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Remotely controlled stroboscope

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FAX interface for PCs

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Elektor Electronics

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- Battery tester
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- Versatile NiCd battery charger
- TTL-level 100 MHz crystal oscillator

Front cover

Computers that recognize sounds are faced with enormous problems, including the fact that there are a very large number of different words in a language and so much variation in the way they are spoken.

This puts huge demands on computers that are required to interpret quickly the information fed in by the electronic 'ear'. A British system, known as Armada, is one of the first large vocabulary speech recognition systems to be based on an array of transputers, each capable of handling 10 million instructions a second. As it works in real time, it can process continuous speech.

Seen in the photograph, a scientist at the Royal Signals and Radar Research Establishment illustrates the science behind the Armada system, which uses a technique known as hidden Markov modelling.

Broadly, the system recognizes words by using statistical models of their constituent sound. Unlike other systems, it is said to be able to recognize words that do not occur in the training data, needs minimal training when new speakers are introduced and can easily be reconfigured for new applications.

Speech Research Unit, Royal Signals & Radar Establishment, St Andrews Road, Malvern WR14 3PS, Fax (0684) 894540.

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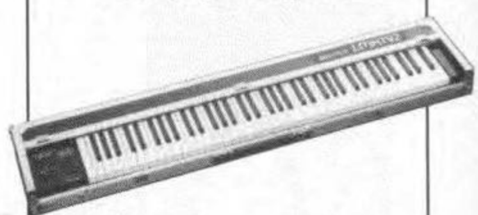

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FACSIMILE INTERFACE FOR IBM PCs AND COMPATIBLES

M. Brochand F6BFX and G. Warin F6DCK

The authors, two dyed-in-the-wool radio amateurs, describe how the facsimile decoder we published early last year for Atari and Archimedes micros can be adapted to run on EGA-based IBM PCs and compatibles. In addition, they present the required software, PCFAXPC, which, as an extra feature, supports fax transmission!

Hardware

The decoder described in Ref. 1 requires a few minor modifications to enable it to work with an IBM PC. Fortunately, most modern PCs accept TTL levels at their RS-232 port, so that the connection between it and the fax decoder need not be more complex than when an Archimedes or Atari computer is used. Apart from its simplicity, the advantage of the RS-232 link is that it ensures the correct pulse timing, irrespective of the clock speed of the PC. In fact, the sampling frequency of the fax board (determined by a quartz crystal plus associated 4060 divider) is in direct relation to the bit rate set on the RS-232 port.

To ensure perfect synchronization, the fax board also provides the central clock in the transmit mode. This clock, 2.4576 MHz, allows a bit rate of 19,200 per second on the RS-232 port. Other quartz crystals may be used (e.g., 4.9252 MHz), provided that the clock at the input of the 4040, pin 10, is 19,200 Hz to give a sampling rate of 19,200/12=1,600 per second. Hence, the 2-lines-per-second fax standard would require an image resolution of 800 pixels per line. Since the EGA card offers only 640 lines, 160 are simply ignored.

When the 2.4576 MHz crystal is retained, connect the 4060 to the 4040 via pin 6 instead of pin 14. This modification is not required when the 2.4576 MHz crystal is replaced by a 4.9252 MHz type.

The authors found that the central VCO frequency of the decoder must not be set too high to keep it within the passband of communications receivers. To centre the pass band at 1800 Hz, change the following component values: R_2=100k, C_2=22 nF and C_4=10nF. Also change P_4 to 4k7, and R_{14} to 470 Ω.

It is worth while to change P_4 to a potentiometer fitted on the front panel of the decoder to enable the adjustment to be changed in accordance with the standard of the received signal.

The Type NE567 synchronization detectors are no longer used and may be omitted, along with their associated passive parts R_7–R_{11}, C_7–C_{14}, D_1 and D_2, P_2 and P_3. The 4011 (IC_8) must remain on the board, however, since gates N_2 and N_3 remain necessary.

Using the decoder modified as described above, the authors have achieved excellent reception results with various facsimile stations, including meteorological and press photo services. In all cases, the LED-based read-out (described as an option in Ref. 1) proved an invaluable tuning aid.

The PC connection

Receive mode

Only two wires are required:

- **ground** of the decoder is connected to pin 7 (GND) of the 25-way D-connector at the PC end (or pin 1 of the AT-style 9-pin version);
- the **signal output** of the decoder (pin 10 of the 74150) is connected to pin 3 (RxD; received data) of the 25-way D-connector at the PC end (or pin 2 of the AT-style 9-pin version).

When a 25-way D-connector is used, connect pin 4 (RTS) to pin 5 (CTS), and pin 8 (DCD) to pin 20 (DTR) to make it permanently high. The connections for the 25-way as well as the 9-way D-connector are shown in Fig. 1.

The program defaults to the use of serial port COM1:. To run the program with COM2:, actuate either NUM LOCK or CAPS LOCK and exit the welcome screen by pressing the '2' key. The PC may run at 4.77 MHz or 8 MHz.

Transmit mode

In this mode, the datawords supplied to the computer at a rate of 1600 per second are ignored and serve to generate hardware interrupts. Each of these is acknowledged by the CPU sending out a tone—via the PC's internal loudspeaker—whose frequency corresponds to the relative intensity of the currently ad-

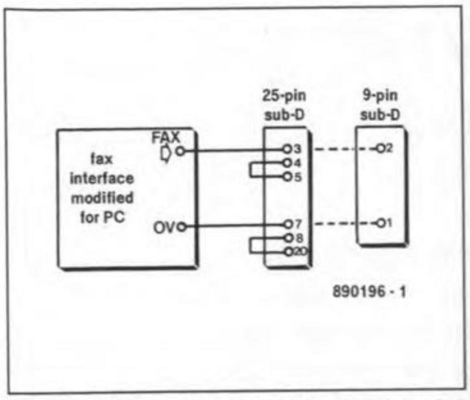
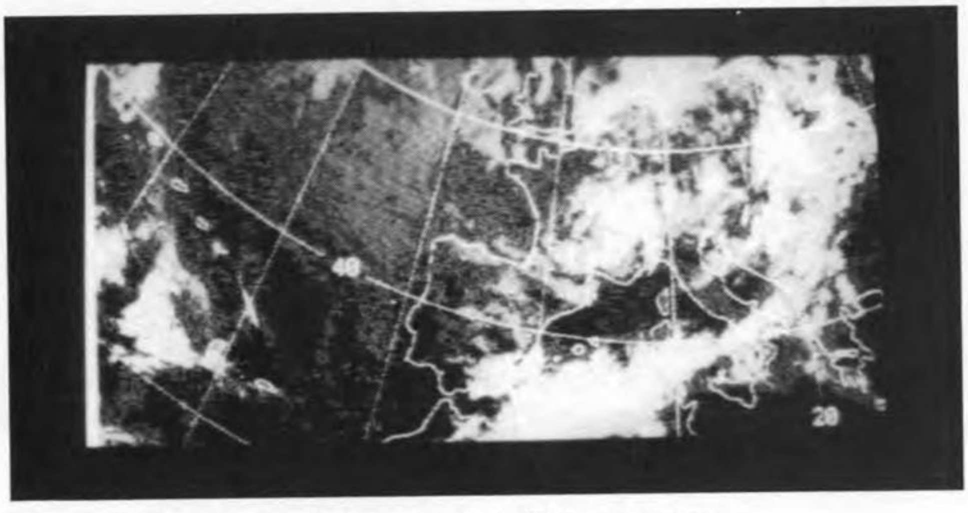


Fig. 1. Connections between the fax decoder and the RS-232 input on the PC.



Screen photograph of a weather chart received with the fax decoder.

dressed pixel on the screen. The necessary connection between the PC and the microphone input of the transceiver or transmitter is shown in Fig. 2. Do not omit the 100-nF capacitor since one terminal of the loudspeaker in the PC is usually at +5 V. Obviously, the transmit mode works only with the decoder board switched on.

Fax transmission is simple: load a picture from disk and press the 'T' (for 'transmit') key. The image is sent following a tone that allows the receiving station to get synchronized.

Control software

The control program, FAXEGA22, written in Turbo Pascal 4, handles the receive as well as the transmit mode with the aid of PC hardware interrupt $0C_H$.

The receive mode brings you pictures transmitted by radio amateurs, VLF utility stations, and meteorology services including low-orbit satellites. The .DOC file provides all the necessary background information on the program (produce your own copy of the manual by sending the file to your printer).

The software is contained on two diskettes. Disk ESS119-1 contains:

- the control program, FAXEGA22.EXE
- a video driver, EGAVGA.BGI
- two example picture files.

Disk ESS119-2 contains three further example image files of 112 Kbytes each.

The PCFAXPC program can be used on any IBM PC or compatible fitted with a colour or monochrome EGA card. Contrary to what might be assumed, the monochrome mode is likely to give the best results since it provides up to 14 levels of grey. In the EGA colour mode, up to 11 shades are available of a single colour. Owners of a VGA-based computer must use a utility program (supplied with your computer or with the VGA card) to switch to EGA emulation.

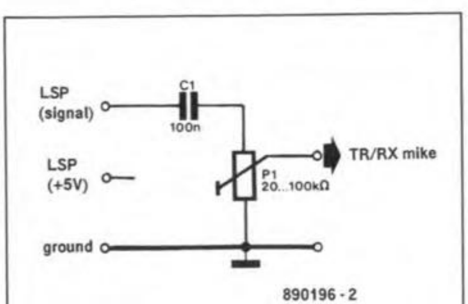


Fig. 2. Fax transmit connection.

Controls summary

The help menu is accessible via the **ESC** key. Press it again to return to the program. The functions of the keys used to control the program are as follows:

F1 to **F10** allow the grey levels or the colour shades to be changed within a certain range. Use **F5** for monochrome operation.

DEL clears the screen without changing the cursor position.

HOME clears the screen and moves the cursor to the top line without changing its horizontal position

PgUp moves the cursor to the first line without changing its horizontal position and without clearing the screen.

← moves the cursor about 10 mm to the left (use this function to centre the image during the run-in and sync periods).

→ moves the cursor about 10 mm to the right (use this function to centre the image during the run-in and sync periods).

CTRL ← moves the cursor about 1 mm to the left.

CTRL → moves the cursor about 1 mm to the right.

M writes the current image to disk. After pressing **M** the screen is cleared and you are prompted to enter the file name. The file is saved and the image appears again.

W retrieves an image file from disk. After pressing **W** you are prompted to enter the name of the requested file. Reception is blocked once the image appears on the screen. Press **Q** to receive a new image.

1, 2, 3 displays 2, 4 or 6 fax lines respectively on one picture line.

END exits PCFAXPC

Q blocks reception and allows an image to be frozen without disconnecting the interface.

S resumes reception after **Q** or **W**.

P writes lines from the right to the left (for press photo stations).

C writes lines from the left to the right (for meterological stations).

I creates a horizontally mirrored image. Useful as a correction to swap the **P** and **Q** modes once the image is on the screen.

T switches to transmit mode.

D go to DOS without leaving the program.

Example image files and PC Paintbrush

The program comes with a number of example files that contain press photographs and weather charts. These files allow the quality of the images to be gauged before commencing the construction of the interface. To load one of these example images, leave the welcome menu by pressing any key except **ESC**. Next, press **W** and enter the name of the requested file.

Note that the file load and display operations take a while, so do not press any keys in the mean time.

The transmit routine of the PCFAXPC program accepts image files made with the well-known drawing program PC PaintBrush and the associated image capture utility, Frieze. These powerful programs allow you to design your own images or capture existing graphics images. ■

Reference:
1. Facsimile interface for Atari ST (and Archimedes). *Elektor Electronics* January 1989.

ELECTRONIC LOAD

G. Boddington

For testing power supplies and transformers, an appropriate, preferably adjustable, load is indispensable. Often, a number of interlinked high-wattage resistors are used for this purpose, but that is not alway satisfactory, feasible or safe. The electronic load described here is a much more flexible and suitable solution.

Dissipation: 300 W (up to 1 kW with forced cooling)
Adjustable from 0.25 Ω to ∞
Voltage range 4–60 V
Maximum current 20 A
Operating modes: constant current and constant resistance
Internal or external modulation
Internal waveform generator (triangular, sinusoidal or rectangular)

The load is based on a number of power transistors. The output current, which is the sum of the collector currents of these transistors, is converted into heat that is lost by convection and radiation through a suitable heat sink.

The base current of the transistors is arranged at a value that results in the required level of emitter-collector current. Since the base-emitter voltage that determines the base current of a transistor varies with temperature (≈ 2–8 mV/°C), an opamp is used to iron out any consequent variations of the base current.

The end result is an adjustable load with a 'resistance' value ranging from almost zero to infinity and a thermal rating that depends solely on the power transistors and the manner in which these are cooled.

The load may operate in either the constant-current mode or the constant-resistance mode. In the first, the current remains constant irrespective of the applied voltage, while in the second it is directly proportional to the applied voltage.

A waveform generator (triangular, sinusoidal and rectangular) enables the 'resistance' to be modulated.

Circuit description

Each of the power transistors, T8–T12, in Fig. 1 can draw a collector current of up to 2 A with an appropriate heat sink. Resistors R24–R33 provide a measure of current feedback, which ensures equalization of the currents drawn by the individual power transistors.

Resistance simulation

Each group of power transistors, T3–T7 and T8–T12 respectively, is driven by one half of dual opamp IC2 via a driver transistor, T1 and T2 respectively. The driver is necessary since the opamp can not provide the required current by itself. Note that the two groups are connected in parallel, since "U" and "T" are strapped together as explained later.

The inverting input of the opamps is connected to the emitter resistor of the first power transistor in each group. The non-inverting inputs are connected in parallel and linked to the pole of switch S1b. This pole receives one of four different control signals via the switch contacts.

When S1 is in position 1 (7), terminal "P" carries part of the voltage, as set by P2-R9, that exists between terminal "U" and earth. The opamp tries to reduce the potential difference between its inputs to virtually zero. It will therefore increase the base currents of the power transistors, and thus the load current, until the voltage drop across R24 (R29) and the input voltage set by P2 are equal. When the input voltage rises, the potential at the non-inverting inputs, and thus the load current, increases. This means that the circuit behaves as a resistance, the value of which may be set with the aid of P2.

Modulation and constant current

Opamps IC1a and IC1b form a simple function generator that produces rectangular waveforms when S1 is in position 2 (8) and triangular waveforms when the switch is in position 3 (9). The frequency is adjustable over the range 5–50 Hz by P3.

The signal from the generator is amplified in IC1d and fed, via S1B and "P", to the non-inverting inputs of control amplifier IC2 where it serves as reference voltage. Since this potential is no longer dependent on the input voltage to the load, an increase in the input level no longer leads to a higher load current. In fact, if the signal, whether triangular or rectangular, is used as the control signal, the circuit functions as a modulated constant-current source.

The load current is modulated in the same way as the control signal. The gain of IC1d, which determines the depth of modulation, is set by P1.

Potentiometer P4 enables an offset to be added to the control sign al. This offset makes it possible to shift the modulation level with respect to zero. In other words, P1 sets the level by which the current varies, while P4 determines between which values modulation is effected, for instance, between 2.5 A and 3.0 A. This assumes, of course, that the unit or device under test can provide currents at those levels.

External modulation

Opamp IC1c serves as an inverting, unity-gain amplifier. Its output signal is available at contact 4 of S1. It may be fed with an external modulating signal via K1. The input must be between 0 V and +10 V. The control characteristic may be set between 3 A/V and 1.5 A/V for each power transistor. If, for example, the voltage at K1 changes by 100 mV, the output current varies by 3 A with P1 set to maximum and by 1.5 A with P1 set to minimum.

Constant-current operation

With K1 connected to ground, it is possible, with the aid of P4, to arrange a constant current to flow through the power transistors.

Lower loads

It is not necessary to use all ten power transistors: the circuit operates perfectly well with just one, but the permissible load is then, of course, reduced to 1/10. If the

POSITIONS OF SWITCH S1

1 (7) Resistance, adjustable with P2.
2 (8) Internally-modulated constant-current source (triangular).
3 (9) Internally modulated constant-current source (rectangular).
4 (10) Constant-current source, adjustable with P1; may be externally modulated via K1.

POTENTIOMETER FUNCTIONS

P1 Sets depth of modulation.
P2 Sets value of resistance.
P3 Sets modulating frequency.
P4 Sets constant-current mode or modulation level.

group T8 – T12 is omitted, T2, C4, R29–R33 and R19–R23 may also be omitted. The maximum load current is then 10 A.

Construction and alignment

The load is best built on the PCB shown in Fig. 2. This figure does not show the part of the board for power transistors T8–T12 and associated components since this is identical to that for T3–T7. Before any start can be made with populating, the board must be cut into four with a fine hacksaw.

Screw each of the two long parts to a 5 mm thick aluminium bracket. Since the collectors of the power transistors will be at the same potential, there is no need for insulating washers, provided that the heat sinks and aluminium brackets are isolated from the enclosure.

Do not fit the emitter resistors too close to the board, because even in normal operation these get fairly hot. The same applies to R22 and R23.

The choice of enclosure depends in the first instance on the heat sinks used. The requirements of these are fairly stringent because they have to dissipate some 300 W. The dissipation may be increased to 1 kW if forced cooling is used.

The only calibration is in the provision of a scale for P2. For that purpose, a laboratory-type power supply with variable output and capable of providing an output current of at least a few amperes is required. Measure the voltage and current for a number of positions of P2 and calculate the corresponding resistance. The resulting scale is linear for input voltages greater than about 4 V.

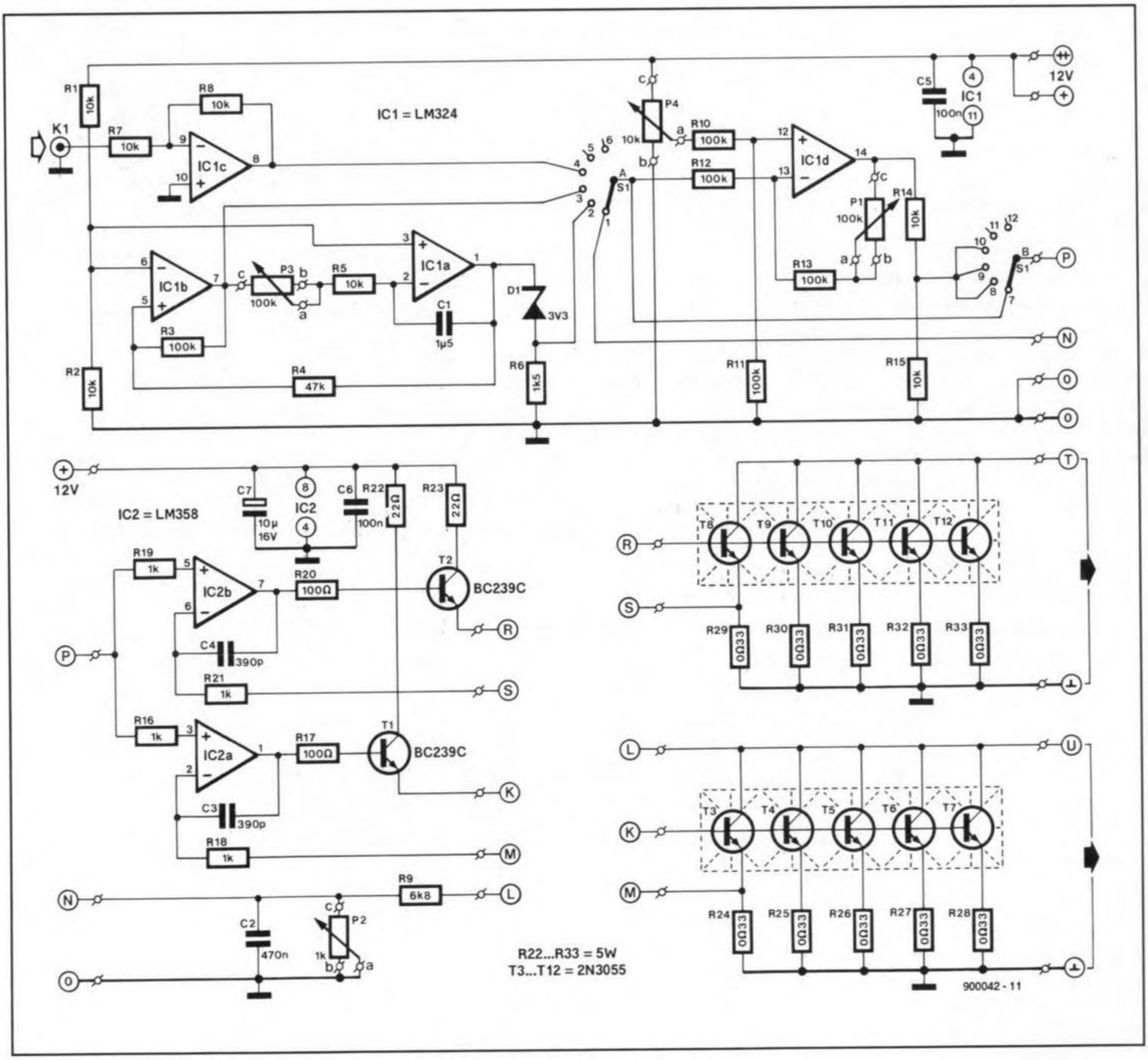


Fig. 1. Circuit diagram of the electronic load.

PARTS LIST

Resistors:
R1,R2,R5,R7,R8,R14,R15 = 10 k
R3,R10–R13 = 100 k
R4 = 47 k
R6 = 1k5
R9 = 6k8
R16,R18,R19,R21 = 1 k
R17,R20 = 100 Ω
R22,R23 = 22 Ω, 5 W
P1,P3 = 100 k, linear potentiometer
P2 = 1 k, linear potentiometer
P4 = 10 k, linear potentiometer

Capacitors:
C1 = 1µ5
C2 = 470 n
C3,C4 = 390 p
C5,C6 = 100 n
C7 = 10 µF, 16 V

Semiconductors:
D1 = 3V3, 400 mW
T3–T12 = 2N3055
IC1 = LM324
IC2 = LM358

Miscellaneous:
K1 = PCB mounting socket
S1 = 2-pole, 6-position rotary switch
2 off aluminium angle bracket, 5 mm thick
2 off heat sink, SK42, 75 mm, 1.5 K/W
(Dau UK Ltd – 0243 553031)

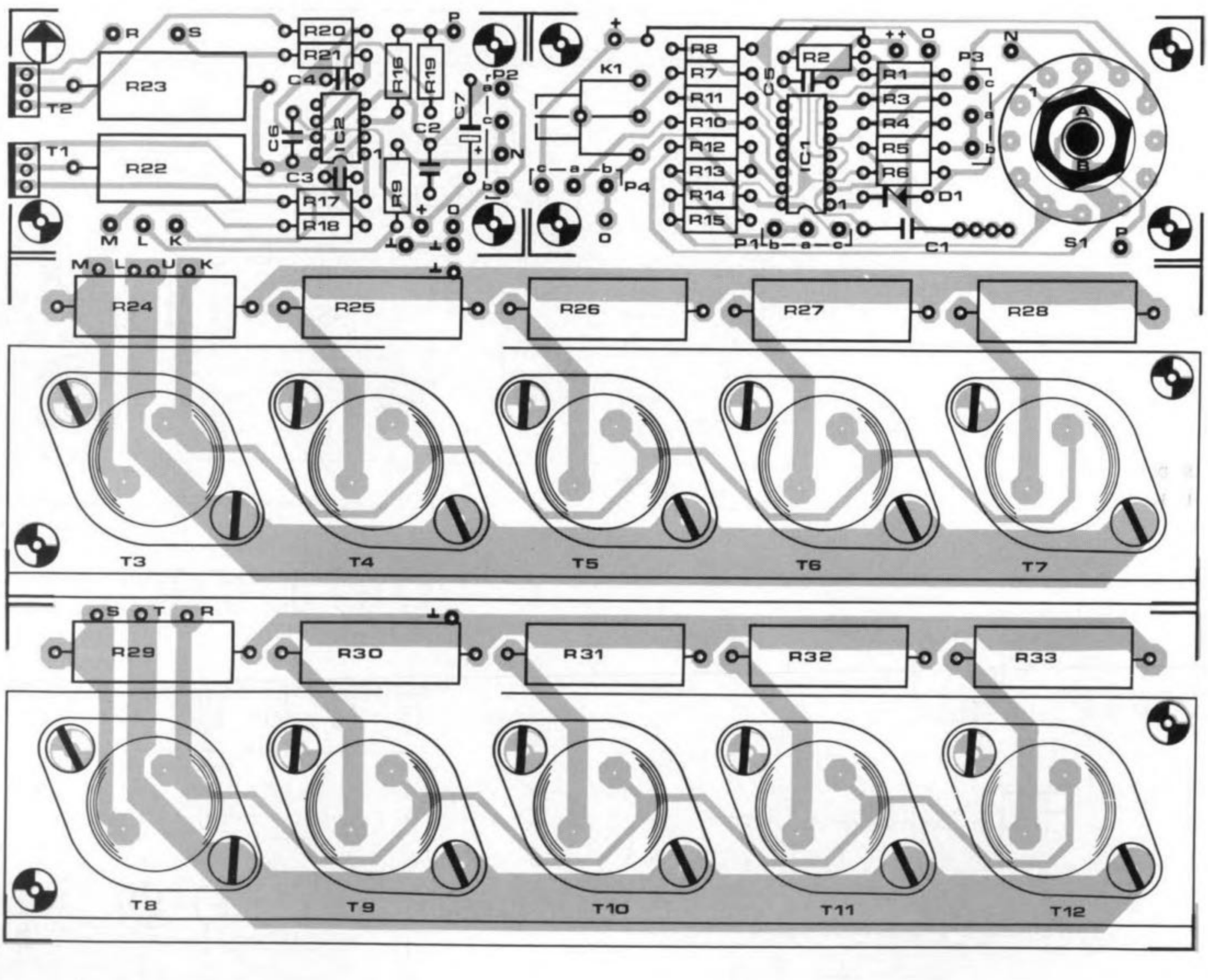


Fig. 2. Printed-circuit board for the electronic load. Note that the part for the second group of power transistors is not shown.

Fit the potiometers ~~and~~ switch on the front panel of the enclosure together with two heavy-duty, spring-loaded, insulated terminals. Connect points "U" and "T" to the positive terminals, and the earth points to the negative terminal.

To ensure that sufficient current flows through the power transistors, the load needs its own 12 V power supply, for which a simple unit with a 500 mA transformer and a Type 7812 voltage regulator will do fine.

Limiting values

The maximum values the power transistors can tolerate are not those found in their data sheet, that is, 60 V, 15 A, 115 W. Instead, the maximum dissipation is determined from the safe operating area (SOA) shown in Fig. 3. This shows that the maximum collector current decreases with a rising collector-emitter voltage. Conversely, when a current of 15 A flows through the transistor, its collector-emitter voltage should not exceed 8 V. It is imperative that at no time the C-E voltage exceed the limit for a given current and vice versa. But that is not all. The SOA characteristic in Fig. 3 refers to a maximum dissipation of 115 W at a case temperature of 25°C. With rising temperature, the dissipation is degraded at a rate of 0.65 W/°C. This means that at a case temperature of 80 °C the maximum dissipation is only 80 W and at 140 °C it is only 40 W.

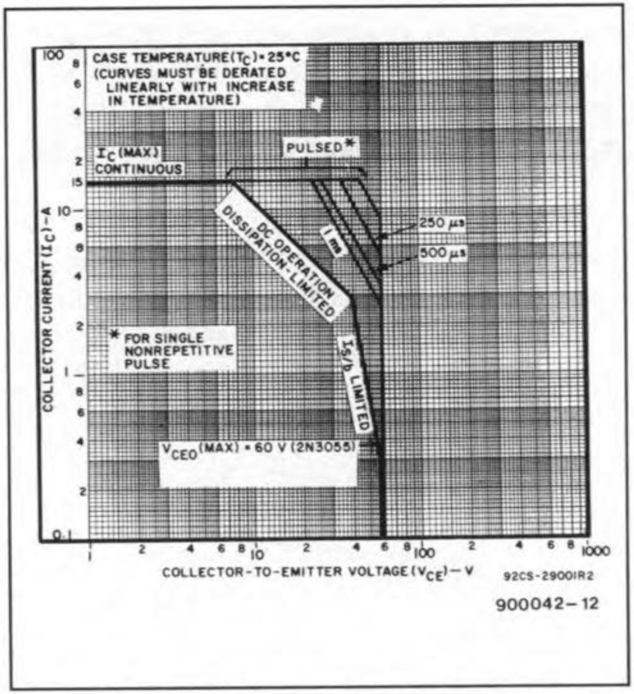


Fig. 3. Safe operating area of Type 2N3055 transistor.

The design of the load allows for each power transistor to tolerate a current of up to 2 A, so that the input voltage can go up to close to 60 V. If that is not sufficient, the 2N3055 transistors should be replaced by types with a higher rating.

Finally

The load is intended for use with direct currents, but alternating currents may be used by inserting a bridge rectifier between the current source and the load. The rectifier must, of course, be rated at the maximum current through the load.

The internal resistance, R_i, of a stabilized voltage source is calculated from:

$$R_i = \Delta U_m / \Delta I_m,$$

where ΔU_m and ΔI_m are the changes in modulating voltage and current respectively. In the "controlled constant-current" mode, the current is set with P_1 and P_4, while the change in voltage at the output terminals of the load may be measured with an oscilloscope.

The internal generator is also useful for determining whether the internal resistance depends on frequency.

In the "preset constant-current" mode, it is possible, for instance, to discharge a battery at a preset current. Measuring the exact capacity of the battery at a number of discharge currents is then a simple matter.

PARTS LIST

Resistors:
R1,R2,R5,R7,R8,R14,R15 = 10 k
R3,R10–R13 = 100 k
R4 = 47 k
R6 = 1k5
R9 = 6k8
R16,R18,R19,R21 = 1 k
R17,R20 = 100 Ω
R22,R23 = 22 Ω, 5 W
P1,P3 = 100 k, linear potentiometer
P2 = 1 k, linear potentiometer
P4 = 10 k, linear potentiometer

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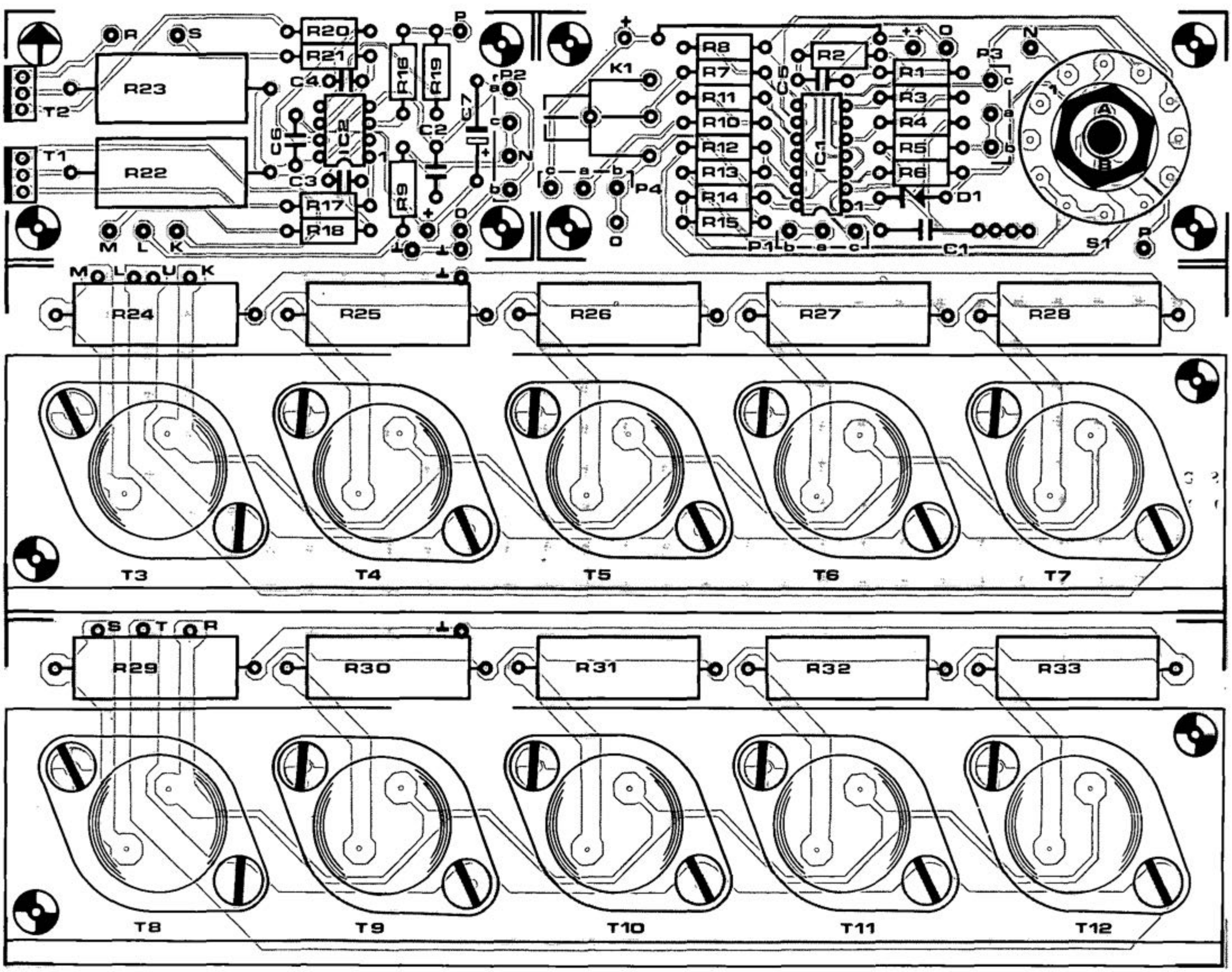

Fig. 2. Printed-circuit board for the electronic load. Note that the part for the second group of power transistors is not shown.

REMOTELY CONTROLLED STROBOSCOPE

This mains-powered stroboscope, designed by ELV, offers a wide range of settings as well as an external trigger feature so that it can be used as a slave flash unit.

A stroboscope is an instrument that produces light flashes with an intensity far greater than may be obtained with common light bulbs. The flash is a brilliant burst of light produced as a result of firing a gas (usually xenon) in a glass envelope, by means of a high-voltage pulse. Since the rate of the light flashes can be controlled accurately, moving objects illuminated by the stroboscope appear to stand still. This effect is obtained when the flash rate of the stroboscope corresponds to the period of the movement of the illuminated object. Useful applications of a stroboscope include the visual examination of rotating or relatively fast moving objects or parts such as flywheels and camshafts. Among the less useful, but certainly interesting, applications are lighting effects on theatre stages, on dancefloors, in discotheques and window sills.

The stroboscope presented here has two basic modes of operation:

- as a continuously operating stand-alone light effects unit with an adjustable flash rate of 0.5 to 5 flashes per second (= 30 to 300 per minute);
- as a slave flash unit with an adjustable trigger delay of up to one second. In this mode, the stroboscope is triggered by a light flash from another unit. After the set delay, the slave stroboscope produces its own flash. Exciting lighting effects may be obtained by using a single (mother-) stroboscope and an array of slave units, each with its own trigger delay.

Operation and controls

The stroboscope is simple to use since the complete circuit is contained in a single ABS enclosure that can be plugged straight into a mains outlet.

Operating the TRIGGER push-button in the lower right-hand corner of the front panel switches the unit from stand-alone (continuous) operation to slave operation, or vice versa (toggle function). A green and a red LED indicate the respective modes of operation.

The functions of the FLASH RATE and DELAY controls are self-evident: both are based on potentiometers and therefore offer a continuously variable setting. The light sensor located in the lower left-hand corner of the front panel has a lens to boost its sensitivity. This sensitivity, and that of the associated circuitry, is such that even relatively weak flashes, or flashes from a distance of 10 m or more, are reliably detected to enable the stroboscope to be triggered. Although the unit is largely insensitive to light from normal bulbs or sound-to-light units, two points should be noted in relation to the external triggering mode:

- the sensor must not be illuminated direct by a constant light source;
- flickering luminescent tubes may cause erroneous triggering owing to the light pulses they emit.

The shape of the reflector behind the xenon tube ensures a light distribution that is particularly suitable for effects applications. Since a straight xenon tube is used, the reflector is U-shaped rather than spherical as in, for instance, a torch.

Circuit description

Power supply and flash tube circuit

The power supply of the circuit consists of mains transformer Tr_1, diodes D_1–D_6 and capacitors C_1–C_3. Note that although a mains transformer is used, the circuit is not isolated from the mains: a path exists via R_1, R_2, D_1, D_2 and C_2. This means that the circuit must **never** be used when it is not enclosed in the ABS case supplied with the kit. After removing the stroboscope from the mains outlet, always wait at least 30 s before opening the enclosure so as to allow the flash capacitors to get rid of their lethal high voltage.

Diodes D_3–D_6 and capacitor C_3 provide voltage regulator IC_1 with its direct input voltage. The output voltage of IC_1 is 15 V.

The mains voltage is applied to a two-phase voltage doubler, D_1-C_1-D_2-C_2, via power series resistors R_1 and R_2. The flash voltage of about 600 V exists between the +terminal of C_1 and the –terminal of C_2. The xenon tube, H_1, is fired by a high-frequency, high-voltage burst at its trigger electrode. This burst is provided by the discharging of C_4 across the primary winding of the firing transformer, Tr_2. Voltages in excess of 10,000 V occur at this point.

The firing capacitor, C_4, is charged via R_{3a}, R_{3b} and the primary winding of Tr_2. When thyristor Thy_1 is fired via R_4, it conducts and enables C_4 to be discharged via the primary winding of Tr_2. The voltage induced in the secondary winding fires the xenon tube. Since the xenon gas in the tube conducts during the flash, C_1 and C_2 are rapidly discharged. The energy stored in these capacitors is thus converted to light. When the high voltage has fallen to about 100 V, the xenon tube turns into a high impedance again, so that the buffer capacitors can be charged again via R_1 and R_2. The firing capacitor, C_4, is also charged again via R_{3a} and R_{3b}. The values of the components used in the firing and supply circuit around the xenon tube are such that up to five flashes per second can be produced.

Continuous operation and mode selection

When the stroboscope is used in the stand-alone mode (continuous operation), the firing pulse for thyristor Thy_1 is provided by an oscillator formed by IC_{3c}-IC_{3d}. This is a fairly conventional two-gate stable multivibrator with potentiometer R_{19} acting as an output frequency control. Resistor R_{17} may have to be adapted to ensure

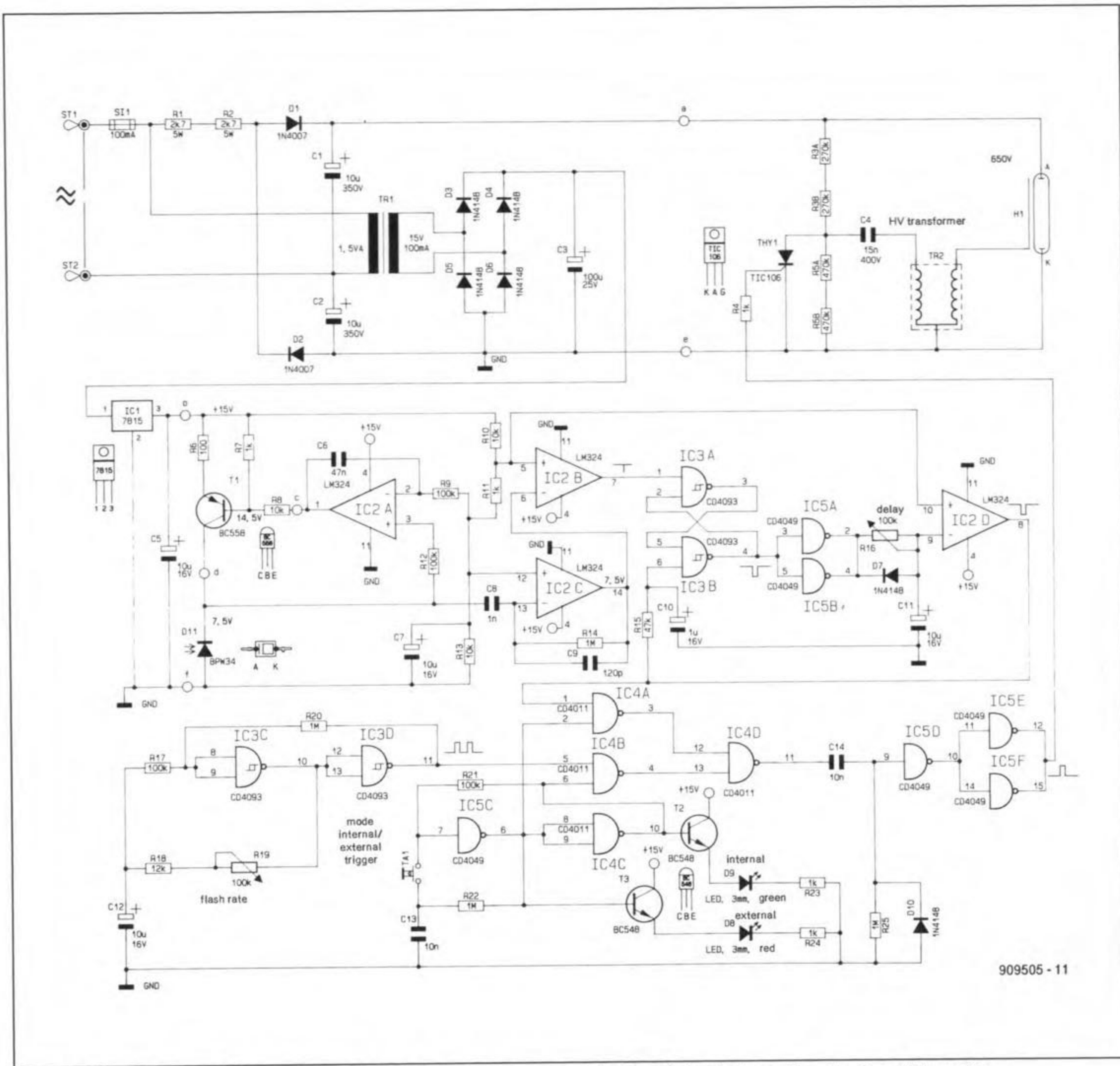

Fig. 1. Circuit diagram of the stroboscope. Note that there is no electrical insulation between the circuit and the mains.

the highest flash rate of 5 per second with R_{19} turned fully counter-clockwise. When this highest flash rate is exceeded, increase R_{17} to 120 kΩ. When it is too low, change R_{17} to 82 kΩ. When R_{19} is turned fully clockwise, the flash rate should be 0.5 per second, i.e., one flash is produced every two seconds.

The oscillator output signal is applied to input pin 5 of NAND gate IC_{4d}. Another NAND gate, IC_{4a}, is provided with external trigger pulses. The bistable composed of IC_{5c}-IC_{4c} and push-button Ta_1 determines whether the oscillator output signal or the external trigger output voltage is passed to IC_{4d}. Each time the push-button is pressed, the selection changes between IC_{4b} (continuous trigger) and IC_{4a} (external trigger).

A differentiating network, C_{14}-R_{25}, changes each level transition at the output of IC_{4d} into a positive-going needle pulse, which is fed to inverter IC_{5d}. The two parallel-connected inverters that follow IC_{5d}, IC_{5e} and IC_{5f}, make this pulse positive again for firing Thy_1 via R_4.

External trigger

When photodiode D_{11} detects externally generated light flashes, amplifier IC_{2c} supplies a positive output pulse, which is converted into a negative-going rectangular signal by comparator IC_{2b}. This signal sets bistable IC_{3a}-IC_{3b} via pin 1. The output, pin 4, changes from high to low so that buffer pair IC_{5a}-IC_{5b} supplies a positive pulse. This results in C_{11} being charged via potentiometer R_{16}. When the delay has lapsed, comparator IC_{2d} toggles and provides IC_{4a} with a negative pulse. Provided the stroboscope is in the continuous trigger mode (selected by Ta_1), the pulse obtained from the external trigger circuit causes the xenon tube to fire as described above. It also causes the rapid discharge of C_{10} via R_{15} so that bistable IC_{3a}-IC_{3b} is reset via its second input, pin 6. The result is that C_{11} is rapidly discharged via IC_{5a}-IC_{5b} and D_7 to prepare this circuit for a new trigger pulse. The short delay introduced by R_{15}-C_{10} is required to prevent the stroboscope being triggered by its own light flash.

The flash sensor circuit around photodiode D_{11} has an automatic sensitivity control function to cope with changing ambient light conditions. Transistor T_1 provides D_{11} with a current that causes the diode to drop about half the supply voltage. This voltage is fed to IC_{2a} via R_{12} for comparison with a reference level applied to pin 2 of the opamp. When the ambient light intensity increases, the voltage across D_{11} drops. The current source, T_1, and the opamp, IC_{2a} form a control loop that ensures a point of optimum sen-

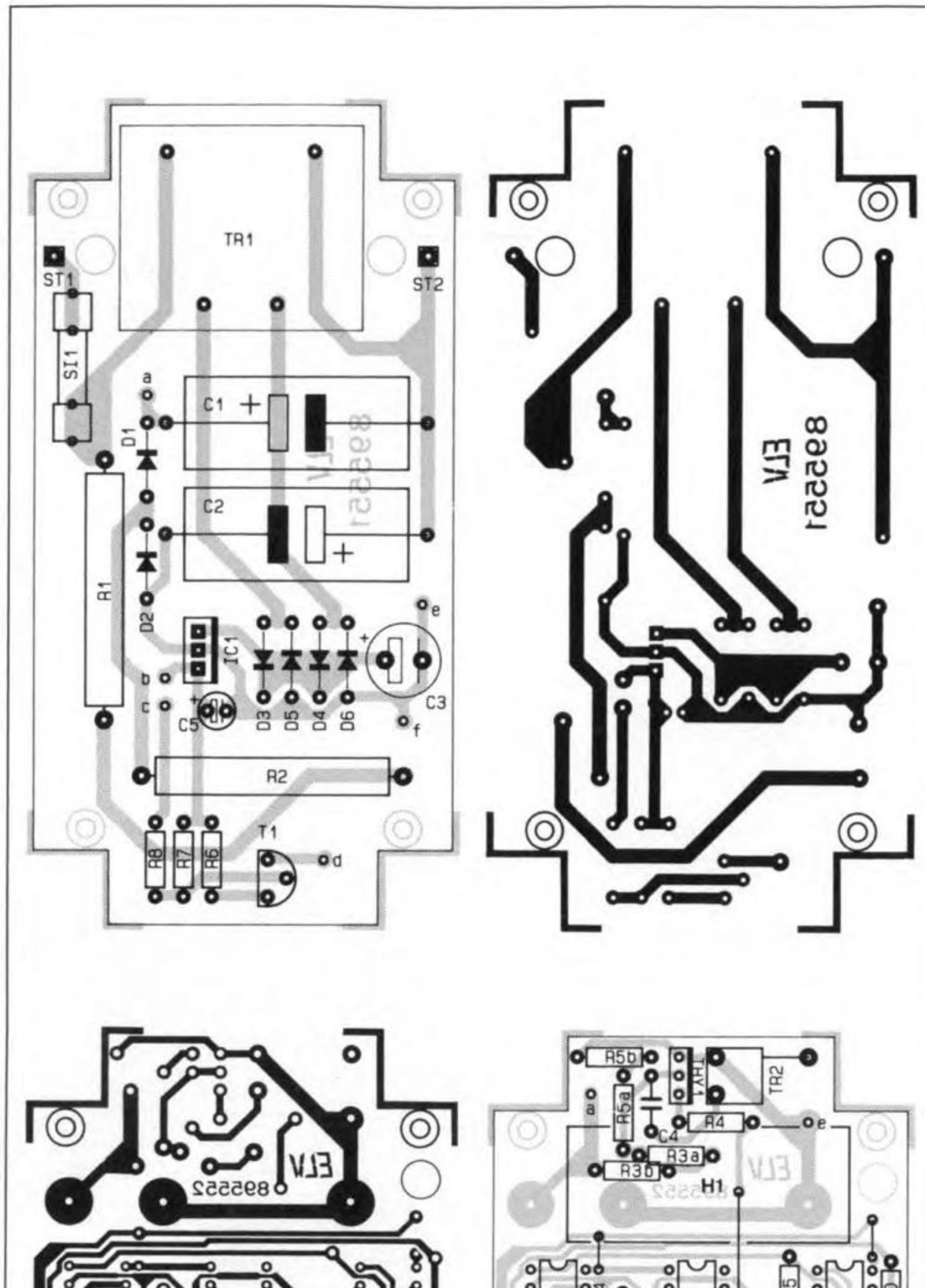

Fig. 2. Track lay-outs and component overlays of the two printed-circuit boards.

COMPONENTS LIST

contents of kit supplied by ELV France

Resistors:

1	100Ω	R6
5	1k0	R4;R7;R11;R23;R24
2	2k7 5W	R1;R2
3	10k	R8;R10;R13
1	12k	R18
1	47k	R15
4	100k	R9;R12;R17;R21
2	270k	R3a;R3b
4	1MΩ	R14;R20;R22;R25
2	100k potentiometer	R16;R19

Capacitors:

1	120pF	C9
1	1nF	C8
2	10nF	C13;C14
1	15nF 400V	C4
1	47nF	C6
1	1μF 16V	C10
4	10μF 16V	C5;C7;C11;C12
2	10μF 350V	C1;C2
1	100μF 25V	C3

Semiconductors:

1	7815	IC1
1	LM324	IC2
1	CD4011	IC4
1	CD4049	IC5
1	CD4093	IC3
2	BC548	T2;T3
1	BC558	T1
1	TIC106	Thy1
2	1N4007	D1;D2
6	1N4148	D3;D4;D5;D6;D7;D10
1	LED 3mm red	D8
1	LED 3mm green	D9
1	BPW34 with lens	D11

Miscellaneous:

1	mains transformer; sec. 15V/100mA	Tr1
1	tube firing transformer	Tr2
1	fuse 100mA slow	Si1
1	xenon flash tube	H1
1	push-button switch	Ta1
2	PCB-mount fuse holder	
1	reflector	
12	solder pin	
4	PCB spacer 30 mm	
1	M3×15 mm screw	
2	M3×20 mm screw	
4	M3×55 mm screw	
11	M3 nut	
2	solder eye 3.2 mm	
30 cm	0.22 mm² flexible wire	
10 cm	0.75 mm² flexible wire	
15 cm	silvered wire	
2	printed-circuit board	
1	ABS enclosure with moulded mains plug	

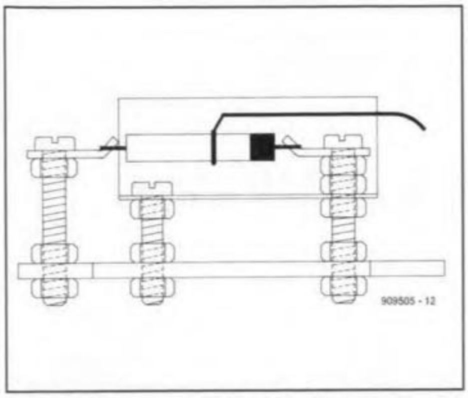

Fig. 3. Construction of the reflector, which is secured to the PCB with the aid of three M3 screws. These also serve to carry the xenon tube voltage.

sitivity of the photodiode in relation to the ambient light intensity.

Construction

The circuit is constructed on two printed-circuit boards —see Fig. 2. Construction is mostly straightforward on the electronic side; the following descriptions therefore detail mainly certain points in the mechanical work.

Start the population of the flash tube board with the nine wire links. Fit the two potentiometers at the track side of the board, and secure them with the nuts provided. Push-button Ta1 is mounted on two solder pins to enable it to protrude from the from panel. The reflector is fitted with the aid of three screws as shown in Fig. 3. The cathode (marked by a black ring) and the anode of the flash tube are connected to solder eyes fitted on M3 screws. Nuts are used to provide the correct mounting height of the reflector.

The high-voltage transformer, Tr2, is mounted on to the board as indicated by the component overlay. The firing voltage is carried by the flexible, insulated wire at the top of the transformer. Carefully remove the insulation material over a distance of about 12 mm at the end of this wire. Wind this wire end around the xenon tube, roughly at the position indicated in Fig. 3, and join the turns of the winding by soldering rapidly and care-

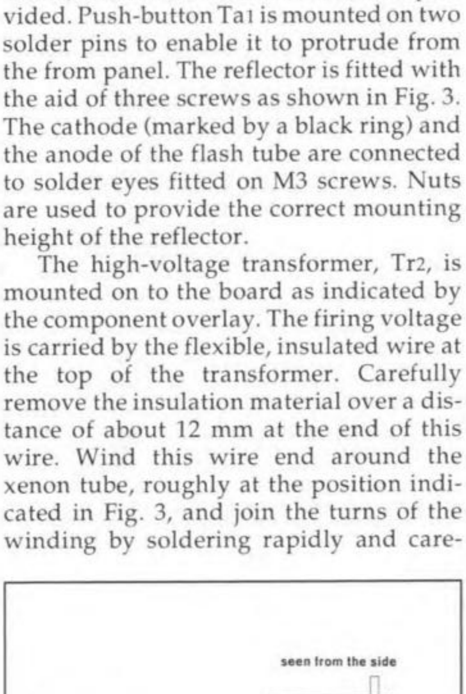

Fig. 4. Bend one pin of the high-voltage pulse transformer and solder it as shown.

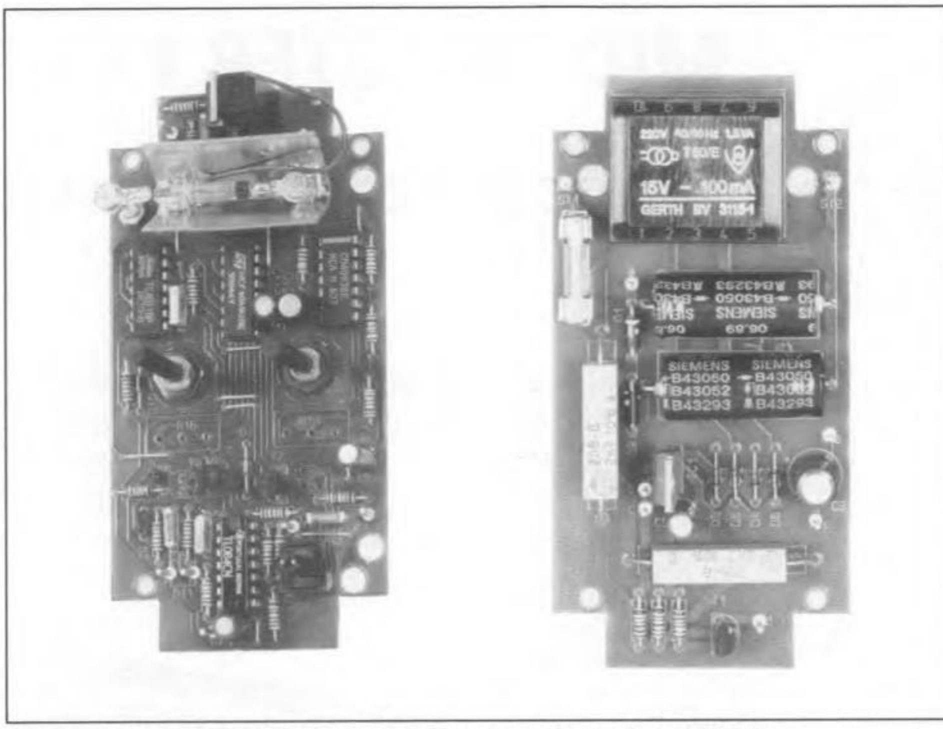

Fig. 5. Completed flash tube board (left) and transformer board (right).

fully. This completes the construction of the flash tube board.

It should be noted that ELV may supply a high-voltage transformer whose connections are different from the type for which the PCB was originally designed. Fortunately, the replacement transformer is readily fitted —see Fig. 4. Solder a short flexible wire to the high-voltage output terminal which is at the top of the transformer after this has been fitted on the PCB. The other end of this wire is secured to the flash tube as described earlier.

The construction of the transformer board is straightforward and requires no description other than that power resistors R1 and R2 are mounted at a height of about 10 mm above the transformer board.

The two boards are connected with 6 flexible wires between the corresponding points 'a' to 'f'. Connect the photodiode to the circuit with two short lengths of wire —the electrical orientation is marked on the component overlay (look at the line in the box printed on the PCB: it marks the anode of the BPW34, which is also visible as a kind of line in the photodiode).

Next, fit the completed boards into the enclosure. Use two short lengths (approx. 50 mm) of insulated 0.75 mm^2 wire to connect input terminals ST1 and ST2 to the moulded mains voltage (live and neutral) pins in the enclosure. The two boards are stacked and secured to the bottom half of the enclosure with the aid of four M3×35 mm screws and associated 30-mm long spacers. Each of the four screws is inserted into a corner hole of the flash tube board. Next, fit the spacer, and insert the screw into the corner hole in the transformer board. Place this 'sandwich' into the enclosure and secure the screws in the moulded threadings.

To prevent the reflector or other parts of the circuit being touched while the stroboscope is operating, the rectangular clearance for the xenon tube is covered by a 2-mm thick perspex plate, which is secured to the inside of the enclosure with a little glue. Also with safety in mind, make sure that the LEDs and the mode control push button, Ta1, are mounted at a height that enables them to protrude 1 to 2 mm from the front panel; they must fully occupy the relevant holes so as to prevent any likelihood of the circuit being touched.

The spindles of the two potentiometers are cut to a length of about 10 mm above the front panel. Finally, screw the top part of the enclosure to the bottom plate, fit the collet knobs on the potentiometer spindles, and fit the coloured caps. ■

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

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

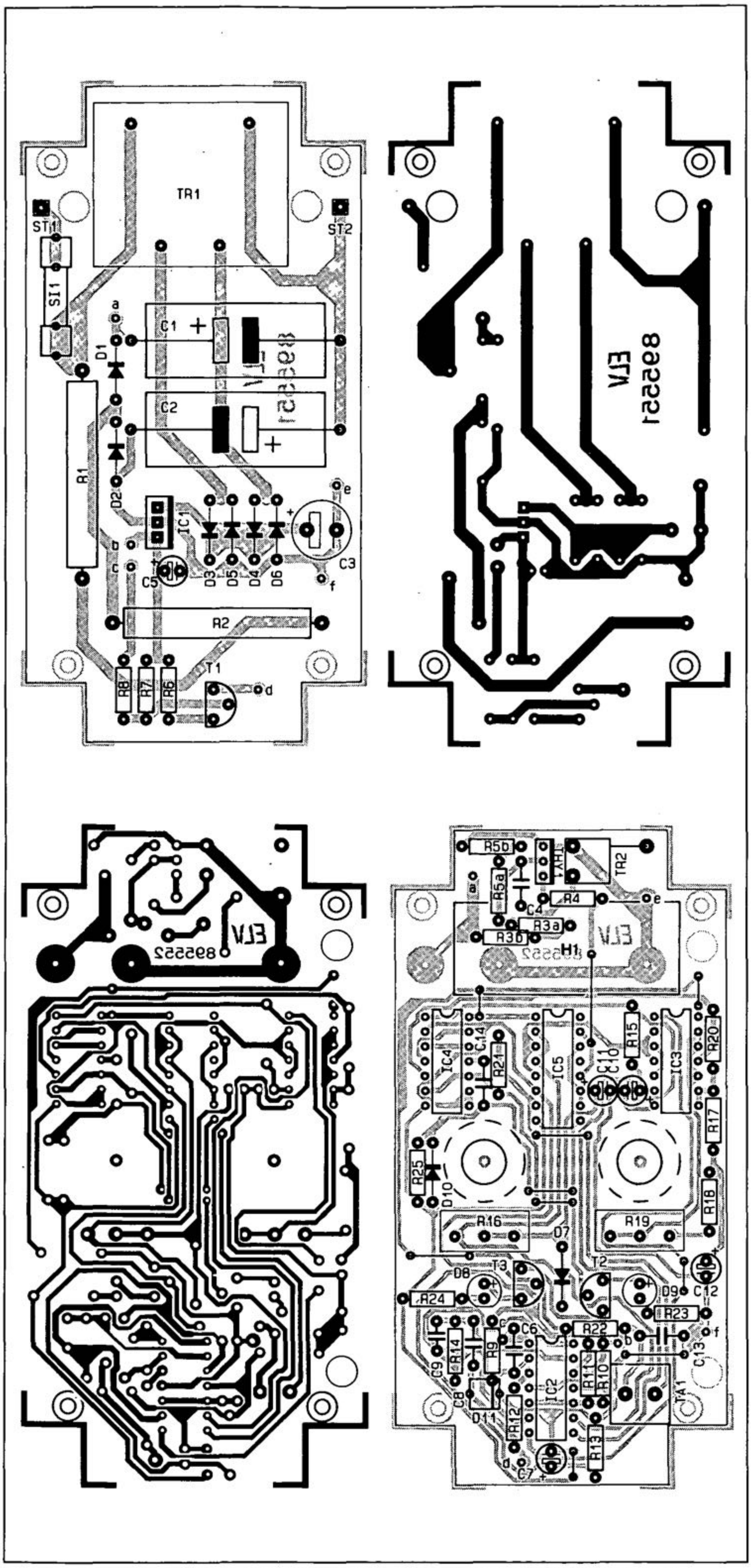

Fig. 2. Track lay-outs and component overlays of the two printed-circuit boards.

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4	100k	R9;R12;R17;R21
2	270k	R3a;R3b
4	1MΩ	R14;R20;R22;R25
2	100k potentiometer	R16;R19

Capacitors:

1	120pF	C9
1	1nF	C8
2	10nF	C13;C14
1	15nF 400V	C4
1	47nF	C6
1	1µF 16V	C10
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1	100µF 25V	C3

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1	TIC106	Thy1
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6	1N4148	D3;D4;D5;D6; D7;D10
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1	LED 3mm green	D9
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Miscellaneous:

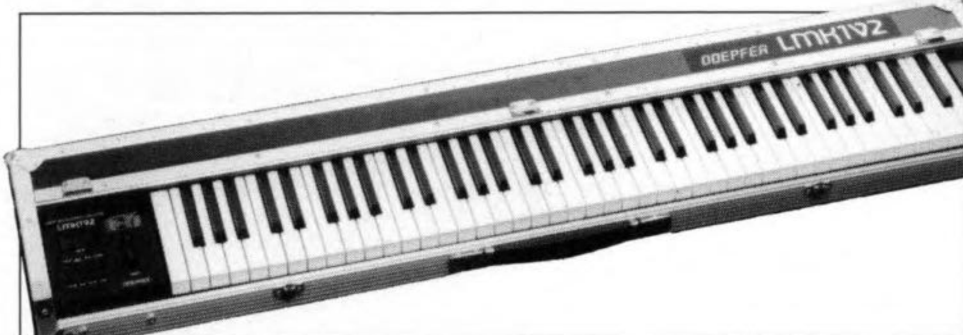
1	mains transformer; sec. 15V/100mA	Tr1
1	tube firing transformer	Tr2
1	fuse 100mA slow	Si1
1	xenon flash tube	H1
1	push-button switch	Ta1
2	PCB-mount fuse holder	
1	reflector	
12	solder pin	
4	PCB spacer 30 mm	
1	M3×15 mm screw	
2	M3×20 mm screw	
4	M3×55 mm screw	
11	M3 nut	
2	solder eye 3.2 mm	
30 cm	0.22 mm² flexible wire	
10 cm	0.75 mm² flexible wire	
15 cm	silvered wire	
2	printed-circuit board	
1	ABS enclosure with moulded mains plug	

MIDI MASTER KEYBOARD

PART 1: CIRCUIT DESCRIPTION AND CONSTRUCTION

D. Doepfer

The multi-purpose MIDI controller described here is based on a microcontroller and a special MIDI keyboard processor, the E510. The circuit is ideal for upgrading many types of existing keyboard, synthesizer, or even grand piano with a MIDI interface.



The microprocessor section in the block diagram, Fig. 1, is based on an 8031 from Intel. This 8-bit processor provides an extensive instruction set, powerful I/O capabilities, an internal 1024-bit RAM and an 8-bit arithmetic logic unit (ALU). Fortunately, the 8031 is inexpensive and widely available. Its main characteristics are:

- four bidirectional 8-bit ports, P0–P3, two of which remain available when P0 and P2 are used for the data and address bus respectively.
- a serial full-duplex I/O channel with programmable bit rate and data format
- two 16-bit timers
- two external interrupt inputs
- five interrupt sources (two external; three internal)
- two interrupt levels
- a 128-byte internal RAM
- up to 64 KByte program memory (with external EPROM)
- up to 64 KByte scratch/data memory (with external RAM)

In the present application, port P0 and bits P2.0–P2.4 of port P2 are used to interface to the program memory, an EPROM (see Fig. 2). Port P1 and bits P3.4–P3.7 of port P3 form a kind of data and address bus for various functions on the controls board. The other four bits of port P3 are used for special applications:

- **P3.0:** serial input for MIDI data from E510
- **P3.1:** MIDI output
- **P3.2:** VCO interrupt for modulation wheel
- **P3.3:** input for pedal status

Terminal **P3.0** (serial input = RxD) of the 8031 is connected to the serial (MIDI) output of the E510 (IC5). The MIDI output signal provided by the microcontroller is taken from port line **P3.1** and fed to two 5-way DIN output sockets, K2 and K3, via buffers IC11a, IC11b and IC11c.

Terminal **P3.2** ($\overline{\text{INT0}}$) is connected to the output of VCO (voltage controlled oscillator) IC4b. The VCO control voltage at input pin FC2 is provided by the WHEEL potentiometer connected to ST3. The VCO frequency is determined by the control voltage and the value of C4. The 8031 measures the relative position of the wheel by counting the bit rate of the signal received at P3.2. The wheel is assigned a MIDI function, e.g., pitch-bend, modulation or volume, with the aid of software.

Terminal **P3.3** ($\overline{\text{INT1}}$) reads the position of the break-type switch in the SUSTAIN pedal connected to K4.

The program memory formed by EPROM IC3 is addressed by the processor via data/address demultiplexer IC2 and port lines P2.0–P2.4. The address latch enable (ALE) signal of the 8031 allows port P0 to

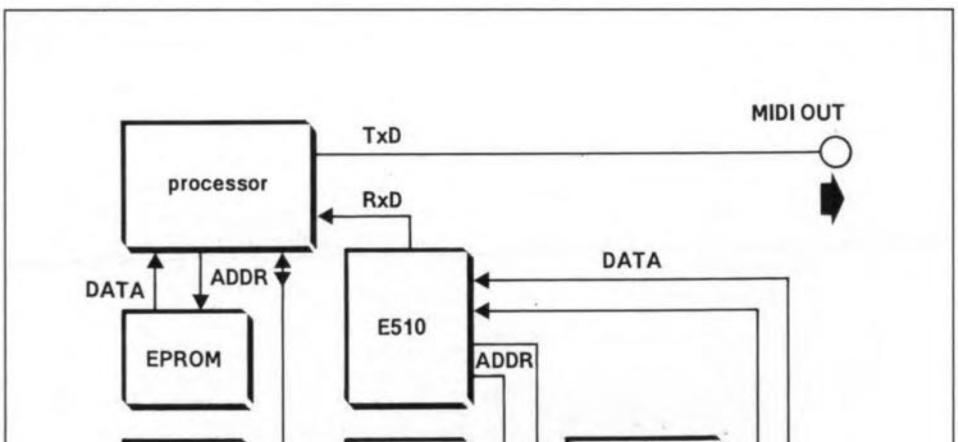


Fig. 1. Block diagram of the MIDI master keyboard.

supply the least-significant address byte (A0–A7) and to act as an 8-bit wide databus (D0–D7). The double function is achieved by time-multiplexing: the LS address byte is latched when ALE is actuated.

Once the LS address byte is latched in IC2, port P0 functions as a databus that allows the processor to fetch instructions from the EPROM. A read operation on part of the 8031 is marked by $\overline{PSEN}$ going low, which actuates the EPROM via its $\overline{OE}$ (output enable) terminal. The EPROM that contains the control program may be a 27128 (16 KByte) or a 2764 (8 KByte).

Quartz crystal Q1 and the associated capacitors, C1 and C2, sets the processor clock frequency at 12 MHz. The standard MIDI bit rate of 31,250 per second is derived from this clock by means of software.

The CPU is reset at power-on by a brief pulse supplied by network R1-C3.

Keyboard interface

The keyboard scanning circuit is based on a dedicated MIDI controller, the E510 (Ref. 1, 2). This chip has been developed for scanning keyboards capable of supplying velocity information. The output of the E510 supplies serial MIDI data with NOTE ON/NOTE OFF and velocity information. It does not, however, supply additional control data such as PROGRAM NUMBER, PITCH BEND or MODULATION. The E510 transmits data on CHANNEL 1 only. Auxiliary functions such as redirection to other channels, transposition and split-assignment are carried out by the 8031.

The E510 has seven address outputs (A0–A6) which select one of 128 keys. When a key is pressed, an internal contact spring is pressed against a gold- or silver-plated contact rail. This establishes a connection between the contact rail and the key contact. Since it has a rest contact and a work contact —which are closed when the key is not pressed or pressed respectively — any one addressed key can assume three states:

- it is not pressed: the rest contact is closed, and the work contact is open;
- it is being pressed: both the rest and the work contact are open;
- it is fully pressed: the rest contact is open, and the work contact is closed.

The time that lapses between the opening of the rest contact and the closing of the work contact is measured by the E510 and provided as the VELOCITY parameter. When the rest contact is opened, an internal counter counts down from 127 and is stopped when the work contact is closed. The counter value thus represents the speed at which the key was pressed.

The key scanning and addressing operations performed by the E510 are fairly unusual since the keyboard is taken up in an 8×8, 8×9; 8×10 or 8×11 matrix, depending on the number of keys. The three least-significant bits, A0–A2, are connected to address multiplexer IC8. When addressed, one of the eight outputs of this device provides a high level on the associated rest contact rail, G1–G8, in the keyboard matrix. The four MS address bits, A3–A6, are applied to demultiplexers IC9 and IC10. Outputs Y0–Y7 of these devices are connected to the associated work contact rail (A1–A11) and the rest contact rail (R1–R11) via pull-up resistors R8–R29.

All 11 rest contact rails and 11 work contact rails are connected to the inputs of NAND gates IC6 and IC7 respectively. The outputs of these gates are connected to the two rail inputs, BS (work contact rail) and BE (rest contact rail), of the E510.

The E510 requires only an external 4-MHz quartz crystal and two small capacitors to obtain a clock from the internal oscillator. Resistor R2 is required to terminate the open-collector MIDI output of the controller. The MIDI data produced by the E510 is fed direct to port line P3.0 of the 8031.

The circuit is powered by an external mains adapter capable of supplying 7 to 12 V at about 300 mA. The on-board 5-V regulator based on a 7805 is conventional. A number of ceramic and tantalum capacitors are fitted at various points on the supply lines to ensure a well-decoupled supply voltage.

Controls board

The circuit diagram of the control/indicator circuit is shown in Fig. 3. The control of this circuit is assumed by the 8-bit databus formed by CPU port P1 and the 4-bit address bus formed by P3.4–P3.7. The databus is connected to all datalines of I/O drivers IC1–IC5. Of these, IC2–IC5 are write-only registers, and IC1 a read-only register. Address lines P3.4–P3.7 are connected to demultiplexer IC6, which arranges the register selection. The selection is accomplished with the aid of the Y0–Y4 outputs, which are connected to the OE (output enable) inputs of the read-only register, and the CLK (clock) inputs of the write-only registers. This creates the following functions of the demultiplexer outputs:

address	Y	IC	Function	Register
0	Y0	IC1	8 keys	read
1	Y1	IC2	8 LEDs	write
2	Y2	IC3	1st display	write
3	Y3	IC4	2nd display	write
4	Y4	IC5	3rd display	write

The status of the eight keys is requested via a non-inverting three-state driver Type 74HC541, whose inputs A1–A8 are fitted with pull-up resistors so that an actuated key produces a logic 0. When the driver is selected via its OE input, the dataword at inputs A1–A8 is transferred to databus P1 from which it can be read by the processor via port 1. When the driver is not selected, its outputs Y1–Y8 are switched to high impedance.

Thus, to be able to read the status of the eight control keys, the processor must first select the '8 keys' register by placing the appropriate address nibble on P3.4–P3.7. The data on the P1 bus is latched into the write-only registers (IC2–IC5) at the leading edge of the relevant clock signal (Y1–Y4 of IC6).

The registers that drive the eight LEDs and the 7-segment LED displays via current-limiting resistors are octal D-bistables Type 74HC574. The display actuation logic is reversed: writing a 0 turns on a LED or display segment.

Many parts for this project, including the programmed EPROM (IC3), the E510 MIDI controller (IC5), the switches, the printed-circuit boards and 61-, 76- and 88-key keyboards are available from

Doepfer Musikelektronik
Lochamerstrasse 63
D-8032 Gräfelfing
West-Germany
Telephone: +49 89 855578
Facsimile: +49 89 8541698

The E510 MIDI controller is also available from **C-I Electronics, P.O. Box 22089, 6360 AB Nuth, Holland**.

Construction of the boards

The two circuits discussed are accommodated on two double-sided, through-plated printed-circuit boards (see Figs. 4 and 5).

All ICs with the exception of the 74LS629 are CMOS types which should be handled with the usual care to prevent damage from static electricity. Since tantalum capacitors are particularly prone to developing internal short-circuits, each device must be checked with an ohmmeter before it is fitted. Note the orientation of the polarized components, which include the ICs, diodes, electrolytic capacitors, the single resistor array, the LEDs and the switches. All capacitors marked Ce on the boards are fitted to afford adequate decoupling of the supply voltage. Each pair of capacitors consists of a 100-nF ceramic type and a 10-µF tantalum type in parallel. The voltage regulator (IC12) is bolted direct to the board, together with its heat-sink.

The controls board has components at both sides. First, fit the capacitors, IC sockets, resistors, the resistor array and pin headers at the side of the printed overlay. Note that the resistor array, RA1, is a polarized part: its common terminal is marked by a dot which is also found back on the component overlay. Next, fit the keys, the displays and the LEDs at the reverse side of the board. These parts are shown in dashed outlines.

The fitting of the switches requires special attention. First, solder one of the four terminals, and check whether the switch rests flat on the PCB surface. If necessary

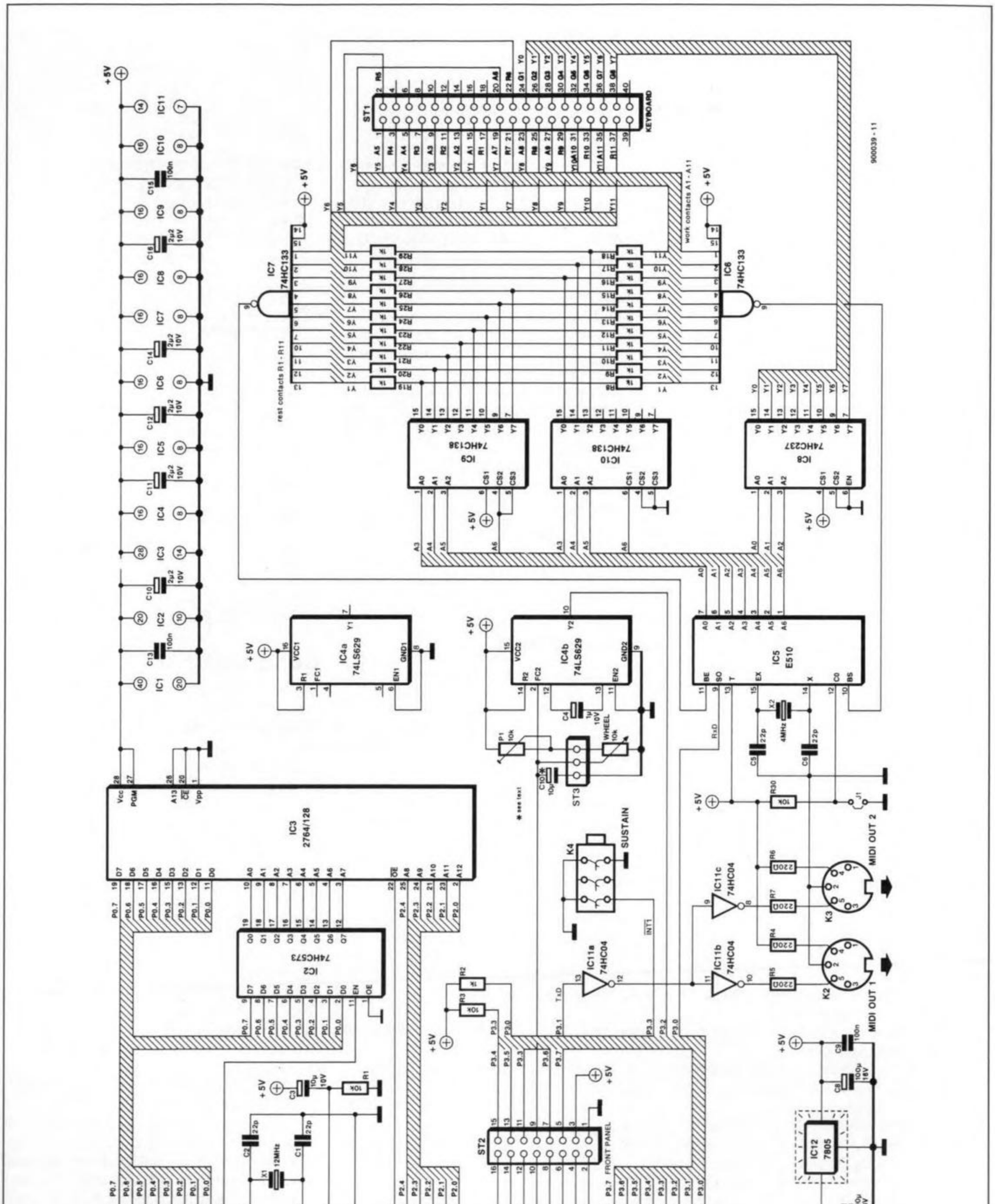


Fig. 2. Circuit diagram of the main controller board. The heart of the circuit is formed by the 8031 microcontroller and the E510 MIDI keyboard controller.

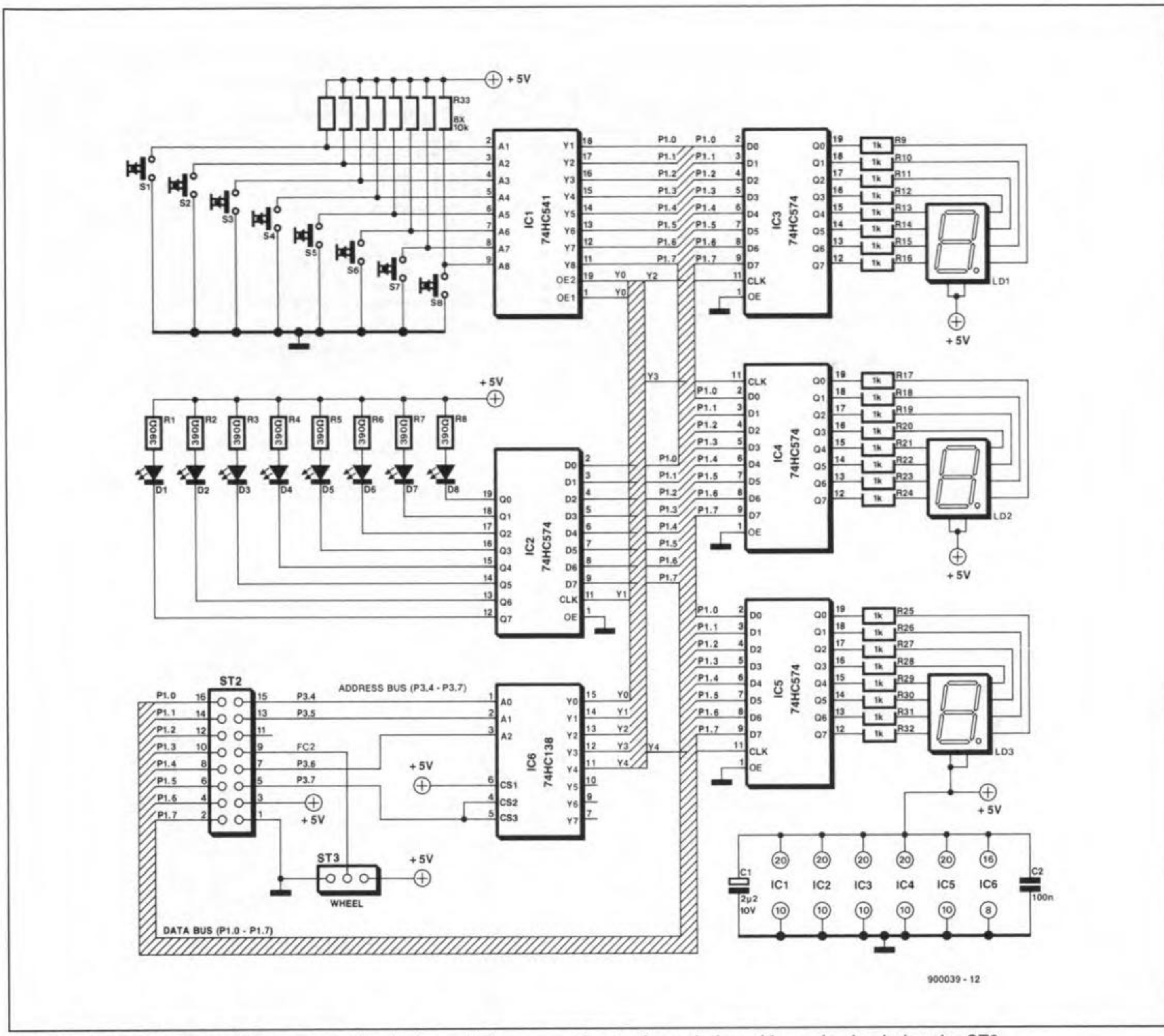


Fig. 3. Circuit diagram of the controls/display board. The connection to the main board is made via pin header ST2.

align the position before soldering the other pins. This simple procedure prevents irregular key positions which spoils the look of the keyboard and causes difficulty with the clearances in the front panel.

Since the faces of the LED displays must be roughly level with the key bases, they are fitted at a certain height above the board surface. The best height is that which results in the faces of the displays being about 1 mm below the top side of the key bases (i.e., not the keytops!).

The rectangular LEDs may be soldered to the board after provisionally fitting the controls panel. This is done to ensure that their height and position align readily with the clearances in the front panel (Fig. 6).

Cables and connections

The main board is connected to the controls board via a 16-way flatcable, and to the 88-key keyboard via a 40-way flatcable. The ends of the two cables are fitted with IDC sockets (note the position of pin 1 of the IDC socket with respect to the cable and pin 1 of the header on the board).

The author can supply an adapter board with mounting instructions to enable the main board to be connected to keyboards with 76, 61 or 49 keys. This adapter is plugged on to ST1 and provides the relevant connections to the keyboard from there.

Modulation wheel

The potentiometer that forms the modulation wheel is connected to header ST3 on the main board. The wheel may be a type which re-adjusts itself by means of a spring (mainly for a pitch-bend function), or one with normal action (for all other control functions, including modulation, volume, panorama and portamento).

Since the full 270° range of the potentiometer will not be covered when the modulation wheel is operated, the required compensation is provided by the control software. The range of the VCO control voltage is defined by preset P1, and the actual value within that range by the modulation wheel.

The wheel potentiometer must not be secured until the complete circuit has been checked for correct operation. Initially, set the associated preset, P1, to minimum resistance.

When a re-adjusting wheel is used (pitch-bend function; this is automatically selected at power-on), the preset is adjusted until the circuit sends a MIDI parameter value of 40_H (64_D) (do not operate the wheel). When a MIDI monitor with parameter selection and a data read-out is not available for this alignment, connect a MIDI receiver (e.g., an expander), and adjust the preset until there is no tone shift with the wheel at its rest position. The software provides a self-adjusting function for the parameter range around 40_H, to afford a kind of compensation for mechanical tolerance on the rest position of the potentiometer.

In case a normal (non-re-adjusting)

wheel is used is used, the VCO preset is adjusted until the modulation potentiometer covers a range of about 00_H to $7F_H$. Small deviations from these values should not cause problems and may be tolerated. Repeat the adjustment with a slightly different value for P1 in case either the lowest (00_H) or the highest ($7F_H$) value is sent long before the end stop of the potentiometer is reached. When the previously mentioned MIDI monitor is not available for checking the relevant data, the modulation wheel is best assigned the function of volume control. This allows you to check whether the volume can be reduced to nought at the corresponding wheel position.

Assembly

The completed and tested controls board is secured to the front panel (Fig. 6) with three M3×15 mm or M3×20 mm screws and plastic PCB spacers. The distance between the panel and the board should be 13 mm. This is achieved by fixing each of the three M3 screws to the front panel with the aid of a 10-mm long spacer and an M3 nut. Next, mount the controls board

COMPONENTS LIST

MAIN BOARD (E510/8031)

Resistors:

3	10k	R1;R3;R30
23	1k0	R2;R8–R29
4	22Ω	R4–R7
1	10k preset V	P1

Capacitors:

4	22p	C1;C2;C5;C6
2	10µF 10V tantalum	C3;C10
1	1µ0 10V tantalum	C4
2	100µF 10V tantalum	C7;C8
6	100n ceramic	C9;C18–C24
7	2µ2 10V tantalum	C11–C17

Semiconductors:

1	1N4001	D1
1	8031 or 8051	IC1
1	74HC573	IC2
1	2764 or 27128	IC3
1	74LS629	IC4
1	E510	IC5
2	74HC133	IC6;IC7
2	74HC138	IC8;IC9
1	74HC237	IC10
1	74HC04	IC11
1	7805	IC12

Miscellaneous:

1	2×20-way pin header	ST1
1	2×8-way pin header	ST2
1	DC supply socket for PCB mounting	BU1
2	5-way DIN socket for PCB mounting	BU2;BU3
1	stereo 6.3-mm jack socket for PCB mounting	BU4
	IC sockets	
1	Heat-sink for IC12	

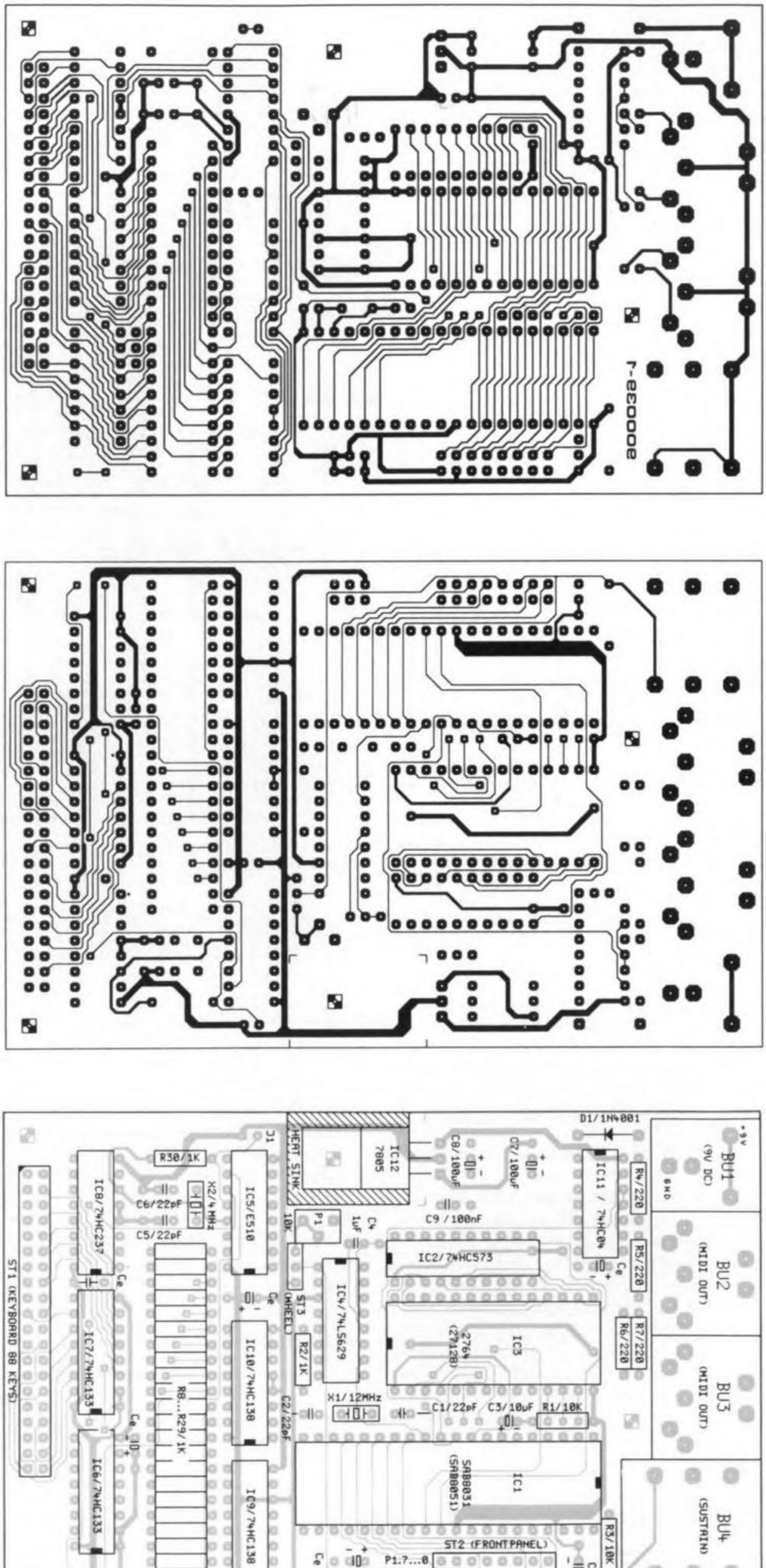


Fig. 5. Track layouts and component mounting plan of main controller board.

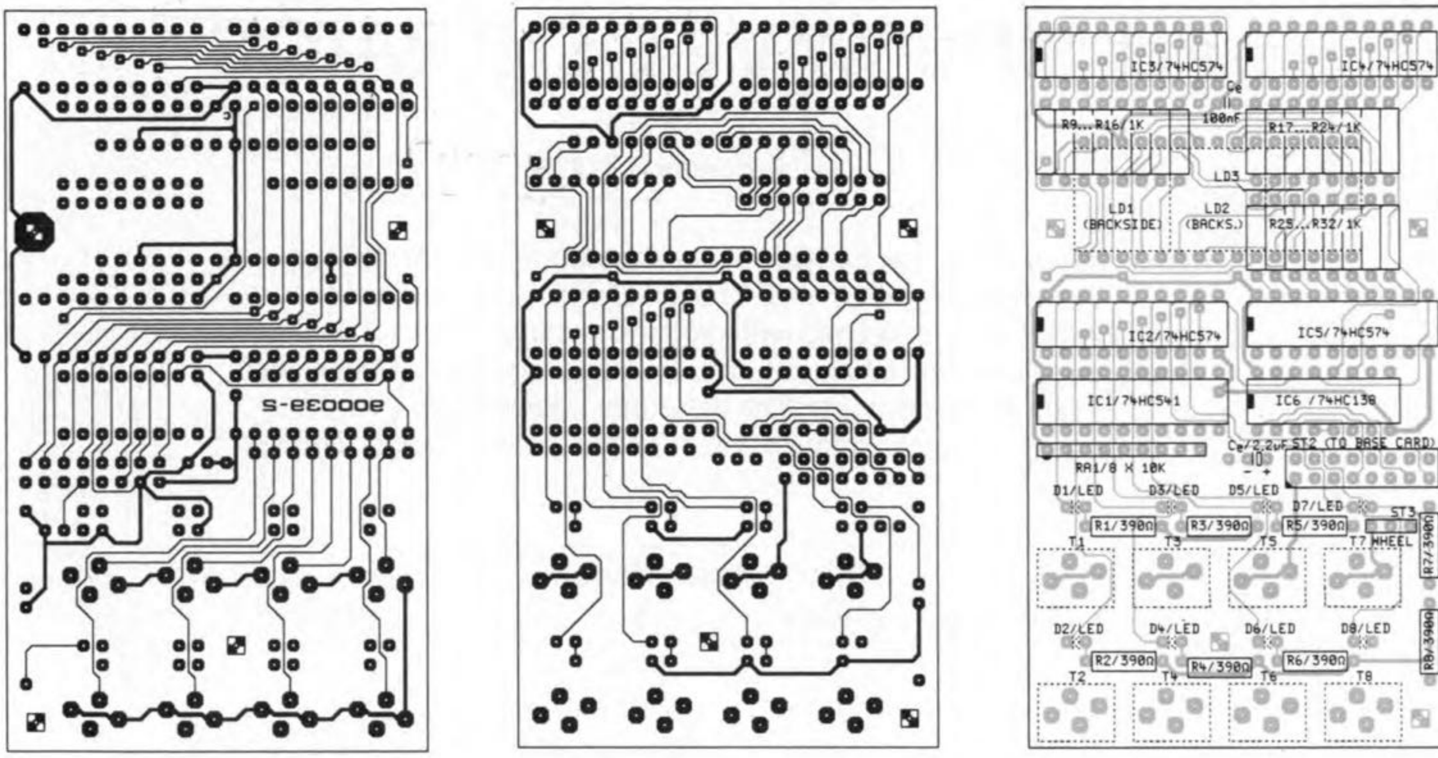


Fig. 5. Track layouts and component mounting plan of the controls/indicator board.

COMPONENTS LIST

CONTROLS/DISPLAY/INDICATOR BOARD

Resistors:

8	390Ω	R1–R8
24	1k0	R9–R32
1	SIL 8×10k	RA1

Capacitors:

1	100n ceramic	C1
1	2µ2 10V	C2

Semiconductors:

8	LED red rectangular	D1–D8
3	LTS546AP	LD1;LD2;LD3
1	74HC541	IC1
4	74HC574	IC2–IC5
1	74HC138	IC6

Miscellaneous:

1	2×8-way pin header	ST2
8	locking SPST push-button	T1–T8

and secure it with the remaining M3 nuts. Cover the displays with a bezel glued to the front panel.

The mounting bracket for the modulation wheel is fitted from below to the front panel with the aid of 5-mm long PCB spacers. The completed and wired front panel is secured to the left side panel of the flight case.

Next time

The concluding part of this article will deal with the function test and troubleshooting procedures. In addition, some attention will be given to the menus offered by the keyboard, as well as to the velocity curves and the default presets. ❑

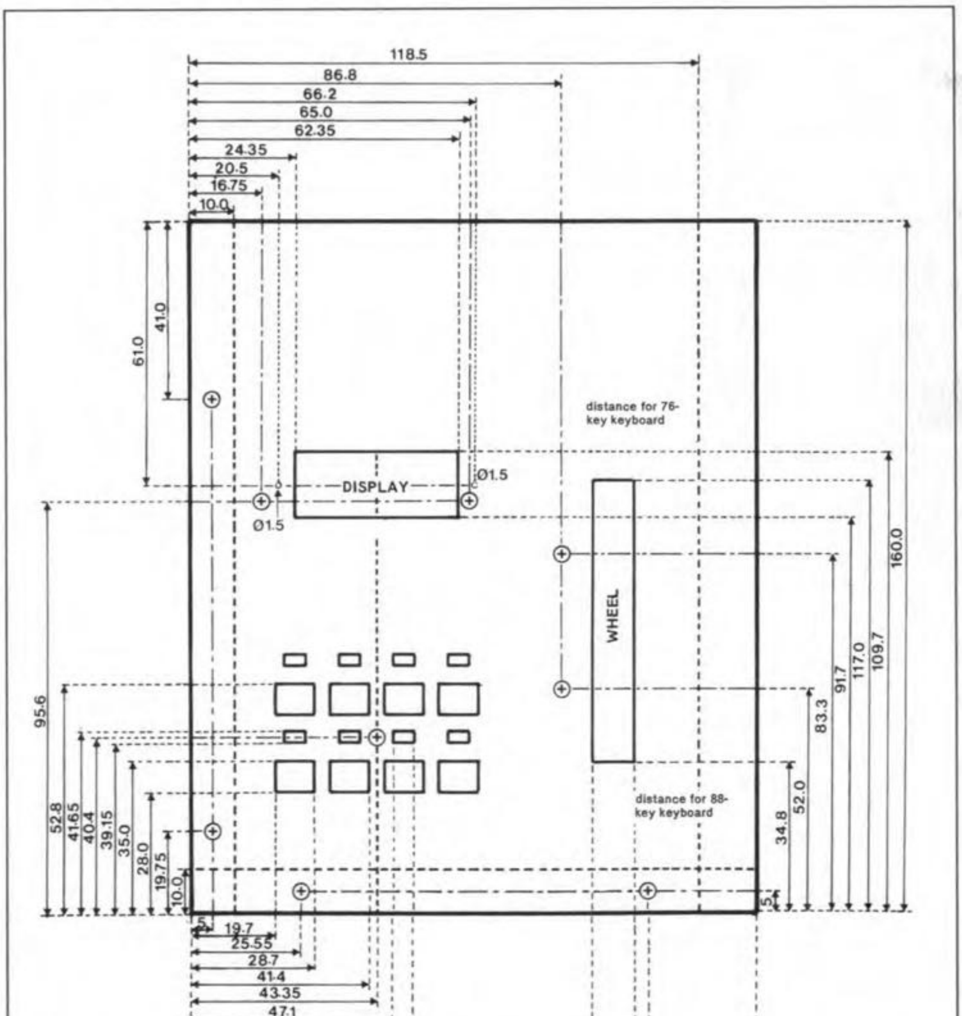


Fig. 6. Cutting and drilling details of the front panel.

wheel is used is used, the VCO preset is adjusted until the modulation potentiometer covers a range of about 00_H to $7F_H$. Small deviations from these values should not cause problems and may be tolerated. Repeat the adjustment with a slightly different value for P1 in case either the lowest (00_H) or the highest ($7F_H$) value is sent long before the end stop of the potentiometer is reached. When the previously mentioned MIDI monitor is not available for checking the relevant data, the modulation wheel is best assigned the function of volume control. This allows you to check whether the volume can be reduced to nought at the corresponding wheel position.

Assembly

The completed and tested controls board is secured to the front panel (Fig. 6) with three M3×15 mm or M3×20 mm screws and plastic PCB spacers. The distance between the panel and the board should be 13 mm. This is achieved by fixing each of the three M3 screws to the front panel with the aid of a 10-mm long spacer and an M3 nut. Next, mount the controls board

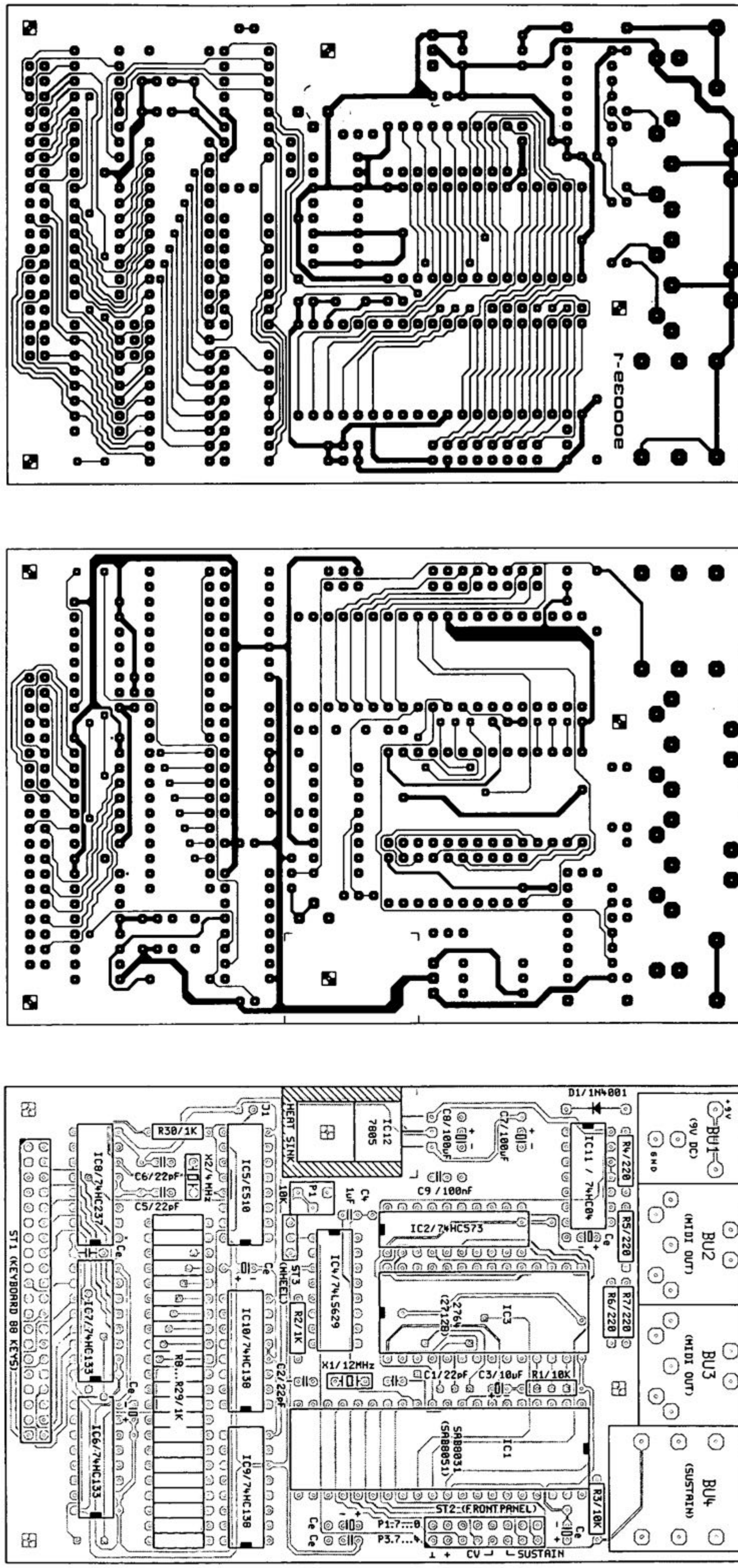


Fig. 5. Track layouts and component mounting plan of main controller board.

COMPONENTS LIST

MAIN BOARD (E510/8031)

Resistors:

3	10k	R1;R3;R30
23	1k0	R2;R8–R29
4	22Ω	R4–R7
1	10k preset V	P1

Capacitors:

4	22p	C1;C2;C5;C6
2	10µF 10V tantalum	C3;C10
1	1µ0 10V tantalum	C4
2	100µF 10V tantalum	C7;C8
6	100n ceramic	C9;C18–C24
7	2µ2 10V tantalum	C11–C17

Semiconductors:

1	1N4001	D1
1	8031 or 8051	IC1
1	74HC573	IC2
1	2764 or 27128	IC3
1	74LS629	IC4
1	E510	IC5
2	74HC133	IC6;IC7
2	74HC138	IC8;IC9
1	74HC237	IC10
1	74HC04	IC11
1	7805	IC12

Miscellaneous:

1	2×20-way pin header	ST1
1	2×8-way pin header	ST2
1	DC supply socket for PCB mounting	BU1
2	5-way DIN socket for PCB mounting	BU2;BU3
1	stereo 6.3-mm jack socket for PCB mounting	BU4
	IC sockets	
1	Heat-sink for IC12	

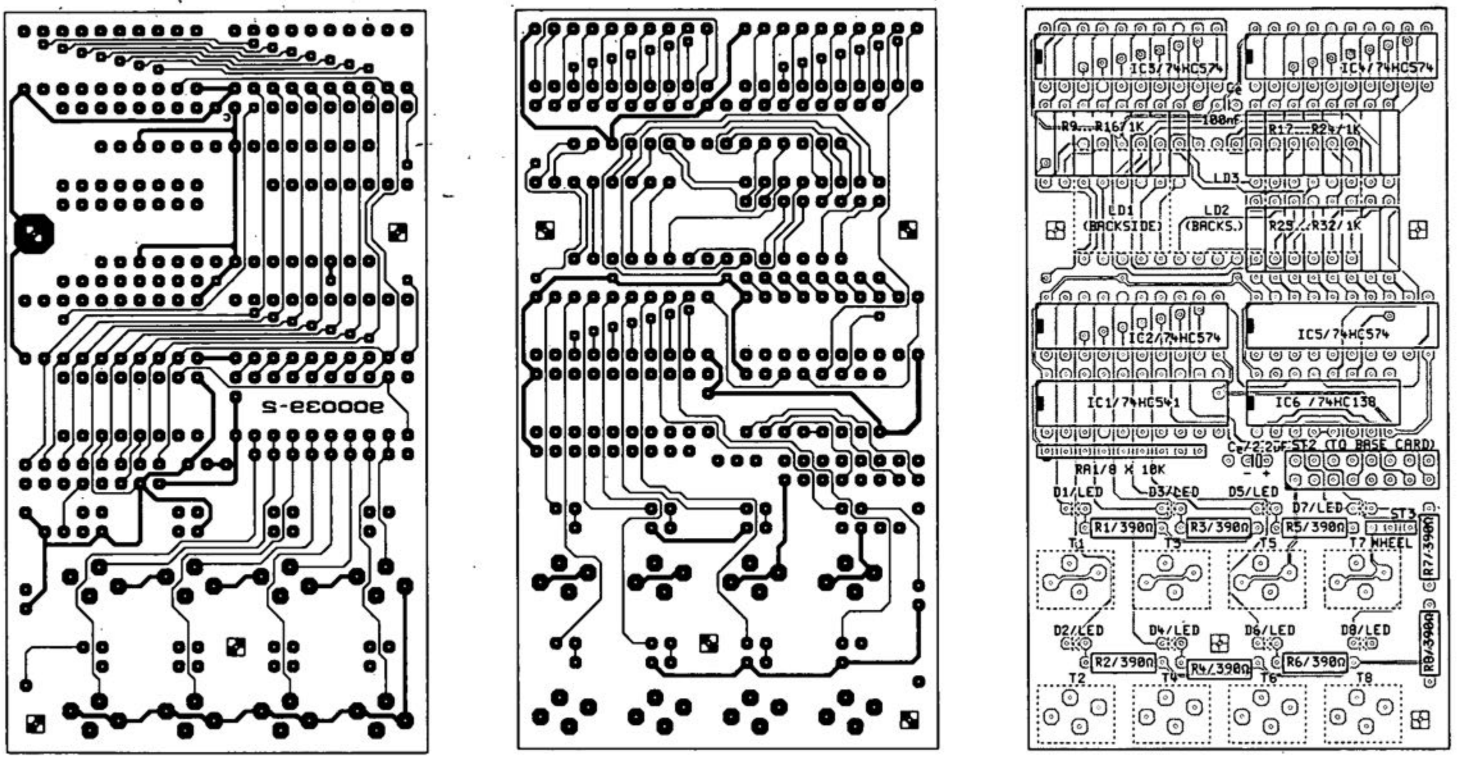


Fig. 5. Track layouts and component mounting plan of the controls/indicator board.

COMPONENTS LIST

CONTROLS/DISPLAY/INDICATOR BOARD

Resistors:		
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24	1k0	R9–R32
1	SIL 8×10k	RA1
Capacitors:		
1	100n ceramic	C1
1	2µ2 10V	C2
Semiconductors:		
8	LED red rectangular	D1–D8
3	LTS546AP	LD1;LD2;LD3
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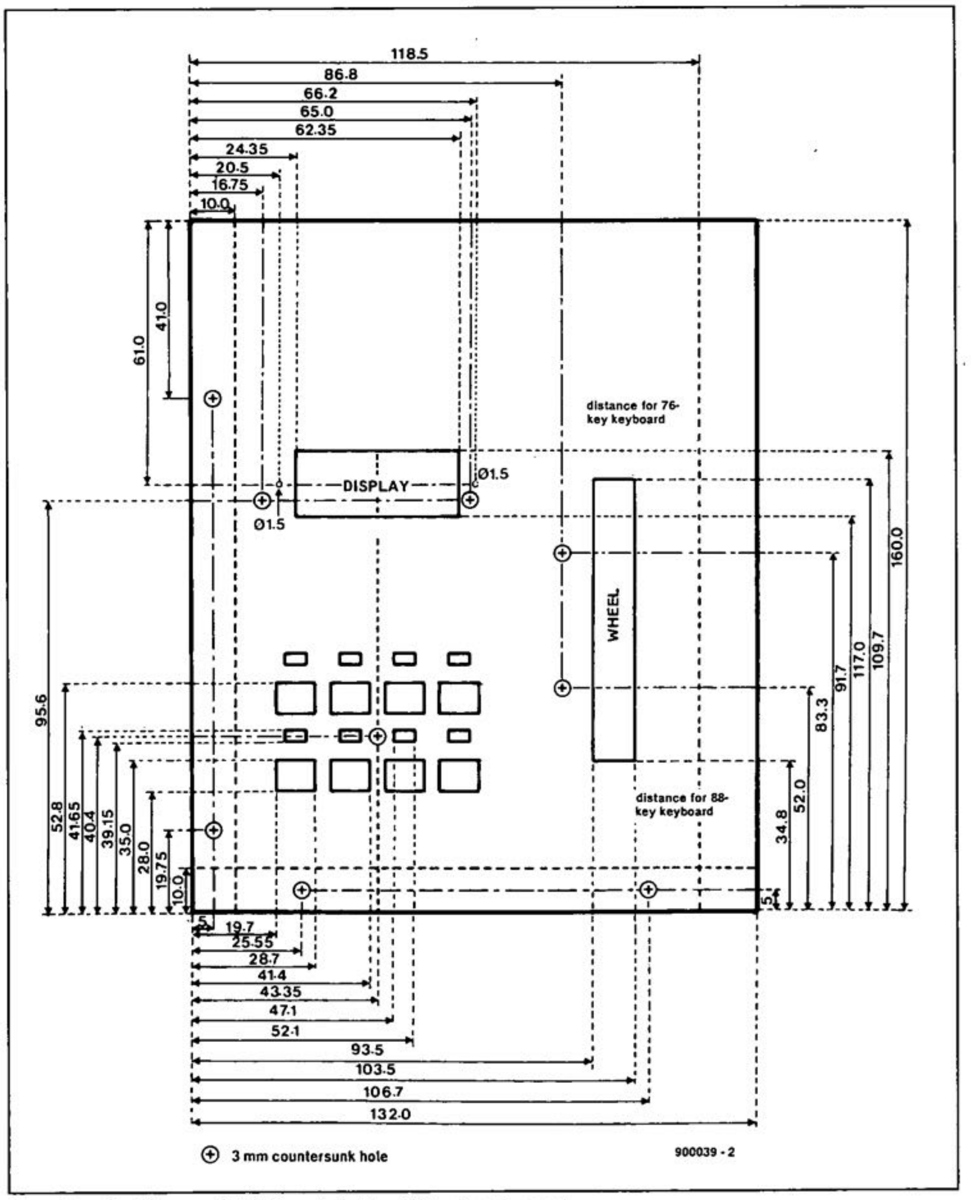


Fig. 6. Cutting and drilling details of the front panel.

PLL SINE WAVE GENERATOR

J. Bareford 890097

Precise frequencies that can be set very accurately are much more easily generated with the aid of digital technology than with analogue techniques. A snag with digital signals is, however, that their shape is rectangular rather than sinusoidal as required for many tests and measurements. The generator described in this article uses digital techniques, yet provides true sinusoidal signals

It is fairly simple to generate frequencies with the aid of, for example, an MS-DOS computer working with GW-BASIC. Without spending any extra money, every PC owner thus has a variable tone generator available. Unfortunately, the signal is rectangular and, therefore, not suitable for a number of tests and measurements.

If the generator is to be used without a computer system, another source of digital reference signal is required. For a number of applications, a simple variable square-wave generator will be perfectly acceptable. Where greater accuracy is required, a stable crystal oscillator with preset scaler must be used.

Technical specification

Input impedance	>10 MΩ
Sensitivity	1–12 $V_{p\text{-}p}$
Frequency range	500 Hz – 100 kHz*
	10 Hz – 1 kHz**
Output impedance	600 Ω
Harmonic distortion	0.5% (typical)
* S1 open	** S1 closed

Control of the VCO

The generator is based on a combination of a phase-locked loop (PLL), here a Type 4046, and an integrated function generator, Type XR2206.

The principle of a PLL is shown in Fig. 1 The two most important parts of a PLL are a phase (Φ) comparator and a voltage-controlled oscillator (VCO).

The VCO will oscillate when a signal at a given frequency is fed to it. This frequency is determined by an external *RC* network. The output of the VCO (2) is fed to one of the inputs of the phase comparator; the other input of this stage is provided with the reference frequency (1).

The output of the phase comparator (3) is the difference between the two input signals. It is invariably a rectangular signal of which the mark-space ratio depends on the phase difference between the two input signals.

A low-pass filter at the output of the comparator integrates the pulses, which results in a signal whose absolute level is directly proportional to the phase difference of the input signals. The output becomes constant at the instant the loop is locked; this normally occurs at a phase difference of 90 °.

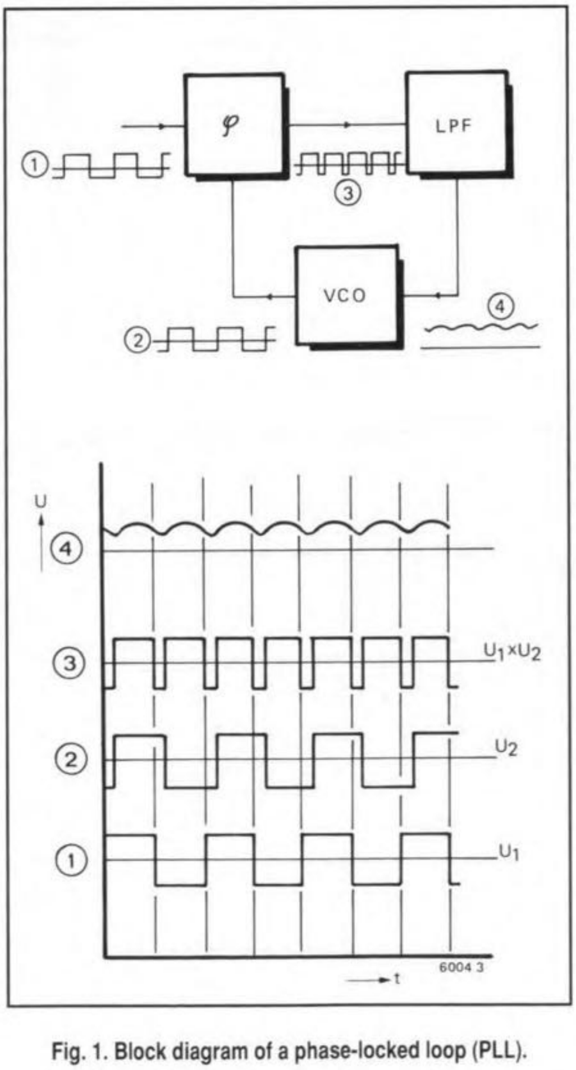


Fig. 1. Block diagram of a phase-locked loop (PLL).

As shown in Fig. 2, the 4046 used in the generator contains two different phase comparators. The first one of these is a single XOR gate, whose output is 1 if the levels of the two input signals are not equal. This comparator is not suitable for the present circuit because it requires symmetrical input signals and it will lock to harmonics of the input signals. Signals associated with this comparator are shown in Figure 3.

The second comparator is rather more complex and perfectly suitable for the present purposes. It is not sensitive to asymmetry of the input signals, since it operates on the edges of these signals. Moreover, it does not lock to harmonics of the input signals. This means that the sinusoidal signal is of exactly the same frequency as the reference signal. Associated signals are shown in Fig. 4.

An additional advantage of this stage is that it allows connexion to an LED that lights when the loop is locked to indicate that the output signal is as stable as the reference signal.

The output of the second comparator has not two but three states, depending on the input signals: 0, 1 or high impedance. The dependence of the output signal on the input signal is clear from Fig. 4. When the output of the comparator is high impedance it has no effect on the circuit following the comparator; it is, as it were, not present.

The period of time that the output is 1 or 0 depends directly on the phase difference between the input signal and the reference signal. If the input signal is first to have a leading edge, that is, is in advance of the VCO signal, the output becomes 1. If the VCO signal is first, the output becomes 0.

The low-pass filter following the comparator converts the output pulses into a direct voltage. This voltage increases when the comparator output goes high; it remains constant when the comparator output is high impedance; and it decreases when the comparator output goes low. This means that the direct voltage is directly proportional to the phase difference between the input signal and the VCO signal.

Since the direct voltage is used to control the VCO, the circuit stabilizes at a given set of conditions. In the present comparator that happens when the phase difference between the input signal and the VCO signal is 0° and not 90° as is usual in PLLs. Because of this, the VCO frequency is exactly the same, and as stable, as the input signal.

From rectangular to sinusoidal waveform

Even a cursory glance at the circuit diagram in Fig. 8 shows that instead of the VCO in the 4046 a separate VCO, formed by IC5, is used. This circuit, strictly speaking a function generator, has the important advantage that it pro-

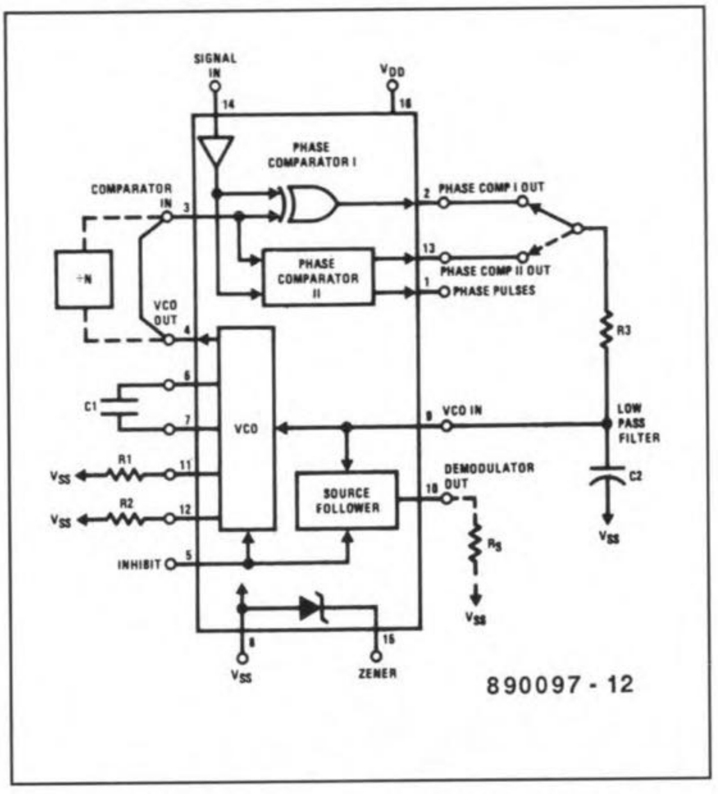


Fig. 2. Block diagram of the Type 4046 phase-locked loop (PLL).

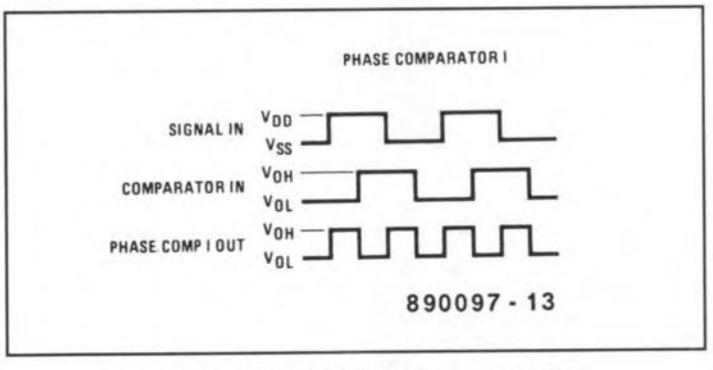


Fig. 3. Signals associated with phase comparator I.

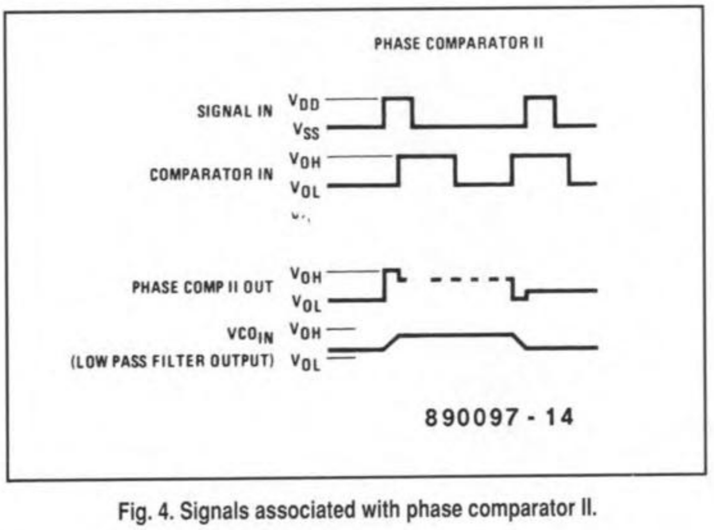


Fig. 4. Signals associated with phase comparator II.

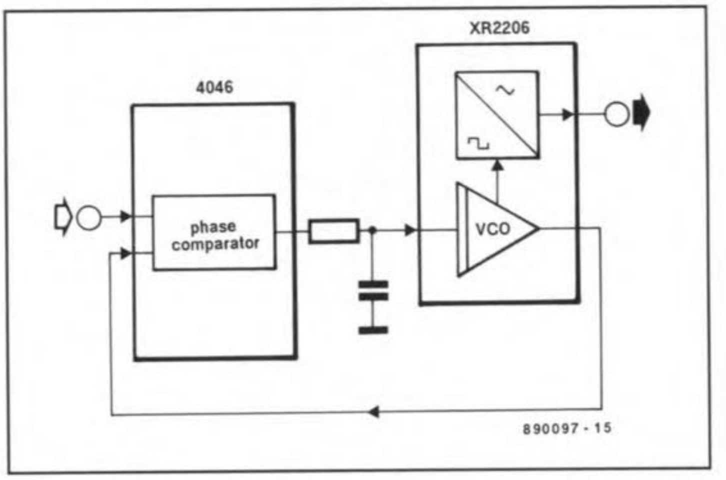


Fig. 5. Phase comparator II in the 4046 and the VCO in the XR2206 form a good-quality PLL.

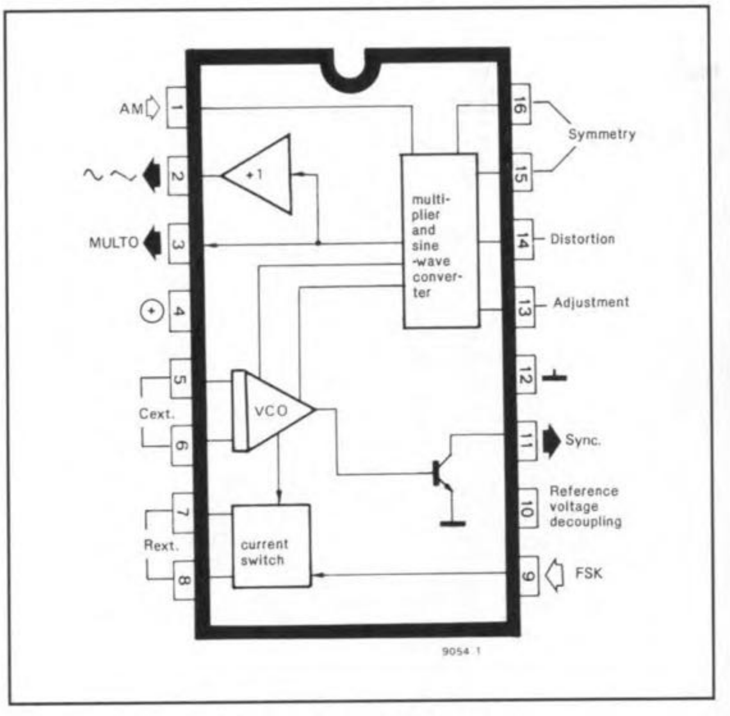


Fig. 6. Circuit diagram of the XR2206.

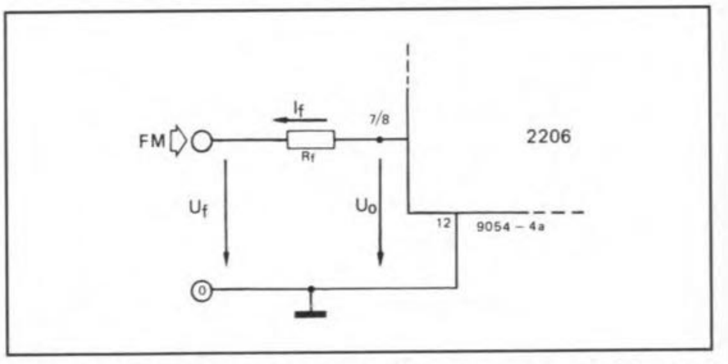


Fig. 7. The VCO input of the XR2206 is, strictly, a current-controlled output across which a fixed potential of 3 V exists. Applying a counter potential U_f via resistance R_f sets the input current.

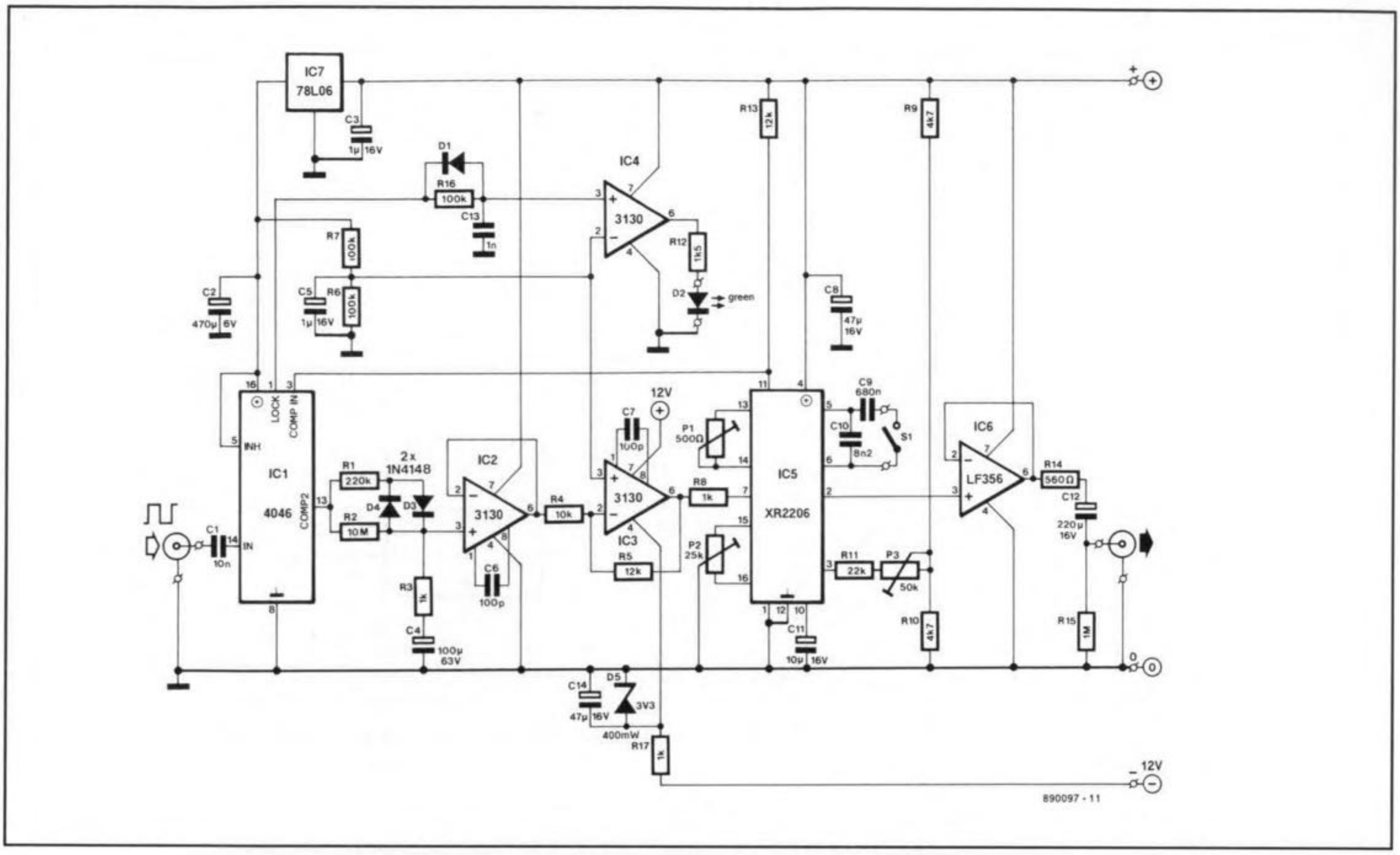


Fig. 8 Circuit diagram of the sine wave generator.

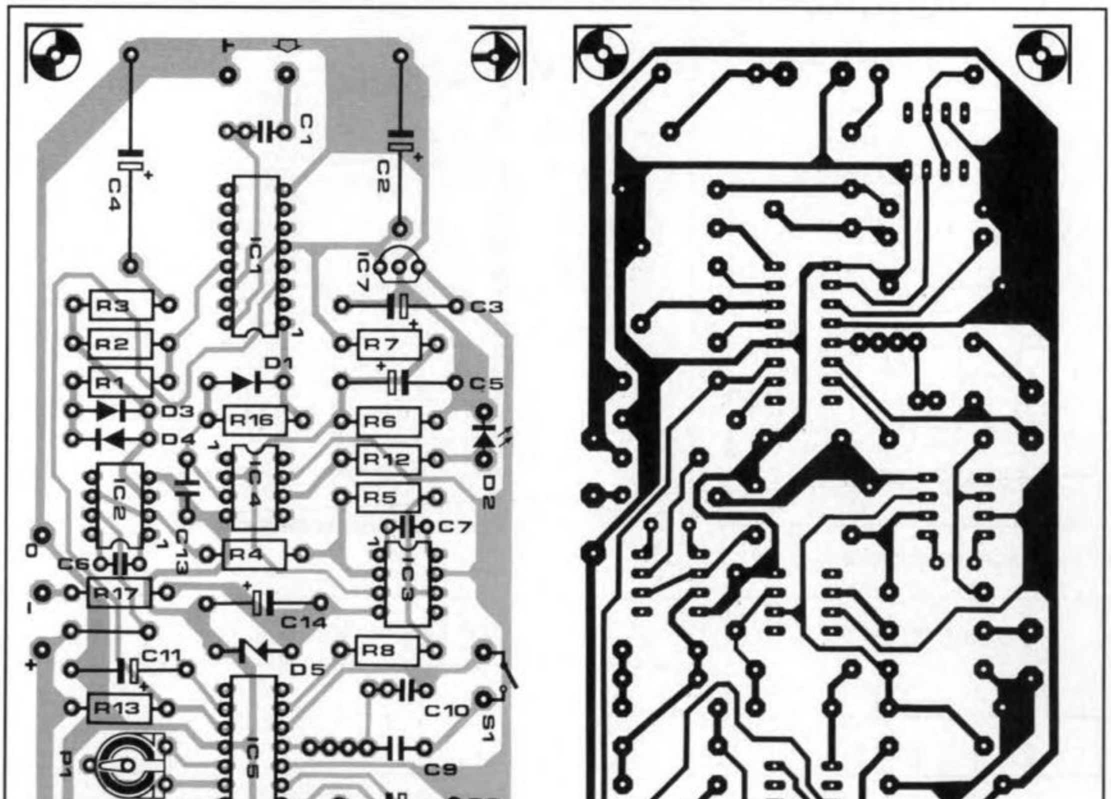


Fig. 9. Printed circuit board for the sine wave generator.

PARTS LIST

Resistors:
R1 = 220 k
R2 = 10 M
R3, R8, R17 = 1 k
R4 = 10 k
R5, R13 = 12 k
R6, R7, R16 = 100 k
R9, R10 = 4k7
R11 = 22 k
R12 = 1k5
R14 = 560 R
R15 = 1 M

Capacitors:
C1 = 10 n
C2 = 470 μ, 16 V
C3, C5 = 1 μ, 16 V
C4 = 100 μ, 16 V
C6, C7 = 100 p
C8, C14 = 47 μ, 16 V
C9 = 680 n
C10 = 8n2
C11 = 10 μ, 16 V
C12 = 200 μ, 16 V
C13 = 1 n

Semiconductors:
IC1 = 4046
IC2 – IC4 = CA3130
IC5 = XR2206
IC6 = LF356
IC7 = 78L05
D1, D3, D4 = 1N4148
D2 = LED (green)

Miscellaneous:
S1 = SPST switch
PCB 890097-1

vides both a rectangular and a sinusoidal signal, since it has a sine wave converter on board. Moreover, the IC remains perfectly reliable at low frequencies, so that it is truly linear over a wide range of frequencies.

The XR2206 is used in exactly the same way as that in the 4046 would have been. Its internal circuit is shown in Fig. 6.

Reverting to Fig. 8, opamp IC3 between the phase comparator and the VCO performs three distinct functions. In the first place, it inverts the control voltage for the VCO, because a decreasing voltage at pin 7 of IC5 would result in an increase in the output frequency. Secondly, its output signal has been arranged to allow it driving the VCO over its full range. Finally, the input of the VCO is controlled by a current rather than by a voltage. In other words, the control input of the VCO is in reality an output across which a fixed potential (U_o) exists and from which a current flows. The output voltage of IC3 determines the potential difference across R8 and thus the value of the current that will flow. The frequency, f, is determined from:

$$f = (U_o - U_f) / (3 \times C \times R8) \quad \text{[Hz]}$$

in which U_o = 3 V; U_f = the output voltage of IC3; C = C10 or, if S1 is closed, C9+C10.

It may be calculated that, depending on the position of switch S1, a frequency range of 6 Hz to 125 kHz is available.

The remainder of the circuit is virtually nothing but the connexions between the various components.

The output of the phase comparator is connected to the VCO via low-pass filter R1-R2-R3-C4 and buffer IC2 in a manner that precludes any feedback to the comparator.

Diodes D3 and D4 shorten the time required by the circuit to lock by enabling the quicker charging and discharging of IC4. This is possible because at higher voltage levels R1 is connected in parallel with T2. This causes the impedance to decrease from 10 MΩ for small signals (PLL locked) to about 220 kΩ for large signals (PLL not yet locked).

Since the PLL may be used over a fairly wide frequency range, the low-pass section has a fairly large time constant to ensure good stability. Because of this, the PLL takes a relatively long time to lock. The 'acceleration' provided by the diodes is therefore welcome.

Diode D2, which is controlled by opamp IC4, lights to indicate that the PLL is locked. The opamp is connected to pin 1 of the PLL, which is specially provided for this purpose. A constant high level exists at this pin as long as the PLL is locked. As soon as the PLL tends to drift, small pulses appear at this pin and these are used by diode D1 to arrange for the fast discharge of C13. The voltage across C13 then drops to a level that causes the output of IC4 to become low. This in turn results in D3 being extinguished to indicate that the frequency is not stable.

Finally, a number of components are necessary for the proper functioning of IC5:

- resistor R8 and capacitors C9 and C10 determine the frequency range; S1 makes it possible to choose between the two ranges;
- resistors R9 and R10 set the d.c. operating level at the output (pin 2);
- preset P3 determines the amplitude of the output signal;
- preset P1 serves to shape the output waveform;
- resistor R13 is the collector resistor for the open-collector output of the VCO;
- circuit IC6 forms the output stage proper.

Construction and alignment

The circuit is intended to be built on the PC board shown in Fig. 9 and Fig. 10.

A number of soldering pins must be fitted on the board to facilitate connexions to other equipment.

The ICs may be soldered direct to the board.

With careful work, and particular attention to the polarity of the diodes and electrolytic capacitors, nothing should go amiss in populating the board

Commence the alignment by setting all potentiometers to the centre of their travel.

Connect the output of the generator to an oscilloscope.

Apply a rectangular signal of 1 kHz, at a level of 5 V, to the input of the generator. As stated at the beginning of this article, this signal may emanate from a computer, simple square-wave generator or crystal oscillator.

Open S1 and connect an external ±12 V supply to the supply lines of the circuit.

If all is well, the oscilloscope should now show a reasonably well-shaped sine wave. If necessary, adjust P1 to make the signal as truly sinusoidal as can be judged. Once that is done, adjust P2 to make the signal look even better. Since the two potentiometers affect one another, the alignment must be carried out a few times. After correct alignment, the harmonic distortion is not greater than 0.5%.

The optimum position of P3 depends on the desired output level: be careful not to produce distortion of the output through overdriving.

If you prefer a continuously variable output level, that can be provided by a small modification. This consists of replacing R14 by two soldering pins. Connect a 1 kΩ potentiometer between the pin that is connected to the output of IC6 and earth. Connect the wiper of the potentiometer to the other pin. A continuously variable output signal is then provided via C12.

The suggested front panel shown in Fig. 11 should be given a calibrated scale to give a (relative) indication of the output signal. ■

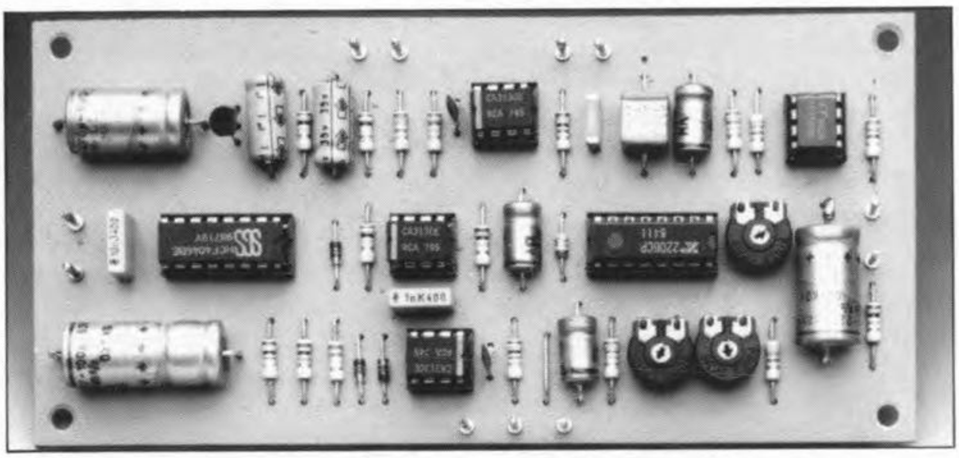


Fig. 10. Photograph of the completed printed-circuit board.

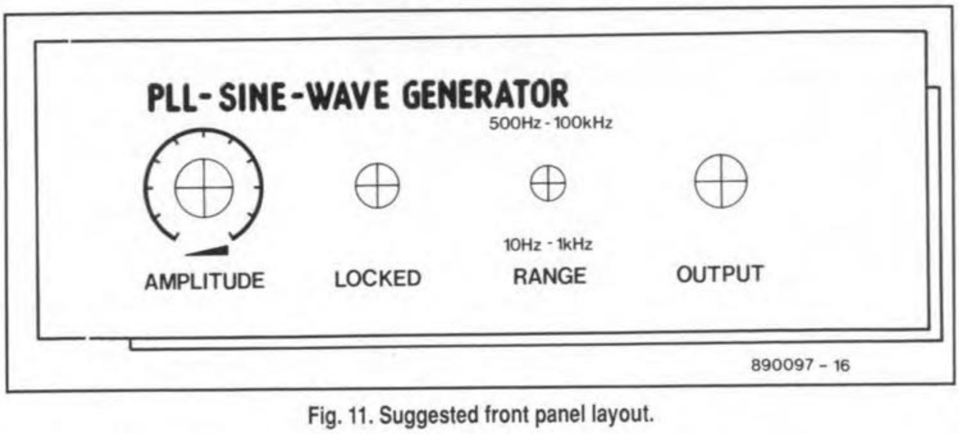
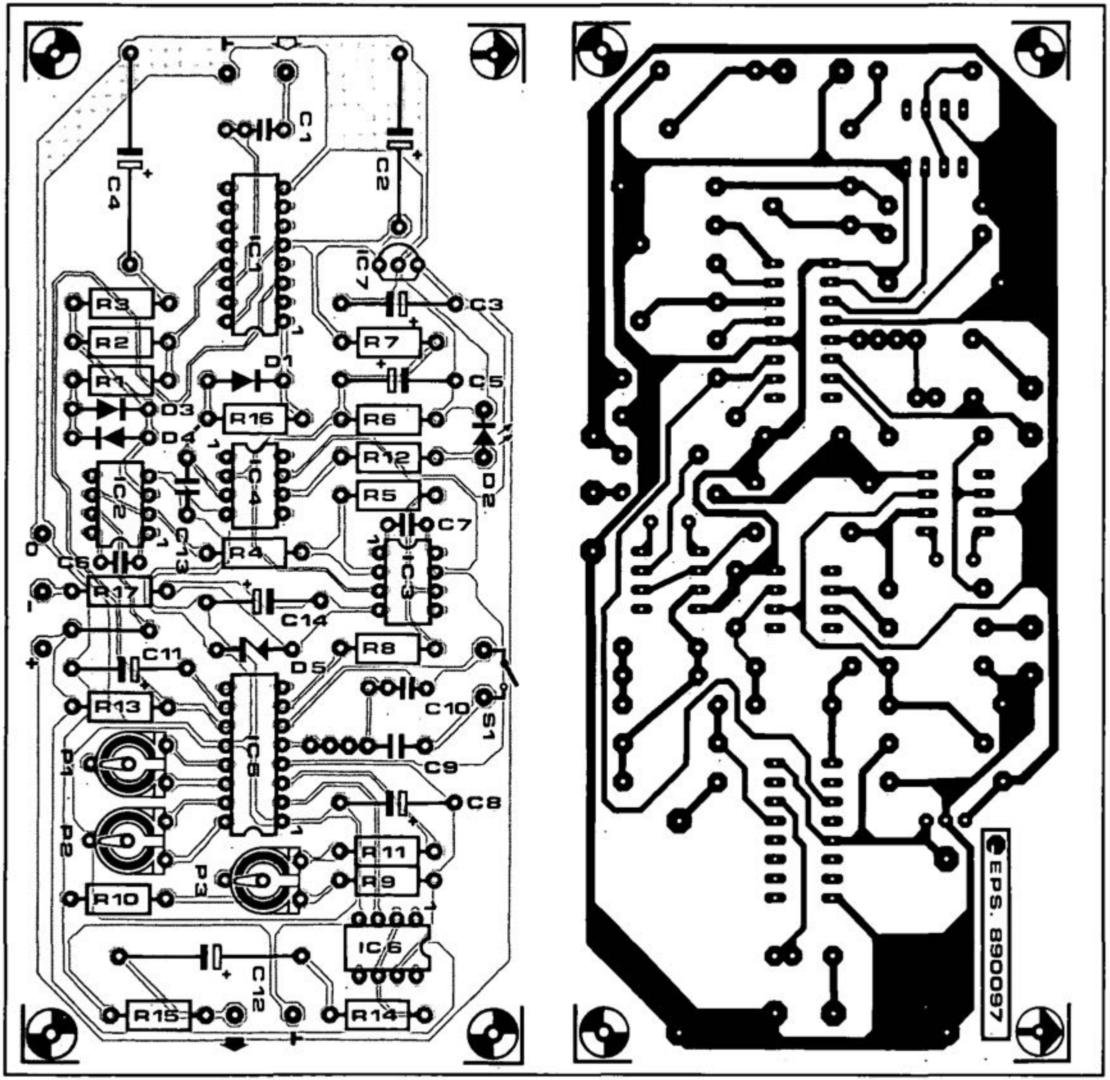


Fig. 11. Suggested front panel layout.



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Miscellaneous:
S1 = SPST switch
PCB 890097-1

ELECTRONIC FUSES

Bourns Electronics has for some time marketed a series of circuit breakers that, in a number of applications, form good alternatives to the usual glass fuses. Known as MultiFuse, the devices are similar in their basic characteristics to positive temperature coefficient (PTC) resistors made from doped barium titanate ceramics, but are of a totally different construction based on conductive polymer composite materials. An important advantage of the MultiFuse is that after it has been tripped by an overload it needs only a short period after the overload has been removed to regain its normal operational characteristics.

Automatic circuit breakers are, of course, not new. Many electric coffee-makers and deep fryers have a thermal trip device that switches off the mains when the appliance gets too hot. After the heating element has had time to cool off, the trip element returns to its original position and the appliance operates normally again.

The MultiFuse has a similar function. As soon as the current through it exceeds a certain value, its resistance increases significantly. This reduces the current to a safe value so that the load does not get damaged.

However, in contrast to the usual electro-mechanical circuit breaker, the MultiFuse is an electronic component that has no moving (mechanical) parts. The advantages of this are clear: a mechanical circuit breaker is sensitive to vibrations, produces sparks when it is operated, and, after a time, presents an increasing resistance owing to corrosion of the contacts.

Ceramic positive temperature coefficient (PTC) resistors that occasionally are used in protection circuits operate in a manner similar to that of the MultiFuse. They have the serious drawback, however, of a much longer switching time. Another snag is that their resistance may decrease significantly, even to the point of a short circuit, when the voltage across them is too high.

Principle of operation

Although the basic characteristics of the MultiFuse are similar to those of a PTC, the construction of the two devices is quite different. Whereas a PTC is made from doped barium titanate ceramics, the MultiFuse is made from conductive polymer. In the past two decades substantial progress has been made in research on conductive polymer materials. In these materials, conductive particles—carbon black in the case of MultiFuse—are dispersed in a polymeric matrix of a suitable plastic material.

The properties of the conductive polymer composites used in MultiFuse lead to improvements in device characteristics that are not obtainable with conventional PTC resistor technologies. In the low resistance state, resistances as low as a few milliohms can be obtained. The PTC effect, which is the basis of the current-limiting function, can increase resistance by typically 5–7 decades. This PTC characteristic is maintained over and beyond the operating range of MultiFuse. Ceramic PTC resistors on the other hand can exhibit a negative temperature characteristic under overvoltage when their temperature rises beyond the anomaly temperature.

Properties of conductive polymers

To better understand some of the characteristics of MultiFuse devices, it is necessary to look at the underlying properties of conductive polymers.

The conductive poymer materials used to produce MultiFuse devices are filled with carbon black. The carbon black particles form themselves into chains instead of randomly dispersed particles. The resulting electrical conductivity of the polymer composition is of the order of 0.01 S ($=1\ \Omega^{-1}\ cm^{-1}$).

The type of carbon black and the volume ratio of carbon black to polymer determine the value of electrical resistivity. Changes in the volume ratio will cause changes in resistivity. A lower ratio of carbon black to polymer means a decreased number of conductive chains and therefore an increase in resistivity.

For MultiFuse devices a crystalline polymer is used. The crystalline structure of this material disappears in the region of 125 °C. The resulting increase in polymer volume reduces the ratio of carbon black to polymer and a very large resistance increase results within a very narrow temperature band. This anomalous positive temperature coefficient of resistance explains the "switching" characteristic of MultiFuse devices.

Figure 1 shows the link between volume and resistance. An increase in resistance of six orders of magnitude over a few tenths of a degree Celsius is typical. The figure also shows that the PTC characteristic is persistent far beyond the anomaly temperature.

With cooling of the device below the anomaly temperature the polymer starts to recrystallize and more and more of the opened carbon black chains are re-established. Although most of the carbon black chains are closed again after minutes, the further crystallization process takes time. This fact leads to the slight increase in resistance after trip: after a cooling period of 1 hour, the value of the resistance is still up to 20% higher than original. However, the resistance value will return within 24 hours to within the tolerances specified by the makers.

MultiFuse devices are manufactured to yield a basic resistance close to its minimum specified value. When they are delivered, they have never been tripped and their volumetric ratio of carbon black to polymer is at a maximum.

Designing in a MultiFuse

The questions you need to answer when designing with MultiFuse devices include:

1. What is the maximum normal current that can pass throught the circuit at the maximum ambient temperature without tripping the device?

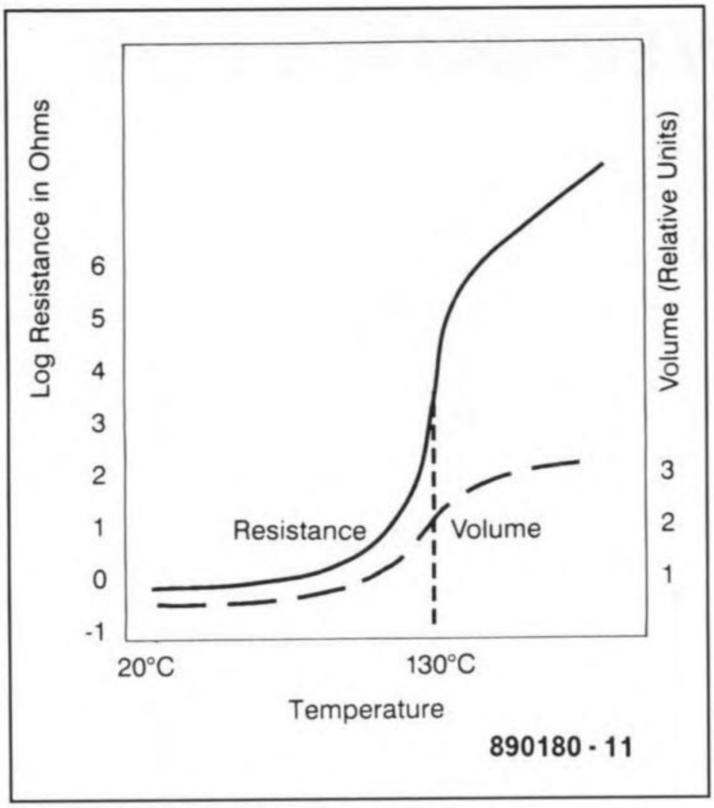

Fig. 1. Resistance and volume as a function of temperature.

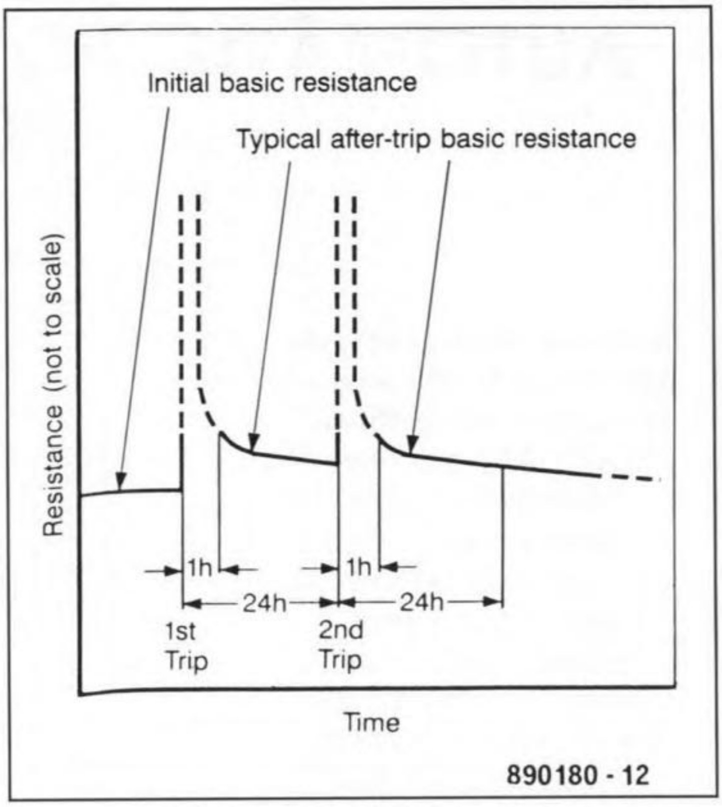

Fig. 2. Recovery of basic resistance after trip.

Part Number	V max V rms	I_{HOLD} A rms	R min Ohms	R nom Ohms	R max Ohms	I trip A rms	P d W	I max A rms
MF-R020		0.20	1.83	2.67	4.50	0.30	0.40	
MF-R025		0.25	1.25	1.83	3.10	0.38	0.45	
MF-R030		0.30	0.87	1.27	2.20	0.45	0.50	
MF-R040	50	0.40	0.55	0.81	1.33	0.60	0.55	40
MF-R050		0.50	0.49	0.75	1.20	0.75	0.75	
MF-R065		0.65	0.30	0.46	0.75	0.98	0.90	
MF-R075		0.75	0.25	0.39	0.62	1.13	0.90	
MF-R090		0.90	0.19	0.34	0.48	1.35	1.00	
MF-R110		1.10	0.08	0.13	0.23	1.87	1.00	
MF-R135		1.35	0.06	0.10	0.17	2.30	1.10	
MF-R160		1.60	0.05	0.08	0.14	2.72	1.20	
MF-R185		1.85	0.04	0.06	0.11	3.15	1.30	
MF-R230	30	2.30	0.03	0.05	0.09	3.91	1.40	40
MF-R250		2.50	0.02	0.04	0.08	4.25	1.50	
MF-R300		3.00	0.02	0.03	0.06	5.10	2.00	
MF-R400		4.00	0.01	0.02	0.04	6.80	2.50	
MF-R600		6.00	0.005	0.01	0.03	10.2	3.50	
MF-R800		8.00	0.005	0.01	0.02	13.6	4.00	
MF-S200	15	2.00	0.03	0.04	0.08	3.00	*	100
MF-S350		3.50	0.02	0.03	0.04	5.25	*	
MF-T110	250	0.11	13	17	26.0	0.17	1.00	3
MF-T145		0.145	7	8.5	13.0	0.22	1.00	

Electrical specifications of a number of types of MultiFuse at an ambient temperature of 20 °C.

2. What is the maximum current that can pass through the circuit at the minimum ambient temperature without causing damage to circuit components?
3. What is the maximum fault current and voltage to which the device will be exposed?

A practical example

Suppose that a resistive load of 33 Ω, for instance, a motor, transformer or heating element, is to be protected against overcurrent with the aid of a MultiFuse.

In normal circumstances, the highest current is 150 mA, the highest ambient temperature is 70 °C and the maximum voltage is 30 V.

On the basis of these data, the most suitable type of MultiFuse is chosen from the table: in this case, the MF-R030. This has a hold current of 160 mA at a temperature of 70 °C and will, therefore, not trip in normal circumstances. Also, it can stand a voltage of 60 V, which is more than adequate for the requirement. At room temperature, the MF-R030 has a minimum resistance of 0.87 Ω, so that if a short-circuit occurs, the maximum current through the device is 34.8 A. This is lower than the maximum allowable current of 40 A. This example makes it clear that the device will not trip spontaneously and will survive the short circuit condition in this application.

Then there is the question whether the device will trip fast enough to protect the load. The relevant datasheet of the makers gives a typical response time at a specified current value at 20 °C: 0.4 s at a current of 2 A, that is an I^2t of 1.6. Therefore, energy equivalent to 52.8 joules ($(I^2t \times R_L)$ may be deposited into the load before the MultiFuse trips. If data on the load indicates that the load can withstand more energy than that amount before it fails, this particular MultiFuse device may be the right one for the application.

To confirm your choice, measure the I^2t failure curve for R_L and compare it with the I^2t trip curve for the MultiFuse device. If the curve for the MultiFuse device is below that for the load, the load will be protected. ■

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90 Park Street
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Surrey GU15 3NY
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AUTOMATIC POWER-DOWN FOR PCs

J. Ruffell

There are many programs that take of lot of precious time, even on relatively fast PC-AT computers. Examples include electronic circuit simulators, sorting routines, disk compression programs, and PCB auto-routers. Eventually, the enjoyment of the PC user sitting back and sipping coffee or tea as the machine crunches its way through masses of data turns into impatience. After a few such sessions, you have seen it all happening, and time-consuming programs are usually banned to night-time hours. Unfortunately, this means that the computer remains on long after finishing the relevant program. The circuit presented here does away with this disadvantage of off-peak, unmanned computing by enabling the computer to shut itself down.

Although the circuit is designed specifically for IBM PCs and compatibles, it may be used with other types of computer as well, provided the user is sufficiently *au fait* with the hardware. In principle, the external part of the automatic power-down circuit could be located in the computer. Apart from the difficulties that may be expected from working on the power supply of the PC, this option is likely to infringe on electrical safety requirements. With that in mind, a separate enclosure that contains a mains-rated relay, a control circuit and a low-voltage power supply will be preferred in many cases.

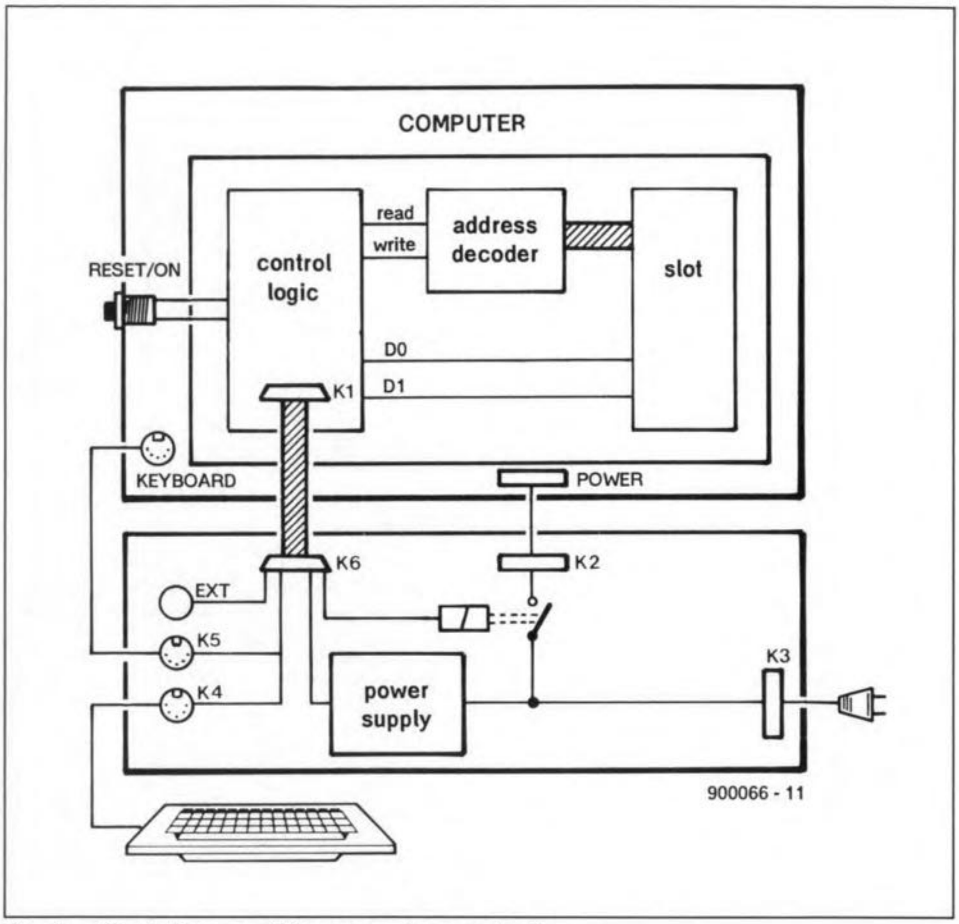


Fig. 1. Block diagram of the power-down controller.

Block diagram

The block diagram of the circuit is shown in Fig. 1. The keyboard is connected to the PC via the auto power-down unit to enable it to function as an on-control. In the external switching box, the keyboard clock signal is 'tapped' and fed to the control circuit fitted in the computer. This arrangement enables the control circuit to switch the computer on when a key is pressed.

The control circuit switches the computer on, and off again when required. A short delay is provided to prevent the computer being switched off by a 'hanging' program. The switch-off procedure starts with a small program making databit D1 of the address at which the control board resides logic high. After a while, this address is read back. If the 1 is still there, the relay is de-energized after a delay of about 10 s (useful for parking the hard disk).

Bit D0 at the control board address determines whether or not the keyboard is to remain on when the computer is switched off. The 'keyboard active' option allows the computer to be switched on by any key action. The other option, 'keyboard disabled' is useful for applications where the computer will remain off for a relatively long period, or where the risk of accidental key actions (cleaners in your office...) must be eliminated. The keyboard remains enabled when bit D0 is high during the write operation that starts the switch-off sequence. A low at D0 causes the keyboard to be switched off with the computer (note that the keyboard is powered separately by the auto power-down unit).

Circuit description

Figure 2 shows the circuit diagram of the auto power-down control. Only a few of the signals on the PC extension bus connector are used. The 10-bit I/O address bus, A0–A9, is connected to an address decoder formed by comparator IC_1 and OR gate IC_{2c}. The I/O address occupied by the circuit can be set between 300_H and $31F_H$ with the aid of DIP switch S_1. The output signal of the address decoder at pin 10 of IC_{2c} is split into I/O read and

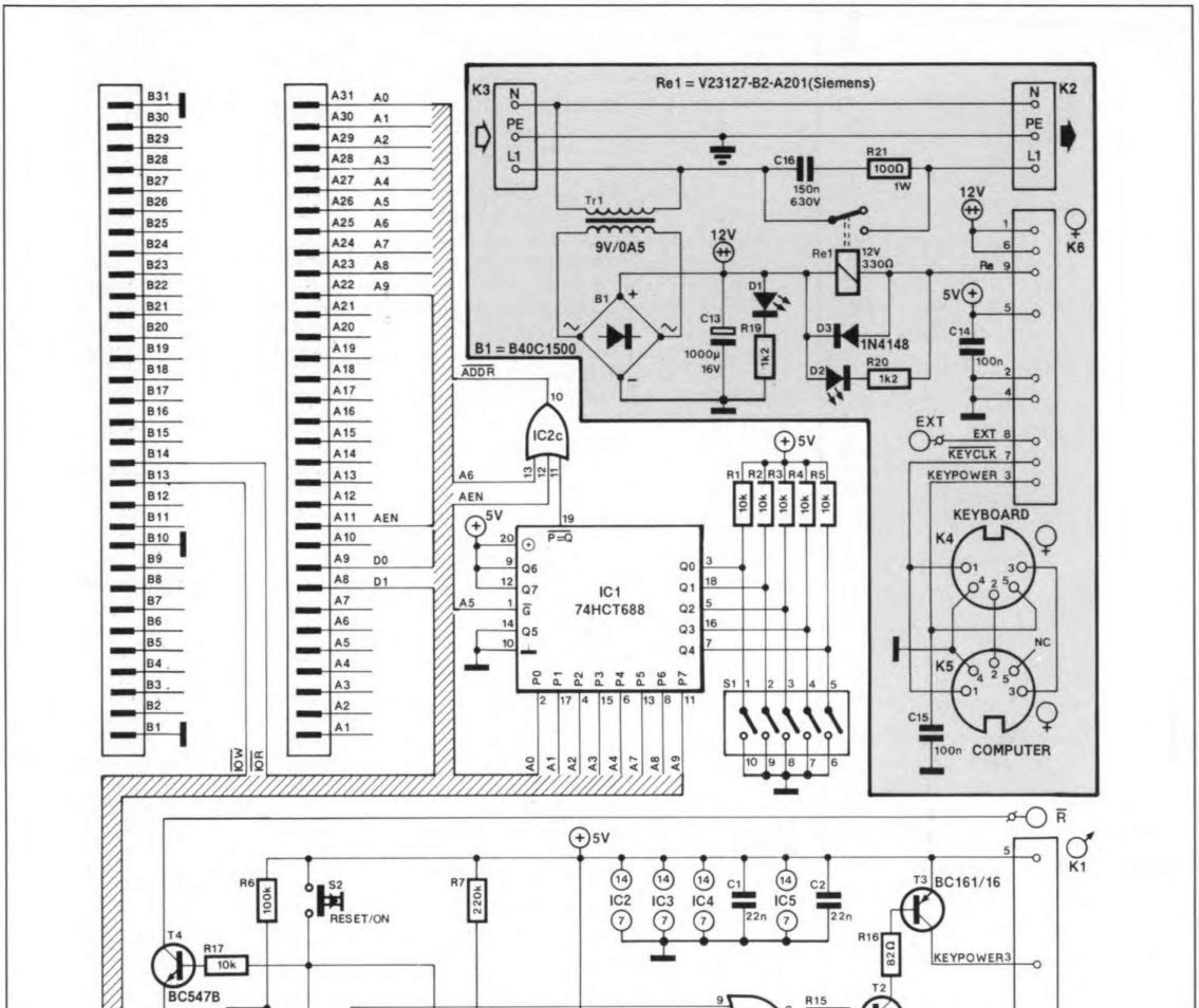


Fig. 2. Circuit diagram of the auto power-down controller. The parts in the shaded block are fitted external to the computer.

Signal name	Pin designation			Signal name
	track side	rear panel of PC →	component side	
GND	B01		A01	$\overline{\text{I/O}}$ CHCK
RESET	B02		A02	D7
+5V	B03		A03	D6
IRQ2	B04		A04	D5
−5V	B05		A05	D4
DREQ2	B06		A06	D3
+12V	B07		A07	D2
reserved	B08		A08	D1
+12V	B09		A09	D0
GND	B10		A10	$\overline{\text{I/O}}$ CHRDY
$\overline{\text{MEMW}}$	B11		A11	AEN
$\overline{\text{MEMR}}$	B12		A12	A19
$\overline{\text{IOWC}}$	B13		A13	A18
$\overline{\text{IORC}}$	B14		A14	A17
$\overline{\text{DACK3}}$	B15		A15	A16
DREQ3	B16		A16	A15
$\overline{\text{DACK1}}$	B17		A17	A14
DREQ1	B18		A18	A13
$\overline{\text{DACK0}}$	B19		A19	A12
CLK	B20		A20	A11
IRQ7	B21		A21	A10
IRQ6	B22		A22	A9
IRQ5	B23		A23	A8
IRQ4	B24		A24	A7
IRQ3	B25		A25	A6
$\overline{\text{DACK2}}$	B26		A26	A5
TC	B27		A27	A6
ALE	B28		A28	A3
+5V	B29		A29	A2
OSC	B30		A30	A1
GND	B31		A31	A0

For easy reference: signal assignment on the PC expansion slot connector. Note that 'track side' and 'component side' refer to insertion cards.

write signals by IC3d and IC3a respectively. When the CPU writes to the address occupied by the circuit, the logic levels on datalines D0 and D1 are latched into bistables IC4a and IC4b respectively. The databit in IC4a indicates whether or not the keyboard must remain on, while that in IC4b indicates that the switch-off sequence has been started. Bistable IC4b is wired as a monostable multivibrator, which can be triggered only when D1 is high. This forms the first security measure against accidental switching off. The second measure is that the auto-power down address must be read within the monotime of IC4b. Only when this is achieved is the monostable formed by IC5b started. Nothing happens in the circuit during the delay of about 10 s introduced by IC5b. However, the delay allows the system shutdown (or hard disk parking) program to be loaded and run from a batch file. After the 10-s delay, IC5b resets bistable IC5a via network R13-C11. As a result, the relay is de-energized via T1 so that the computer is switched off.

The computer may be switched on in three ways. All three have in common that IC5a is set so that the relay is energized.

The first way is by means of the RESET switch, S2. This is the RESET key fitted on most PCs. When S2 is pressed, the clock input of IC5a is made high via OR gate IC2a (the data input of IC5a is permanently high). To retain the original function of the RESET key, it is connected to the computer via T4. This allows it to be used as before when the computer is on.

The second way to turn on the computer is by means of a pulse applied to the EXT input (pin 8 of connector K1). The leading edge of the pulse clocks IC5a and switches the computer on. The pulse may be supplied by external equipment or a computer peripheral that requires a relatively long period of computer activity. It should not be used for short intervals since frequent switching on and off may reduce the lifetime of the computer.

The third way to switch the computer on is via the keyboard. As already discussed, this works only if databit D0 is high when the CPU writes to the I/O address occupied by the circuit. Since it is powered by the auto power-off circuit, the keyboard is on irrespective of whether the computer is on or off. In virtually all cases, a key code is sent when any key is pressed. The accompanying clock signal is fed to OR gate IC2b. This gate passes the clock pulses only when the keyboard is actually on (this is done to prevent interference from a disabled keyboard), and when at least a second has lapsed since the computer was switched off. The delay is provided by R11 and C12 which maintain a brief high level after the switching off.

The keyboard interface brings us to a part of the circuit that is fitted into an external enclosure instead of into the computer. In the circuit diagram, this section is shown with a shaded background. The keyboard connection consists of two DIN sockets to establish the connection to the computer. All terminals of K4 and K5 are interconnected with the exception of the keyboard supply pin 5 (remember that the keyboard is powered by the auto power-down supply). The keyboard clock signal is 'tapped' at K4.

The power supply is conventional with two LEDs, D1 and D2, to indicate the presence of mains voltage and the state of the relay respectively. The unregulated voltage of about 12 V across C13 is brought down to 5 V by IC6 on the control board in the computer.

Programming and adjusting

The programs that start the power-down sequence can be as simple as shown below. The examples are written in Turbo Pascal. First, a routine to switch off the computer only:

```
VAR DUMMY: BYTE;

BEGIN
PORT[$300] := 3;
DUMMY := PORT[$300]
END.
```

This works provided the auto power-down circuit is assigned I/O address 300_H by appropriate setting of the DIP switches. The second example switches the computer off but leaves the keyboard enabled:

```
VAR DUMMY: BYTE;

BEGIN
PORT[$300] := 2;
DUMMY := PORT[$300]
END.
```

The executable programs made in this way may be included in a batch file, just before the park (or shutdown) routine for the hard disk.

Preset P1 is relatively simple to adjust. Use the program that leaves the keyboard enabled. The computer is **not** switched on via the relay (it is in normal use). Initially, set P1 to minimum resistance, then run the auto power-down program. If necessary adjust P1 until LED D2 goes out; this indicates that the circuit works. Press any key; the computer must be switched on again. Try a few settings of P1 until the circuit responds reliably to the software. In case the relay is de-energized with P1 set to minimum resistance, or when nothing happens with P1 set to maximum resistance, the value of C9 may have to be reduced or increased respectively. ■

REAR WINDOW WIPER COUPLER

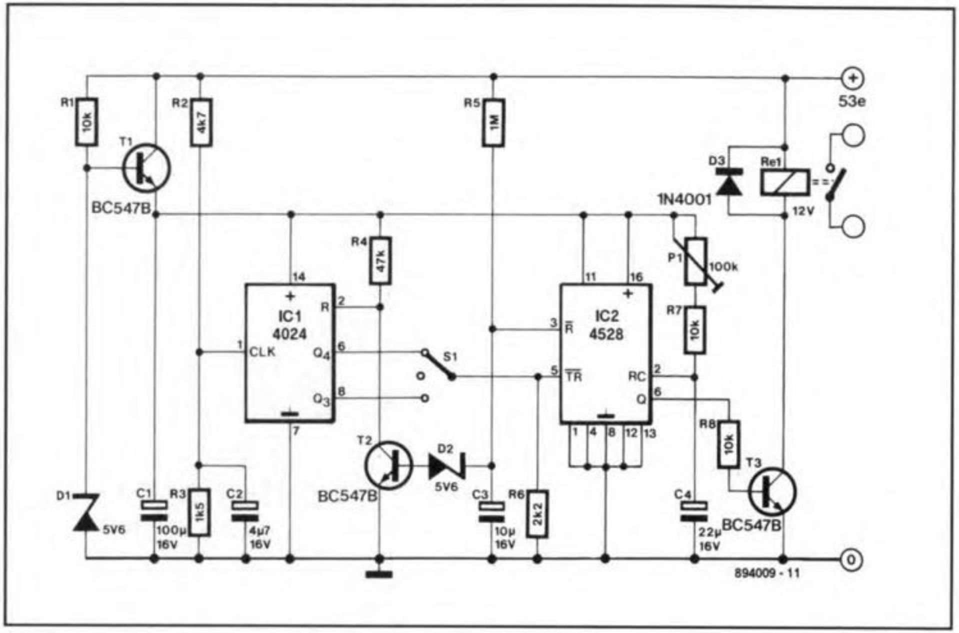
G. Kleine

From any point of view, it is convenient to couple the rear window wiper on our cars to the windscreen wiper. Since the rear window of a moving car does not get nearly as wet as the windscreen, it suffices if the rear wiper operates only once for every 8, 16 or 32 wipes of the windscreen.

At terminal 53e (green/black wire) on most cars, which is the return of the windscreen wiper motor, the clock signal for that motor is present. This signal, which is a square wave, is applied to the clock input of counter IC1 via R2–R3. The Q3 output of the counter goes high every eighth clock pulse, and the Q4 output once every sixteenth clock pulse. The output pulse is applied to monostable IC2 via switch S1. The monostable may be a Type 4528 (as drawn) or a Type 4548. At each trailing edge at pin 5, the monostable output (pin 6) goes high at a frequency of 1.5–16 Hz. When that happens, T3 is switched on, the relay is energized and the rear wiper operates.

The supply for the circuit is taken from terminal 53e also: during the intervals between the clock pulses, this terminal is at a potential of 12 V. In that state, T1 is switched on and C1 charges. When the clock signal is present at the terminal, T1 is off so that C1 cannot discharge. Diode D1 limits the supply to 5 V.

Network R4–R5–C3–T2–D2 ensures a regular reset of the circuit during the windscreen wiper interval. ■



INTERMEDIATE PROJECT

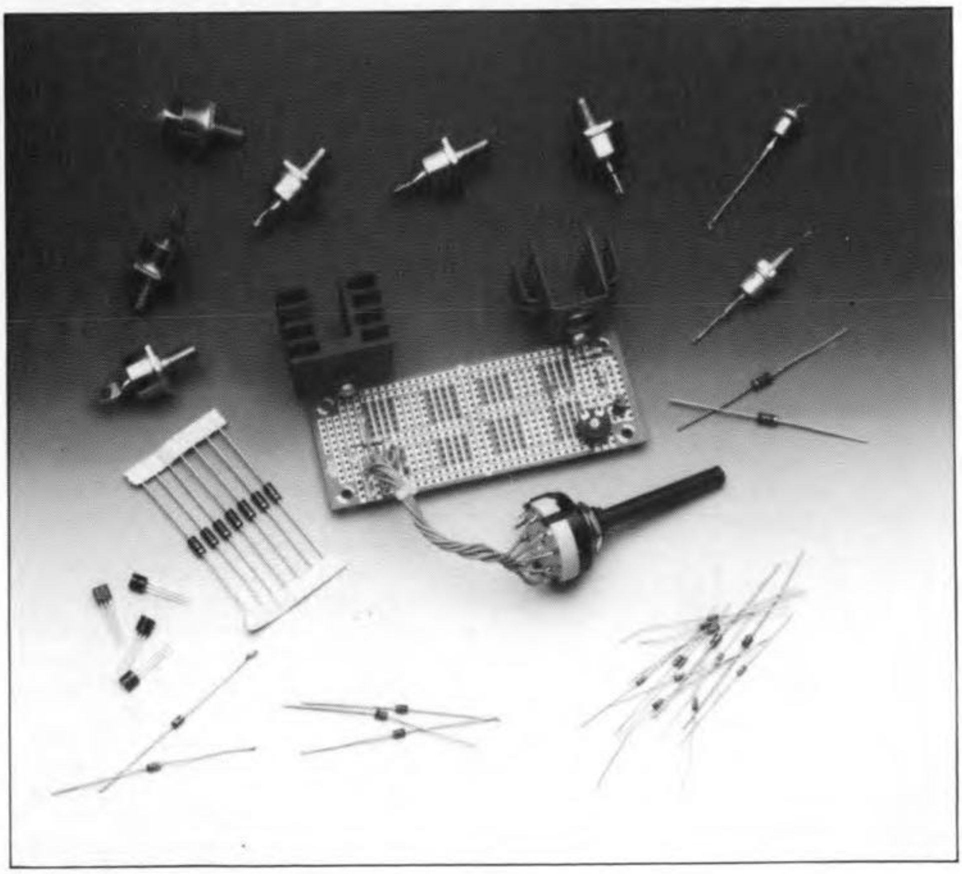
A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project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

POWER ZENER DIODE

K. Walters

Power zener diodes are rare, hard-to-obtain and quite expensive electronic devices. The circuit described here is based on a handful of common discrete parts that simulate a high-power zener diode. Apart from the saving in cost, the discrete equivalent has an additional advantage in that it can be set up to meet a wide range of output voltage and power requirements without the need of changing components.

Many voltages, whether direct or alternating, are far from constant. Examples of fluctuating alternating voltages include those supplied by the mains and many types of generator. In the case of direct voltages, sources like car batteries, inexpensive DC adapters and solar cells come to mind. Fortunately, most electronic equipment can cope reasonably well with too low a supply voltage, which is not usually worth worrying about because it is unlikely to cause damage. Overvoltage, however, often has disastrous effects even when the nominal supply voltage is exceeded by as little as 15%. Examples of easily damaged components include certain types of battery and chips from the now obsolete 74-TTL series.



In many cases, such damage may be prevented by inserting a regulator between the voltage supply and the equipment or circuit to be powered. Two types of regulator are available for this purpose as discussed later.

Although they may consume power from an alternating voltage source, most electronic circuits work from direct voltages, that are usually regulated (alternating voltages are fairly difficult to regulate with simple means).

Most of you will be familiar with the 78xx and 79xx series of three-pin fixed voltage regulators (Fig. 1). These devices are widely used in many circuits because they are both inexpensive and simple to use: all they require is an input and an output decoupling capacitor, a small heatsink and, importantly, an input voltage

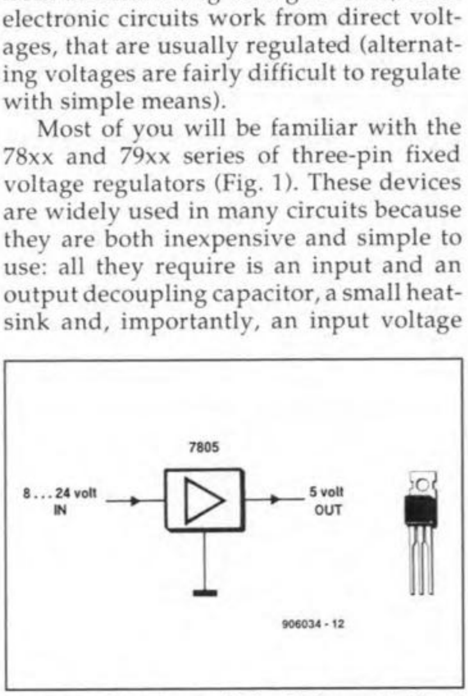


Fig. 1. Circuit symbol of three-pin fixed voltage regulator.

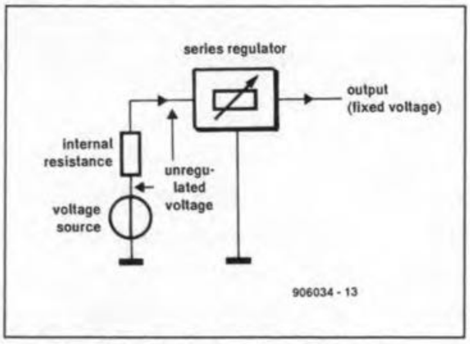


Fig. 2. Principle of series regulation.

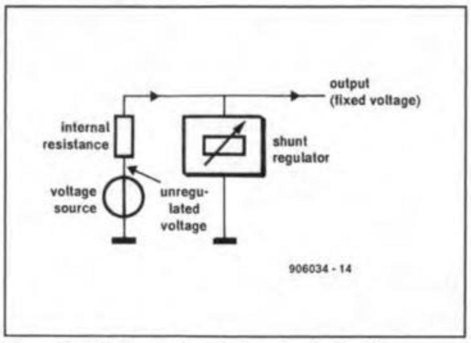


Fig. 3. Principle of shunt regulation.

that is at least 3 V higher than the output voltage, but not so high as to exceed the maximum specification (which is usually between 30 V and 40 V). These regulators are available in a wide range of output voltages (both negative and positive) and output currents (100 mA to 2.5 A).

Series or parallel?

The basic circuit shown in Fig. 1 is a series regulator. The regulating device is essentially an adjustable resistor connected between the voltage source (input) and the load (output) —see Fig. 2. In practice, the series resistor nearly always takes the form of a power transistor. The circuit compares the voltage across the load with a reference level. The difference between these voltages is used to provide the series transistor with a proportional base voltage. When the output voltage drops, the difference voltage rises so that the series transistor is driven harder. This, in turn, causes a lower series resistance so that the circuit counteracts the load variation by keeping its output voltage constant.

A series regulator based on an 78xx or 79xx integrated circuit typically has a relatively wide input voltage range with the minimum level defined as 3 V above the output level, and the maximum level defined by the permissible dissipation, which is the product of the voltage across the regulator and the output current. Unfortunately, the minimum voltage drop requirement may be a problem in a number of applications. For these, a different type of regulator is required.

Parallel regulation

In a parallel (or shunt-) regulator (Fig. 3), the regulating element prevents the voltage across the load exceeding a certain value. The zener diode is the simplest shunt regulator, although transistor are also commonly used for this purpose.

The operation of the shunt regulator is based on the internal resistance, R_i, of the voltage source. Assuming that the voltage rises above the level established by the regulator, R_i drops the resultant voltage difference. In more practical terms, the shunt lowers its resistance so that the voltage source sees a heavier load. The resulting current increase thus counteracts the voltage rise.

The shunt regulator is relatively inefficient since it passes more current as the voltage rises above the maximum level. Its advantage, however, lies in the absence of a voltage drop between the voltage source and the load.

Returning for a moment to the previously mentioned example of the solar cells, it will be evident that series regulation is of little use in that application: the maximum output voltage of the cell will seldom be high enough to cater for the 3-V drop caused by a series regulator used to charge a battery. The reasons for applying a shunt regulator are fairly obvious here.

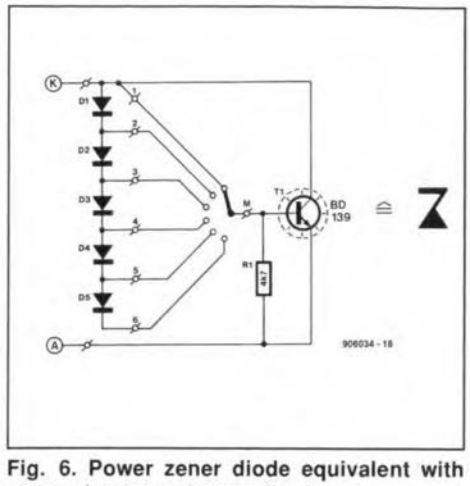


Fig. 6. Power zener diode equivalent with stepped zener voltage adjustment.

Zener diode

The voltage/current characteristic of a normal diode and that of a zener diode are shown in Figs. 4 and 5 respectively. In the case of the normal diode, the current increases sharply when the forward voltage rises above 0.6 V. The diode blocks when the voltage is reversed (negative), resulting in a small leakage current.

The curve in Fig. 6 shows that the forward voltage characteristic of a zener diode is almost the same as that of a normal diode. When the voltage is reversed, the current remains small initially, but then rises suddenly at a particular value. At this value, the zener voltage, the device prevents the reverse voltage increasing by passing more current. The effect is, of course, bound by practical limits: the zener diode must be capable of handling the current that flows when the voltage rises.

When applied to our example of the solar cells, a zener diode would afford sufficient protection against overvoltage on the batteries when there is intense sunshine. Unfortunately, it is not that simple in practice.

Power zener diode

It may well happen that the zener diode can not handle the current supplied by the solar cells because its maximum dissipation is exceeded. Most zener diodes have a maximum dissipation of 400 mW or 1.3 W, while 5-W types are also found occasionally. Zener diodes with a higher power rating are expensive and hard to find, whence the idea to make one from discrete components. This has a further advantage in that it allows the zener voltage to be made adjustable.

Figure 6 shows the principle. The base of a medium-power transistor is connected to the junctions of a series of diodes. Since a normal (silicon) diode starts to conduct at a forward voltage of about 0.6 V, the rotary switch allows the collector-emitter voltage at which the transistor starts to conduct to be set in steps of 0.6 V. When the switch is set to position 1, the circuit simulates a 0.6-V zener diode, or 3.6 V when the switch is

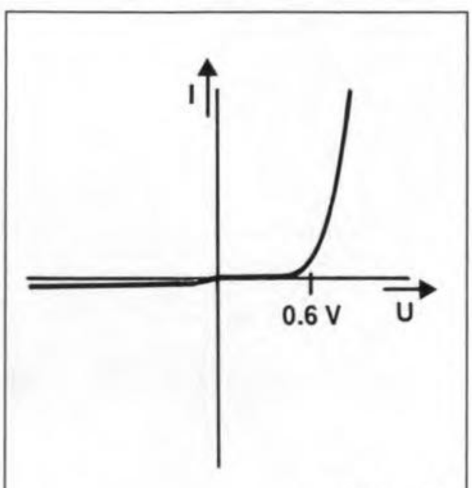


Fig. 4. Typical silicon diode U-I characteristic.

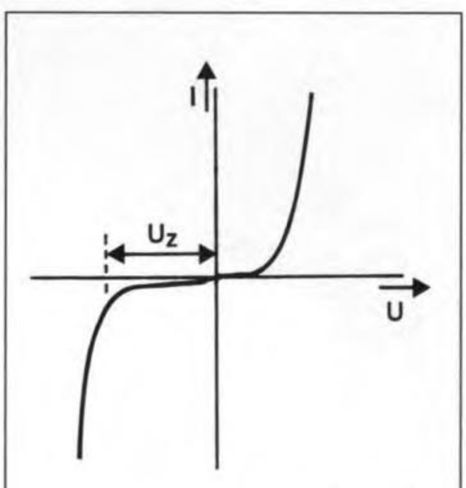


Fig. 5. Typical zener diode U-I characteristic.

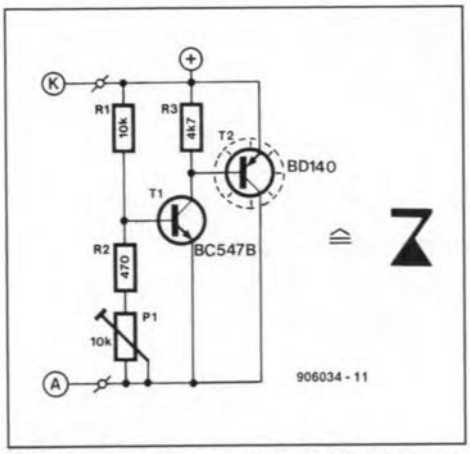


Fig. 7. Power zener diode equivalent with continuously adjustable zener voltage.

set to position 6. The maximum current is about 1 A in both cases.

Figure 7 shows an alternative circuit which provides a continuously adjustable zener voltage. The base-emitter voltage of transistor T1 rises with the voltage across the circuit. When this voltage reaches the level at which T1 starts to conduct, T2 conducts also since its base is made negative with respect to its emitter. Hence, T2 is driven harder as the base voltage of T1 rises so that any increase in the voltage across the circuit is counteracted by a larger load current.

Since the base of T1 is connected to a potential divider, R1-(R2+P1), the point at which the base voltage of 0.6 V is reached is determined by the input voltage level and the setting of P1. This creates a zener diode with an adjustable zener voltage between 2 V and 12 V. Provided T2 is fitted with a reasonably sized heat-sink, the circuit is capable of passing up to 1 A. The shunt current may be increased to 8 A by replacing the BD139 in Fig. 6 with a darlington transistor Type TIP130 (note that this results in a zener voltage which is 0.6 V higher than that of the BD139-based circuit). The same current boost may be achieved in the circuit of Fig. 7 by replacing the BD140 by a TIP135. ■

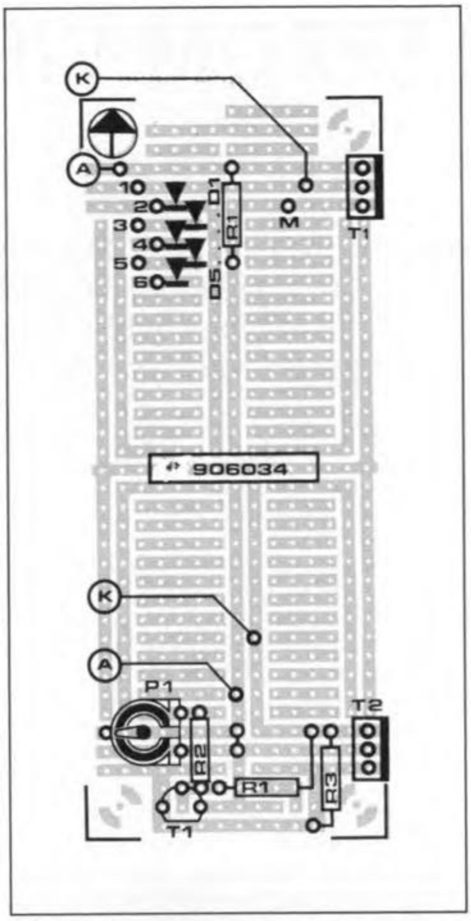


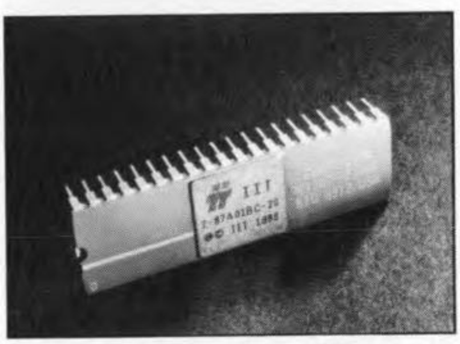
Fig. 8. Suggested construction of the power zener circuits (Figs. 6 and 7) on universal prototyping board size-1 (UPBS-1) Do not forget to fit the power transistors with heat-sinks.

NEW PRODUCTS

Ultra-fast co-processors break price barriers

IIT Inc. can now supply the 2C87, a high-performance numerics co-processor that is plug and object-code compatible with 80287. The IIT 2C87 is a low-power CMOS device capable of operating at clock rates up to 20 MHz. Its low current demand makes it ideally suited for lap-top computer implementation.

The 2C87 is less expensive than a 80287-16, yet performs most of its functions in far fewer clock cycles. When combined with the faster clock frequency (the 2C87 can operate on the same clock as the 80286), the floating point processor achieves performance at least two times faster than the 80287. When used with an 80286 processor the computing system fully conforms to the IEEE Floating Point Standard. The new co-processor includes a built-in instruction to calculate a 4×4 matrix transformation. This results in the capability to perform matrix transformations 6 to 8 times faster than the 80287. In addition, the 2C87 provides extra functions which are not available on the 80287. The benchmark test and the table compare the performance of the 2C87 with the 80287. The IIT 2C87 is housed in a 40-pin ceramic package.



Also new from IIT is the 3C87, which is claimed to operate up to 50% faster than a 80387 in the same application.

Integrated Information Technology Inc. (IIT) • 2540 Mission College Blvd. • Santa Clara, Califirnia 95054 • U.S.A. Telephone: (408) 727-1885. Fax: (408) 980-0432.

UK representative:

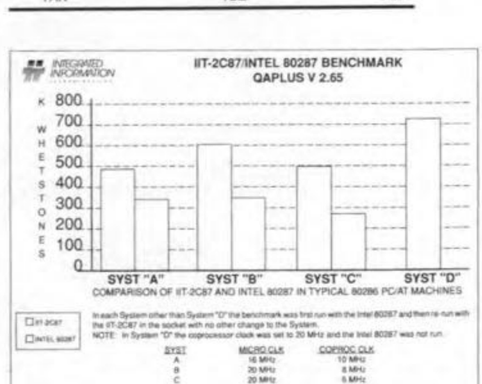
DMST Ltd. • Mead House • Pinkneys Drive • Maidenhead • Berkshire SL6 6QD. Telephone: (0628) 777700. Fax: (0628) 27237.

CLOCK CYCLES REQUIRED

INSTRUCTION	IIT-2C87	80287
ADD	15-17	70-100
MPY	19	90-145
DIV	48	193-203
SQRT	49	180-186
COMPARE	17	40-50
REM	58	15-190
TAN	196	30-540
LOG	235	900-1100

TABLE 2. CLOCK CYCLES REQUIRED

INSTRUCTION	IIT-3C87	80387
ADD	11	31
MPY	15	57
DIV	44	88
SQRT	45	125
REM	54	155
TAN	192	726



MINI EPROM VIEWER

J. Ruffell

This circuit complements the mini EPROM programmer featured a few months ago (Ref. 1). The simple-to-build unit allows data loaded into most of the currently used types of EPROM to be examined in a straightforward manner.

It may be argued that a full-blown EPROM programmer with PC control is, in many cases, superfluous luxury when only a small part of the contents of an EPROM is to be examined. The EPROM viewer discussed here allows this to be done rapidly and in a simple manner with the aid of three keys for address selection and EPROM type, and, of course, a clear display to indicate the relevant data.

Simplified circuit diagram

Integrated circuit IC5 is a suitable starting point for the description of the operation of the circuit shown in Fig. 1. For the sake of clarity, this diagram is a simplified version of the full circuit diagram, which is discussed later.

When the circuit is switched on, IC5, a decade counter, and dividers IC1–IC4, are reset by an *R-C* network. The reset condition is indicated by all LEDs (D6, D7 and D8) and all decimal points in the displays LD1–LD4 remaining off. Circuit IC5 controls all functions of the EPROM viewer. By pressing the SET key, it is provided with a single clock pulse, so that counter IC4 is enabled via an electronic switch. The other counters, IC1, IC2 and IC3, remain disabled. Thus, only IC4 is set by actuating the UP and DOWN keys. By pressing the SET key, IC3 is actuated. The next key actions actuate IC2, and, lastly, IC1. This selection allows the four nibbles that forms the EPROM address to be set (one nibble = four bits = one hexadecimal number). The decimal points on the 7-segment LED displays indicate the current nibble selection.

The next three states of IC5 determine the EPROM type selection. The types are divided into three groups: xx512, xx256 or xx128, and xx64. When IC5 reaches one of the last three states, IC1–IC4 are transformed from four discrete 4-bit counters into a single 16-bit synchronous counter. At the same time, the display is switched from an address read-out to a combined data/address read-out: the data byte in the EPROM is shown on the two least-significant displays, and the least-significant EPROM address byte on the two most-significant displays. This combined display mode is indicated by the decimal points on the 'data' displays, LD1 and LD2.

The EPROM address may be set by pressing the UP or DOWN key. It is possible to cross page boundaries (one page = 256 addresses). This is achieved with the aid of the CARRY output of IC2, which enables IC3 and in addition causes IC16 to switch via a monostable, so that the full address is briefly displayed again. After the mono time has lapsed, the display reverts to the combined LSB-address/databyte indication mode with the two actuated decimal points. The next action on the SET key causes IC5 to reset itself, allowing a new function to be set.

MAIN SPECIFICATIONS

- **Reads 8-KByte, 16-KByte, 32-KByte and 64-KByte EPROMs.**
- **Full address entry or one-address up/down mode.**
- **Combined address/data read-out mode.**
- **Multiplexed 7-segment LED display.**
- **Three-button control.**
- **Optional computer interface.**
- **Ideal for small-scale EPROM applications.**

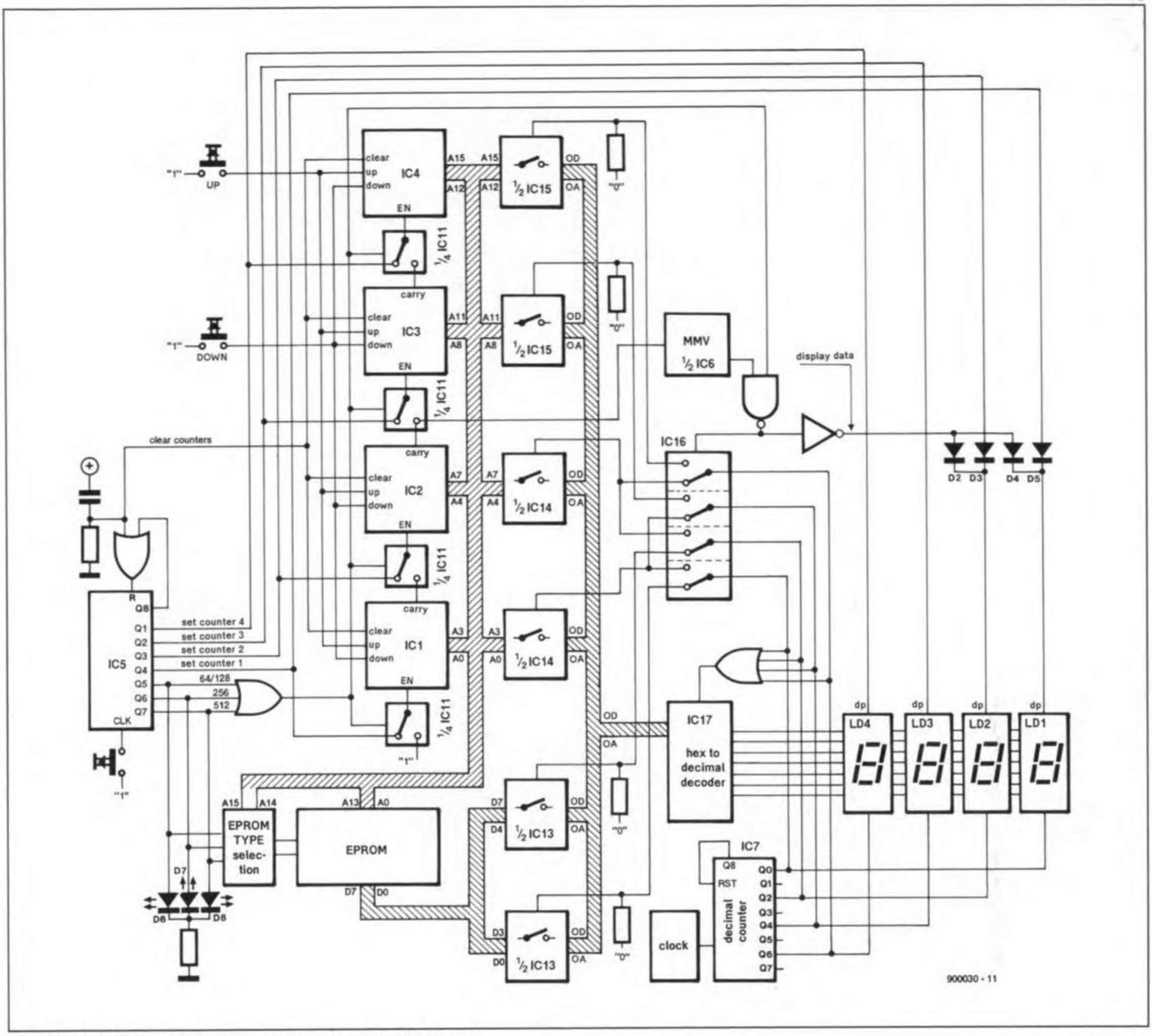


Fig. 1. Simplified circuit diagram of the EPROM viewer.

The 4-digit LED display is multiplexed under the control of counter IC7. The displays are rapidly actuated and turned off again by the respective outputs of IC7. The data associated with a particular address is simultaneously fed to the 7-segment decoder, IC17, via data switches IC13, IC14 and IC15. Since the outputs of IC7 are alternately connected and not connected, a short delay is introduced during which the data switches are allowed to change state. The delay effectively prevents 'ghosting' of the display owing to the settling time of the switches and the display driver transistors, T1–T4.

Circuit IC16 distributes the multiplex signals in two ways across the switches. The distribution shown in the circuit diagram enables the combined 'LSB-address/data' read-out, the other distribution the 'full address' read-out.

The real thing

The full circuit diagram, shown in Fig. 2, differs from the previously discussed simplified version in a number of ways. This is mostly due to practical realizations of target functions set out in Fig. 1. Take, for instance, the OR gate for the reset function of IC5 in Fig. 1: in the actual circuit, this takes the form of a diode-resistor combination, D1-R5. Similarly, counters IC1–IC4 are not actually reset but preset with value 0000, which has the same effect. The simple up/down control suggested by Fig. 1 is actually realized by a clock pulse generator, NAND gate IC9b, a delay, R9-C5, a re-triggerable monostable, IC6a, and IC9c. The latter controls the level of the UP/$\overline{\text{DOWN}}$ inputs of the counters. The clock pulses for the counters are provided by IC6a, whose monotime is set to about 2 ms to cope with fast actions on the keys. The 2-ms delay must be taken into account if the circuit is controlled by a computer via terminals A, B and C.

The EPROM type selection circuit must ensure the correct signals at pins 1, 26 and 27 of the EPROM used. Pin 26 is either not connected (xx64 types) or forms the A13 input (all other types). This allows A13 to be connected to EPROM pin 26 with impunity. The situation is a little more complex in the case of pin 1. Fortunately, a single three-input OR gate, IC12b, does it all. If an xx512 is selected, this gate passes address signal A15. In all other cases, it holds pin 1 logic high. Pin 27 is controlled in a similar manner: with xx64 and xx128 types, it is logic high, while with xx256 and xx512 types signal A14 is applied via IC12c. The valid address ranges for the four EPROM types that may be used are as follows:

EPROM	size	address range
xx64	8 KByte	0000 – 1FFF
xx128	16 KByte	0000 – 3FFF
xx256	32 KByte	0000 – 7FFF
xx512	64 KByte	0000 – FFFF

The circuit is powered by a single 5 V rail. The power supply used must be capable

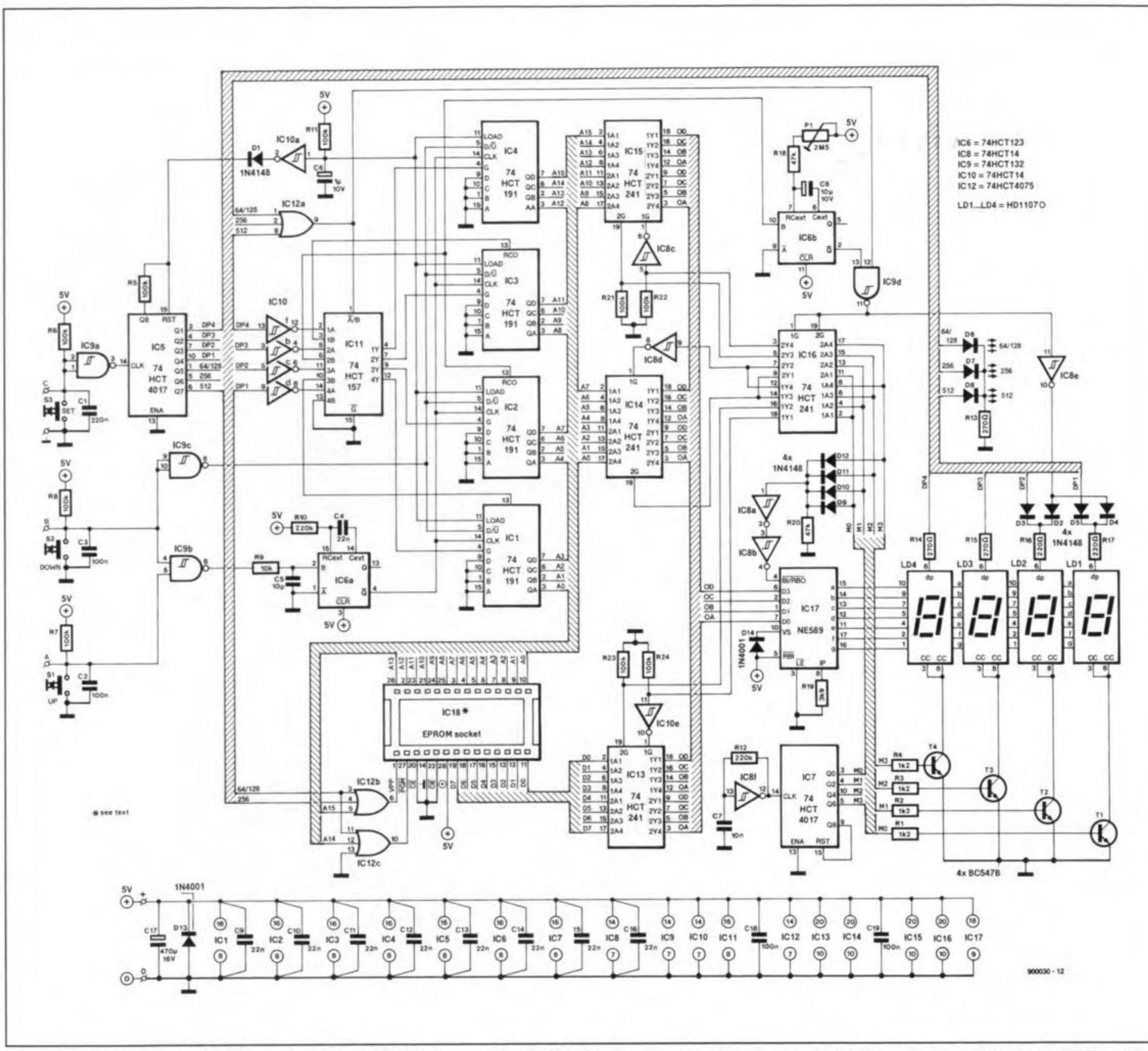


Fig. 2. Complete circuit diagram of the EPROM viewer. The EPROM to be examined is plugged into the zero-insertion force (ZIF) socket marked IC18.

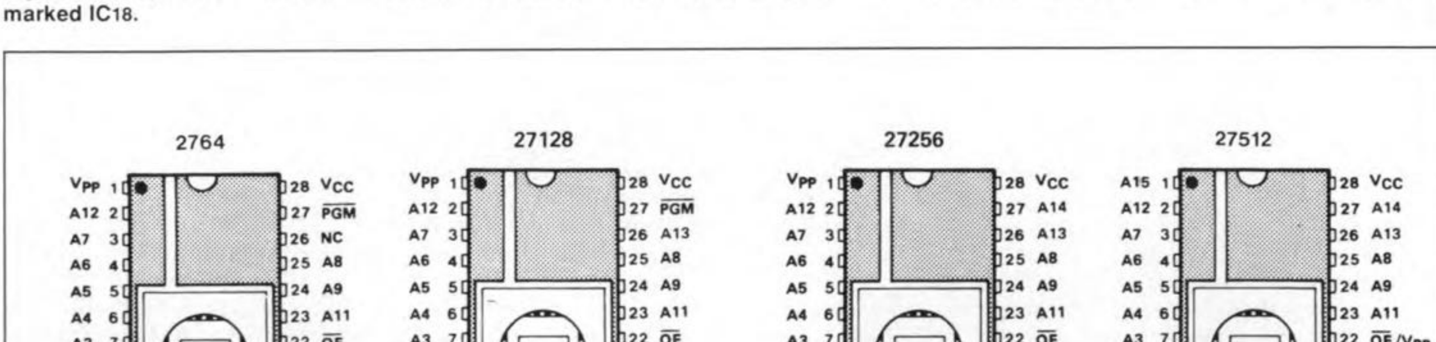


Fig. 3. Pin-outs of the EPROMs that can be handled by the viewer.

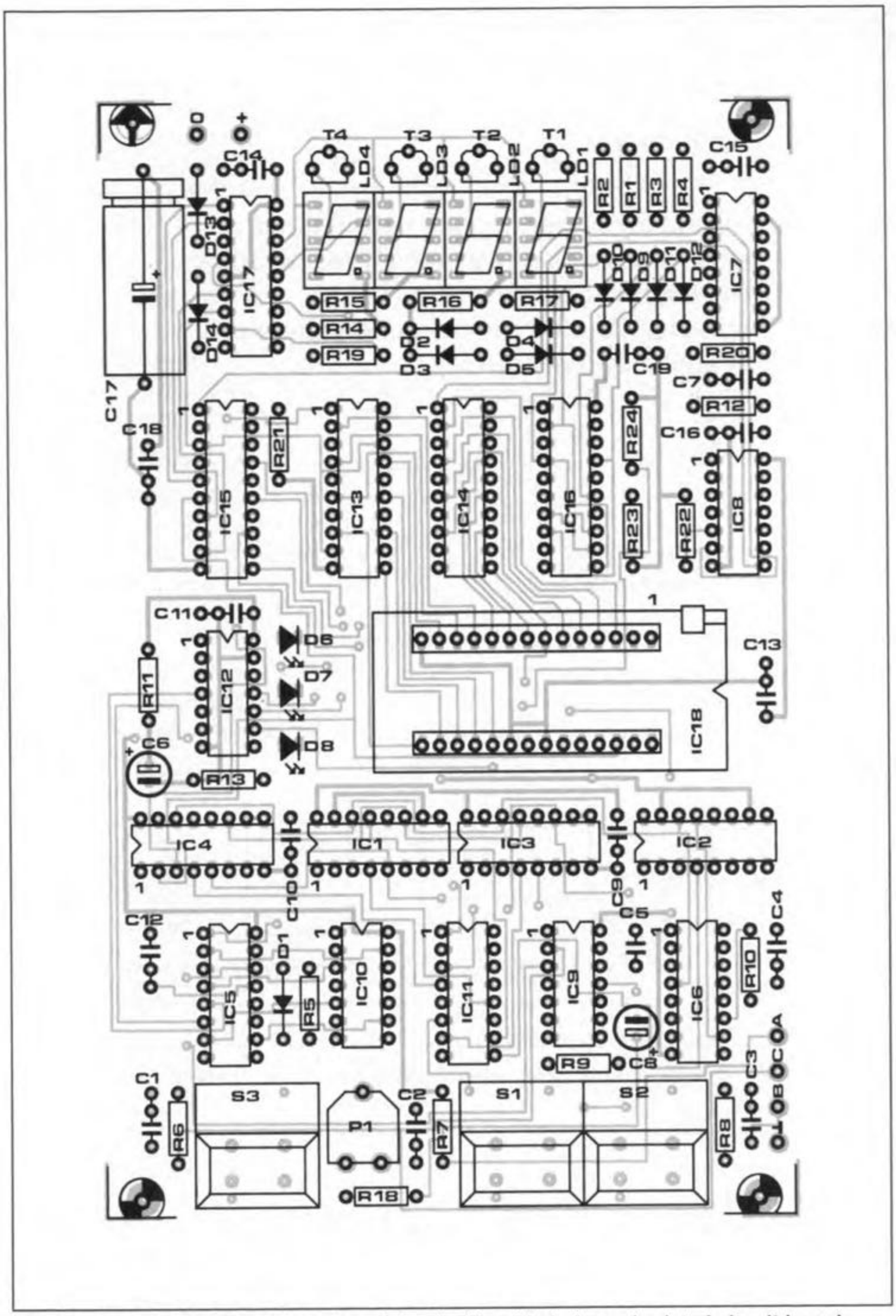


Fig. 4. Component mounting plan of the double-sided, through-plated circuit board.

COMPONENTS LIST

Resistors:

4	1k2	R1–R4
9	100k	R5–R8;R11; R21–R24
1	10k	R9
2	220k	R10;R12
3	270Ω	R13;R14;R15
2	220Ω	R16;R17
2	47k	R18;R20
1	3k9	R19
1	2M5 preset H	P1

Capacitors:

1	220nF	C1
4	100nF	C2;C3;C18;C19
9	22nF	C4;C9–C16
1	10p	C5
1	1µ0 10 V radial	C6
1	10n	C7
1	10µF 10 V radial	C8
1	470µF 16 V axial	C17

Semiconductors:

9	1N4148	D1–D5;D9–D12
3	LED red 3mm	D6;D7;D8
2	1N4001	D13;D14
4	HD1107O (common cathode)	LD1–LD4
4	BC337	T1–T4
4	74HCT191	IC1–IC4
2	74HCT4017	IC5;IC7
1	74HCT123	IC6
1	74HCT14	IC8
1	74HCT132	IC9
1	74HCT14	IC10
1	74HCT157	IC11
1	74HCT4075	IC12
4	74HCT241	IC13–IC16
1	NE589N (Signetics)	IC17

Miscellaneous:

1	28-way ZIF-socket (e.g. Textool)	IC18
3	push-button ITW type 61-10204000	S1;S2;S3
1	enclosure (200×110×30 mm) e.g., Bopla type EG2030	
1	printed-circuit board	900030

of providing a regulated 5 V at up to 0.5 A to cater for the relatively high current consumption of the LED displays.

Construction

The printed-circuit board for the project, shown in Fig. 3, is double-sided and through-plated. Construction is mostly straightforward soldering work with reference to the parts list and the component marks printed on the board. Position IC18 represents a 28-way zero-insertion force (ZIF) socket to ensure that the EPROM to be examined can be fitted and removed from the viewer without its pins being deformed. Depending on the height of the enclosure used, it may be necessary to mount the LEDs, the displays, the ZIF socket and the switches at a certain height above the board. The displays are conveniently given the appropriate height by stacking a few IC sockets. A bezel may be used to improve the readability of the displays.

The circuit has only one adjustment: P1. The preset defines the time during which the full address is shown when the most-significant address byte changes. This time may be set to a value between 0.5 s and about 5 s.

Finally, always switch off the circuit before inserting or removing an EPROM. ■

Reference:

1. Mini EPROM programmer. *Elektor Electronics* January 1990.

CRO CALIBRATOR

D. McBright

A true-square-wave signal is useful for measuring the performance of the Y-amplifier of an oscilloscope. In particular, it enables three parameters to be checked: gain, amount of tilt produced by the amplifier and the frequency response. The calibrator described here provides such a signal.

The calibrator delivers a true-square-wave signal to facilitate (see Fig. 1):

- (a) measurement of the gain of the Y-amplifier by displaying a signal whose peak-to-peak level has been set accurately;
- (b) measurement of the amount of 'tilt' produced by capacitive coupling circuits along the flat top, at a frequency of about 50 Hz;
- (c) checking the frequency response of the Y-amplifier (but not its rise time) set by the frequency correcting trimmers in the coarse gain switch and in the probe at a frequency of 500–1500 Hz.

The calibrator is also useful when the oscilloscope is used in the 'variable gain' mode to measure accurately the level of a displayed signal (especially if the scope is home-built!).

The principle

The principle used in the calibrator is a simple but effective one for which only a handful of components are needed to produce a good square wave.

First, the level of a direct voltage is set with the aid of a DC voltmeter. The voltage is then 'chopped' to produce a square wave of known amplitude, and this is used as an AC signal to calibrate the oscilloscope.

Since the calibrator is connected directly across the output terminals, the accuracy of the system is determined chiefly by the accuracy of the voltmeter used. For level measurements, the voltmeter remains connected all the time, but is only read when the calibrator is switched to SET UP, since the square wave in the CAL position produces a reading of approximately half the peak-to-peak value. However, when the calibrator is used for checking the frequency response, the voltmeter should be disconnected to reduce any shunt capacity that might degrade the signal.

Level measurements may be made in two ways:

- first, the DC level is set with the aid of a voltmeter, after which the oscilloscope is adjusted to the resultant square wave;
- the square wave, as viewed on the oscilloscope, is adjusted to the required level, after which the calibrator is switched to SET UP and the voltmeter read to give the peak-to-peak level.

The output impedance, with terminating switch S4 open, is of the order of 600 Ω, but this will be modified by the resistance of the connected voltmeter. With S4 open, the output voltage range extends from over 4 V_{p-p} to less than 2 mV_{p-p} when both the coarse (S2) and fine (P2) are set correctly. When S4 is closed, R21 is connected in parallel with the output, which causes the voltage range to be halved: this makes setting up easier at some levels.

Since the meter terminals are wired in

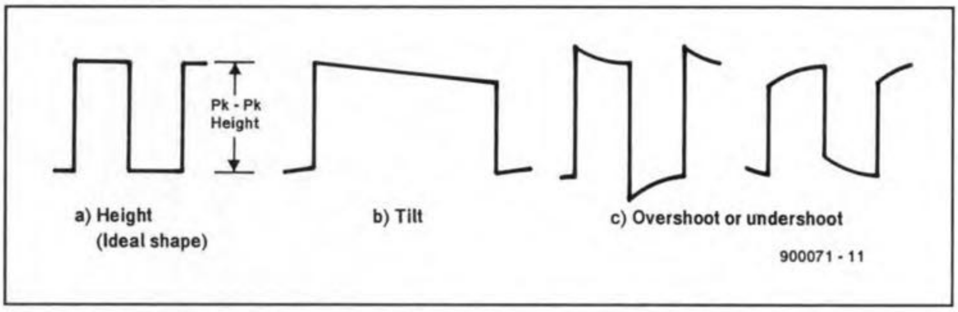


Fig. 1. The calibrator enables three parameters to be checked.

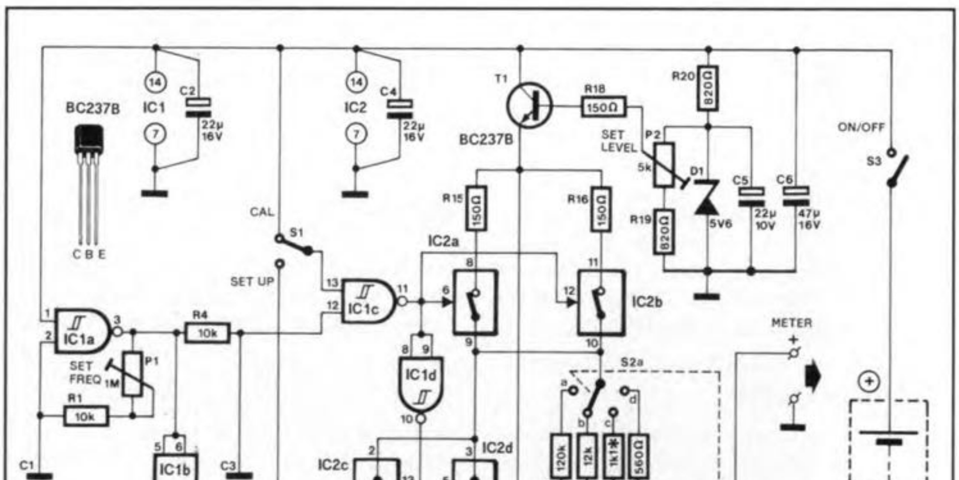


Fig. 2. Circuit diagram of the CRO calibrator.

parallel with the coaxial outlet, the connexion to the scope may be taken from either. However, since the output impedance is relatively high, connexions made in screened cable should be kept as short as possible.

A separate output of about 1 V_{p-p} is provided for use as a triggering signal. The calibrating square wave is delayed some 20 µs relative to this output.

The frequency range is of the order of 30 Hz to 3 kHz.

The circuit

The variable oscillator is based on Schmidt NAND gate IC1a: oscillations result from the 'toggle' action of the gate on network R1-P1-C1. The output signal of the oscillator is divided between IC1b and IC1c.

Circuit IC1b provides an output of about 1 V across 600 Ω for triggering purposes.

After a delay of some 20 µs provided by network R4-C3, circuits IC1c and IC1d produce a square wave and an inverted square wave respectively. These signals are used to drive the control circuits of the analogue switches.

One input of IC1c is connected to the mother contact of CAL/SET UP switch S1. This switch, when in position SET UP (GND), keeps the analogue switches connected to the emitter of T1; in position CAL (+9 V), it allows control of IC1c and IC1d by the square wave.

Circuit IC2 is connected as a change-over switch with its mother (common) contact, connected to coarse level control S2a, switching between the reference voltage at the emitter of T1 and GND. The switches in IC2 are connected as paralleled pairs in series with small isolating resistors: R5, R6, R15 and R16. The base voltage of T1 is set by 'fine level' control P2. The potential across P2-R19 is stabilized by zener diode D1.

The choice of output impedance was limited by the series resistance of the analogue switches and their inability to handle high currents. Ideally, 75 Ω or 50 Ω would have been preferred, but since this was not possible without greatly reducing the maximum output voltage, 600 Ω was chosen.

The square-wave output of the calibrator has a rise time of about 0.08 µs and, at some levels, small overshoots at the same speed. These are, however, fast enough to be ignored in any measurements for which the calibrator is designed.

The current drawn from the 9 V battery lies between 5 mA and 11 mA.

Construction

All components, except the potentiometers, switches and terminals which should be fitted to the front panel, may be mounted conveniently on a piece of veroboard of about 75×75 mm.

Stand-off pins should be used for connecting the wiring to the various controls and the battery. The board, together with the battery, is fitted in a 150×100×60 mm enclosure.

Only two precautions are necessary: decoupling capacitors C2 and C4 should be mounted as close as possible to the relevant ICs, and, because of the presence of 'fast edges', connexions to coarse level switch S2, the output terminals, and frequency control P1 must be kept as short as possible. ■

AZTEX 24 CM FM ATV TRANSMITTER AND 24 CM GaAs FET PREAMPLIFIER

900077

A review by Mike Wooding, G6IQM

Ever since it was first exhibited at last year's Leicester Show, the Aztex 24 cm FM ATV transmitter has created a lot of interest and I have often been asked for details of this unit. These requests prompted me to do a detailed review to satisfy the many enquirers.

Before commencing the review, I had some correspondence with designer Ken Stevens, G4BVK, from which I quote: "In the design of a transmitter, the need of a stable output is one of the governing factors and in the 24 cm model the Type SP5060 phase-locked loop was specified. The use of surface-mounted components was another way of maintaining stability."

"The video pre-emphasis network, whilst based on the standard CCIR circuit was designed to give an HF lift in addition to that given by a standard CCIR network. This overcomes the HF losses in the modulator in addition to providing the normal HF lift. Introducing some kind of DC restoration on the signal was also deemed necessary before it was injected into the modulator, and the circuit adopted to achieve this is very effective in stopping the video content from altering the black-level position."

"The two sound inputs are mixed actively with the aid of a Type TL072 opamp before they are fed to the modulator. There is a separate PCB-mounted preset for adjusting the line input only: the front panel sound control adjusts the composite level of both inputs. A sub-carrier injection level preset is also provided on the PCB."

TECHNICAL SPECIFICATION	
Frequency - channel 1	1249 MHz
channel 2	1255 MHz
RF power output	2.5 W (typical)
Harmonics	<50 dBc
Modulation system	FM with built-in emphasis
Audio sub-carrier	Preset to 6 MHz; ≥17 dB below carrier (variable with deviation setting)
Video input	1 V p-p into 75 Ω
Audio inputs	Dynamic microphone Adjustable line input (VCR, etc.)
Power consumption	1.6 A at 13.8 V

Description

The unit comes fully assembled in a die-cast box measuring 188×120×57 mm and has a removable lid secured by six cross-head screws. The professionally produced front panel contains the mains on-off switch; channel selector; two potentiometers—one for the sound and the other for the video deviation; and four LEDs:one indicating the connection of the DC supply, one indicating transmit, and two indicating which channel is selected.

The rear panel contains an N-type socket for connecting the aerial; a BNC socket for the video input; two sockets for the audio inputs; a phono socket for the line input and a 1/4 inch socket for the microphone. The DC input is via a 3-pin plug: a lead with a matching line socket is provided.

My only criticism of the outward appearance of the unit as supplied was the lack of identification of the connection points at the rear of the unit. Whilst no one is likely to confuse the aerial socket or the power plug with any other, confusion could arise between the two audio inputs and, perhaps, the video input. I understand, however, that these deficiencies are attended to in production models (the review model was only the fourth or fifth made).

Internally, the transmitter is laid out neatly, with the mother board housing the audio amplifier, modulator and sub-carrier generator circuits, and the video circuits. The board is held in place with four nuts, bolts and spacers, so that removal for any servicing will be an easy matter.

All the RF circuits are contained in a small die-cast box, occupying about one third of the main case at the right-hand side. This has the advantage of providing a further level of screening between the baseband and RF sections of the transmitter. The N-type RF output socket is mounted through the case directly into the inner box and is soldered direct to the PCB, so that no RF cables float around inside. Interconnections between the RF box and the mother board is effected by several feed-through terminals carrying the baseband signal, power supply and frequency switching control signals.

The RF assembly is bolted to the main case with the same bolts securing the internal PCB via spacers. The N-type aerial socket is bolted to the RF box and a clearance has been drilled through the main case. Thus, although a little intricate, removal of the RF assembly and circuit board for servicing can be effected fairly easily, although some care is required. The circuitry is designed in state-of-the-art surface mount technology (SMT).

Three user adjustment points are provided inside the transmitter: a preset potmeter for audio sub-carrier injection level; a trimmer capacitor for setting the audio sub-carrier frequency; and a preset potmeter for the line audio level. The supply is protected by a miniature wire-ended fuse soldered between two posts on the mother board.

Bench tests

The test equipment used to carry out the laboratory tests is listed below. I would like to thank Roland Hall, G0GSA, for his assistance with the analyser tests and plots.

- Marconi 2383 Spectrum Analyser & Tracking Generator.
- Racal Dana 9087 Signal Generator.
- Philips PM5646 Television Pattern Generator.
- Hewlett Packard 435A Power Meter.
- Hewlett Packard 8481A Power Sensor.
- Racal Dana 1998 Frequency Counter.
- Philips PM3226 Oscilloscope.
- Fluke 8050A Digital Multimeter.
- Racal Dana 9232 Bench Power Supply.

Frequency stability.

Because of the crystal-controlled PLL exciter, the frequency stability is good as shown in Table 1. It is seen that the total

TIME	FREQUENCY
switch on	1249.02623 MHz
10 sec	1249.02611 MHz
20 sec	1249.02581 MHz
30 sec	1249.02569 MHz
40 sec	1249.02559 MHz
50 sec	1249.02558 MHz
60 sec	1249.02549 MHz
70 sec	1249.02543 MHz
80 sec	1249.02541 MHz
90 sec	1249.02538 MHz
100 sec	1249.02537 MHz
110 sec	1249.02535 MHz
120 sec	1249.02534 MHz
3 min	1249.02503 MHz
10 min	1249.02403 MHz
15 min	1249.02411 MHz
20 min	1249.02401 MHz
30 min	1249.02389 MHz

TIME	POWER
switch on	2.72 W
5 min	2.54 W
10 min	2.46 W
15 min	2.37 W
20 min	2.35 W
25 min	2.34 W
30 min	2.34 W

Table 2.

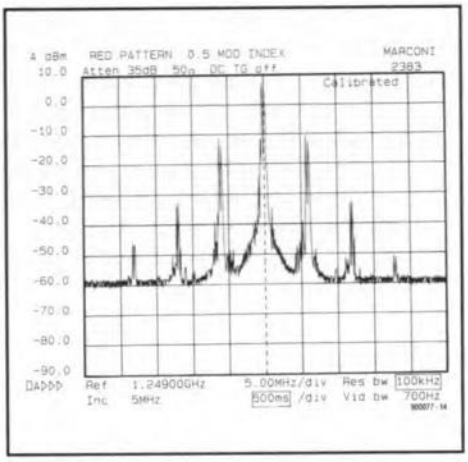


Fig. 1. Spectrum obtained using a plain red pattern.

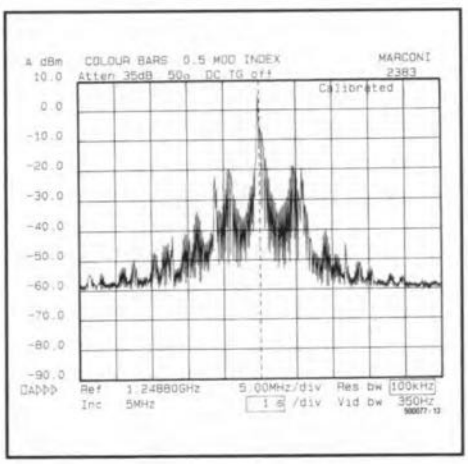


Fig. 2. Spectrum obtained using 100% saturated colour bars.

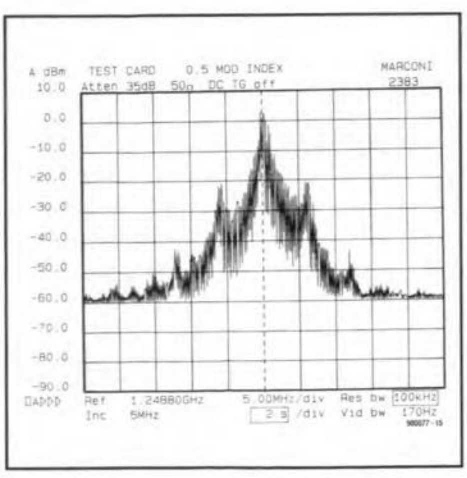


Fig. 3. Spectrum obtained with a Philips PM5534 test card.

drift of the review unit amounted to only 2200 Hz over a 30-minute period from switch-on.

Power output and harmonics.
The RF power output was monitored over a 30-minute period at 1249 MHz: the results are shown in Table 2. After the initial 20-minute period during which the power dropped by 0.37 W (0.65 dB), the output, to all intents and purposes, remained constant at 2.3 W. A similar check was carried out at 1255 MHz: here the initial power output was slightly higher at 2.84 W, with the final output settling at 2.54 W.

The harmonic content of the unmodulated output was very low, indeed, probably thanks to the out-of-band rejection characteristic of the SC1043 PA output device. The second harmonic was measured at slightly less than 50 dB down on the carrier. Third and subsequent harmonics were not detectable above the –75 dB noise floor of the analyser.

Video and audio characteristics.
A modulation index of 0.5 was ascertained by applying a 5 MHz sine wave to the video input and, viewing the output on a spectrum analyser, adjusting the video amplitude (deviation) from the signal generator so that the sidebands coincided with the recommended modulation index of 0.5. The output from the signal generator was then measured and this level used as the reference output level from the Philips TV pattern generator for the plots shown in Figures 1, 2 and 3.

The audio sub-carrier generator is normally set for UK use at 5.9996 MHz, but can be easily reset to the 5.5 MHz European standard with the sub-carrier oscillator trimmer, which is accessible when the top cover is removed from the transmitter.

At maximum video deviation, the sub-carrier level was measured at 17 dB below carrier, but this relative difference becomes greater as the video deviation is reduced with the VIDEO control on the front panel. At minimum deviation, the sub-carrier was measured at 32 dB below carrier. With a standard video input level of 1 V_{p-p}, it was necessary to set the VIDEO control at approximately 50% to achieve a normally deviated picture, and at this setting, the audio sub-carrier was measured at about 24 dBc, which proved to be quite sufficient for good audio with P5 contacts. Nevertheless, I did adjust the sub-carrier injection control on the main PCB and brought the relative level back to 17 dBc at this VIDEO control setting. This provided very good audio fidelity commensurate with picture reception.

In all fairness, I should point out that if the input video level to the transmitter is adjusted so that the VIDEO control on the transmitter is set towards fully clockwise in operation, the audio sub-carrier level is satisfactory without internal adjustment.

On-air tests and conclusions

Overall, I was very impressed with the workmanship and presentation of the transmitter. Upon receipt, it was simply a matter of connecting 13.8 V, plugging in the camera and microphone, connecting the aerial, switching on, and adjusting the audio and video controls. This is ideal for those of us who are not of the home-brewing fraternity. In my opinion, this is currently the only unit available that satisfies this need.

Furthermore, the very useful output level of around 2.5 W is enough to drive a 2C39A valve linear to quite a useful output (in my case to about 60 W). The colour handling characteristics of the unit gave excellent results, as did the audio response, when tested over a P5 path.

Apart from the criticism mentioned earlier on, I do not like the use of a soldered-in fuse. Although to some of us obtaining and changing this should failure occur would not be much of a problem, at some stations it could be a daunting task.

The Aztex ULNA 23–24 GaAs FET preamplifier

MANUFACTURER'S SPECIFICATION	
Typical gain	17 dB
Noise figure	1 dB
Bandwidth	1250–1350 MHz ±1 dB
Rejection	8 dB at 700 MHz
DC supply	7–18 V

The ultra-low-noise preamplifier comes in a blue hammer-finished die-cast case measuring 110×60×30 mm, with N-type input and output sockets provided at the sides. The top cover of the case is secured by four cross-head screws. NOTE: this enclosure is not water-proof, as stated by the manufacturers, and therefore needs to be mounted inside another weather-sealed enclosure for external/mast mounting.

The printed-circuit board is secured by two of the N-type socket's fixing bolts on each side. The DC supply is fed into the box by two insulated solder terminals and is provided with a reverse polarity protection diode.

The GaAs FET is a Type ATF10135, one of the latest Avantek products. It is mounted on a vertical PCB screen soldered to the main circuit board. A brass horizontal top screen is soldered to the vertical PCB screen and is also clamped to the side of the box under two of the retaining screws of the N-type output socket.

Input and output tuning trimmer capacitors are mounted at each socket and a bias preset potmeter is located on the main PCB.

Bench tests

The test equipment used to carry out the laboratory tests consisted of:

- Marconi 2383 Spectrum Analyser and Tracking Generator.
- Fluke 8050 Digital Multimeter.
- Racal Dana 9232 Bench Power Supply.

Air tests and conclusions

The flat, even response of the preamp over the entire 23–24 cm band meant that I was able to tune to signals at the RMT2 repeater input and output frequencies (1249 MHz and 1318.5 MHz respectively) without any loss of preamplification. This was a new experience for me as my own homebrew GaAs FET preamp is a half-band unit, which requires adjusting when tuning from one end of the band to the other.

Also, the very low noise figure of the preamp means that the 15.5 dB or so of gain appears to be a great deal more when compared with results obtained from other preamps with perhaps higher gain figures (not the least my own unit!). Noise figures are complex and I do not intend to go into these, but I can say that this unit with its noise figure of around 1 dB will take a lot of beating.

My overall impression of the preamp is of a sound, well-made unit that provides very good performance. I wholeheartedly recommend anyone requiring a preamp for the 23–24 cm band (which, to my mind, just about includes everyone working 24 cm FM ATV) should seriously consider this unit.

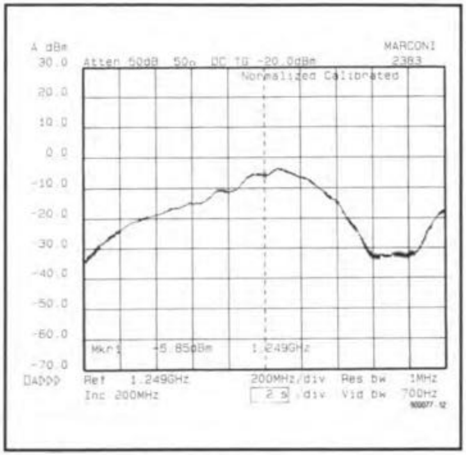


Fig. 4. Gain vs frequency characteristic over the band 249–2240 MHz. The reference input level was –20 dB and the centre frequency of the characteristic is 1249 MHz. It can be seen that the 0 dB gain points are at about 600 MHz and 1700 MHz (frequencies above 2200 MHz are ignored). The –3 dB band extends roughly from 1318 MHz to 1500 MHz.

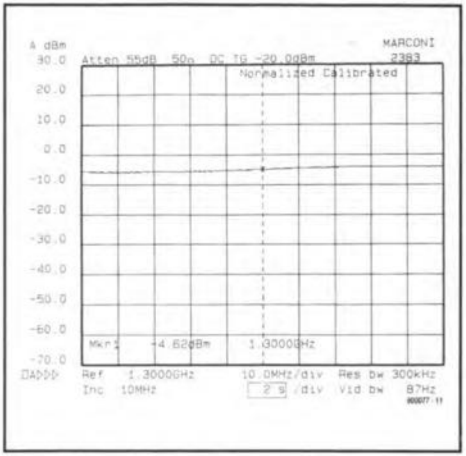


Fig. 5. Gain vs frequency characteristic over the band 1250–1350 MHz with the centre frequency at 1300 MHz. It is seen that the response in the 23–24 cm band is very flat with a positive gain slope. The reference input level was –20 dB and the gain at 1300 MHz was measured as 16.1 dB. The gain at 1249 MHz was measured as 15.5 dB, while at 1318 MHz it was 16.4 dB.

The Aztex TVTX 24 cm FM ATV transmitter is priced at £220 plus £2.50 p&p, and the Aztex ULNA 23-24 GaAs FET preamp retails at £52 plus £1.50 p&p.

Both units are available only from the Severnside Television Group, 15 Witney Close, Saltford, BRISTOL BS18 3DX, Telephone (0255) 873098. ■

FOUR-SENSOR SUNSHINE RECORDER

J. Ruffell

Although it is almost impossible to imagine present-day meteorology without electronic equipment, sunshine is still widely measured with the aid of the classic, all-mechanical, Campbell-Stokes recorder. The electronic instrument described in this article offers far better accuracy than the Campbell-Stokes meter. It measures the total time during which sunshine is detected, and records this information with the aid of a computer.

The recording of the duration of sunshine time over a certain period is among the daily routines of meteorologists. The measurement is usually carried out with the aid of the Campbell-Stokes recorder, a drawing of which is shown in Fig. 1. The recorder consist of a sphere arranged to focus the sun's image on to a bent strip of card on which the hours are marked. When the sun shines with sufficient intensity, the sphere focuses the light on the paper, which is burned at that spot. At the end of the day, the length of the burnt track on the paper enables the sunshine duration to be calculated. Interruptions in the track indicate passing clouds. Although the inaccuracy of the Campbell-Stokes recorder is fairly high at 10 to 20%, the instrument is still widely used for lack of an better alternative.

Like the original Campbell-Stokes recorder, the all-electronic version discussed here has no moving parts. Its measuring principle does, however, require a computer for data recording purposes. In principle, any computer may be used for this application, provided the right interface and control program are available. In this article we will assume that a 8052-BASIC computer board is used (see Refs. 1 and 2).

Four sensors: a new principle

The electronic sunshine recorder is based on four photodiodes Type BPW21 which measure the ambient light intensity. Since meteorologists hold that there is sunshine if there is a shadow, the position of each of the four sensors enables it to measure a different light intensity when the sun shines. When the illumination is homogeneous, as is the case with an overcast sky, all sensors measure the same light intensity. When the sun shines, the sensor that is best aimed at the sun receives the highest light intensity. The computer connected to the sensor assembly runs a continuous calculation on the position of the sun, while accounting for the geographical coordinates, the season and the local time. In this manner, the system always knows which sensor faces the sun and thus receives more light than the others. Next, the average light intensity on all four sensors is measured. The system detects sunshine when the sensor aimed at the sun measures a substantially higher light intensity than the sensor with the weakest illumination. This approach makes the accuracy with which the number of sunshine hours is recorded dependent on the measurement frequency only. While an accuracy of about 30 minutes per day is quite good for a Campbell-Stokes recorder, the present circuit is capable of resolving seconds of sunshine —quite an improvement!

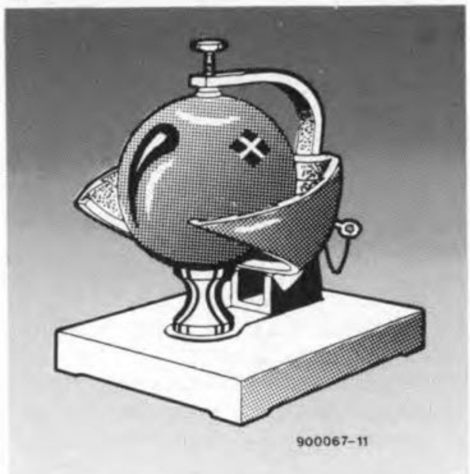
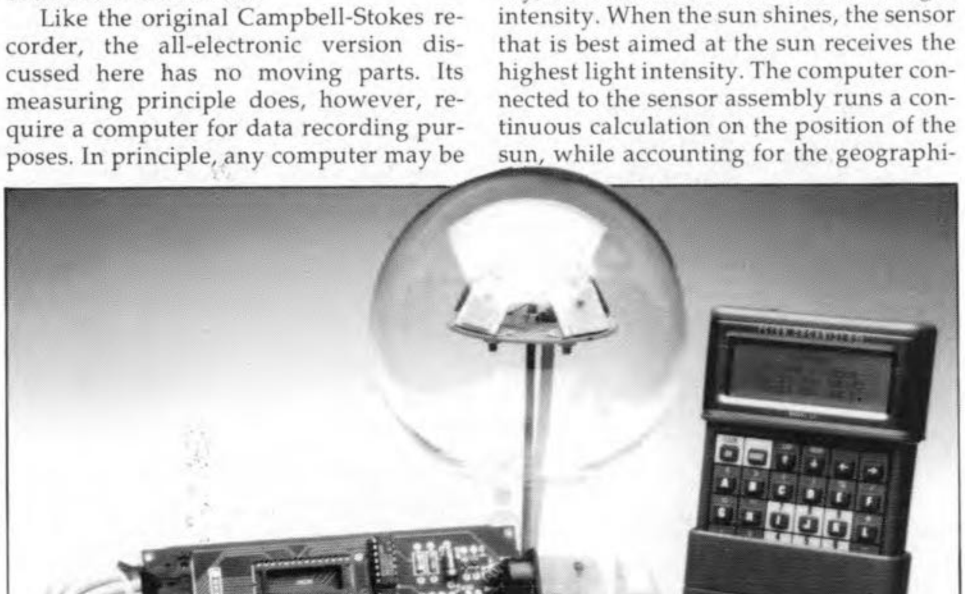


Fig. 1. Artist's impression of a Campbell-Stokes sunshine recorder.



Sun position calculations

The fact that the movements of the planets in the solar system are regular makes it fairly easy to provide reasonably accurate predictions for the position of the moon, the sun and planets in relation to the earth. This means that the sunrise and sunset times on earth may be calculated. In his article "Calculating the position of the sun" (Ref. 3), R. Walraven proposed a

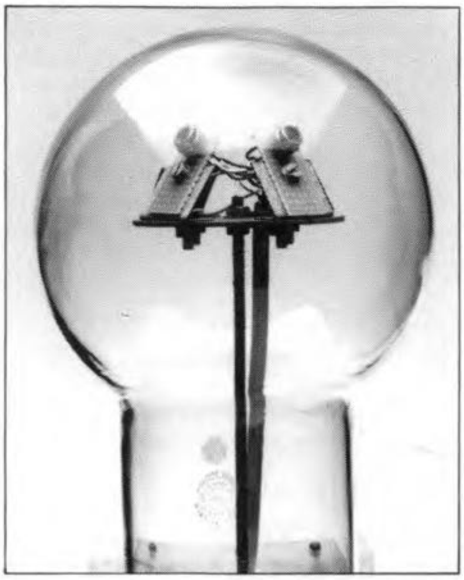


Fig. 2. Prototype sensor assembly enclosed in a glass vessel. The photodiodes are mounted on small pieces of stripboard.

Fortran computer program for calculating the position of the sun in the sky above any location on earth, with an accuracy of 0.01°. This program has been used as the basis for developing a BASIC version that works in conjunction with the real-time clock function available in most PCs.

The program has been adapted further for use on the single-board BASIC computer. Since this has a real-time clock, the system may be used as a stand-alone application that works around the clock for sunshine recording. In this set-up, the PC may collect measured data when required by connecting it to the BASIC computer via its RS-232 interface. The program that runs on the BASIC computer is available as a ready-programmed EPROM.

Sensor assembly

The photograph in Fig. 2 shows that the four photodiodes are fitted together to form a in a transparent sensor. Each photodiode faces one of the compass directions, north, west, south and east. Since they are enclosed in a perspex, plastic or glass cover, they may be used for many years without the need of being serviced.

The control program that runs on the BASIC computer performs light intensity measurements at regular intervals. The data so obtained is collected and processed statistically by the PC at a later time. The sun position calculations enable the computer to know which photodiode (measurement sensor) is expected to measure the highest light intensity. The photodiode that measures the lowest light intensity forms the reference. The system decides that there is sunshine if the light intensity on the measurement sensor is four times greater than that on the reference sensor.

The sensitivity of the photodiode used in the sensor assembly is highest for light incident at right angles to the face of the diode. The altitude, α, of the sun varies between 0° and 90° in the tropics, but this reduces the further north or south one is from that region; in the UK (and most of north and central Europe, the southern parts of Canada and the northern states of the USA) it varies from 0° to 65°. In this article that value is used, and the sensor assembly is therefore constructed in a manner that ensures that the maximum

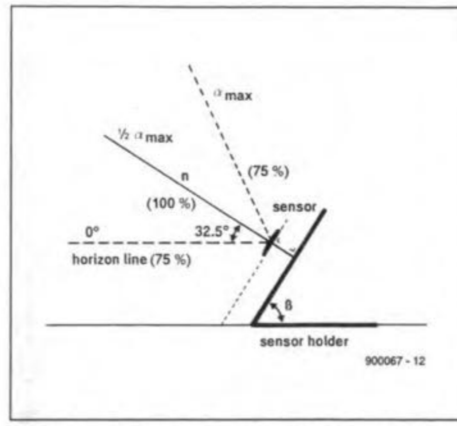


Fig. 3. Determining the optimum sensor position with respect to the earth surface.

deflection of the direction of the light with respect to the normal, n (the line perpendicular to the plane of the photodiode), is 32.5°. In other regions of the earth that angle needs to be adapted as appropriate. The maximum deflection of 32.5° causes a sensitivity reduction of 75%, which is still acceptable in this application.

The computer calculates the sensitivity as a function of the position of the sun and that of the sensor. This is done to make the measurement independent of the position of the sensor relative to the sun, so that the angle at which lights falls upon the sensors (angle of incidence) has virtually no effect on the collected data. The drawing in Fig. 3 shows that each photodiode is mounted on a post or plate that is at an angle, β, of 57.5° (90°–32.5°) with respect to the earth surface (this angle needs to be adapted in other regions). The assembly must be mounted such that photodiode D_0 faces north, D_1 west, D_2 south and D_3 east.

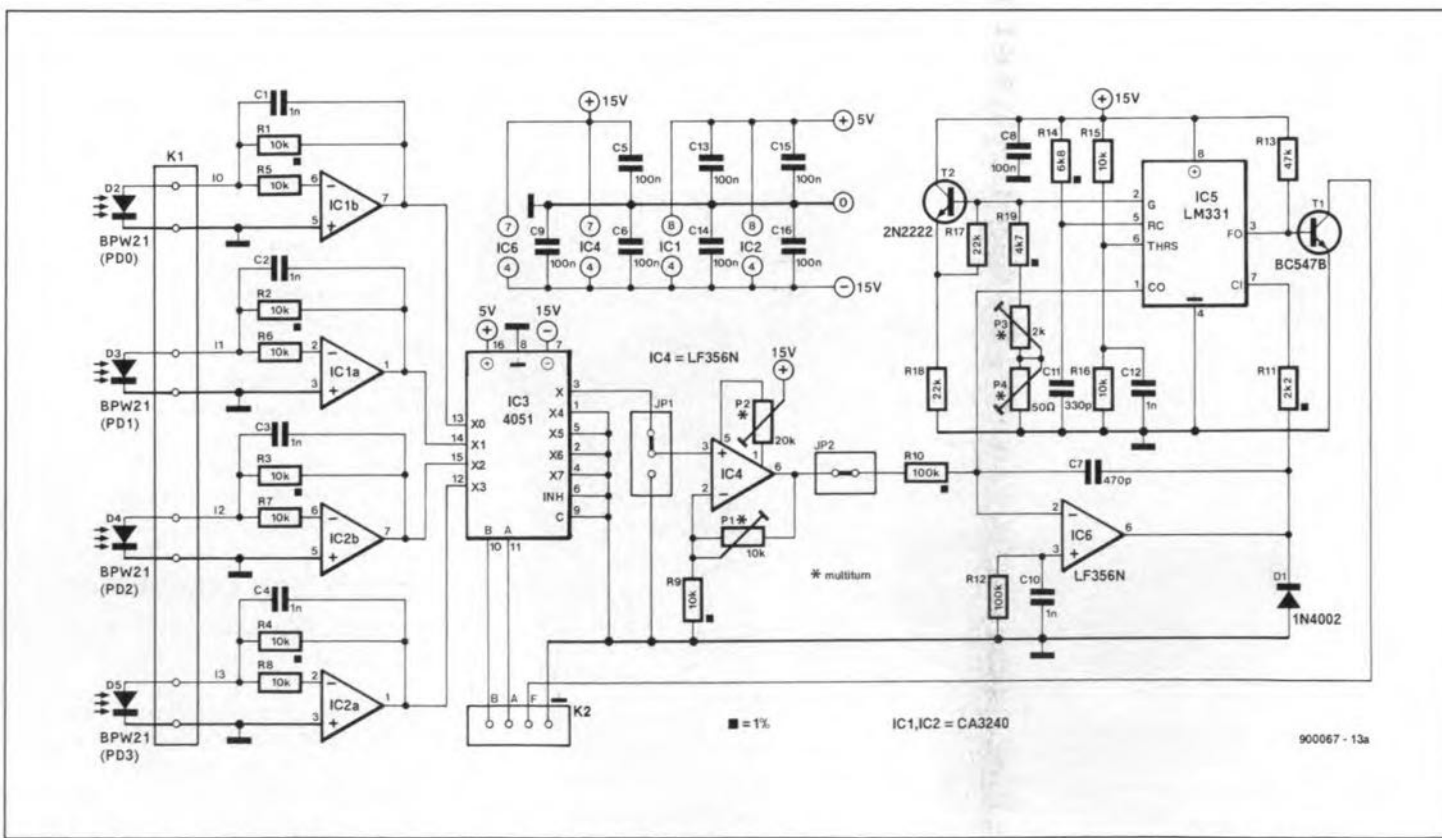


Fig. 4. Circuit diagram of the sensor amplifiers and the VCO that enables the measured values to be recorded by the BASIC computer.

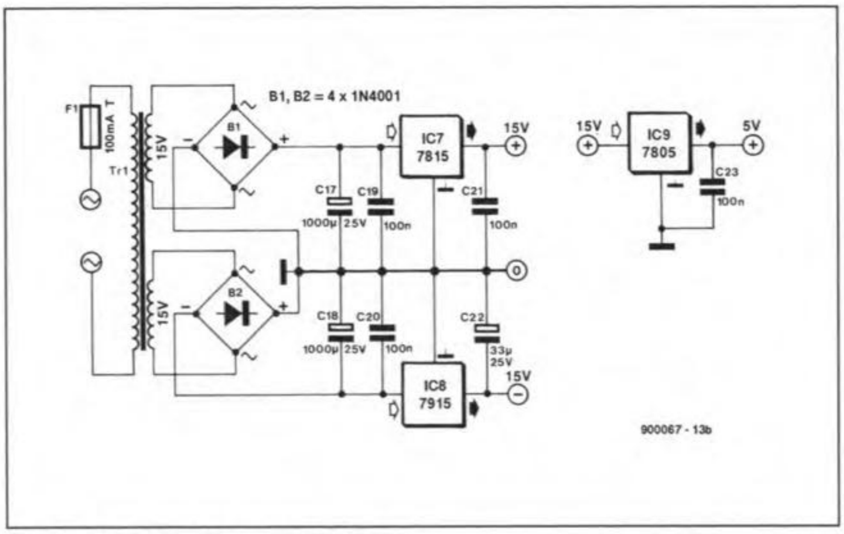


Fig. 5. Suggested power supply.

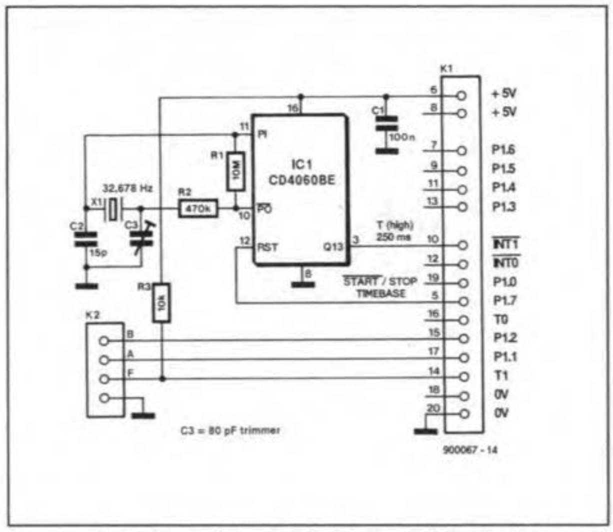


Fig. 6. Timebase for use with the BASIC computer.

Electronics!

The circuit diagram of the sunshine recorder is shown in Fig. 4. The whole circuit is in fact a peripheral device whose output data is applied to the BASIC computer.

The currents supplied by photodiodes D_2–D_5 are first converted to voltages by IC_{1a}, IC_{1b}, IC_{2a} and IC_{2b}. Next, the voltages are applied to analogue multiplexer IC_3, of which the channel selection inputs, A and B, are controlled by the computer. A communication protocol is required to ensure correct data transfer between the sensor and the BASIC computer, which may be separated by several meters. Voltage-to-frequency (v-f) conversion is used for this purpose: the sensor assembly supplies a signal whose frequency is determined by the light intensity on the selected photodiode. Opamp IC_4 buffers the multiplexer output voltage and controls the VCO (voltage-controlled oscillator) built around IC_5. The VCO is a relatively complex circuit to ensure the required accuracy.

The VCO output signal is sent to the BASIC computer via output connector K_2. The advantage of v-f conversion is that the BASIC computer can accept the measured signal direct without the need of additional hardware.

A simple time-base circuit (Fig. 6) provides 250-ms long enable pulses to the timer on board the 8052AH-BASIC processor. The number of pulses read from the count register indicates the frequency of the measurement signal and thus the intensity of the light that falls upon the selected sensor.

The EPROM-resident control program alternately selects the measuring sensor and the reference sensor by addressing the analogue multiplexer, IC_3, via control lines A and B. Immediately after each selection, the frequency of the signal on the output line, F, is measured. Next, the program corrects the sensitivity of the sensors and converts the measured frequency into a corresponding illumination intensity. The result of this process forms the basis of the sunshine/no sunshine decision.

Construction and test

Both the interface circuit and the time-base circuit are so small that they are easily constructed on veroboard or prototyping board. Wiring is not critical since only low frequencies are involved. Care should be taken, however, to fit all the decoupling capacitors indicated in the circuit diagram as close as possible to the relevant ICs.

The interface has four adjustments, the time-base circuit only one. If a frequency meter is available, adjust the preset on the time-base board until 32.768 kHz is measured. If you do not have access to a frequency meter, set the preset to the centre of its travel.

Initially, presets P_2, P_3 and P_4 on the interface board are also set to the centre of their travel. Turn P_1 to maximum resistance and install jumper J_1 to connect the non-inverting input of IC_4 to ground. Adjust P_2 until the output of IC_4 is at 0 V.

Preset P_1 determines the sensitivity of the VCO. This sensitivity, in turn, determines the full-scale measurement value, which may lie between 500 µA and 1000 µA for each sensor. The prototype was set for 700 µA, which corresponds to an illumination of about 100 lux. This value is achieved by setting the amplification of IC_4 to $^{10}/_7$ times (= variable 'Au' in the control program).

The adjustment is fairly simple: apply 1 V to the input of the opamp and adjust P_1 until the output voltage is 1.429 V.

Next, adjust the VCO. Remove jumper J_2 and apply –10 V from an external supply to the VCO input resistor, R_{10}. Adjust P_3 and P_4 until an output frequency of 100 kHz is obtained at terminal F on connector K_2. In some cases, this requires a 10-kΩ resistor to be fitted between the positive supply line and the collector of T_1. This adjustment results in a VCO sensitivity of 10 kHz/V and completes the setting-up procedure.

The control program

Insert the EPROM (order number 5921) with the control program into the relevant socket on the BASIC computer board, and connect the interface to the BASIC computer as shown in Fig. 4. Apply power, and point D_0 to the north. Press the space bar on the external terminal hooked up to the serial port of the computer. Wait for the 'ready' message and type RUN. The control program prompts you to enter the time zone with respect to Greenwich Mean Time (GMT), the latitude and longitude of your location (consult an atlas!) and, finally, the height above or below sea level. After accepting the geographical data, the program asks you whether or not summer/winter (daylight savings) time is used. The last prompts are the local time and the date.

The calculations and other program operations that follow the data entry phase take a minute or two. On their completion, the system is ready to start the actual measurements. Since it makes no sense to measure at night, the recording period covers the time between 30 minutes before sunrise and 30 minutes after sunset. Nighttime readings are simply not processed. The instructions on the terminal tell you how to retrieve measured data.

As already stated, the terminal may be replaced by a PC running a communications program. A log and a file store function are valuable in that case. One program that was found well suited to the application is Procomm. Alternatives to a full-blown PC include the hand-held Psion Organiser shown in the introductory photograph. ■

References:

1. BASIC computer. *Elektor Electronics* November 1987.

2. Peripheral modules for BASIC computer. *Elektor Electronics* October 1988.

3. Calculating the position of the sun. *Solar Energy* Vol. 20; p. 393–197. By R. Walraven, Dept. of Land, Air and Water Resources, University of California, Davis, CA 95616, U.S.A.

SCIENCE & TECHNOLOGY

'INTELLIGENT' ROAD SYSTEMS

by Professor Peter Hills, Director of the Transport Operations Research Group, University of Newcastle upon Tyne

In spite of periodic oil 'crises' and fitful bouts of environmental concern, in almost every country the dominance of road vehicles over other forms of transport cannot be denied. Traffic congestion in Europe is set to go on rising well into the next century. In the UK, latest revisions to official forecasts by the Department of Transport, published early in 1989, acknowledge that the persistent upward trend of car ownership could double the amount of road traffic within 20 years.

Building roads can not of itself prevent the spread of congestion; indeed, it is argued that building more roads in cities may often make matters worse. Even if the obvious response to rising demand for road space were to be increased supply, it could not be provided fast enough to catch up with the predicted growth of traffic. Moreover, vehicle users have a poor perception of the actual costs that their use of the congested road system is causing to others while significant components of these costs, notably pollution and accidents, are borne by the community at large. This inefficiency of pricing is compounded by the inefficiency of routeing because, at the outset of many journeys, drivers have only a hazy idea of what traffic levels they are likely to meet. A surprisingly high proportion of them do not kn ow the shortest route to their destination, let alone the quickest; most decisions about journeys are based on a mixture of hunch and hearsay. To add to all this, finding somewhere to park at the destination will get progressively harder as congestion grows.

In spite of this gloomy prediction, one way forward lies in the harnessing of information by making the road system more than just the infrastructure for vehicles to reach their destinations. Making it also a reliable source of up-to-date information about the service it can provide, given that traffic conditions are changing all the time, requires that each driver should be able to communicate with the road system in an 'intelligent' two-way conversation. That would improve efficiency through better decisions about journeys, better traffic control, fewer accidents and less pollution. I shall go on to describe work on a technological means of attaining this two-way communication between moving vehicles and the roadside.

The DRIVE programme

Launched in January 1989, the European Commission's DRIVE programme is one of imaginative and wide-ranging research, costing some £40 million over three years. It brings together, in 73 different consortia, multidisciplinary groups from universities, government agencies, industry and potential users to explore ways of applying information technology to the problems of road traffic. A further programme, DRIVE II, to run from 1992 to 1997, is being planned to develop the ideas and apply them in practice. It is called Advanced Road Transport Telematics and will compete for funds with other research programmes in the EC.

One of the consortia in the present programme is co-ordinated from Newcastle. It comprises the University, jointly with the Polytechnic, three separate Philips companies (in the UK, Sweden and Federal Germany) and two potential users, namely Compagnie de Signaux et Equipe-ments Electroniques (CSEE) in France and Empresa de Investigaçao e Desenvolvimento de Electronica (EID) in Portugal. The consortium is working on an automatic two-way communications system for traffic

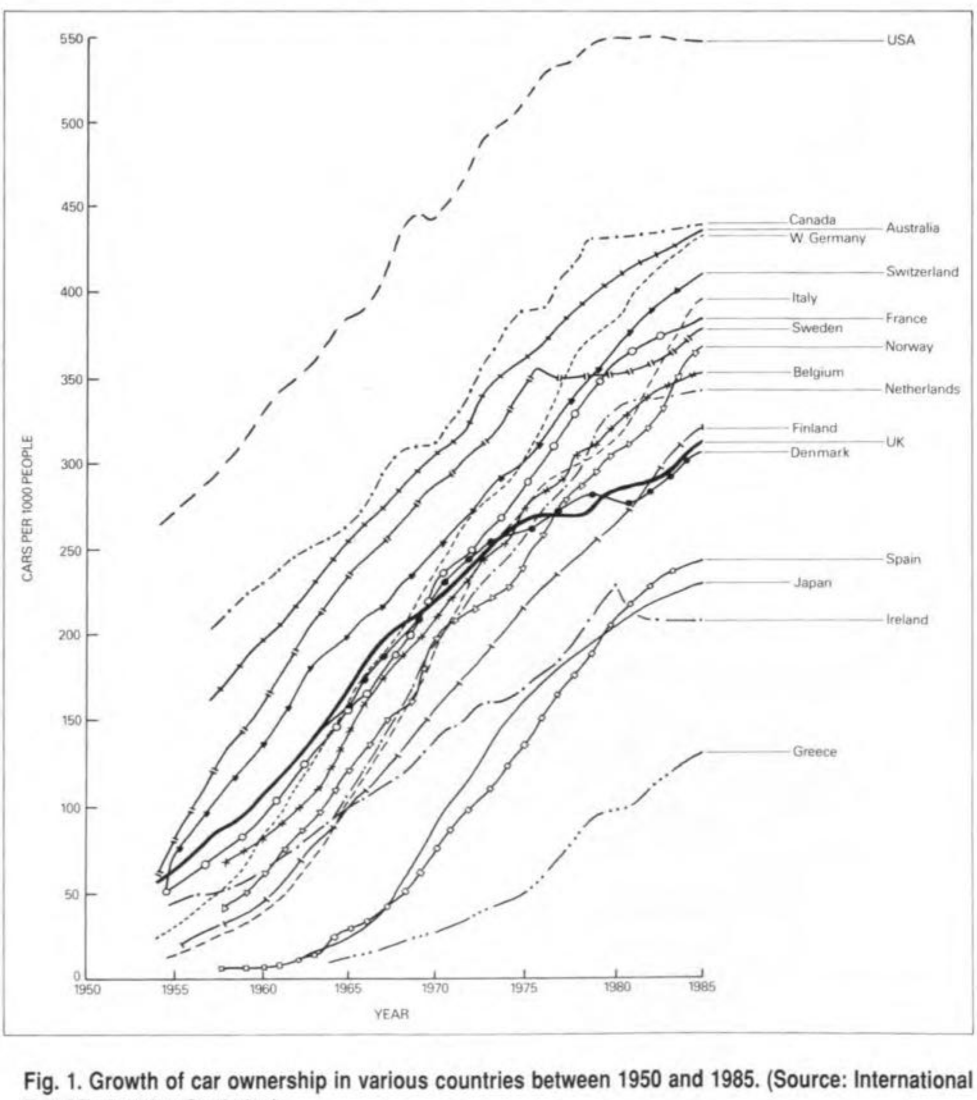


Fig. 1. Growth of car ownership in various countries between 1950 and 1985. (Source: International Road Federation Statistics).

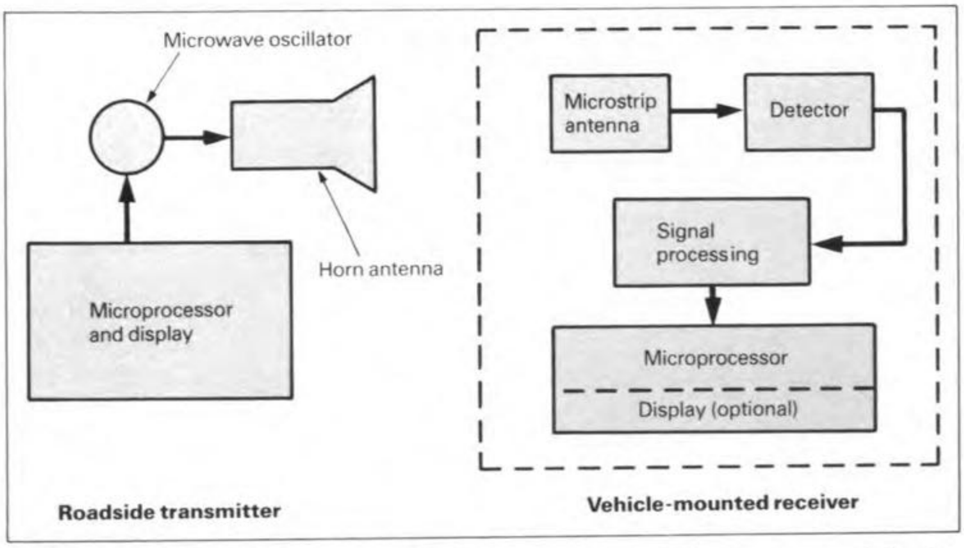


Fig. 2. Block diagram of an overall roadside-to-vehicle data communications system now being developed under the European DRIVE programme.

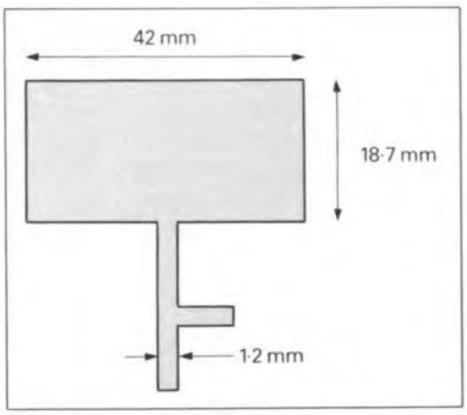


Fig. 3. Rectangular microstrip antenna with stub matched feeder.

monitoring and pricing, with the aid of microwave technology. The intention is to apply the system to the automation of motorway tolls in France and to a parking information and management scheme in Lisbon. The following is a description of the work.

Microwave system

The aims of the research work at Newcastle have been to investigate the feasibility of using a low-energy microwave link for two-way data communications between a short-range transmitting beacon at the roadside and moving vehicles, and to develop and test the equipment necessary for the system.

The transponder unit for the vehicle was developed in microstrip circuit technology to make all the microwave-frequency components needed. Microstrip technology is highly suited to traffic applications because the circuits have a low profile, are small and may be made cheaply. The associated logic and auxiliary circuits for the vehicle unit will be made compatible in size and profile with the transponder microstrip circuitry by the use of custom-built very large scale integrated (VLSI) circuitss. This has not yet been done, so the signal processing and logic circuits described are built from discrete components.

The roadside transmitting unit has been built from off-the-shelf components. It is battery powered, small enough to be portable and it can be set up at the roadside on, say, a lamp-post or gantry.

Field trials to find out how the transponder performs when mounted in a vehicle have been very encouraging. Substantial amounts of error-free data have been received by the device when mounted either on the windscreen or on the side window of a vehicle moving at up to 40 km/h. More field trials are planned for this year at vehicle speeds up to 120 km/h.

Background to development

Work began at the University and the Polytechnic in September 1985, with two surveys of the literature to investigate (a) what systems were already available for traffic monitoring and control applications, and (b) the various types of microstrip antenna that might be used and the methods of analysis used to predict their performance.

Existing systems use a variety of techniques to communicate with vehicles, including inductive loops in the road surface and local radio broadcast systems. Most are restricted to a low data transfer rate of a few hundred bytes per second. This indicated the need of a system with a high data rate. At the proposed 2.45 GHz operating frequency of the system under development, it is expected that the microwave communications link can transfer several thousand bytes of data from the roadside to a passing vehicle.

Figure 2 shows a block diagram of the component parts of the roadside unit and the transponder. The vehicle-mounted transponder under development comprises four distinct components: an antenna; a detector with associated circuits; a signal processing unit; and a microprocessor and display unit. The prototype roadside unit was built from off-the-shelf components

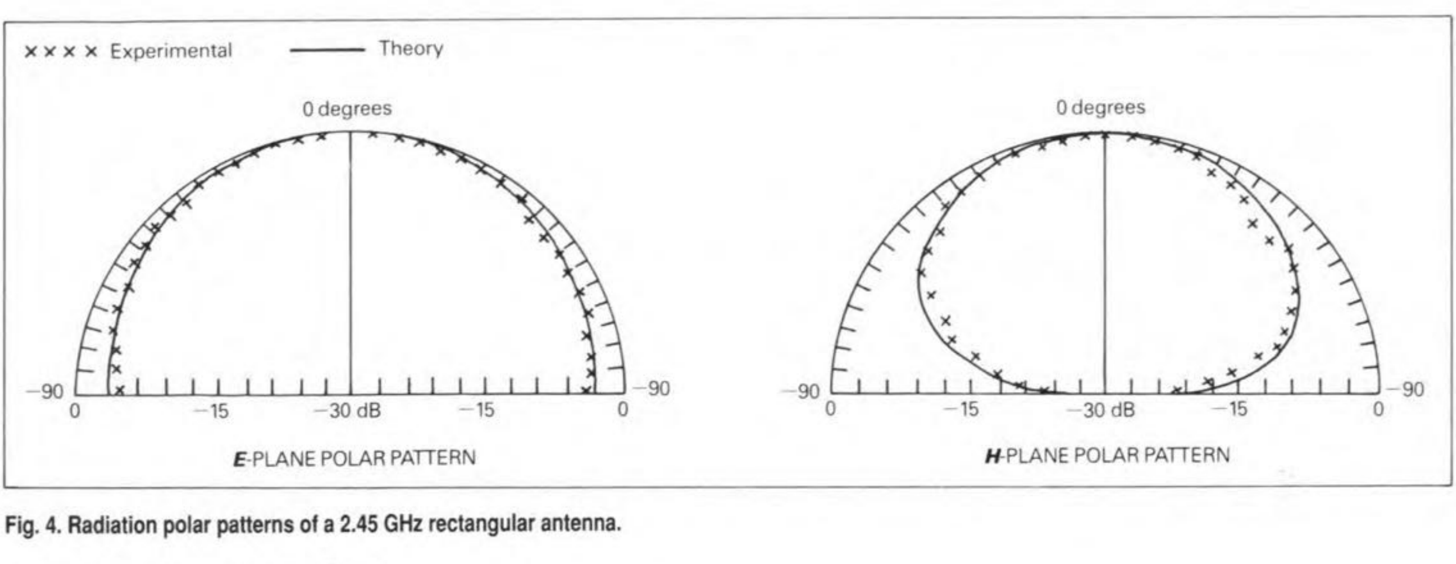


Fig. 4. Radiation polar patterns of a 2.45 GHz rectangular antenna.

and comprised a BBC microcomputer to generate the data to be transmitted over the microwave link, a microwave oscillator and modulator, and a pyramidal horn antenna.

Prior to transmission over the link, the digital data stream generated by the roadside microprocessor has to modulate the microwave-frequency carrier generated by the roadside unit. Several techniques of digital modulation were considered, including frequency shift keying (FSK), phase shift keying (PSK) and amplitude shift keying (ASK). Although FSK and PSK have better performances in terms of bit error-rate, ASK was chosen for the prototype system because it simplified the design of the transponder demodulator.

Vehicle transponder

Microstrip circuit technology has been used for all the microwave-frequency components of the vehicle transponder circuit. Microstrip circuits are made by a photo-etching process, similar to that used in the manufacture of printed-circuit boards. A conducting pattern is etched on the upper surface of an insulating substrate, about 1–2 mm thick, with a metallic ground-plane on its lower surface. The microstrip structure may be made cheaply; because of the low profile and small size of the circuits it is very attractive for vehicle-mounted applications where low cost and small size are at a premium. It is expected that the final version of the vehicle unit will use a single custom-built VLSI circuit to perform all the signal processing and logic functions, thereby maintaining the low profile of the transponder. The prototype, however, used discrete components for these circuits.

Several geometric shapes were considered for the microstrip antenna, including circular ring, annular ring and rectangular. Theoretical and experimental study showed that their performance characteristics were similar. We selected a rectangular shaped antenna because it could be fed from a microstrip feeder line, thereby retaining the low profile of the circuit. To ensure maximum energy transfer to or from an antenna via its feeder is is necessary to match the characteristic impedance of the feeder to the impedance of the antenna itself. This was done in two ways: the first by a quarter-wave matching transformer, and the other by stub matching. In this, a short length, or stub, of feeder is attached to the feeder itself and is short-circuited at its far end. By adjusting the point of attachment of the stub to the feeder and the length of the stub itself, it is possible to cancel out any effects of a mismatch and thus effectively match the feeder to the antenna. Both methods of matching may be employed in practice with the use of microstrip lines, and it was found that any loss through mismatch was kept to not less than 25 dB below the signal level. Figure 3 shows the arrangement of using a stub on a 50 Ω microstrip feeder. The dimensions relate to an antenna operating at 2.54 GHz.

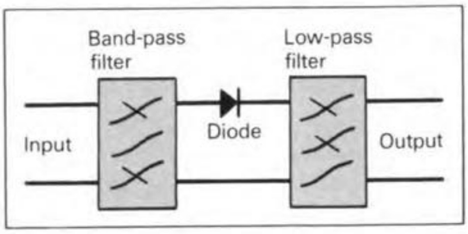


Fig. 5. Vehicle-borne diode detector circuit with filters.

A detector circuit is necessary to recover the digital data signal from the ASK modulated microwave signal received by the antenna. The circuit is shown in Fig. 5, where a Schottky barrier diode and filter network based on microstrip were employed. A band-pass filter was used at the input and a low-pass filter at the output to remove residual carrier and unwanted products generated by the diode in the demodulation process. The radio-frequency power from the receiving antenna may vary over the range 0.1 μW to 10 μW, depending on the distance from the transmitter, while the demodulated digital waveform output from the detector has a peak value of between 2 mV and 250 mV.

Processing and logic

In the final design, the signal processing and logic circuits will be put together in custom-built VLSI circuits. For testing the prototype, discrete components were used. Signal processing is needed to increase the voltage amplitude of the digital pulses from the detector and to reshape them to compensate for any distortion to the signal that may have taken place during transmission and demodulation. To do this, a direct-current stabilized amplifier and a level detection circuit were used.

To perform the logic functions on board the receiver, a microprocessor was used. A Type 6502 was chosen because of its compatibility with existing 'smart card' technology. The microprocessor may be a stand-alone item and perform all the processing needed at the receiver or, for more advanced applications, it could be interfaced with a more powerful microprocessor and display unit fitted in the vehicle.

Roadside beacon

As shown in Fig. 2, the roadside unit consists of a 6502 microprocessor, a microwave oscillator, a modulator and a horn antenna. The antenna has its radiation pattern shaped to cover transmission to either a single lane or a multi-lane road. The transmitter should be designed to feed several antennas to cover different approaches to a road junction, thereby reducing the cost of a large number of units. The microprocessor may work as an autonomous unit or be coupled to a central control computer governing all the beacons in a certain area. The prototype unit is fully portable, battery powered and measures 200×300×400 mm.

System performance

Field trials were carried out to test the performance of the transponder in a moving vehicle, mounted either in the side window or on the windscreen to find out the best place to fit it. Results showed that, while the vehicle is in range of the roadside beacon, substantial amounts of error-free data can be received. No interference was observed either from sources in the vehicle, such as the engine, or externally from other vehicles or objects blocking or reflecting the microwave transmissions.

At a speed of 19.2 kilobytes per second, over 1000 bytes of data were transferred to the vehicle. This amount is enough to send a small digitized map or real-time information relating to prevailing traffic and road conditions.

So far, the serial ports of the roadside and vehicle microprocessors have been used as the data input and output ports. This has restricted the data rate to 19.2 kilobits/second. However, if the processors' parallel ports were used to generate the data, much higher rates would be available to test the prototype system, though a parallel-to-series data conversion would be needed prior to the transmission or reception by the microprocessors. Preliminary tests have shown that error-free data can be received successfully at rates of up to 200 kilobits/second.

To improve the flexibility of the system, the present stage of development under the DRIVE Programme is to make the system into a full two-way data communications link that would enable a vehicle transponder to 'talk back' interactively to the roadside beacon. That would greatly enhance the system capability, especially in the fields of vehicle fleet control, automatic toll collection and route guidance. This in turn would herald the beginnings of an 'intelligent' road system capable of reducing congestion substantially. ■

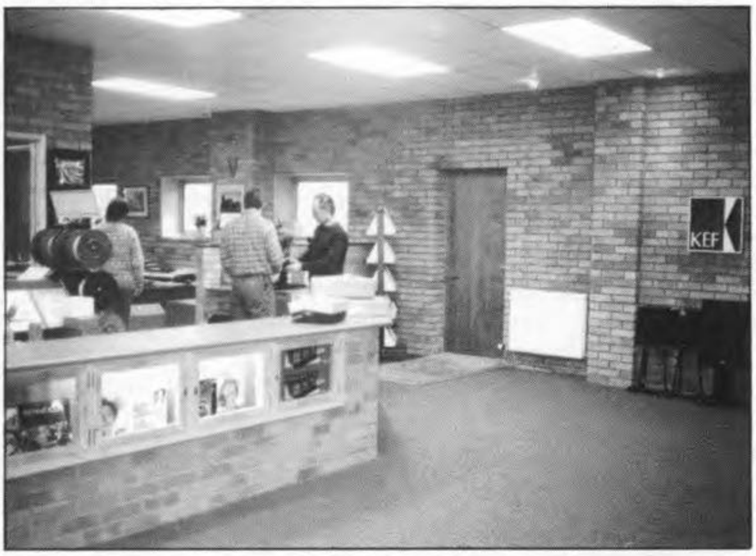
PROFILE: WILMSLOW AUDIO

by Bernard Hubbard

Wilmslow Audio of Knutsford in Cheshire offers one of the widest ranges of loudspeaker kits and drive units in the United Kingdom. Founded some thirty years ago, the company recently moved into new, 10,000 sq.ft. premises at Wellington Close, Parkgate Estate, Knutsford.

In the new showroom, Mike Aldington, one of the directors, told *Elektor Electronics*: "We offer very competitive prices and nowhere else are you likely to find the interest, the product knowledge and the sound advice that is based on so many years in this specialist market".

"In addition to loudspeaker kits and drive units, we offer a wide choice of hi-fi products and public-address equipment: all models selected for sound design and reliability and in keeping with the Wilmslow Audio philosophy of value for money".



Mike stresses: "Our orders come from all over the world. We have orders from Japan, France, Germany and the United States, to name but a few".

Drive units range in price from £4.95 to £350.

Most of the kits are created from bought-in components, but the company does manufacture cross-over networks.

The company dispatches the majority of orders on the same day of receipt of instructions, depending on the availability of the equipment, of course.

Mike continued: "DIY loudspeakers can be as good as ready-made units with considerable savings. We offer the most popular models in three forms".

- "The BASIC kit, which includes all drive units and cross-over filters (although the filters are sometimes in easily assembled kit form). The BASIC kit usually provides all that is necessary if you're upgrading a pair of speaker cabinets".
- "The PLUS kit, which includes all the items in the BASIC kit, as well as wadding, terminals, T nuts and bolts, reflex tubes (where applicable), etc. This is the kit to buy if you're building the cabinets from your own materials".
- "The TOTAL kit, which includes all items in the PLUS kit, with the addition of cabinets in flat-pack form. These are accurately machined from smooth MDF board and are very easy to assemble. Baffle apertures, etc., are cut and rebated where applicable and it is a simple job to construct a pair of loudspeakers to a professional standard".

"Using the Wilsmlow Audio iron-on veneer," he says, "it is possible, with a little care and attention to detail, to achieve a finish that is at least the equal of commercially produced cabinets. Only the very best of ready-made loudspeakers come close to the solid construction achieved with the Wilmslow Audio TOTAL kit. Power handling of a loudspeaker is difficult to quantify and often misunderstood. While there is an obvious limit to the power handling of a particular design, do not worry too much about using an amplifier that has a nominal output that is greater than the quoted power handling of your speakers. In domestic use, you are more likely to damage the speaker with too small an amplifier than with one that appears to have too great an output".

"A high level of efficiency and sensitivity is not a measure of quality—it means that you get a lot of sound per watt of power, but that has no significance inasfar as the quality of the sound is concerned, although a low powered amplifier will more easily be driven into clipping when used with speakers of low efficiency. In general, the larger the cabinet, the more extended the bass response whatever cabinet design is employed. However, this does not mean that all speakers of the same size will have a similar bass extension as this results from several aspects of the individual speaker design. Do not confuse good bass response with extended bass response. Extended but 'boomy', ill-controlled bass is by no means as pleasant to listen to as a tightly controlled but less extended bass response. Room dimensions inhibit perceived bass response. A small, relatively insensitive speaker, such as our micro monitor design, will provide all the sound quality, volume and bass response that you could desire in a small room and yet would be quite incapable of providing the bass response, power handling and volume that you might require for a noisy party in a large room".

The company is looking forward to 1992, because it knows that the prices for such kits are high in continental Europe and once various restrictions have been removed and the barriers are down, Wilmslow Audio believe that they are in for an even rosier business than they are currently enjoying. Whatever, the company and its team of six people are very much in sound heart. ■

Wilmslow Audio's advertisement appears on page 9 of this issue.

ACL: ADVANCED CMOS LOGIC

F.P. Zantis

Almost ten years ago, the CMOS logic technology took a giant step forward with the introduction of the HC and HCT series. However, in the intervening years, it has become clear that even these fast logic devices can not replace all types of bipolar (TTL) circuit. The introduction of the fourth CMOS generation, Advanced CMOS Logic or ACL, means that now even the fastest TTL logic circuits, ALS, may be replaced by CMOS devices.

For the efficient operation of fast bus and transfer systems, it is essential that short transit times are combined with a large fan-out. This was not possible with CMOS logic until the arrival of the ACL series. This new CMOS technology also takes us a step further on the way to removing the disparity between high switching speeds and low power dissipation. Another drawback of HC/HCT circuits is their sensitivity to electrostatic charges: ACL devices have all but lost this sensitivity.

The essential characteristics of ACL, TTL, HC/HCT, ECL, and I^2L logic circuits are compared in Table 1.

	Power dissipation at f = 1 MHz	Transit time	Signal/noise ratio (Ub = 5 V)	Integration density
TTL	medium	short	small	medium
HC/HCT	small	long	large	very high
ECL	high	very short	very small	small
I^2L	very small	medium		very high
ACL	small	very short	large	very high

Table 1. A comparison of the most important families of logic circuits.

ACL technology

ACL devices are manufactured in time-proven CMOS technology. The thickness of a single wafer has been reduced to 1 µm, however. A number of production improvements make it possible for certain properties of the circuits to be optimized. For instance, the transfer resistance of not only the output terminals, but also of the internal junctions, has been reduced significantly. Moreover, the relatively large parasitic capacitances found in HC/HCT devices are virtually absent from ACL circuits. All these improvements make it possible for ACL circuits to be used in applications that until now were the preserve of TTL logic.

Characteristic data

The transit delay in ACL chips is about 3 ns per gate, so that, in theory at least, operating frequencies of up to 160 MHz are possible. The maximum permissible output current is 24 mA per gate. It is of no relevance whether the gate functions as a source or as a sink. This level of current is high enough to enable direct driving of high-gain transistors, such as darlingtons.

Low power dissipation is of great importance in many circuits, because it allows greater packing density, extends the duty cycle in battery operation, and increases the reliability of the system because of the lower temperature.

Thanks to CMOS technology, the power dissipation in ACL circuits is very low: in quiescent operation, it is only a few nW per gate, which rises to 1–2 mW during operation at 1 MHz. As an example, Fig. 1 shows the relation between dissipation and operating frequency for an ACL11008 and a similar TTL circuit (four AND gates).

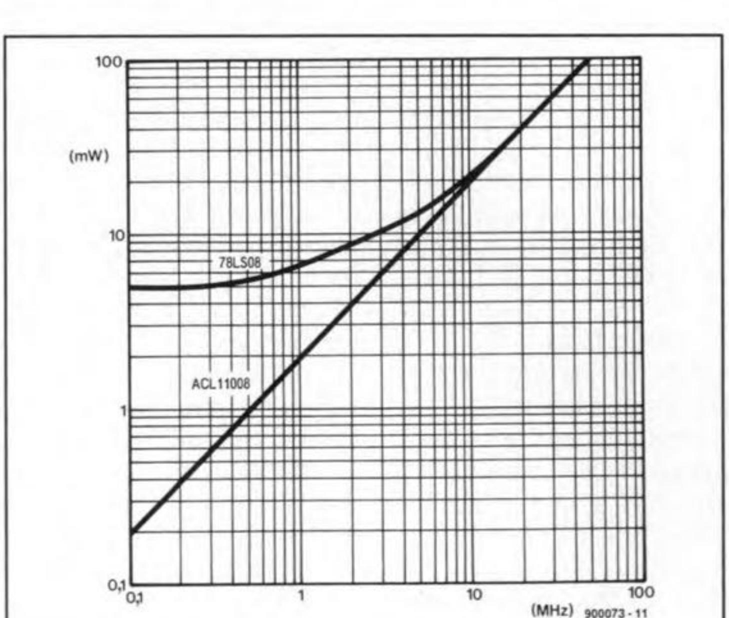


Fig. 1. Power dissipation in an ACL 11008 (4 AND gates) and a 74LS08. All outputs switched simultaneously. Ambient temperature = 25 °C. Supply voltage = 5 V. Output load = 50 pF.

The signal-to-noise ratio is also very good. The output varies from 0.1 V (low level) to U_b – 0.1 V (high level) at a load current of 50 µA (equivalent to 50 CMOS inputs). At the full load current of 24 mA, the low level is 0.5 V and the high level, U_b – 0.8 V.

The change-over thresholds at the inputs are type-dependent. In CMOS compatible devices, they lie between 30% and 70% of the supply voltage. In TTL compatible circuits, they are 0.8 V (low level) and 2 V (high level).

The operating temperature of standard ACL versions extends from –40 °C to +55 °C; types with an extended range of –55 °C to +125 °C are also available.

Circuit design

Very fast logic devices switch with very steep edges and this creates a few difficulties. For instance, parasitic reactances and line reflections in the circuit may cause functional interference. To obviate such difficulties with ACL chips, no attempt was made to achieve pin compatibility with HC/HCT or TTL devices. Instead, ACL circuits are designed in flow-through architecture, which

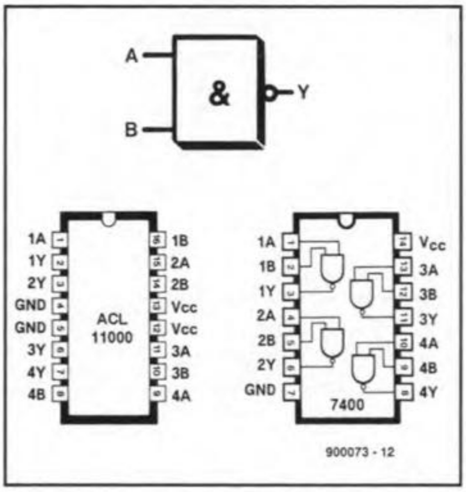


Fig. 2. Comparison of the pin-outs of an ACL circuit and a similar TTL chip.

means that the internal structure, as well as the pin-out, is as much as feasible in line with the signal flow. This results in a number of supply connections being paralleled, inputs and outputs of gates being well isolated, and the less frequently used control connections being grouped at the upper or lower part of the package. These measures result in an average reduction of the inductive layer in and around the chip by a factor of 3.8 in standard DIL-packaged circuits and higher in other housings.

When ACL devices are used in bus-oriented systems, notable simplications are possible in the wiring of the printed-circuit board. However, the great advantages of the unorthodox pin-out become really conspicuous when ACL circuits are used in conjunction with multilayer PCBs.

Figure 2 gives a comparision between the pinouts of a Type ACT 11000 ACL circuit and a TTL chip Type 7400. Note that the ACL device needs two more pins.

The pin-outs of a number of important ACL circuits in a standard DIL package are shown in Fig. 3.

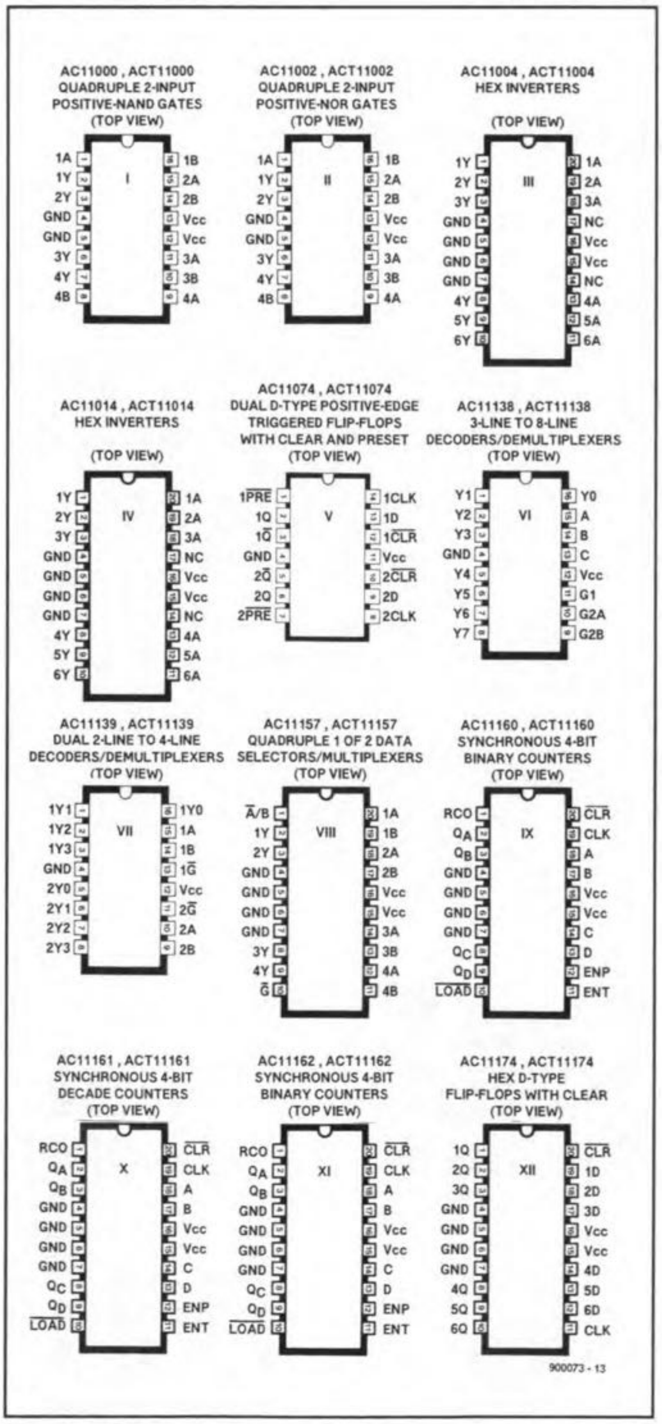


Fig. 3. Pin-outs of a number of important ACL devices.

Type coding

ACL circuits are available in two different versions:

1. The 74AC series with CMOS compatible high/low thresholds ($U_t = U_b / 2$) for operation from supply voltages of 2–5.5 V.
2. The 74ACT series with TTL compatible high/low thresholds (U_t = 1.5 V) for operation from 4.5–5.5 V.

The type coding consists of six groups, the meaning of each of which is shown in Table 2. Note that group 5 defines the logic function in the same way as for TTL and HC/HCT circuits. Also, group 4 is always 11, indicating that the supply connections are located at the centre of the housing (and thus there is no pin compatibility with HC/HCT or TTL devices).

Protection against electrostatic charges

All switching inputs and outputs are protected against static charges. This holds good for potentials of whatever polarity at levels up to 2 kV. The protection circuits, consisting of two diodes connected as shown in Fig. 5, are equivalent to a 100 pF capacitor being earthed via a 1.5 kΩ resistor across the input or output. Proper operation of the protection circuit may therefore be verified by the circuit shown in Fig. 4.

The difficulty with efficient protection is that the protection elements, integrated on the chip, can, owing to their small size, cope only with relatively low levels of energy. In ACL circuits, taking the values stated earlier on, protection is guaranteed up to an energy, E, of

$$E = CU^2/2 = 10^{-10} \times 4 \times 10^6/2 = 0.2 \text{ mJ}$$

The diodes in the protection circuits have a forward foltage of about 0.9 V and a reverse (zener) voltage of around 17 V.

Group 1 —	blank = standard version SNJ = meets MIL STD 883 JANB = meets MIL STD 38510
Group 2 —	74 = temperature range of –40 °C to +85 °C 54 = temperature range of –55 °C to +125 °C
Group 3 —	AC = CMOS compatible thresholds ACT = TTL compatible thresholds
Group 4 —	11 = supply connection in centre
Group 5 —	conventional functional type coding, e.g.: 000 = quadruple NAND gate 074 = dual D-type bistable
Group 6 —	N = plastic DIL case NT = plastic slim-line DIL case J = ceramic DIL case JT = ceramic slim-line DIL case D = plastic miniature case

Table. 2. ACL circuits are type-coded by these six groups.

Summary

The ACL series gives the designer the opportunity of putting to use all the advantages of CMOS technology, such as high signal-to-noise ratio and small power dissipation, at fairly high operating frequencies.

The pin-out is unorthodox and not compatible with either TTL or HC/HCT circuits. It is, however, of paramount importance in realizing the excellent properties of the ACL devices.

ACL circuits may replace TTL as well as HC/HCT chips, and are already widely used in new designs. There is no doubt that ACL technology will become the new standard for integrated digital circuits. ■

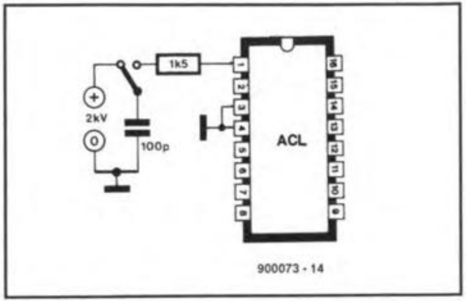


Fig. 4. Test circuit to verify the proper operation of the protection circuits in an ACL device.

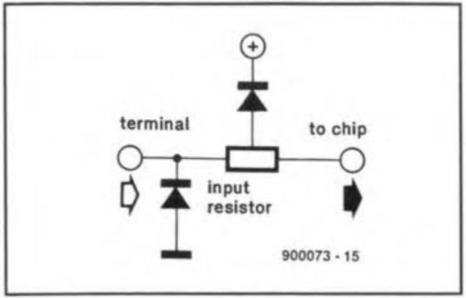


Fig. 5. The protection diodes in an ACL circuit are connected as shown here.

R–2R RESISTANCE NETWORK IN SMT

H. Bierwith

The SIL *R–2R* network intended for use in, say, a digital-to-analogue converter, is constructed in surface-mount technology (SMT). It is inexpensive, precise, space-saving and allows unusual values to be obtained. It is described here in a digital-to-analogue converter that uses a CMOS latch Type 4042.

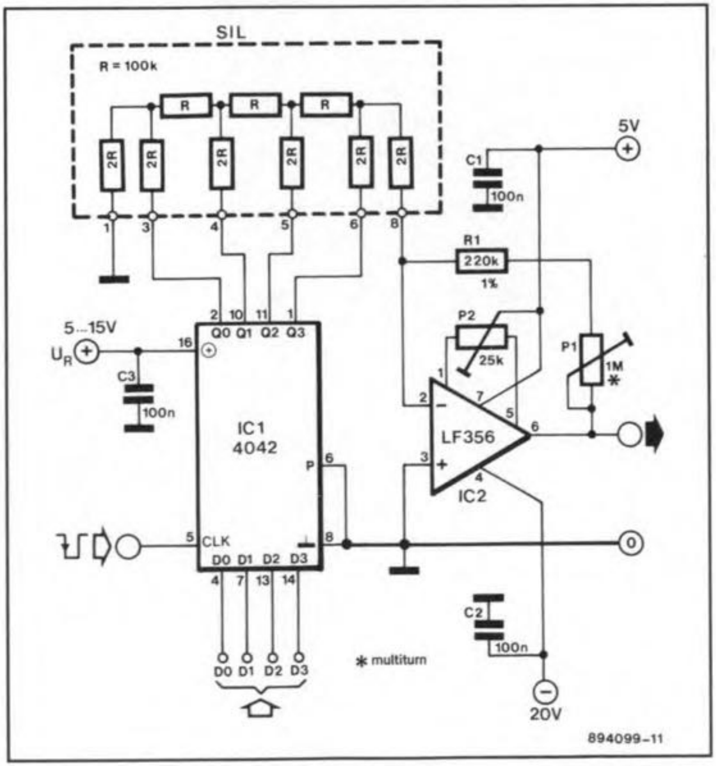
The latch and the associated *R–2R* network are connected in the feedback loop of an LF356. The network uses 100 kΩ resistors (the 2*R* resistors are made up of two series-connected 100 kΩ resistors).

The output voltage of the converter is calculated from:

$$U_o = U_r(R_1+P_1)/6R(2^0Q_3+2^{-1}Q_2+2^{-2}Q_1+2^{-3}Q_0)$$

The Qs in the formula have a value of 0 or 1, depending on the state of the latch output. The factor $(R_1+P_1)/6R$ is the amplification, *A*, which may be set between 0.4 and 2.0 by P_1.

The supply voltage for the latch, which is also the reference voltage for the network, U_r, may be between 5 V and 15 V: the logic levels at latch inputs D0–D3 and at CLK must be in accord with that voltage. The reference voltage must be decoupled direct at the input of the IC.



The circuit is calibrated by making the inputs of the latch 0 and adjusting P2 for an output of 0 V from the opamp. Then, load the data inputs with F_{hex} and adjust P_1 to obtain maximum output voltage.

If a larger number of bits is to be processed, two or more PCBs may be connected back-to-back.

Moreover, the PCB may be used with other components, such as diodes, capacitors, combinations of resistors and diodes or simple potential dividers for measuring instruments.

The current drawn from the U_r supply is 75 μA; that from the 5 V supply is about 7.5 mA ■