectly flat system; b) practical bandwidth limited system; c) phase shift vs. frequency.

Another example exists in optoelectronics. When a small variation in light intensity must be detected, and the signal recovered, in the presence of a strong ambient light, then the situation is similar to the ECG case: a millivolt level signal riding on a high value DC offset signal.

Finally, when all errors are accounted, and corrected, it is sometimes the case that analogue circuits have less error. An example seen in a logarithmic amplifier circuit from a commercial supplier provides some insight. When the analogue logarithmic amplifier is used without adjustment, the error is of the order of 1.5 to 3 per cent. But this error can be trimmed to less than 0.5 per cent with external components. Direct digitization, however, can see 0.2 per cent amplitude error, 0.03 per cent sampling error, and 0.6 per cent A-D digitization error, or about 0.83 per cent.

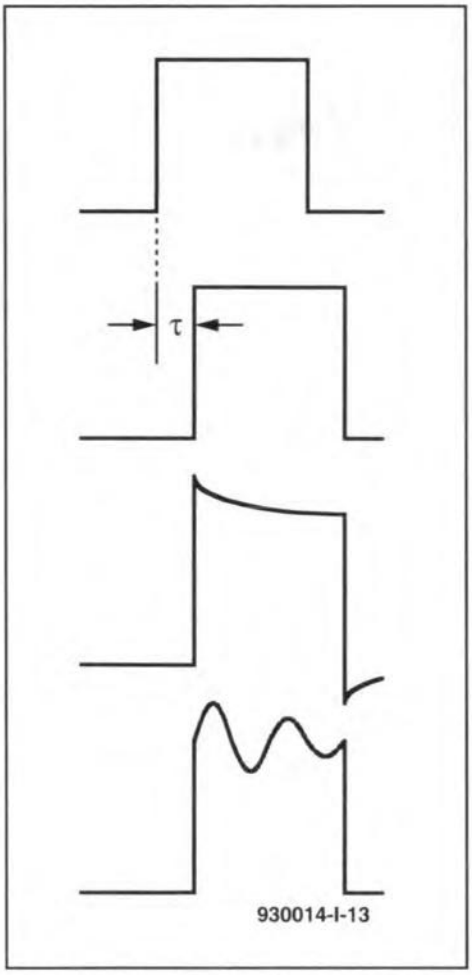


Fig. 3. Effects of bandwidth limitations on a square wave signal.

Electronic instrument systems

Figure 1 shows a block diagram for a generic computerized electronic instrument. While this figure is merely a 'mind example', it could easily serve for a large class of actual analogue and digital instruments presently seen on the market. The principal stages or functions are: **input parameter** (stimulus), **sensor** or **transducer**, **analogue signal conditioning** (input functions, amplification, level translation and signals processing), **output display** or **recording** or other means for presenting the data to the 'external world'.

Physical stimulus. The physical stimulus that is sensed in an instrumentation system may be temperature, light, displacement, fluid or gas flow, electrical resistance, electrical potential or any of a host of other physical parameters. Which particular stimulus is intended is not important in a discussion of generic instruments, but becomes important when specifics are being discussed.

Sensors and transducers. The **sensor** or **transducer** is a device that is capable of responding to the applied stimulus, and producing an electrical output signal that corresponds to the value of the applied stimulus. A **sensor**, or **transducer** (the words are roughly equivalent for our purposes), is a device that converts energy derived from a physical phenomenon into a representative electrical current or voltage, for purposes of measurement, control or information.

There is often some ambiguity in the use of the words **sensor** and **transducer**, and in many cases they are properly used interchangeably. A transducer is a device that converts energy from one form to another (e.g. pressure to an electrical potential), while a sensor may or may not make some sort of conversion, at least in an obvious way (e.g. a biopotentials electrode). Thus, an electrode used in medical electrocardiograph (ECG) recording is a sensor but not a transducer, while a pressure transducer is both a sensor and a transducer. A loudspeaker transduces an electrical audio frequency signal to a mechanical acoustical vibration. It is an **output transducer** and not a sensor at all [a].

Some physical parameter or stimulus (e.g. temperature, flow, pressure, or displacement) affects the output state of a sensor; the sensor produces an output signal that is proportional to the applied parameter. Thus, the output of the sensor will be either a voltage or current that represents the parameter being measured (e.g., a temperature sensor that outputs a voltage, U_o, of 10 mV/K). More often than not, the magnitude of the voltage or current from the sensor represents the magnitude of the parameter at the instant of measurement. Over time, this voltage or current represents the time history of the stimulus parameter.

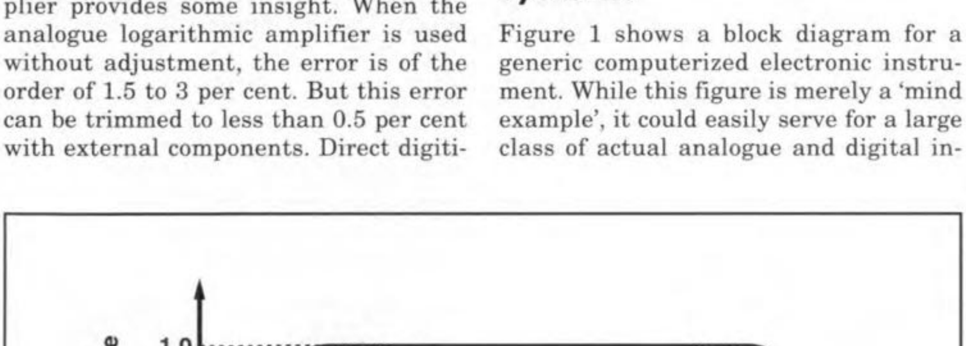


Fig. 4. Three different bandwidth systems.

(a). For those who know more than is necessary for this discussion, it is recognized that certain loudspeakers, notably those of the dynamic PMMC design, can also serve as a microphone — so are both sensors and output devices.

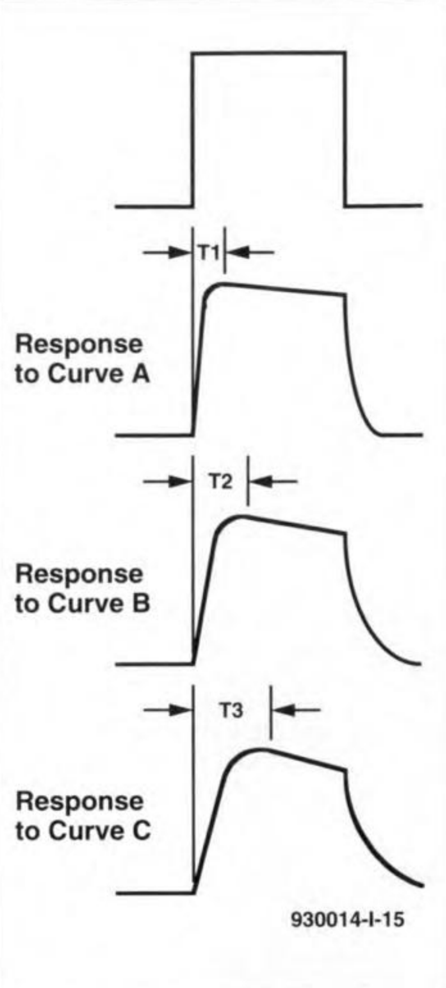


Fig. 5. Responses to the curves of Fig. 4.

The most desirable sensors have an output signal characteristic that is linear with respect to the stimulus parameter. However, there are also many useful transducers that are either quasilinear (i.e. linear only over a portion of their total range) or even nonlinear. Such transducers are often used over a limited range, or must be artificially linearized.

Input functions. The purpose of the input circuit is to receive the signal from the transducer, and convert it into whatever form (usually a voltage) that is required by the circuits to follow. In this section of the instrument, interfacing becomes terribly important. The input functions usually include amplification, but can also include an AC or DC excitation voltage (especially in the case of Wheatstone bridge sensors), DC level shifting, and isolation of the input circuit from the remainder of the instrument (common in medical instruments because of patient safety considerations).

Signals Processing/Amplification. The output signal from most sensors is not usually suitable for immediate display. Rather, some form of signal conditioning is usually needed. This conditioning may be only amplification, or it might also include frequency selective filtering, mathematical operations such as differentiation, integration, 'logging' or 'antilogging', or it may include something simple like dc level translation. In other cases, the signals processing uses the analogue circuit, in effect, as a fixed-program, dedicated analogue computer to solve for a mathematical expression. Some of these functions are most reasonably apportioned to analogue circuits, while others are most reasonably apportioned to either digital circuits or computer software. In each case, the system designer must decide which is the proper engineering choice for the problem at hand.

Output Functions. The output of the instrument must often be processed in some manner before it can be displayed. The output functions may include power amplification (as in the case of another control system motor driver), digitization for input to a computer, or voltage scaling so that the display is easily read by a human operator.

Output and Display. Finally, for an instrument to be useful there must be a display, data storage or control function to perform. Various different forms of output and display function are provided, depending on the need. An analogue output might be needed, especially if the signal must be passed to some other instrument for further processing or application. Alternatively, the signal might be digitized (or, indeed, originally generated as a digital signal), and applied to some form of digital display. A Y-time or X-Y oscillographic display can be provided with either a strip-chart recorder or a cathode ray oscilloscope. Various forms of visual (sometimes called alphanumeric) display devices are sometimes used, especially when the output information is in the form of a numerical value (e.g., '100 Torr of pressure').

The visual display or recording devices might be a DC meter movement, an oscilloscope, a strip-chart recorder, digital printer, video terminal, or even a simple 'GO-NO GO' lamp.

Tactics and signals processing for improved sensing

The selection of sensors and the circuits that are connected to them can go a long way towards ensuring that the data acquired will accurately represent the physical phenomenon or event being detected. For proper operation in a dynamic input environment, the sensor selected should have a flat response curve, i.e., one that is free of amplitude distortion, phase distortion (which often causes amplitude distortion), 'ringing' or resonances.

An implication of these problems concerns the **frequency response** of the sensor and its signals processing system. Figure 2a shows a perfectly linear system in which the gain is constant over the entire spectrum of frequencies, i.e. in an ideal theoretical system from 'DC to daylight' and beyond. But real systems do not have such characteristics. The specification that determines the ability of the system to handle the frequency spectrum of any given signal or physical parameter is the **bandwidth**. It is usually given in the form of a difference between the maximum and minimum

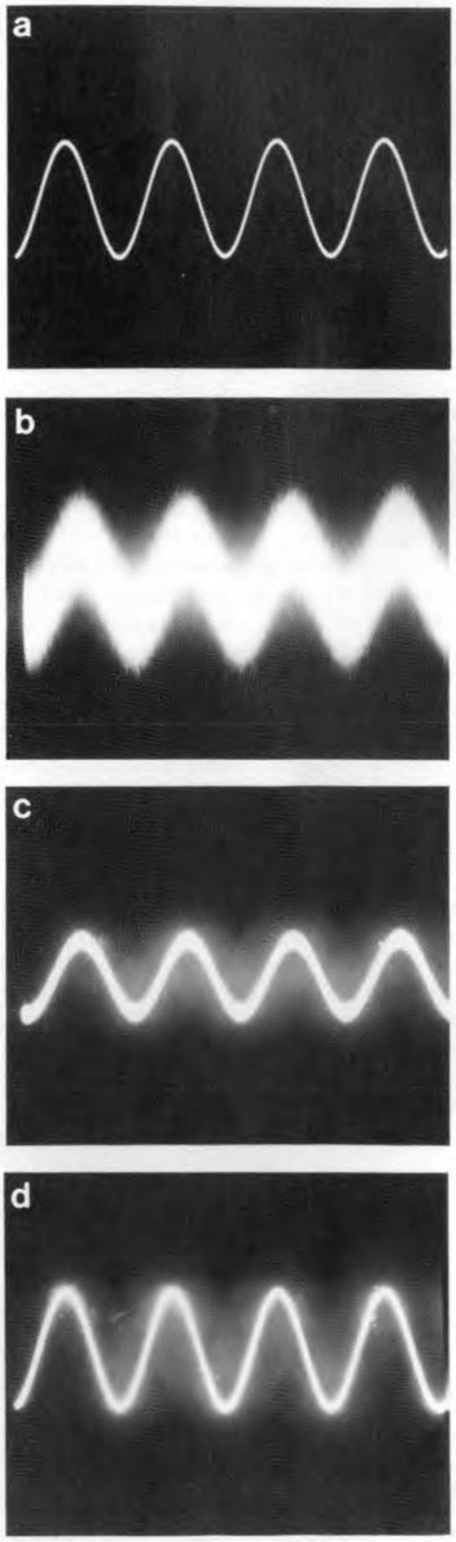


Fig. 6. a) small amplitude signal; b) combined with noise; c) with some filtering; d) with more filtering.

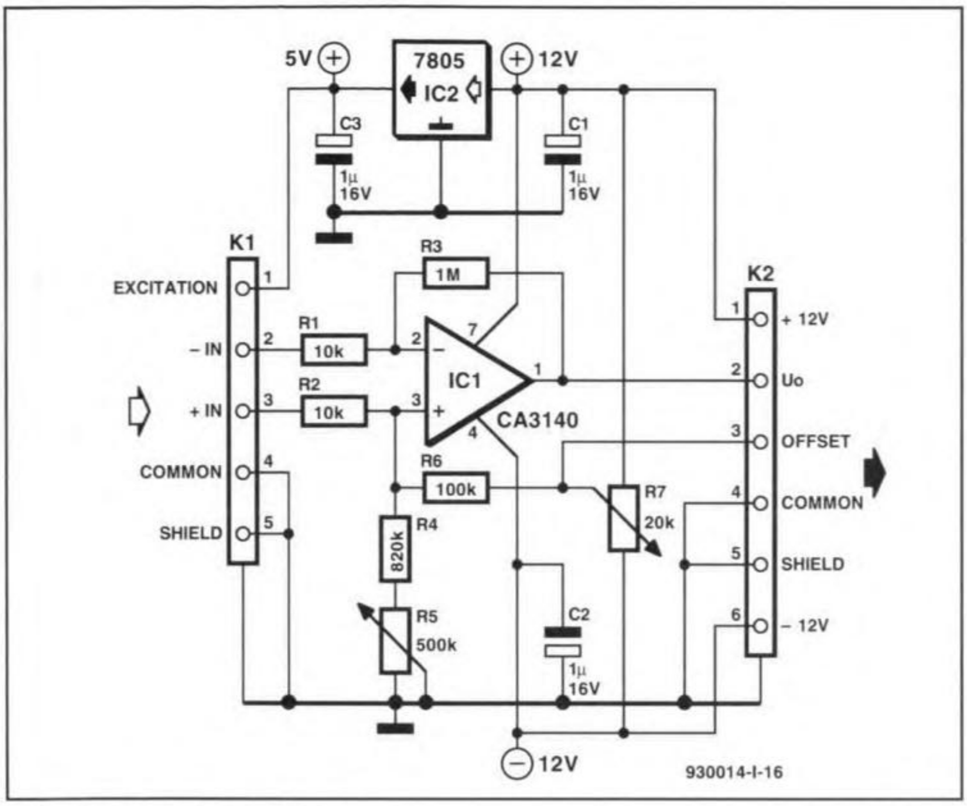


Fig. 7. Signal boosting amplifier.

frequencies that must be handled, and has units of frequency.

Figure 2b shows the type of frequency response that might be found on real systems. In this example, the gain is flat between two frequencies, and over this region the performance is similar to the ideal case. But beyond these points, the gain falls off at a given slope. The breakpoint that defines the flat region is, by convention, taken to be the frequencies (F_L and F_H) at which the gain falls of to 70.7 percent of its gain in the flat region. These points are known as the –6 dB points in voltage systems, and the –3 dB points in power systems.

When the frequency response is not entirely flat, one can expect to find phase distortion. Figure 2c shows the situations where the phase shift of the system is a linear function of frequency (solid line), and also where it is a nonlinear function of frequency (dotted line).

We can see the effects of phase distortion in a somewhat simplistic sense in Fig. 3. Figure 3a is the applied signal, e.g., the output of an ideal sensor in response to step-function changes of the measured input parameter. If the signals processing electronics and the sensor mechanism itself are perfectly ideal, then the only effect of the change will be displacement in time (t), as shown in Fig. 3b. There will be no distortion of the shape of the wave. But in the presence of phase distortion, however, the wave will not only be time displaced, but also distorted. Figures 3c and 3d show two forms of distortion that can occur with phase non-linearity.

A slightly different view of the same phenomenon is shown in Figs. 4 and 5. Consider a system in which the bandwidth can be varied across several limits, represented by curves **a**, **b** and **c** in Fig. 4. Curve **c** represents the most restrictive of the three possibilities because it sharply limits both low and high frequency response, while curve **a** is the least restrictive. Note in Fig. 5 the various responses to the three bandwidths represented in Fig. 4. These curves can be simulated by examining the response to square waves in *R-C* filter networks. In fact, one of the problems that one must consider when using electronic filters is the effects of the –6 dB points on the applied waveform.

One might erroneously assume from the discussion above that the instrument designer should select amplifiers with as wide a bandwidth as possible. That is not the case, however, because bandwidth can cause other problems at least as severe as those that are solved. Noise, for example, is proportional to bandwidth. It is possible to eliminate the problems of noise, plus certain input signal problems such as ringing or resonances by proper selection of the frequency response cut-off points. Thus, the selection of amplifier bandwidth and phase distortion characteristics is a trade-off between the need to make a high fidelity recording of the input event and the other problems that can occur in the system.

Very small signals are often obscured in noise of similar amplitude. If this signal is amplified and then applied directly to the A-D converter for input to a computer, then large errors can be expected. Consider a contrived but representative example in Fig. 6. The signal in Fig. 6a is a 100-Hz, 1-mV sine wave signal. In Fig. 6b, however, the signal is mixed linearly with the output of a zener diode pink noise generator to produce a realistic analogue signal as might be found in practical instrumentation applications. Clearly, direct digitization would result in a large error, except in expensive, high performance computer systems. Fig. 6c and 6d show one and two stages of analogue filtering, one a simple passive *R-C* filter and the other a single-stage operational amplifier filter. Figure 6d is the recovered 1-mV sine wave with only a small error term from noise sources.

Even when digitization occurs relatively rapidly in the system, a little analogue circuitry can produce significant improvement in the signal. Figure 7 shows the circuit diagram of an analogue amplifier box that I built for a Grass FT-3 force-displacement sensor used by a medical physiologist. This device produces an output signal of about 180 µV/gram-force[b], and was experiencing severe interference from the 50/60 Hz AC power mains radiated signal. By preamplifying the signal ×200, and providing a small degree of low-pass filtering in the electronic circuit box mounted to the sensor, the interference was reduced to a negligible amount.

Next month...

In Part 2 of this article we will take a look at the problem of interfacing sensors to amplifiers at the front-end of the analogue subsystem. ❑

Acknowledgement

I am indebted to Mr. John M. Brown of Burr-Brown Corporation, Tucson, AZ, USA, for some of the examples used in this article.

Recommendations for further reading

1. Richard S.C. Cobbold, *Transducers for Biomedical Applications*, John Wiley & Sons, Inc. (New York, 1974).
2. L.A. Geddes and L.E. Baker, *Principles of Applied Biomedical Instrumentation*, John Wiley & Sons, Inc. (New York, 1968).
3. Joseph J. Carr and John M. Brown, *Introduction to Biomedical Equipment Technology*, John Wiley & Sons, Inc. (New York,1981); acquired and republished by Prentice-Hall (1989).

References and Notes

1. Note: Gain is defined as the ratio of the output function to the input function. Because voltage gain (A_v) is used here for illustration purposes, A_v equals U_o/U_{in}.

(b). One gram-force is about 980 dynes.

APPLICATION NOTES

The contents of this article are based on information supplied by electronic component manufacturers or their representatives, and do not imply practical experience by *Elektor Electronics*.

INFRA-RED RECEIVER/DEMODULATOR SFH505A (Siemens Components)

An increasing number of infra-red remote control circuits in video recorders and TV sets have an integrated infra-red receiver/demodulator. The advantages of such modules are clear: a small, less complex, circuit, and lower production costs. Furthermore, the combination of an IR receiver and a demodulator in one package increases the reliability of the remote control. This Siemens application note takes a further look at the SFH505A.

THE SFH505A from Siemens Components is an integrated circuit that contains an infra-red receiver and a demodulator. The SFH505A is a hybrid combination of the SFH205 silicon PIN photodiode and the TDA4065 IR demodulator (Fig. 1). Together with the SDA2208-3 IR transmitter, the SFH505A offers a modern, cost-effective, concept for infra-red remote control.

The SFH505A guarantees excellent technical characteristics, and outperforms similar products in a number of respects:

- Photodiode and receiver/demodulator in one;
- Very low external components count (supply voltage decoupling only);
- On-chip receiver bandpass matches transmitter carrier frequency (30 kHz);
- Maximum photosensitivity at 950 nm IR wavelength;
- Switching threshold of 40 nW/cm^2 at 950 nm/30 kHz (no ambient light, T_a=25°C; see Fig. 2);
- High sensitivity yields large range. Test transmitter: SDA2208-3 with 3 × LD271 switched at I_f=0.8 A.
  - no ambient light: ≥33 m
  - Weak ambient light: ≥31 m
  - Strong ambient light (distance from light source 1 m at 45°):
  - 100-W light bulb: ≥21 m
  - switched fluorescent tube (Osram Dulux EL15 W): ≥12 m
- High noise immunity;
- Supply voltage 5 V ±10%, current consumption only 0.65 mA (output high);
- Small outline (T0220 like) plastic package with only three terminals (+5 V, GND, OUT). Package optimized for 50° half-power points (Fig. 3);
- Internal metal screening between

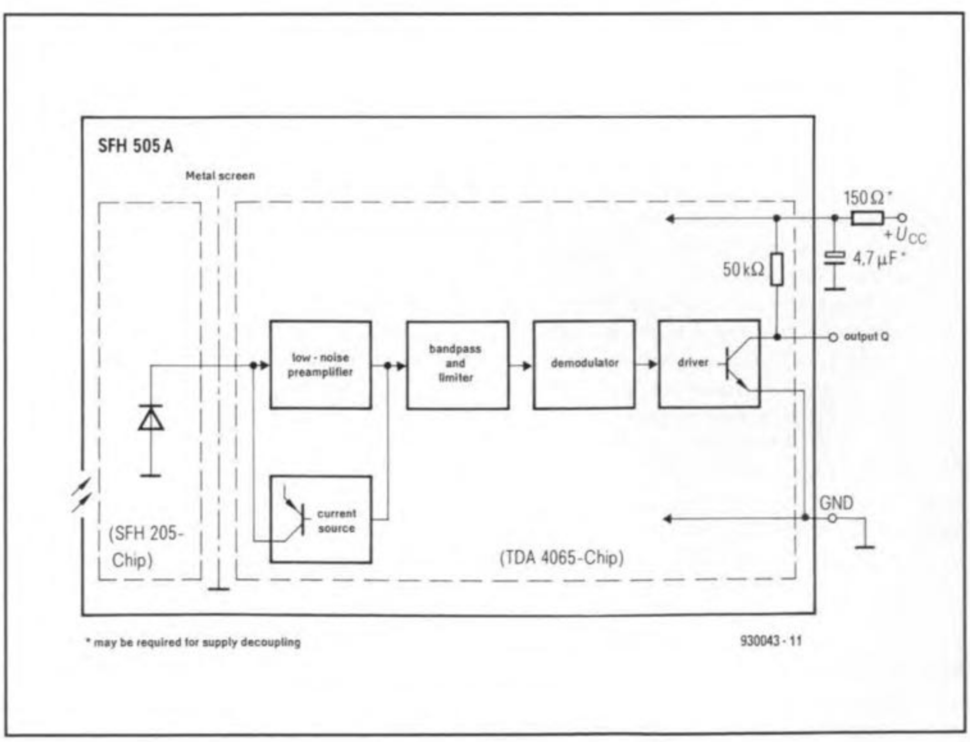


Fig. 1. The SFH505A consists of a photodiode (SFH205) and a demodulator (TDA4065).

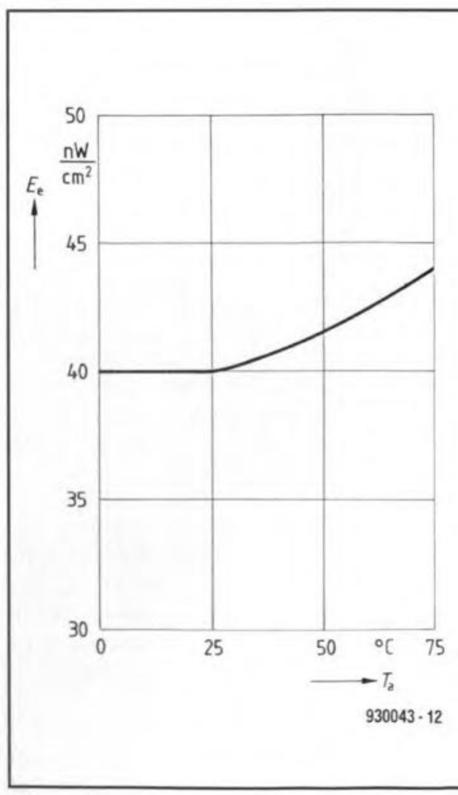


Fig. 2. The switching threshold luminous intensity, E_e, is constant over a temperature range from 0 to 25°C.

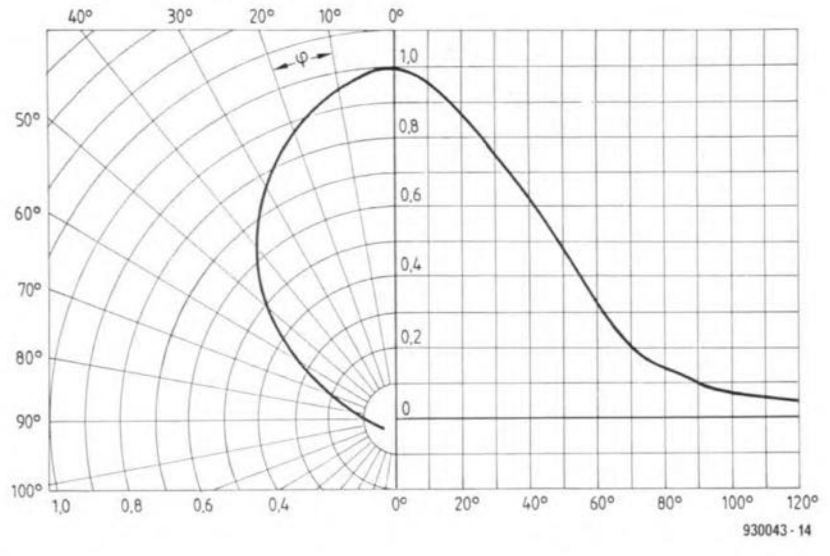


Fig. 3. SFH505A directivity beam pattern with half-power points at about 50°.

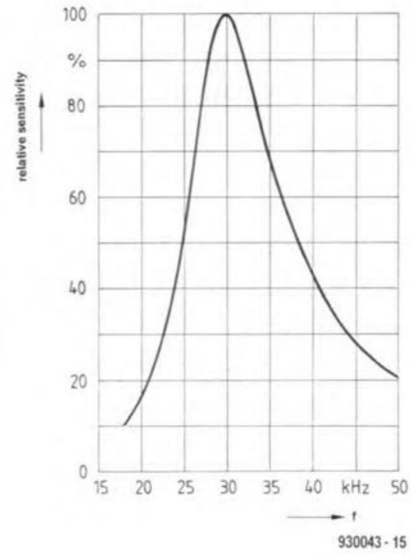


Fig. 4. The selectivity achieved with the aid of the integrated 30-kHz bandpass filter adds considerably to the noise immunity of the complete IR remote control system.

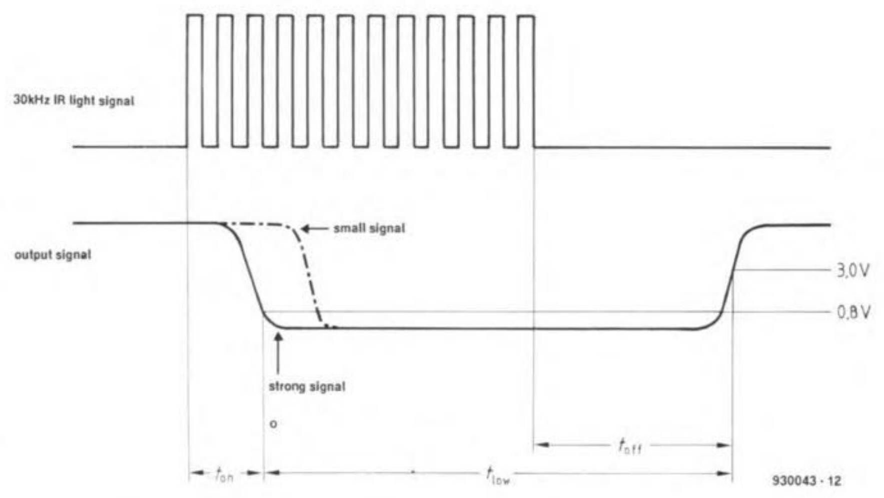


Fig. 5. The demodulator contained in the SFH505A converts the modulated IR carrier into logic pulses, which are available at the buffered output of the device.

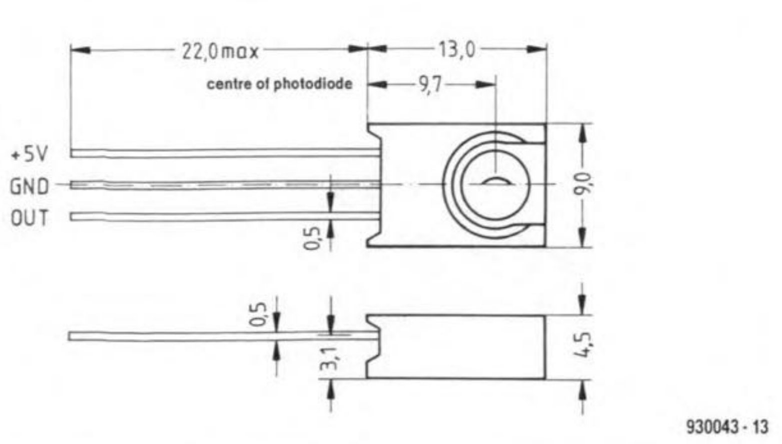


Fig. 6. SFH505A package outline and pinout.

photodiode and demodulator.

In the receiver, the infra-red carrier signal supplied by the photodiode is amplified by a low-noise preamplifier. The bias setting of the preamplifier is stabilized with the aid of a current source, which also serves to compensate low-frequency components in the photodiode current, caused by ambient light sources. The bandpass filter that follows the preamplifier is tuned to 30 kHz, and filters the signal before it is applied to the demodulator. The demodulated pulses appear at the output of the on-chip buffer stage (Fig. 5). ■

Source:
Siemens Components issue 1/91.

ACTIVE PROBE

An active probe is required wherever the capacitance presented by a test instrument (oscilloscope or voltmeter) must be kept as small as possible to prevent the test circuit being unduly loaded. The active probe described here has a maximum input frequency of about 150 MHz, and an input capacitance of only 2 pF.

By R.G. Payne

THE cost of a commercial active high-frequency probe puts this highly desirable workshop tool well out of the reach of the average electronics enthusiast. By contrast, the 180-MHz active probe described here can be built for about £20, which is 10 to 50 times cheaper than a commercial equivalent.

As shown by the circuit diagram (Fig. 1), the probe is based on Maxim's MAX405 precision video buffer/amplifier. This device was chosen mainly because of its large (180-MHz) bandwidth, 650 V/µs slew rate, and 35-ns settling time. Also not to be waved aside are its very small differential phase error (0.01%) and its ability to drive fairly heavy loads (60 mA) at low impedances.

The attenuation of the active probe is 2 times, which is five times smaller than a conventional ×10 passive probe. Hence, the active probe allows much smaller signals to be measured. Although Maxim in the MAX405 datasheets state a typical device input capacitance of 0.7 pF, about 2 pF was measured on a prototype of the active probe. The following measurement method was used to establish this value: first, a 10-kΩ SMA (surface mount assembly) resistor was connected to the probe input. Next, a pulse generator with an output impedance of 50 Ω and a signal rise time of 5 ns was connected to the SAM resistor. The output waveform was measured at 1τ, where $\tau = RC$, hence $C = \tau / R$. Evaluation is simple: the smaller 1τ, the smaller the input capacitance.

The active probe is powered by a ±6 V (symmetrical) supply. The author used a Farnell bench power supply

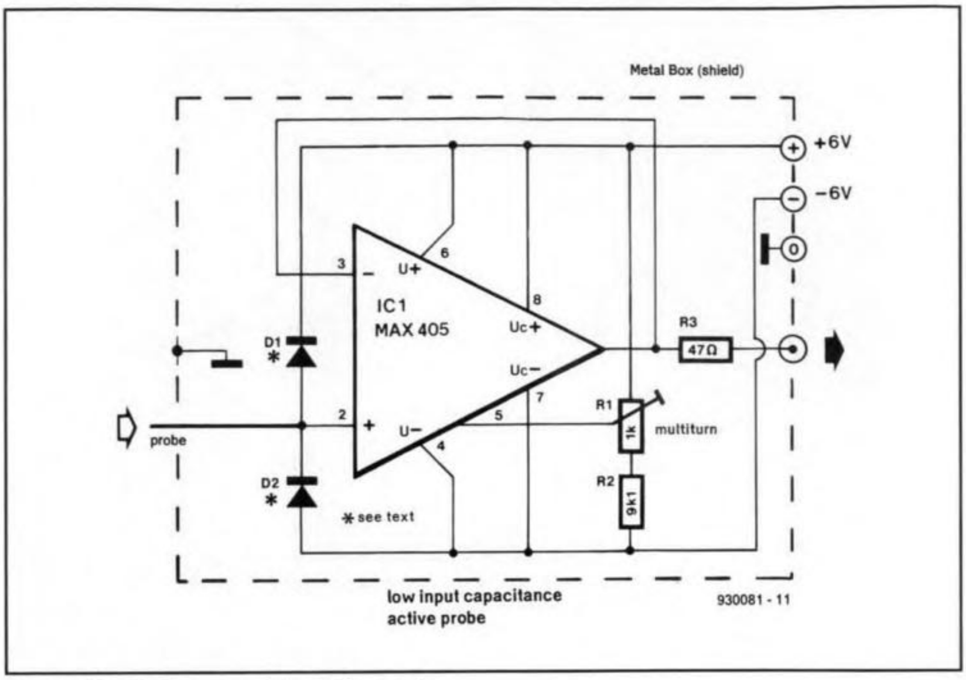


Fig. 1. Circuit diagram of the active probe.

(see Fig. 2), connected to the active probe via a screened 4-core cable.

In view of the very small signals measured, and, of course, the safety of the test set-up, great attention must be paid to proper earthing and screening. Also, the PCB on which the active probe is built must be as small as possible, and surface-mount components are essential to keep internal induc-

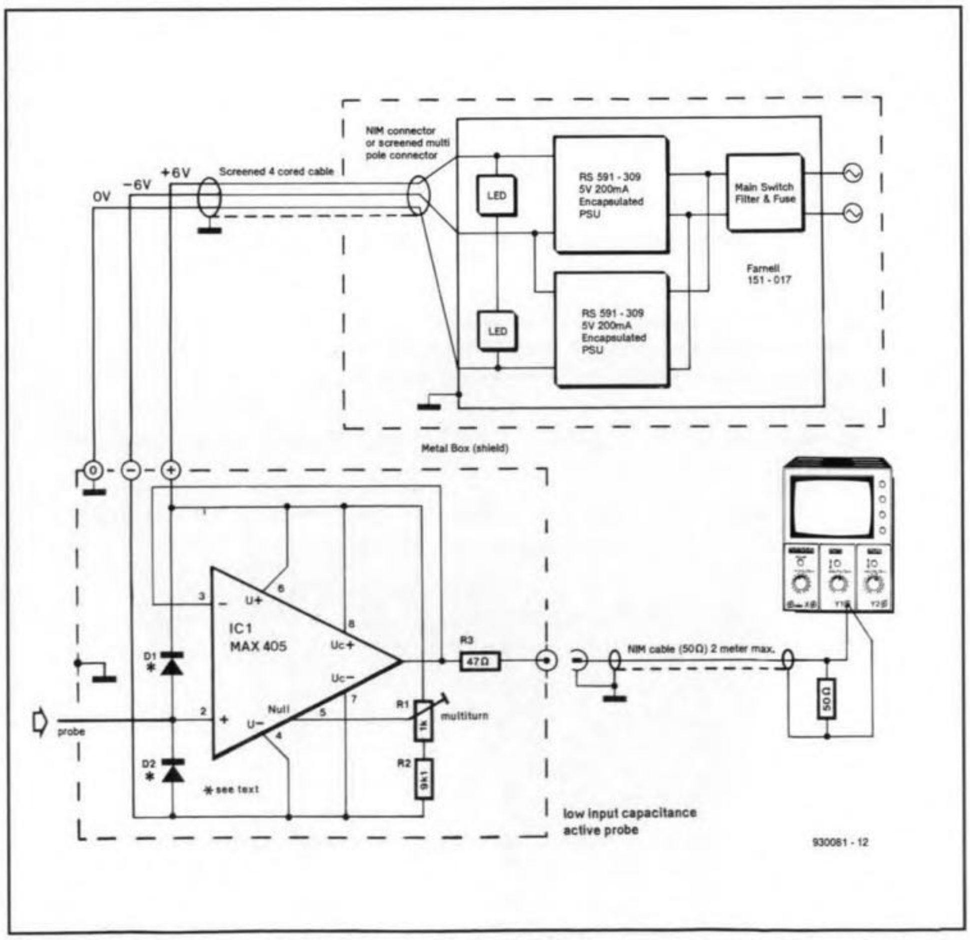


Fig. 2. Use of the probe in conjunction with a workbench power supply and an oscilloscope.

tance to a minimum. For optimum EMI screening the probe is best housed in a small metal tube, with a BNC output socket to connect the coax cable to the scope input.

The input voltage to the probe must be limited to the PSU rails, i.e., ±6 V. Low-capacitance protection diodes are fitted at the opamp input to afford input overload protection.

Finally, some additional technical data measured on the author's prototype:

Accuracy: ±5% max.
Input leakage: ±5 μA max.
Output d.c. off-set: ±1 mV (adjustable)
Linear input voltage range: ±4 V

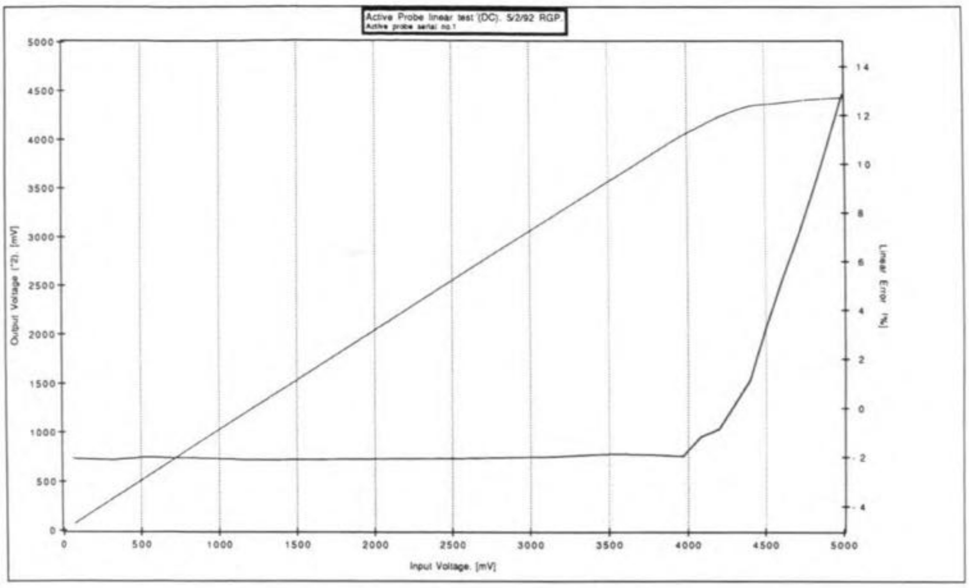


Fig. 3. Input and output linearity test result.

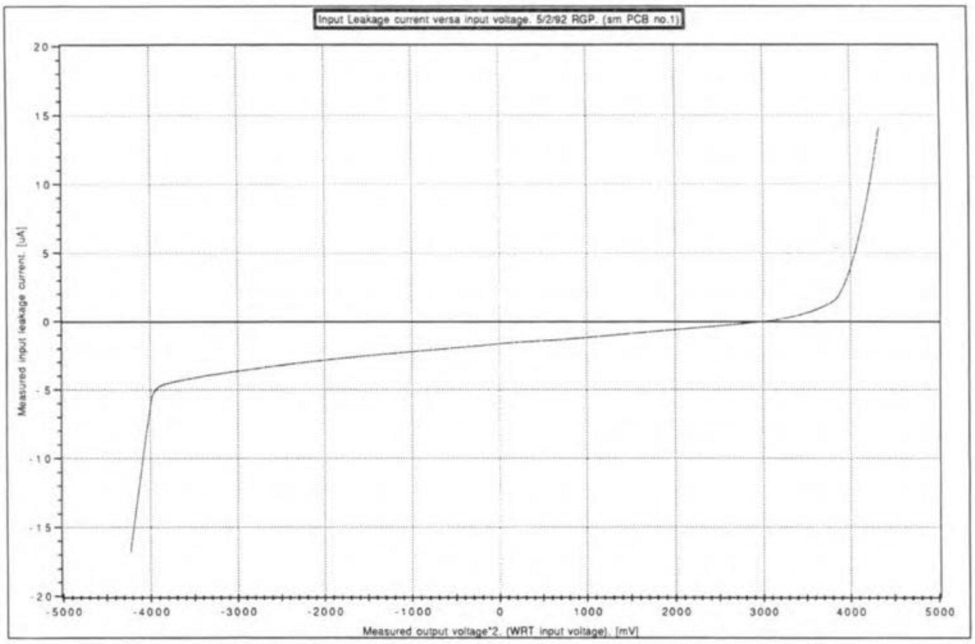


Fig. 4. Input leakage test result.

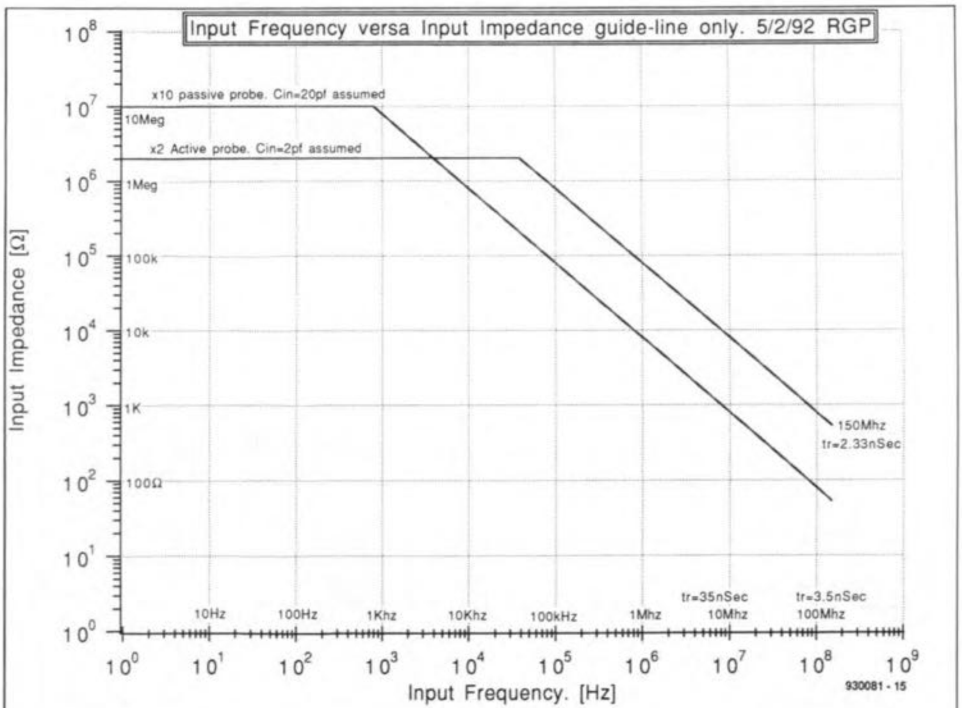


Fig. 5. Input impedance *vs* frequency.

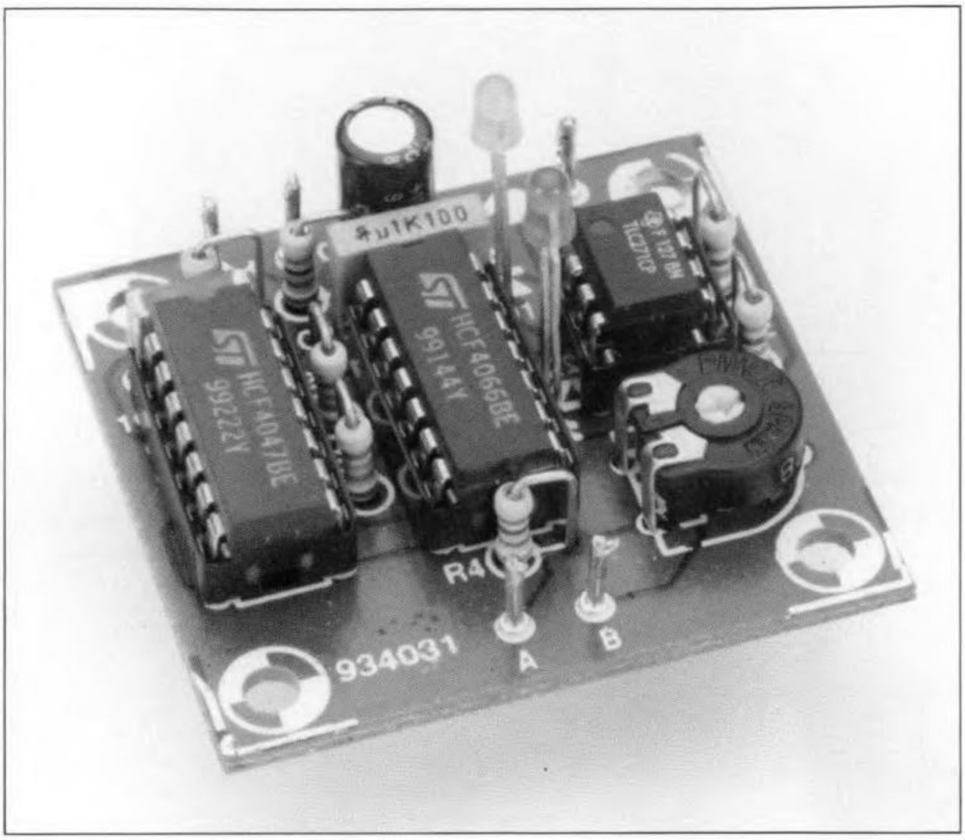
PLANT HUMIDITY MONITOR

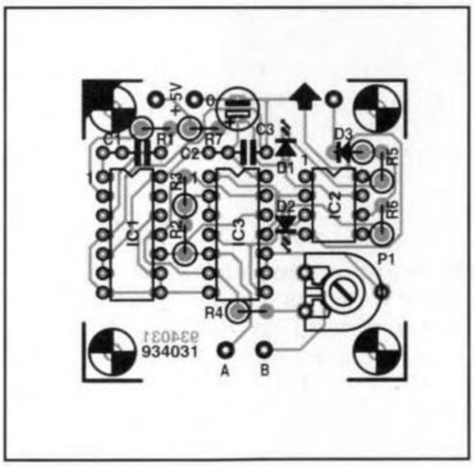
This little circuit monitors the humidity of plant soil, and should be a welcome help for forgetful readers who occasionally find their plants on the window sill suffering from acute drought.

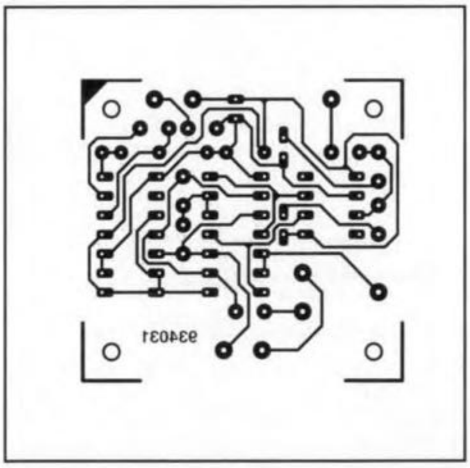
The plant humidity sensor consists of a handful of inexpensive parts, and gives a reliable indication when the soil is too dry. Circuit IC_1 is arranged as an oscillator that supplies two complementary switching signals, Q and Q\, at a frequency of about 58 Hz. Theoretically, the Q and Q\ signals are never active at the same time and thus form a perfect alternating current source that prevents electrolysis on electrodes A and B.

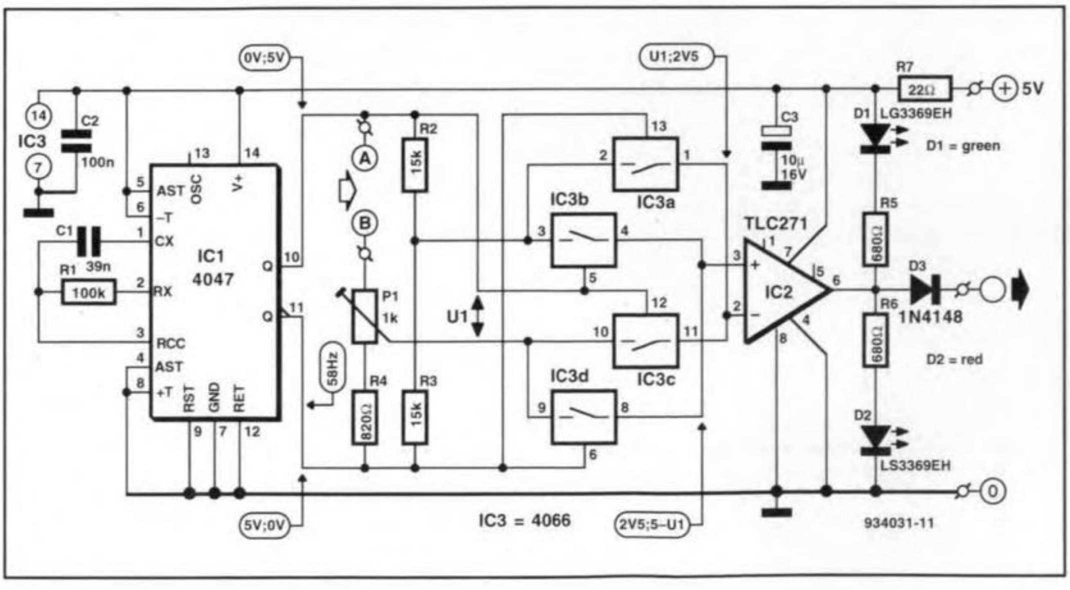
The potential at the wiper of P_1 is a function of the electrode resistance (R_{A-B}), and thus depends on the soil humidity. By comparing this potential with a reference, IC_2 indicates whether the soil humidity falls below a certain level, when the plant needs watering. The reference is derived from the symmetrical electrode current, and is taken from junction R_2-R_3, which is at a constant potential of 2.5 V with respect to ground.

The operation of the comparator is straightforward in spite of the alternat-









COMPONENTS LIST

Resistors:

1	100kΩ	R1
2	15kΩ	R2;R3
1	820Ω	R4
2	680Ω	R5;R6
1	22Ω	R7
1	1kΩ preset H	P1

Capacitors:

1	39nF	C1
1	100nF	C2
1	10µF 16V radial	C3

Semiconductors:

1	LG3369EH* (3mm low current green LED)	D1
1	LS3369EH* (3mm red low current LED)	D2
1	1N4148	D1
1	4047	IC1
1	TLC271	IC2
1	4066	IC3

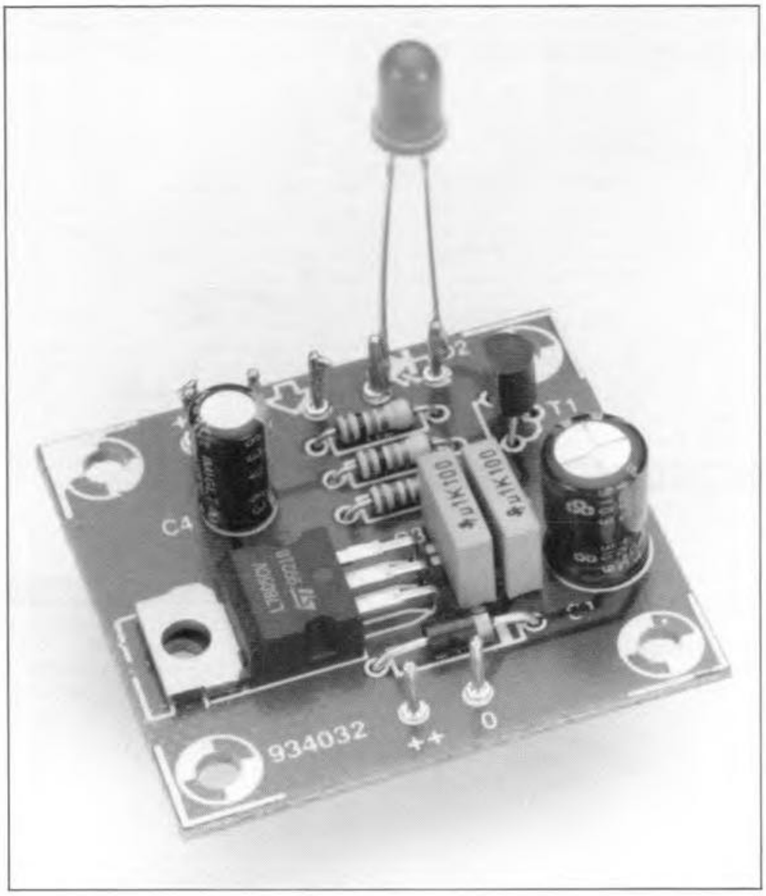
Miscellaneous:

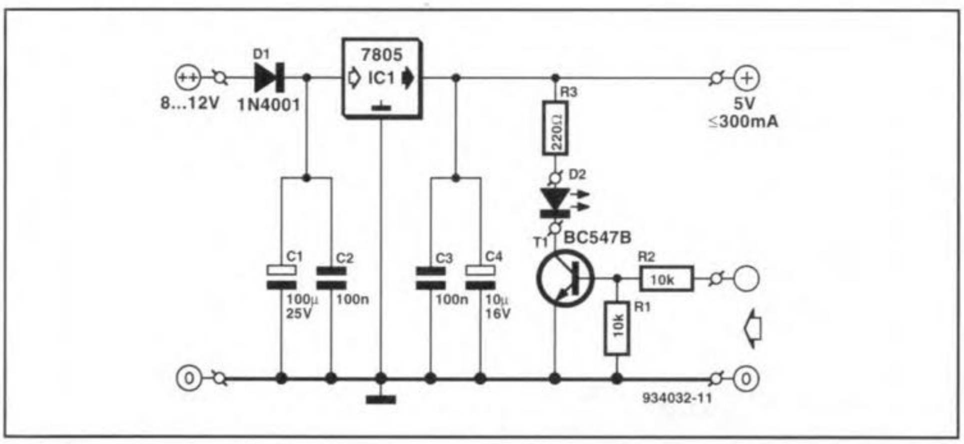
1 Enclosure dim. 65×50×30mm, e.g. Bopla** EG406

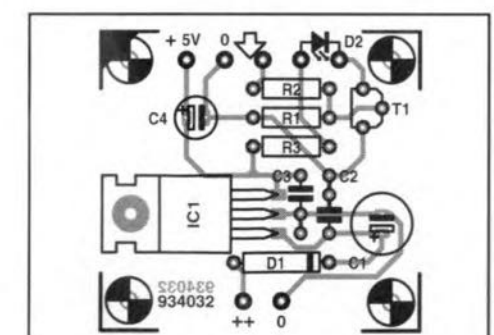
1 Printed circuit board 934031 (see page 110)

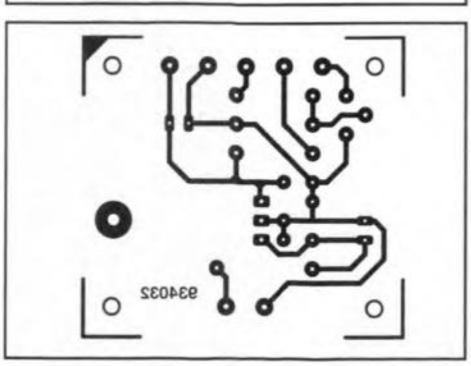
* Siemens product. ElectroValue (0784) 433603.

** Phoenix Mecano Ltd. (0296) 398853









COMPONENTS LIST

Resistors:

2	10kΩ	R1;R2
1	220Ω	R3

Capacitors:

1	100μF 25V radial	C1
2	100nF	C2;C3
1	10μF 16V radial	C4

Semiconductors:

1	1N4001	D1
1	LED 5m red	D2
1	BC547B	T1
1	7805	IC1

Miscellaneous:

1 Enclosure dim. 65¥50¥30mm, e.g. Bopla** EG406

1 Printed circuit board 934032 (see page 110)

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ing electrode current. A simple d.c. voltage comparison is, therefore, out of the question. Assuming that Q=0 V and Q\=+5 V, the wiper of P_1 is at a potential U_1 with respect to ground. Consequently, the wiper will be at $5-U_1$ when the electrode current is reversed (Q=+5 V; Q\=0 V). If, in the first situation (Q=0 V and Q\=+5 V), the wiper potential is, say, higher than the reference, it follows that it is lower than the reference in the other situation. To make sure that the comparator output is actuated in both cases, the input signals are swapped by the Q and Q\ outputs of IC_1. This is achieved with the aid of the four electronic switches contained in IC_3.

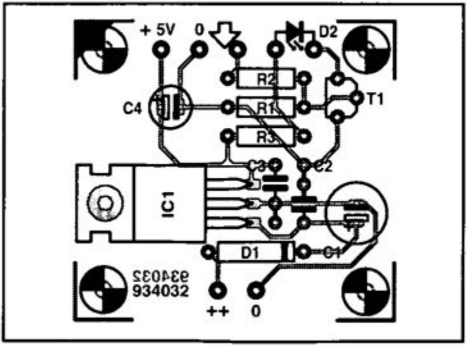
The red LED, D_2, lights when the soil is too dry. This happens when U_1 is greater than 2.5 V. Depending on the setting of preset P_1, this corresponds to a soil (inter-electrode) resistance of 0–1.82 kΩ. The further P_1 is turned towards the electrode connection, the higher the soil resistance (i.e., the drier the soil) required for the green LED to go out and the red LED to light. Interestingly, the capacitance of the soil may cause the red and the green LED to light at the same time, which gives a useful indication of 'between wet and dry'.

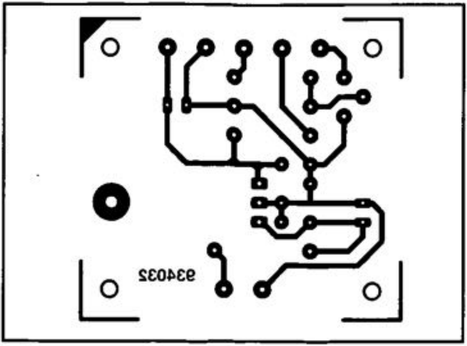
The electrodes are best made from the carbon rods salvaged from used batteries. This is an inexpensive solution, which also prevents corrosion. The electrodes are pushed into the plant soil about 4 cm apart. The exact adjustment of P_1 will depend on the type of plant to be monitored, and has to be established empirically. In most cases, however, good results will be obtained with P_1 set to mid-travel.

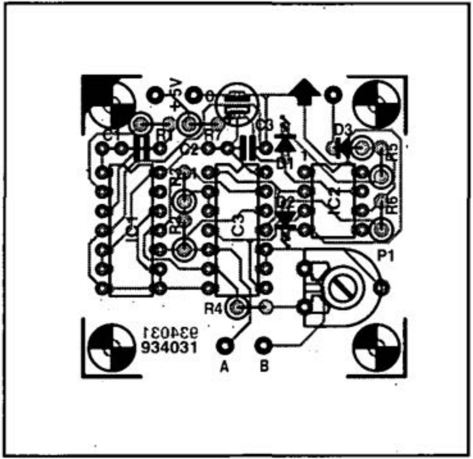
Since the sensor requires a stable 5 V supply, power is derived from an inexpensive mains adaptor whose direct output voltage is 'cleaned' and stabilized with the aid of an 7805 regulator. As the sensor draws a current of only about 5 mA, the supply shown here may be used to power a number of sensors.

In addition to providing the supply voltage to the sensor units, the power supply also functions as a central (remote) indicator. LED D_1 lights when one of the sensor units connected reports 'dry soil'. If none of the sensors reports 'dry soil', but at least one of them 'between dry and wet', LED D_1 lights at reduced intensity, since the voltage applied to the LED control input is then between 2 V and 3 V. Clearly, this means that all sensor outputs must be connected to the LED control input on the supply (wired-OR configuration).

(J. Ruiters — 934031/32)


+ 5V
0
D2
R2
R1
R3
T1
C4
C3
C2
IC1
D1
C1
934032
934032
++
0


934032


+5V
0
C1
R1
R7
C3
D3
C2
D1
R5
IC1
IC2
C3
D2
R6
P1
R4
934031
A
B