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Elektor Electronics

I/O interface for IBM PCs

Laser - Part 1

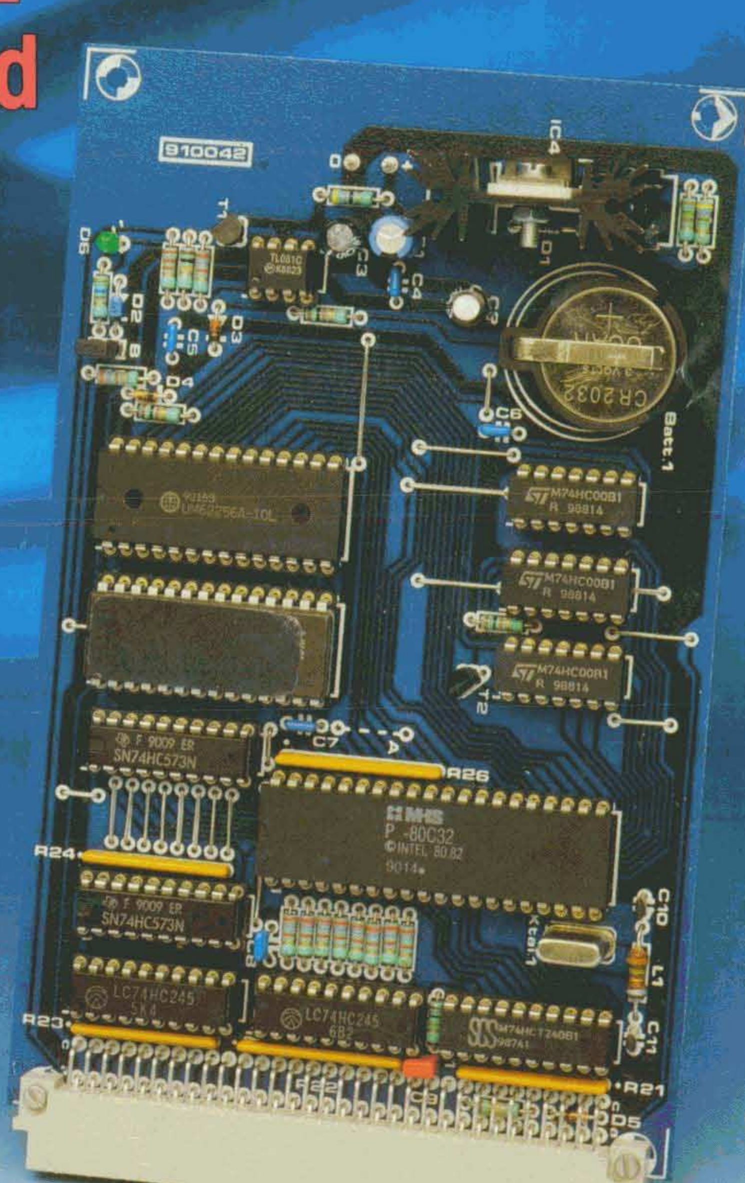
Battery tester

Conductance meter

Video A-D and D-A

Augmented A-matrices

8032/8052 single-board computer



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ELEKTOR Electronics
The international magazine for electronics enthusiasts

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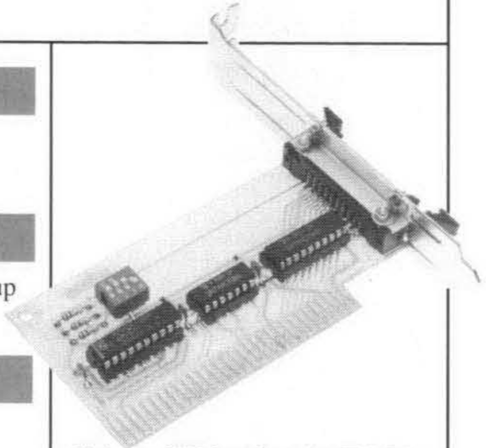
Front cover
Seen here is the Mark II version of the popular BASIC computer we published in November 1987. The Mark II is a more powerful system with more RAM, more ROM, an on-board EPROM programmer, and several other additional facilities on a single-sided Euro-size PCB.
The computer is ideally suitable for small control applications and software development. It can work with Intel's powerful 8032, 80C32 or 8052AH–BASIC processor. The latter has an on–chip interpreter that allows you to program in BASIC with full access to machine code.



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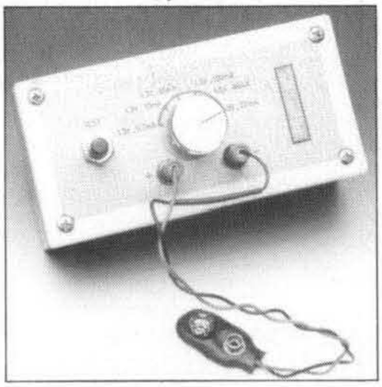

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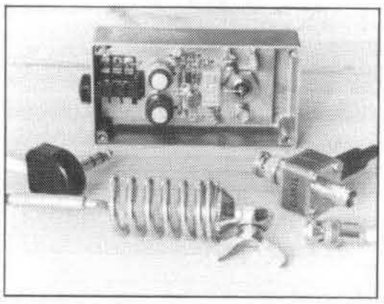

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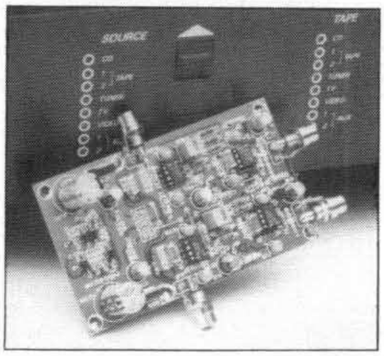

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SPEED CONTROL OF LARGE DC MOTORS

by K.A. Nigim, B.Sc., Ph.D., MIEE

The power circuit described in this article is used for obtaining a variable direct voltage source from the 50 Hz a.c. mains supply. The variable direct voltage is used, among other applications, for controlling the speed of large d.c. motors.

ALTHOUGH d.c. motors are more expensive than asynchronous a.c. motors, their simplicity and controllability of speed over a wide range puts them in a strong, competitive position for use as electrical drive units in a variety of industrial processing plants.

There are several types of d.c. motor, depending on how the motor windings are connected and excited. The most popular and easiest to handle is the separately excited type. In this, the magnetic field circuit is powered by a separate d.c. power source. This establishes the necessary magnetic flux, Φ, in the windings. When d.c. power is supplied to the armature windings (the coils are embedded in the rotating part of the motor), torque is developed and the motor starts to rotate.

To fully understand the behaviour of the motor when it is powered by different levels of direct voltage, the relationship between the produced torque, current and speed is simplified in the following way. The torque, T, developed by the armature windings when the supply is switched on is given by:

$$T = K_b \, \Phi \, I_a \quad [\mathrm{N}]$$

where K_b is a motor speed constant that depends on the number of turns the armature windings consist of, how the windings are wound, and the number of magnetic poles; Φ is the flux per magnetic pole produced by the field windings; I_a is the current (in amperes) flowing in the armature windings and N is the symbol for newton.

The current I_a produces a back electromotive force, E_b, which is given by:

$$E_b = K_b \, \Phi N \quad [\mathrm{V}]$$

in which N is the steady-state motor speed.

The performance of a motor may be analysed with the aid of its equivalent circuit shown in Fig. 1. Note that the armature resistance, R_a, is in series with E_b, because, although the motor has a high inductance, this is ignored during steady-state motor operations.

The field voltage, V_a, applied to the ar-

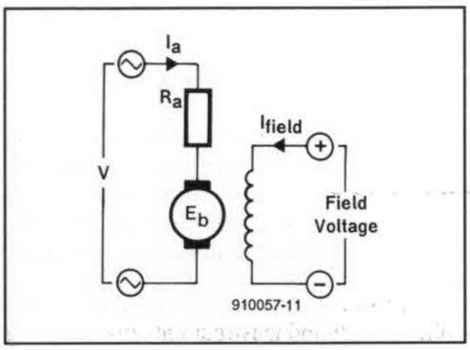


Figure 1.

mature windings is balanced by the sum of the voltage drop I_aR_a and E_b:

$$V_a = I_aR_a + E_b$$
$$= I_aR_a + K_b \, \Phi N$$

from which

$$N = (V_a - I_aR_a)/K_b \, \Phi.$$

This is the motor speed which, when ohmic losses are small, may be approximated for practical purposes to

$$N = V_a/K_b \, \Phi.$$

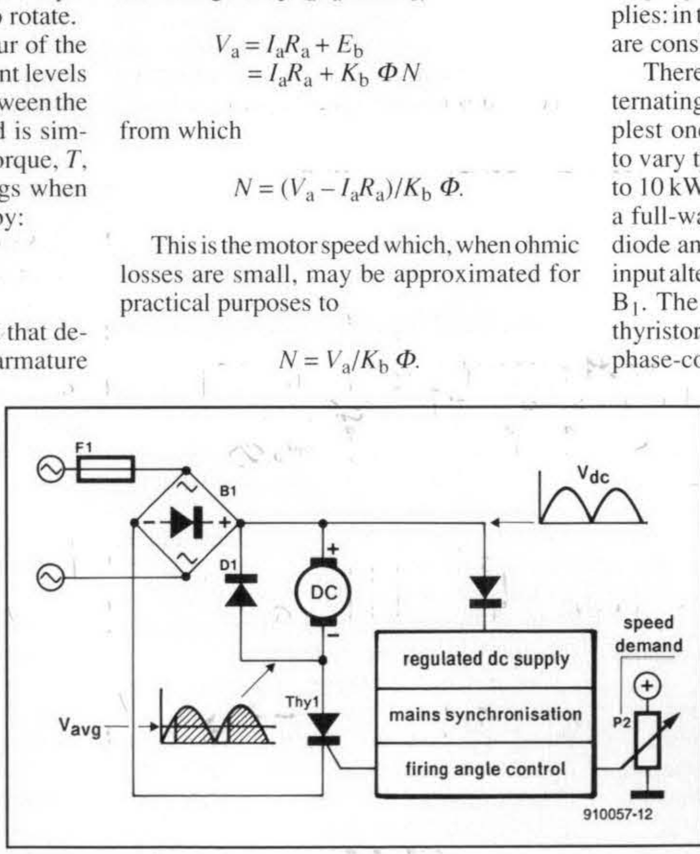


Figure 2

From this equation, it is seen that two methods of speed control are possible. The first one is to alter the supply voltage, V_a. The speed may then varied from zero to its maximum rating. The second method is to vary the magnetic flux, Φ, produced by the magnetic circuit. This method is restricted since the speed range will always be above the specified rated speed, because the magnetic field can not be made any stronger, only weaker. Furthermore, failure of the control circuit would cause very high speeds which might be catastrophic. Generally, therefore, the first method is used to maintain a constant output torque over the entire speed range, while the second method is used where constant shaft power with speed is required.

Variable d.c. power source

Variable d.c. power may be derived from single-phase or three-phase a.c. mains supplies: in this article only single-phase sources are considered.

There are various ways of converting alternating voltage to direct voltage; the simplest one that can be adopted successfully to vary the speed of a d.c. motor rated at up to 10 kW is shown in Fig. 2. The circuit uses a full-wave rectifier bridge, a freewheeling diode and a thyristor. Each half cycle of the input alternating voltage is rectified by bridge B_1. The rectified voltage is controlled by thyristor Thy_1. The thyristor is triggered by phase-controlled signals at its gate. The firing signals are synchronized with the waveform of the mains voltage to provide the required firing delay angle, α. The thyristor is switched off every time the supply voltage goes through its zero crossing point.

The various levels of the average direct voltage across a resistive load and the motor are shown in Fig. 3. Two sets of delay angles are shown; one for high power or speed with a short delay angle and the second for low power or speed with a long delay angle. The relation between the produced voltage and the delay firing angle is given by

$$V_{avg} = E_{s(peak)} \, (1+\cos\alpha)\pi.$$

In practice, the motor can not be represented by a resistive load. As the motor rotates faster, a back e.m.f. is produced for constant magnetic flux. The armature wind-

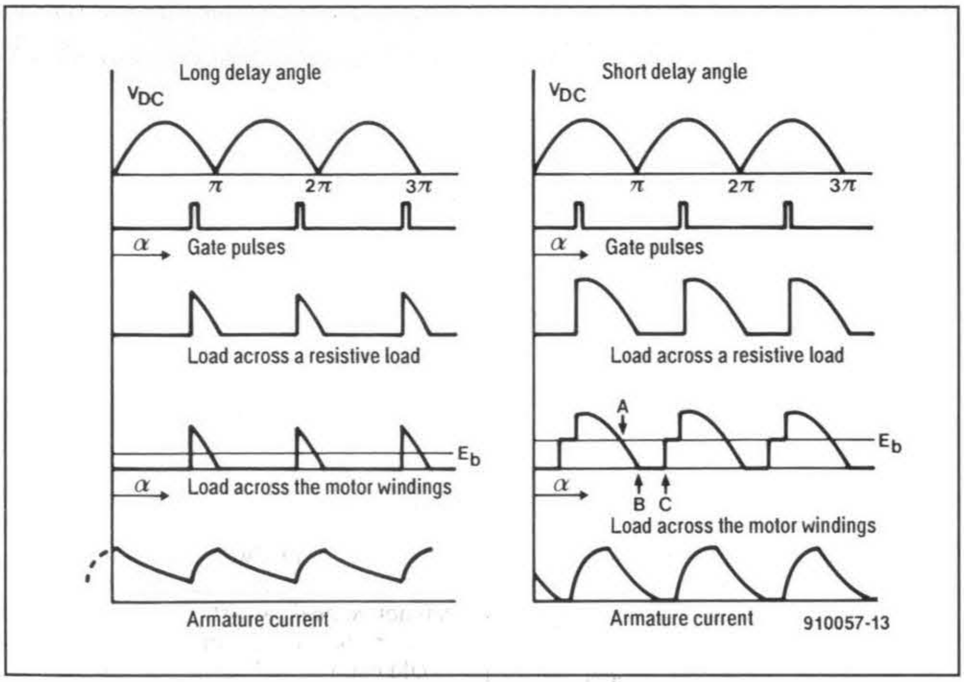


Figure 3.

ings also possess a large inductance and some resistance: the motor will thus truly behave as a complex load.

The reason for the motor current continuing to flow in spite of the supply voltage dropping below the back e.m.f., point A, is the presence of the armature inductance. This forces the voltage across the armature windings to follow the supply voltage until it becomes negative at point B. At that point, freewheeling diode D_1 diverts the current away from the thyristor which then switches off. When the inductive current flowing through the armature windings and D_1 has decayed to zero, point C, the freewheeling diode ceases to conduct and the voltage across the armature windings returns to the level of E_b. From this, it is clear that D_1 allows the circulation of stored energy and maintains a continuous load current. This is a healthy process, since a continuous flow of current through the load is essential for proper control of a d.c. motor.

Thyristor synchronized firing logic

The diagram of the electronic circuit that provides the required phase control (delay angles) over the range 0–180° is shown in Fig. 4. It may be divided into three sections: that producing the synchronizing signal, the comparator, and the gate drive. It is powered by a 12 V regulated power source based on a Type 7812 regulator (IC_3).

A portion of the rectified, and as yet unsmoothed, voltage at points A and B is applied to the base of transistor T_1. The waveform of these voltages is shown in Fig. 5. A pulse is produced at the collector of T_1, point C, every time the base voltage drops below a certain level. This pulse is integrated by the combination of amplifier IC_{1a}, capacitor C_1, and variable resistor P_1.

Transistor T_2 is switched on at the onset of each cyle and then short-circuits C_1 and IC_{1a}. This results in a sawtooth voltage at point D. This voltage is compared with a variable d.c. control signal in comparator IC_{1b}. The intersection of these two voltages determines the pulse at the output of the comparator, point F. The variable width of this pulse is, therefore, in step with the waveform of the mains voltage.

The controlled pulse is applied to monostable IC_2, which produces a narrow pulse (point G) at the leading edge of the pulse at the comparator output. The width of the pulse, t, at point G is determined by R_{10} and C_3 and is kept wide enough to sustain gate drive. The width is given by

$$t = 0.5R_{10}C_3. \quad [s]$$

An emitter follower, T_3, interfaces the output of the monostable and the gate of the thyristor. To avoid gate isolation, the zero voltage rail is tied to the negative rail of the power circuit shown in Fig. 2. Therefore, the cathode of the thyristor must be tied to the negative rail.

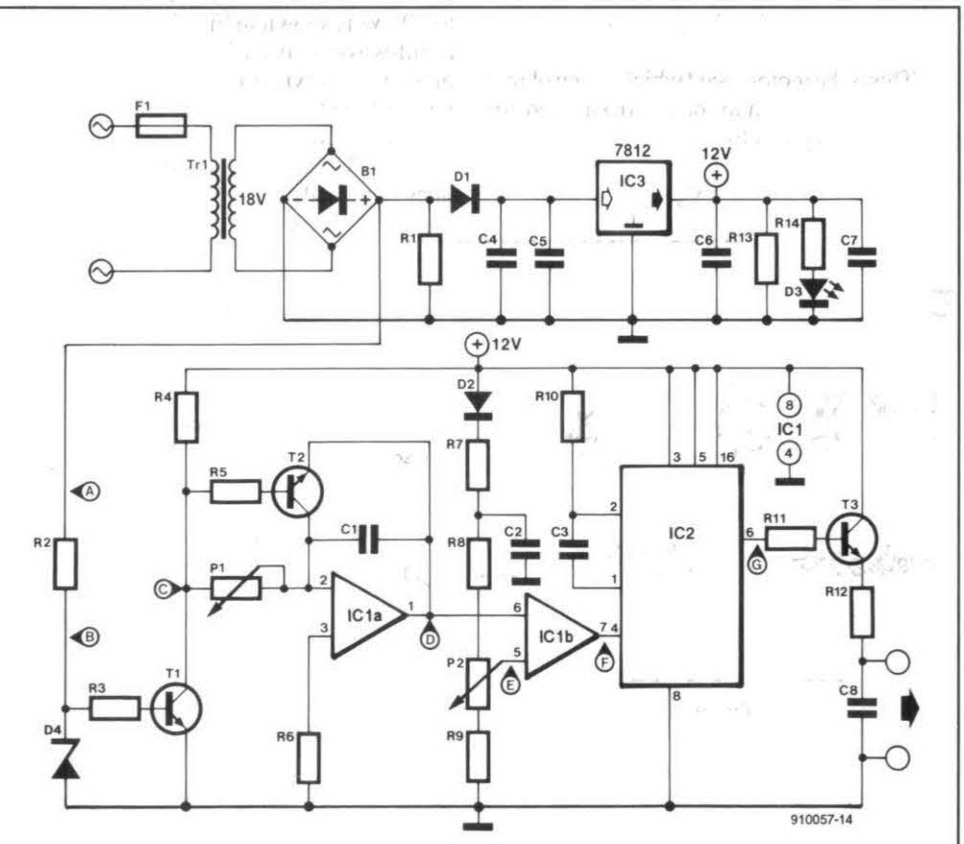


Figure 4.

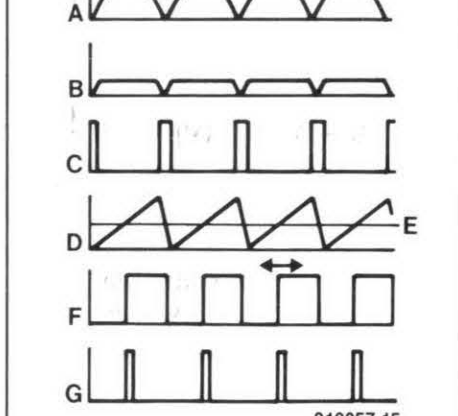


Figure 5.

Construction and testing

Firing control circuit.
The firing control circuit, including the 12 V regulated power supply, should be assembled preferably on a printed-circuit board (for which no design is offered) or on a suitable prototyping board.

Owing to the sensitive nature of the circuit, an oscilloscope is needed to check the relative position of the synchronized pulse with respect to the mains voltage.

Connect the channel X probe of the oscilloscope to test point A and the ground clip to the negative supply rail. With reference to Fig. 5, check the waveforms at points A, B, and C, with the channel Y probe. When the d.c. level at point E is changed, a rectangular wave with variable width should be obtained at point F. At point G, a 2 ms wide pulse should be produced at the leading edge of the pulse at F. Changing the d.c. level with P_2 should shift the pulse over the range 0–180° with respect to the waveform at A.

Network R_7-C_2 across P_2-R_8-R_9 ensures a soft start of the control voltage applied to IC_{1b}, and sets the minimum and maximum delay angles, thus determining the minimum and maximum speeds of the motor.

Connect a 100–500 Ω resistor between the emitter of T_3 and the negative rail to check that an output pulse is produced.

Network R_{12}-C_8 provides current and dv/dt gate protection.

Power circuit and motor.
The ratings of bridge B_1, diode D_1 and the thyristor depend on the load current. If, for instance, it is required to control a 3.75 kW d.c. motor, the current rating for a mains voltage of 240 V is close to 16 A. Both the diode and thyristor, mounted on a suitable heat sink, should then be rated at 35 A.

To test the power circuit, connect the cathode of the thyristor to the negative rail of the bridge rectifier. To play safe, first connect a 100 W light bulb instead of the motor windings across D_1. Connect the output of the control circuit to the thyristor gate and the cathode of the thyristor to the negative rail. When the delay angle is varied from minimum to maximum, the light bulb should behave exactly as if it were controlled by a light dimmer.

Next, connect the armature windings of the motor (low resistance on a multimeter) across D_1. Before switching the supply to the motor, make sure that there is a voltage supply to the field windings (high resistance on a multimeter). Refer to the motor name plate for the rating of the field circuit. When the supply to the motor is switched on and the motor rotates in the wrong direction, switch off and swap the connections to the armature windings.

Finally, protect the thyristor by connecting a series *RC* network between its anode and cathode. Suitable values are $R = 30\ \Omega$ (3 W) and $C = 0.1\ \mu F$ (1000 V). ■

COMPONENTS LIST (FIG. 4)

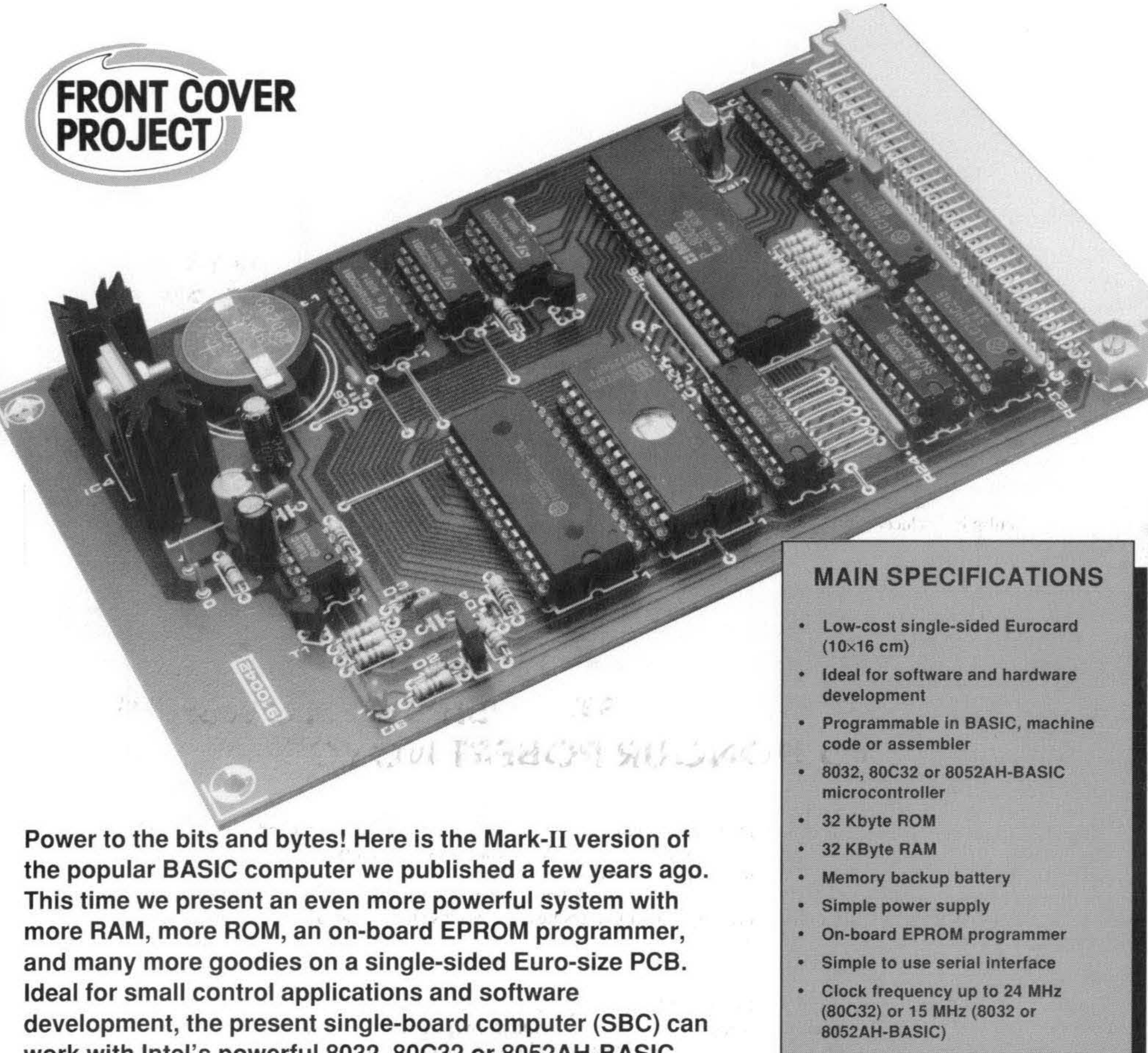
Resistors:
R1, R4 = 10 kΩ
R2 = 470 Ω, 0.5 W
R3, R5, R14 = 1 kΩ
R6 = 22 kΩ
R7 = 47 kΩ
R8 = 3.3 kΩ
R9 = 2.7 kΩ
R10 = 15 kΩ
R11, R13 = 6.8 kΩ
R12 = 100 Ω
P1 = 10 kΩ cermet potentiometer
P2 = 5 kΩ multiturn potentiometer

Capacitors:
C1 = 1 µF, 63 V
C2, C8 = 0.1 µF, 100 V polypropylene
C3 = 0.15 µF polypropylene
C4 = 2200 µF, 25 V, electrolytic
C5, C6 = 22 nF, ceramic
C7 = 10 µF, 35 V, tantalum

Semiconductors:
B1 = bridge rectifier, 200 V, 0.5 A
D1, D2 = 1N4001
D3 = LED, green
D4 = zener diode, 4.7 V
T1, T2 = BC107
T3 = BFY51
IC1 = LM358
IC2 = CD4098
IC3 = 7812

8032/8052 SINGLE-BOARD COMPUTER

FRONT COVER PROJECT



MAIN SPECIFICATIONS

- Low-cost single-sided Eurocard (10×16 cm)
- Ideal for software and hardware development
- Programmable in BASIC, machine code or assembler
- 8032, 80C32 or 8052AH-BASIC microcontroller
- 32 Kbyte ROM
- 32 KByte RAM
- Memory backup battery
- Simple power supply
- On-board EPROM programmer
- Simple to use serial interface
- Clock frequency up to 24 MHz (80C32) or 15 MHz (8032 or 8052AH-BASIC)
- Low-cost a-c row 64-way DIN connector for serial interface and hardware extensions

Power to the bits and bytes! Here is the Mark-II version of the popular BASIC computer we published a few years ago. This time we present an even more powerful system with more RAM, more ROM, an on-board EPROM programmer, and many more goodies on a single-sided Euro-size PCB. Ideal for small control applications and software development, the present single-board computer (SBC) can work with Intel's powerful 8032, 80C32 or 8052AH-BASIC processor. The latter has an on-chip interpreter that allows you to program in BASIC with full access to machine code.

H. Reelsen

MICROCONTROLLERS these days are silent workers in many apparatus, ranging from the washing machine to the video recorder, to mention but two examples in the home. Nearly all of these controllers are mask-programmed and therefore of very little use for hobby applications since the program they execute can not be altered. And even if we could alter the program, the information necessary to do so — an instruction set, an assembler language description and some basic hardware information — is often not available or very difficult to obtain. Also, the last resort of many programmers, a cross-assembler, is long sought but never found. In short, many microcontrollers, powerful as they may be, are not accessible (but bear in mind that they were not designed to be so).

An marked exception to the above 'misery' is the 8052AH-BASIC from Intel. This microcontroller has features that seem to make it more accessible than any other single-chip microcontroller with a reasonable price tag. Consider, for instance, its bit manipulation instructions, its internal BASIC interpreter, its ability to program

EPROMs, or load a BASIC program into memory via a three-wire RS-232C link.

In 1987 we published a single-board computer based on the 8052AH-BASIC (Ref. 1), and it has been a popular project ever since. The computer was not only built and programmed by vast numbers of computer enthusiasts, its design concept was also taken up commercially, witness the sudden influx of 8052-based controller boards and development systems following our publication.

The single-board computer described here is an upgraded version of the 8052AH-BASIC computer. The improvements are basically a larger RAM and ROM space of 32 KByte each, a memory backup circuit with 2 μA RAM data retention current, a 12.5-V EPROM programming voltage supply, and, last but not least, the possibility to use the ROM-less (inexpensive) 8032 or 80C32 microcontroller. A further boon for the home constructor is that the printed-circuit board for the present computer is single-sided and designed with 0.4-mm wide copper tracks for easy reproduction. This allows the cost of the computer to be kept to a minimum. If a double-sided board had been used, it would probably have cost more than all the components on it to build the 8032 version of the computer.

The circuit

The heart of the circuit shown in Fig. 1 is either the ROM-less 80(C)32 CPU or the 8052AH-BASIC processor. The LS (least-significant) group of address lines, A0-A7, is multiplexed with the data, and extracted with the aid of octal latch IC8 when the ALE (address latch enable) signal is logic high. The latched databits are fed direct to the data inputs of the system RAM and ROM, as well as to buffer IC9. The MS (most-significant) group of address lines, A8-A15, is not multiplexed and connected direct to the memories and the address decoders, IC1, IC2 and IC3.

The memory map of the system has the following structure:

- two ROM ranges, one from 0000_H to $3FFF_H$ and one from 8000_H to $BFFF_H$;
- one RAM range from 0000_H to $7FFF_H$.
- An I/O range mapped between addresses $C000_H$ and $C0FF_H$.

The structure of the memory map is illustrated in Fig. 2. Wire jumper Br1 when fitted allocates the range from 0000_H to $1FFF_H$ (a part of the ROM range) to the internal BASIC interpreter ROM in the 8052AH-BASIC processor. The wire jumper is removed when an external ROM (or EPROM) is used, as required in most cases with the 80(C)32.

The ROM and RAM ranges have an overlapping area between 0000_H and $1FFF_H$. The PSEN signal allows the processor to use this 'shared' area either as program memory or data memory. This means that the memory structure of the 80(C)32 and 8052 is not based on the 'classic' Von Neumann model in which the memory areas allocated to program and data are arranged as contiguous blocks in the address space. Fortunately, the

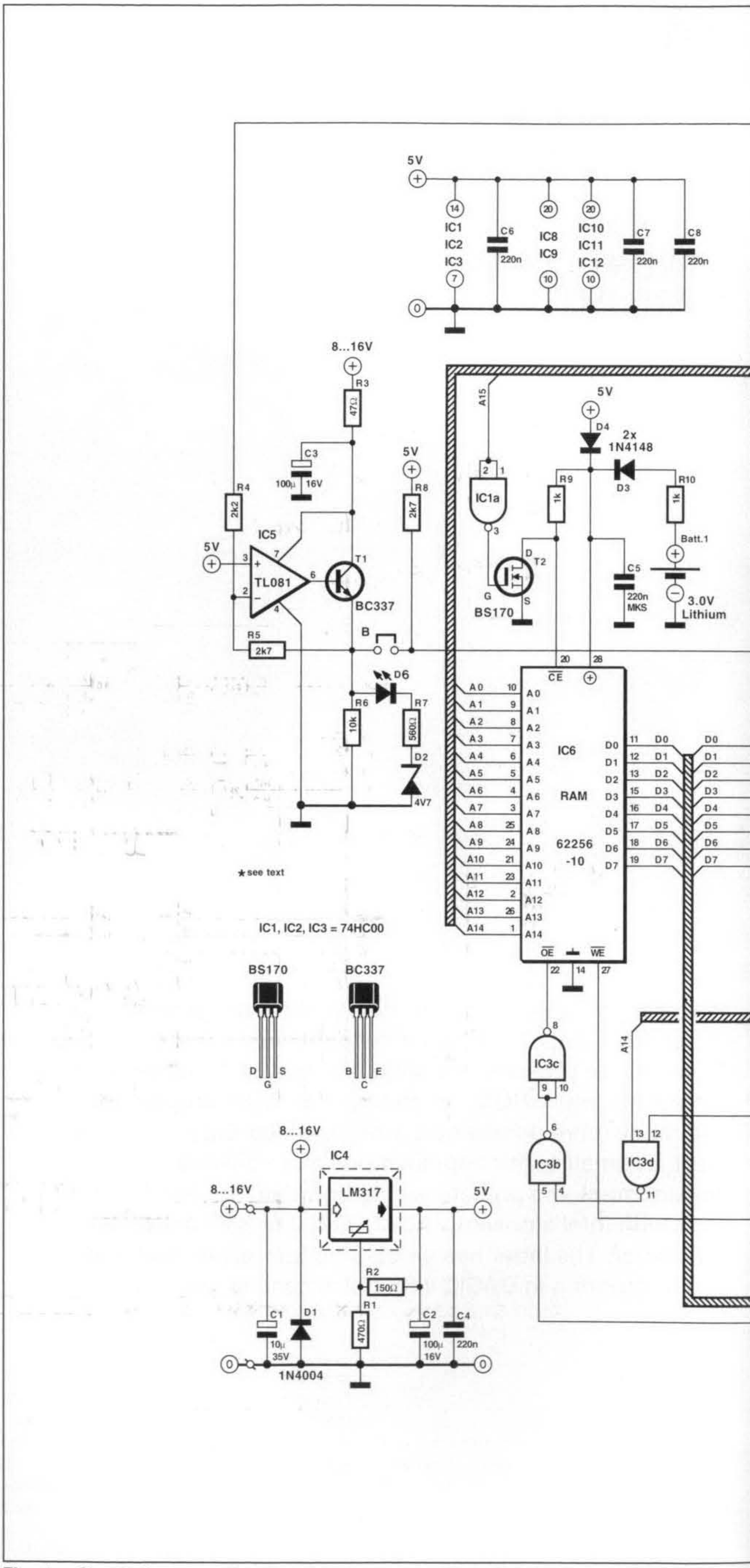
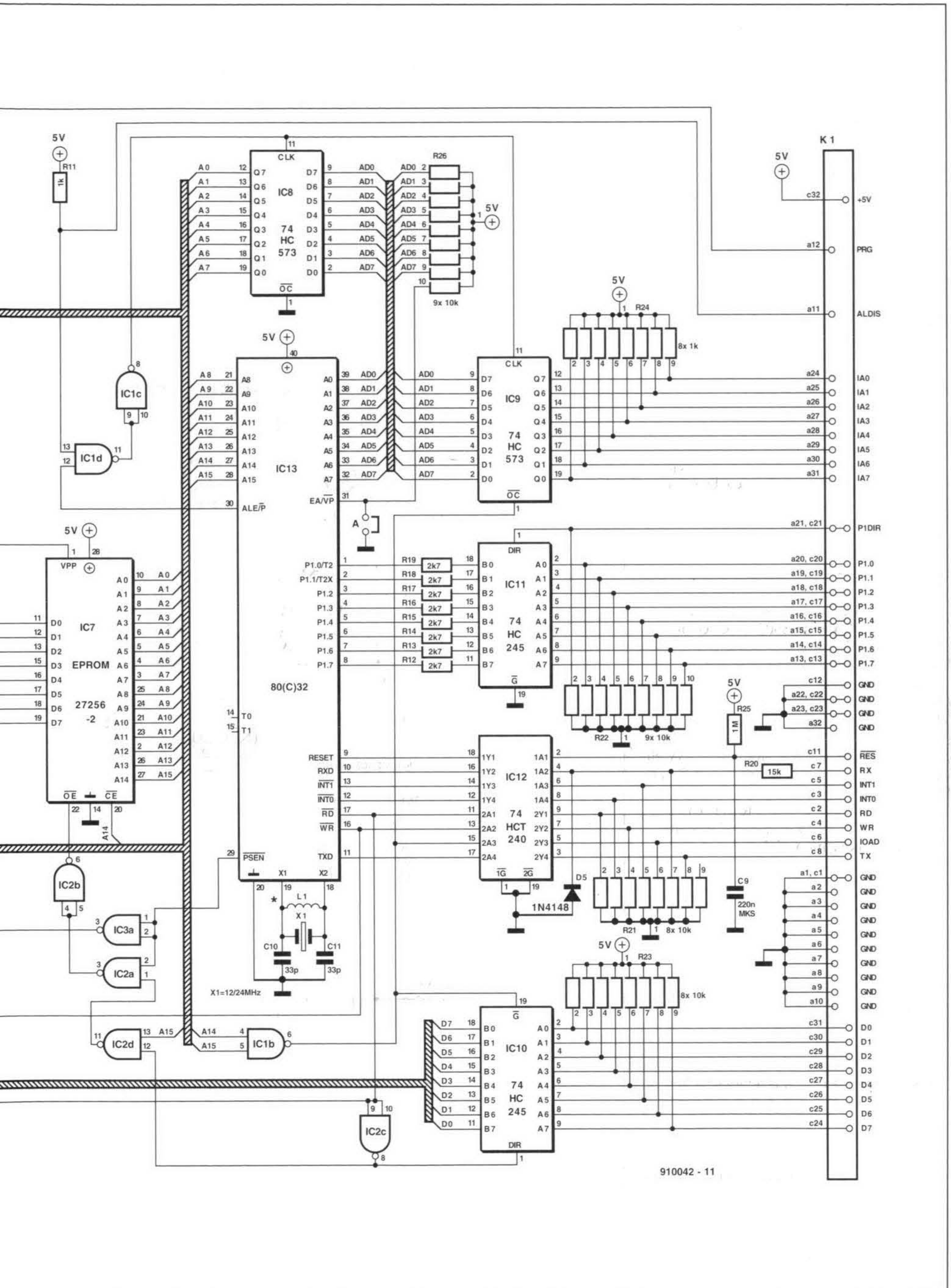


Fig. 1. Circuit diagram of the single-board computer. Note that although a 80C32 CPU is sh



u may also fit a 8032 or 8052AH-BASIC. The latter brings programming in BASIC within easy reach if you have a PC emulating a terminal.

present system does allow for a 'classic' address division. This is achieved as follows: access to the lower 16 Kbytes in the ROM (EPROM) area is only allowed when the CPU supplies the appropriate address, and actuates the PSEN line. The upper 16 KByte range can be selected with the appropriate address and the $\overline{RD}$ (read) or the PSEN signal. This means that the upper 16 KByte range may function as a data memory or a program memory, which is a requirement for the storage of BASIC programs. The memory division is realized by a 74HC00 (IC2) whose gates are combined to form a kind of decoder.

The lower 16 KByte block in the RAM range between 0000_H and $7FFF_H$ can only be used as data memory, while the upper 16 KByte block can be used as data memory or program memory because it does not overlap the ROM range. The upper block is therefore suitable for writing and debugging programs written in machine code.

The address decoding of the RAMs is arranged by IC3. Gate IC1b combines address lines A14 and A15 to provide an input/output address strobe, $\overline{IOAD}$, required for the signalling of access to the address range above $C000_H$, i.e., the range reserved for I/O operations.

Although IC9 latches the lower address byte like IC8, it is controlled differently via its $\overline{OE}$ input which is connected to the previously mentioned $\overline{IOAD}$ line. This means that the 8-bit address is latched only when $\overline{IOAD}$ is actuated. Otherwise, the de-actuated IC outputs are held at +5 V by a resistor array and so represent a value FF_H.

The datalines are buffered by an octal three-state bus driver, IC10. The $\overline{RD}$ signal supplied by the controller is inverted by IC2c and determines the data direction (read/write) of IC10. To keep the external bus free from all of the data transfer operations performed by the processor, IC10 is enabled by $\overline{IOAD}$ via its $\overline{G}$ input.

Port 1 is freely available and may be used to convey outgoing as well as incoming data. A bidirectional bus driver Type 74HC245 protects the controller and increases the drive capacity when the port is used as an output. The logic level applied to the P1DIR connection determines the direction of the data. When P1DIR is not connected, or connected to ground, the port functions as an output (write). The input function (read) is selected by making P1DIR logic high. The 2.7 kΩ resistors between IC11 and the controller prevent overloading when port 1 is programmed as an output while IC11 supplies data as a result of incorrect programming. The resistors limit the processor output current to a safe 2 mA or so in this output-against-output conflict.

A memory backup circuit retains the RAM data when the power is removed from the computer. A PCB-mount 3-V lithium cell supplies the required data retention current to RAM IC6 via resistor R10 and diode D3. When the computer is in normal use, diode D4 conducts and D3 isolates the backup battery. If you have a sensitive voltmeter, the data retention current may be measured indirectly as the voltage across resistor R10: 1 mV corresponds to about 1 µA of RAM current.

The RAM is automatically switched to its low-power standby state when the power is removed. This function is effected by a MOSFET, T2. As long as the computer is powered by the mains supply, the FET takes the $\overline{CE}$ input of the RAM low. When the power is removed, the FET arranges the $\overline{CE}$ input to be effectively connected to the retention voltage via resistor R9. In this manner, the RAM is never allowed to be in a non-defined state (i.e., $\overline{CE}$ 'floating' while a voltage exists across the supply terminals) which results in a relatively high current drawn from the battery.

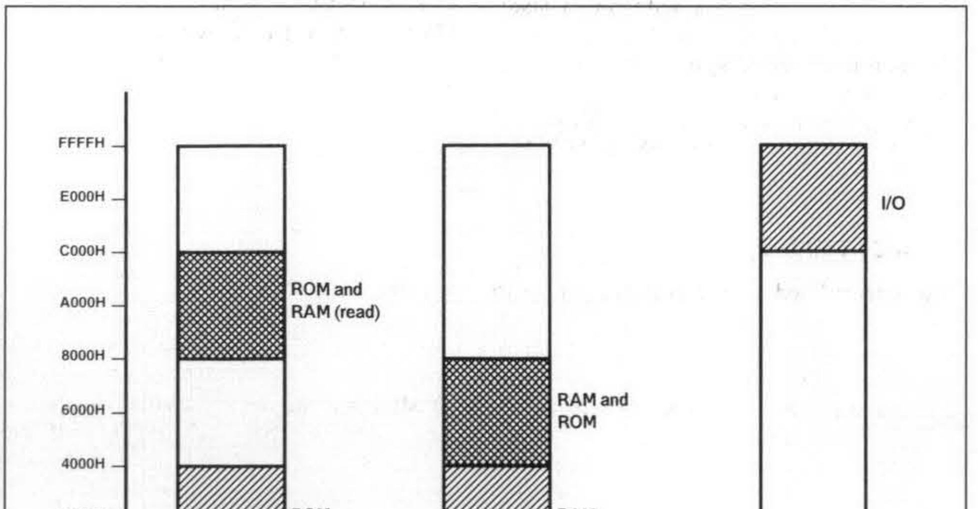


Fig. 2. Parts of the memory space in the system are shared by ROM and RAM. A special kind of memory addressing is therefore required to access the right data or program.

Interfaces

Programming the board in BASIC essentially requires a serial keyboard as an input device, and a serial display as an output device. The two functions are probably best combined into a terminal or a PC running a terminal emulation or general-purpose communication program (see Refs. 2 and 3).

The SBC has an on-board serial interface that works with TTL levels. The RX (received data) and TX (transmitted data) terminals of this interface are available on the DIN-VG64 connector, K1. Although the serial interface of the SBC works basically with TTL levels, there should be no problem in hooking it up to a terminal or PC with an RS232C-compatible input/output. Designed to work with ±12-V signals, most of these interfaces work happily with TTL (0 V/+5 V) signals, although it must be noted that the noise immunity suffers considerably when a relatively long cable is used between the SBC and the PC or terminal.

At the side of the SBC, the signal received at terminal RX of connector K1 is made positive only and limited to a swing of 5 V by R20, a resistors in 8-way array R21, and diode D5. The RS232C signal is inverted as required by that standard by IC12. The serial signal at the TX output of the SBC has a swing of 5 V, and is connected to the RxD input of the terminal. The RX input of the SBC is connected to the TxD output of the terminal or PC.

The two interrupt inputs of the board, INT0 and INT1 (pins c3 and c5 respectively on K1) enable the processor to respond quickly to external events. The use of interrupts is of particular interest when the SBC is at the heart of a computer-controlled system. A software interrupt is acknowledged by the processor with an appropriate servicing routine written by the user.

The SBC may be reset with an external signal via the $\overline{RES}$ terminal, pin c11 on K1. Since IC12 contains inverting buffers, the write (WR), read (RD) and I/O address (IOAD) signals are active-high on connector K1. An example of the use of these signals for a data input/output decoder is shown in Fig. 4.

On-board EPROM programmer

The 8052AH-BASIC processor is capable of storing a BASIC program in EPROM starting from address 8000H. The hardware to do so consists of opamp IC5, transistor T1 and a handful of passive parts. The non-inverting input of the opamp is held at +5 V, while the inverting input is connected to pin a12 (PRG) of connector K1 via a 2.2 kΩ resistor. A low level on PRG takes the opamp output high. Transistor T1 then supplies a programming voltage of about 12.5 V to the Vpp pin of the system EPROM, IC7. When the PRG line is at +5 V, the Vpp pin is at 5 V also.

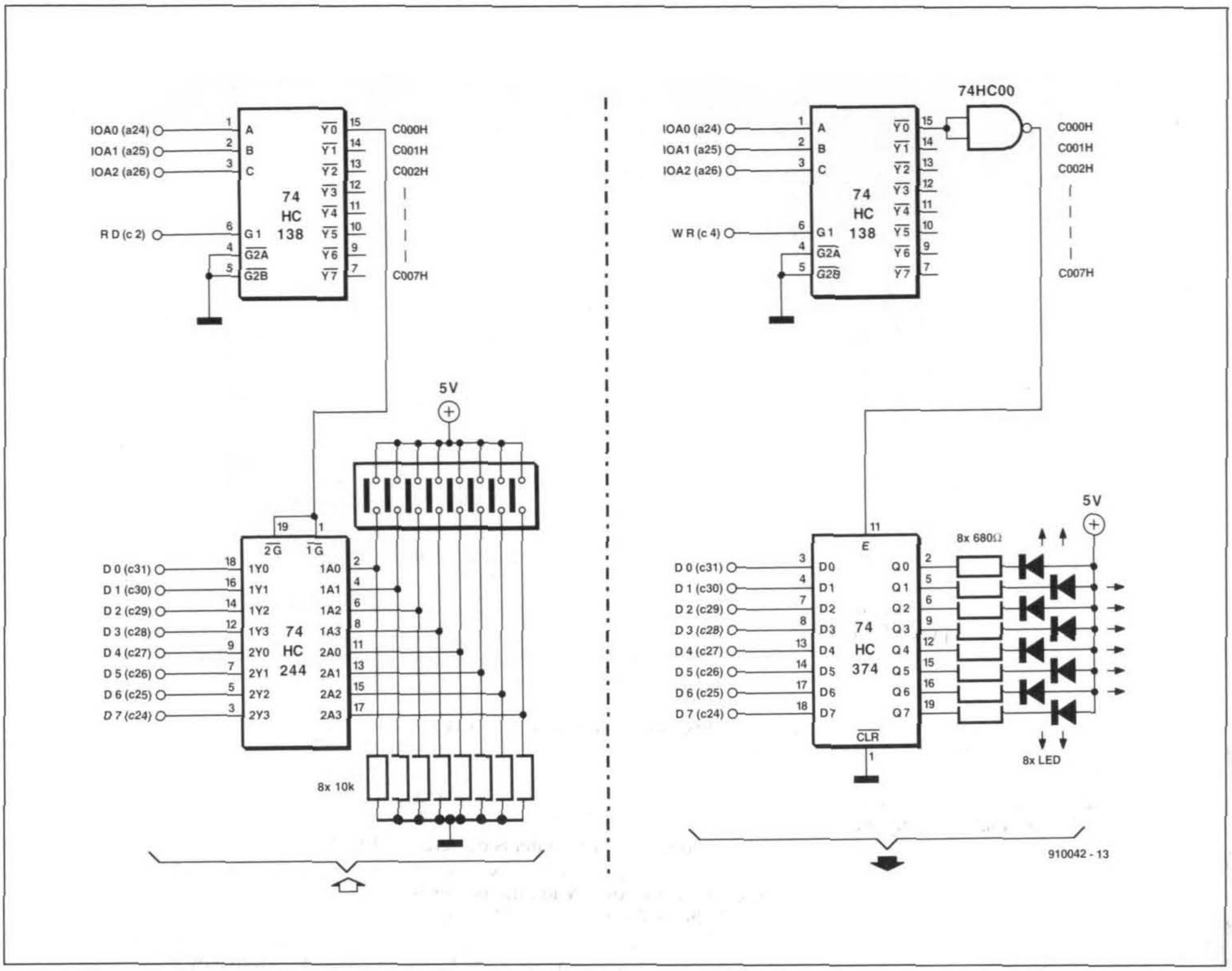


Fig. 4. Illustrating byte-wide input/output for the SBC. Shown here are two suggested external circuits to effect memory-mapped input/output operations with the single-board computer. Both circuits are wired to the system extension connector. The left-hand circuit reads an array of eight push-buttons in a keyboard, while the right-hand circuit is a basic output device that controls eight LEDs.

The ALE signal must be disabled while the EPROM is being programmed. This is achieved by the processor making the ALDIS line logic low. To program an EPROM, connect the PRG terminal to port line P1.4, and the ALDIS line to port line P1.3. The P1DIR terminal is not connected, and jumper Br2 is best removed.

The programming sequence is indicated by LED D6, which lights only when the programming voltage is higher than about 6.5 V. Wire jumper Br2 must not be fitted while D6 lights.

Which processor?

The choice between a 8052AH-BASIC, a 80C32 or a 80C32 processor is up to you. The latter two do not have an on-chip BASIC interpreter, and can only be programmed in machine code. It is, however, possible to unload the BASIC interpreter from the 8052AH-BASIC, transfer it to EPROM, and run it with a 80(C)32. How this can be achieved is detailed in Refs. 2 and 3. In addition to the information provided in these earlier publications we print another EPROM downloader — see Fig. 4.

The 80C32 can work with much faster clocks than the 8052AH-BASIC: according to the manufacturer, it is capable of operating at a clock of 16 MHz. A couple of our prototypes however worked fine at a clock of 24 MHz. By contrast, the HMOS 8052AH-BASIC threw in the towel at about 15 MHz.

Power supply

The unregulated power supply voltage to the SBC should normally lie between 8 V and 12 V. To enable the 12.5-V stabilizer around IC1 and T5 to operate correctly, a minimum supply voltage of about 16 V is required when an EPROM is to be programmed. Provided IC4 is adequately cooled, the SBC may be powered permanently with 16 V obtained from a simple mains supply: a 12-V transformer, a bridge rectifier and a 1,000 μF capacitor should do the job.

The current consumption of the SBC depends on the ICs fitted. When the CMOS 80C32 is used, you can expect a current consumption between 50 mA and 150 mA. The HMOS CPUs (8032 and 8052AH-BASIC) will require the power supply to deliver more than 300 mA. Depending on the type of EPROM installed (CMOS or non-CMOS), add another 100 mA or so to the current requirement.

Practical use

The microcontroller has its own clock generator which operates in conjunction with a quartz crystal. Remove inductor L1 when a quartz crystal specified for fundamental frequency resonance is used, as with the 8052AH-BASIC, which will not go faster than 15 MHz or so. Since most crystals of 20 MHz and higher are overtone types, L1 must be fitted when a 24-MHz type is used with a 80C32 processor. A 1.5-μH inductor then prevents the crystal resonating at its fundamental frequency of 8 MHz.

In case the computer does not function spot on, the power-on reset capacitor, C9, may have to be changed from 220 nF into a 4.7 μF tantalum type. This may be necessary when the supply voltage rises too slow at power on.

You will need software to program the computer, and to use the computer in conjunction with a terminal. MCS-52 assemblers

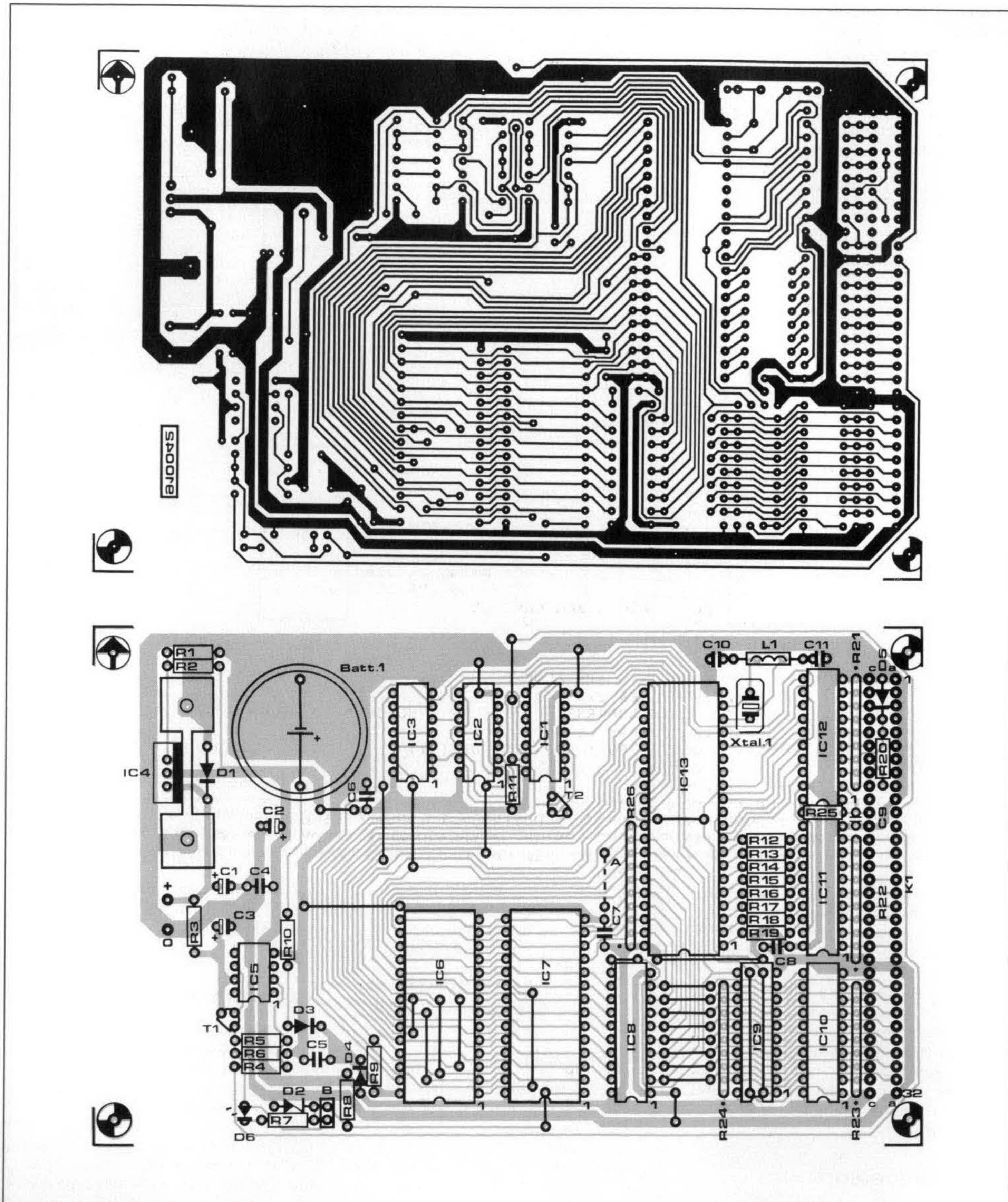


Fig. 5. Track layout (mirror image) and component mounting plan of the *single-sided* printed circuit board for the computer.

and cross-assemblers are available for use on a PC, and the object code may be transferred to the SBC by means of an EPROM. If you want to avoid the hassle of burning EPROMs, debugging the program, erasing the EPROM and programming it again, consider the use of an EPROM emulator which forms a direct link between the PC used to develop the program and the target system, in this case, the SBC.

By virtue of its internal BASIC interpreter, the 8052AH-BASIC is much simpler to get going than the 80C32: connect the terminal or PC to the SBC via RX, TX and ground, set a baud rate of anything between 300 and 19,200, reset the SBC and press the space bar to initiate the automatic baud rate timing. The message

***MCS-51 (tm) BASIC V1.1**
READY

should appear on the console display, and you are ready to start entering or downloading a BASIC program.

One final note: when the 8052AH-BASIC is used, do not suspect a malfunction of the memory backup circuit if you find that your program has disappeared. The BASIC inter-

COMPONENTS LIST

Resistors:

1	470Ω	R1
1	150Ω	R2
1	47Ω	R3
1	2kΩ2	R4
10	2kΩ7	R5;R8;R12-R19
1	10kΩ	R6
1	560Ω	R7
3	1kΩ	R9;R10;R11
1	15kΩ	R20
2	9-way 10kΩ SIL	R22;R26
3	8-way 10kΩ SIL	R21;R23;R24
1	1MΩ	R25

Capacitors:

1	10µF 35V radial	C1
2	100µF 16V radial	C2;C3
6	220nF	C4-C9
2	33pF ceramic	C10;C11

Semiconductors:

1	BC337	T1
1	BS170	T2
1	1N4004	D1
1	4V7 0.4 W zener diode	D2
3	1N4148	D3;D4;D5
1	green LED	D6
3	74HC00	IC1;IC2;IC3
1	LM317	IC4
1	TL081	IC5
1	62256-LP2	IC6
1	27C256-2	IC7
2	74HC573	IC8;IC9
2	74HC245	IC10;IC11
1	74HCT240	IC12
1	80C32 or 8052AH-BASIC	IC13

Miscellaneous:

1	1µH5 choke	L1
1	64-way a-c row DIN connector with angled solder pins	K1
1	12MHz, 15MHz or 24MHz quartz crystal	X1
1	3V lithium battery (button cell) for PCB mounting	Bat1
1	TO220 style heat-sink	
1	printed-circuit board	910042

```
1    FOR I=0 TO 8191 : XBY(I+8192)=CBY(I) : NEXT I
10   PRINT "UNIVERSAL PROM PROGRAMMER" : PRINT "WHAT TYPE OF DEVICE ?"
20   PRINT : PRINT "1 = EEPROM" : PRINT "2 = INTELLIGENT EPROM"
30   PRINT "3 = NORMAL (50 MS) EPROM" : PRINT : INPUT "TYPE (1,2,3) - ",T
40   ON (T-1) GOSUB 340,350,360
50   REM this sets up intelligent programming if needed
60   IF W=.001 THEN DBY(26)=DBY(26).OR.8 ELSE DBY(26)=DBY(26).AND.0F7H
70   REM calculate pulse width and save it
80   PUSH (65536-(W*XTAL/12)) : GOSUB 380
90   POP G1 : DBY(40H)=G1 : POP G1 : DBY(41H)=G1 : PRINT
100  INPUT " STARTING DATA ADDRESS - ",S : IF S<512.OR.S>0FFFFH THEN 100
110  PRINT : INPUT " ENDING DATA ADDRESS - ",E
120  IF E<S.OR.E>OFFFFH THEN 110
130  PRINT : INPUT " PROM ADDRESS - ",P : IF P<8000H.OR.P>OFFFFH THEN 130
140  REM calculate the number of bytes to program
150  PUSH (E-S)+1 : GOSUB 380 : POP G1 : DBY(31)=G1 : POP G1 : DBY(30) = G1
160  REM set up the eprom address
170  PUSH (P-1) : GOSUB 380 : POP G1 : DBY(26)=G1 : POP G1 : DBY(24)=G1
180  REM set up the source address
190  PUSH S : GOSUB 380 : POP G1 : DBY(27)=G1 : POP G1 : DBY(25)=G1
200  PRINT : PRINT "TYPE A 'CR' ON THE KEYBOARD WHEN READY TO PROGRAM"
210  REM wait for a 'cr' then program the eprom
220 X=GET : IF X<>ODH THEN 220
230  REM program the eprom
240  PGM
250  REM see if any errors
260  IF (DBY(30).OR.DBY(31))=0 THEN PRINT "PROGRAMMING COMPLETE" : END
270  PRINT : PRINT "***ERROR***ERROR***ERROR***" : PRINT
280  REM these routines calculate the address of the source and
290  REM eprom location that failed to program
300 S1=DBY(25)+256*DBY(27) : S1=S1-1 : D1=DBY(24)+256*DBY(26)
310  PH0. "THE VALUE ",XBY(S1), : PH1. " WAS READ AT LOCATION ",S1 : PRINT
320  PH0. "THE EPROM READ ",XBY(D1), : PH1. " AT LOCATION ",D1 : END
330  REM these subroutines set up the pulse width
340 W=.0005 : RETURN
350 W=.001 : RETURN
360 W=.05 : RETURN
370  REM this routine takes the top of stack and returns high, low bytes
380  POP G1 : PUSH (G1.AND.OFFH) : PUSH (INT(G1/256)) : RETURN
```

Fig. 6. Type in this listing if you want to unload the BASIC interpreter from the 8052AH-BASIC CPU, transfer it to EPROM and run it with a 8032 or 80C32. The advantages: lower cost and higher speed. The pulse timing in this program is based on a clock of 11.0592 MHz.

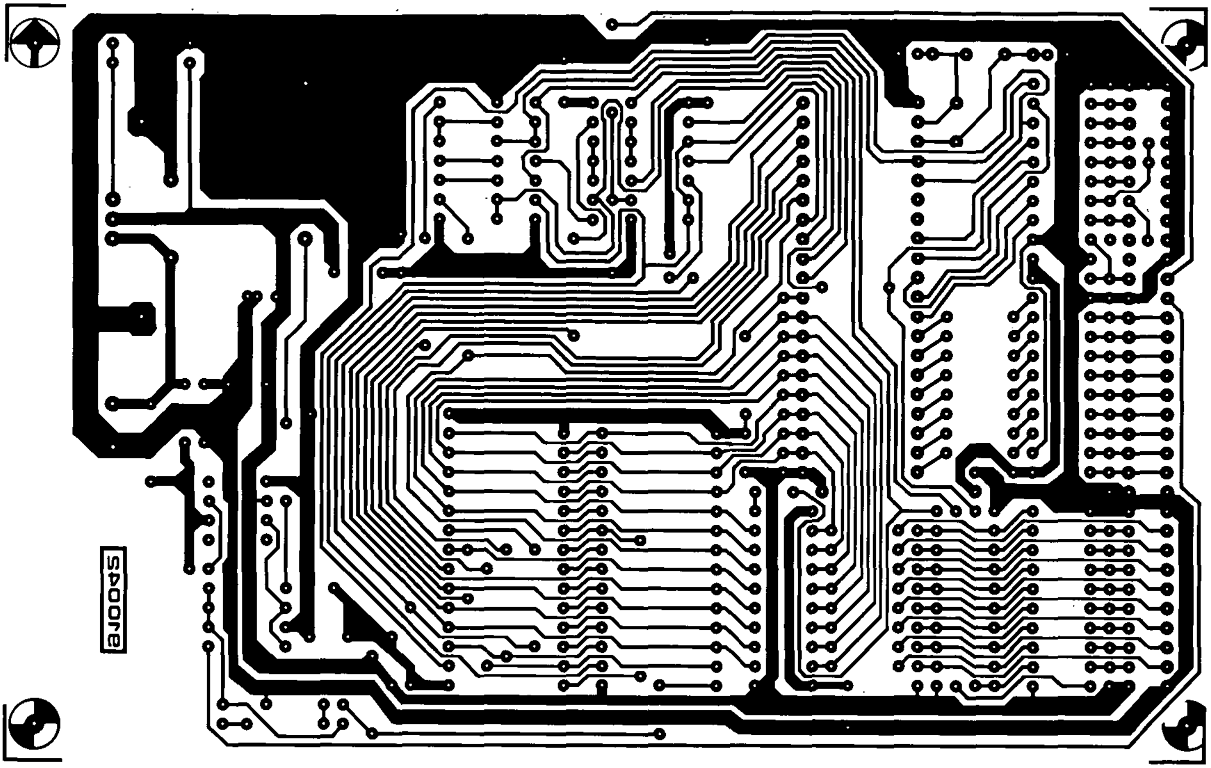
preter on reset runs a memory test that clears all of the RAM content. When the program is transferred to EPROM, you have the option to use the programming mode PROG3, which forces the interpreter to clear the memory up to the value indicated by MTOP. All data above MTOP is retained. ■

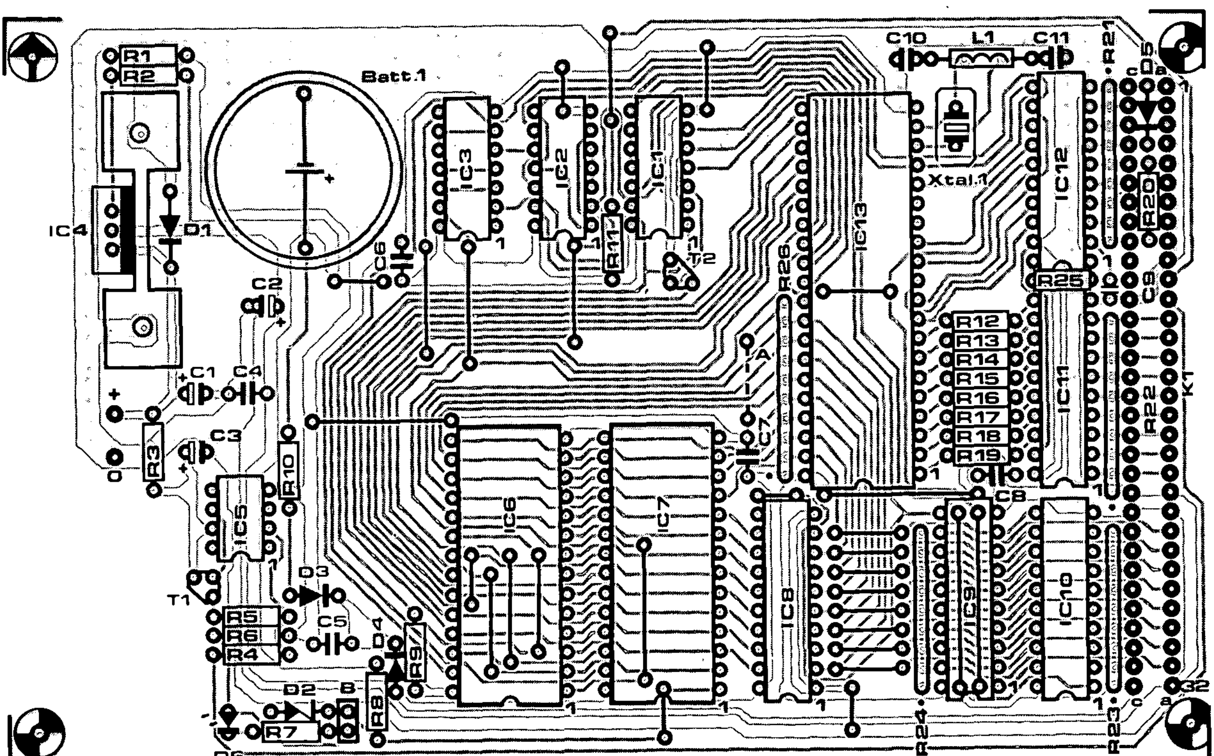
References:

1. "BASIC computer". *Elektor Electronics* November 1987.
2. "CMOS replacement for 8052AH-BASIC". *Elektor Electronics* January 1990.
3. "ROM-copy for 8052-BASIC computer". *Elektor Electronics* September 1990.

For further reading:

1. Microcontroller Handbook, Intel Corp. 1984.
2. MCS® BASIC-52 users manual, Intel Corp. 1984, order no. 270010-001.
3. "LCD for 8052 microcontroller". *Elektor Electronics* Supplement July/August 1990.





R1
R2
Batt.1
IC4
D1
IC3
IC2
IC1
R11
T2
C6
C2
C1
C4
C3
R3
R10
IC5
T1
D3
R5
R6
R4
C5
D4
R9
R8
D2
R7
D6
B
A
IC6
IC7
C7
R26
IC13
C10
L1
C11
Xtal.1
R21
D5
IC12
R20
R25
C9
R12
R13
R14
R15
R16
R17
R18
R19
IC11
R22
K1
C8
IC8
IC9
IC10
R24
R23
32

CORRECTIONS

Wattmeter

April 1991, p. 32-35

With reference the circuit diagram, Fig. 1, the right-hand terminal of the lower section of switch S2 should be connected to the circuit ground. This point is indicated by a dot.

In the adjustment procedure given on page 35, the references to presets P4 and P5 have been transposed. Contrary to what is stated, P4 sets the VY offset, and P5 the VX offset. The functions of the presets are shown correctly in the circuit diagram, Fig. 1.

To improve the accuracy of the instrument, connect R5 direct to the circuit ground instead of junction R6-R7. Finally, all circuit board tracks carrying mains current must be strengthened with 2.5-mm^2 cross-sectional area solid copper wire if currents higher than about 5 A are measured.

80C32/8052 Single-board computer

May 1991, p. 17-23

When a CPU type 8031 or 8052AH-BASIC is used, IC1, IC2, IC3, and IC8-IC12 must be 74HC**T** types. Jumper B is erroneously reffered to as Br2 in the text under "On-board EPROM programmer". Contrary to what is stated, this jumper must be fitted only when an EPROM is to be programmed — for all other use of the SBC, it must be removed. Also note that jumper B may only be fitted when the programming LED is out.

Sequential control

July/August 1991, p. 61

Motor M should be a d.c. type, not an a.c. type as shown in the circuit diagram.

Digital phase meter

June 1991, p. 32-39

In Fig. 5, the switch between input 'A' and IC1 should be identified 'S1', and that between input 'B' and IC2 'S2'. Switch S4 is an on/off type, not a push-button as shown in the diagram. Capacitors C3 and C6 are shown with the wrong polarity. The component overlay of the relevant printed-circuit board (Fig. 8) is all right.

Universal NiCd battery charger

June 1991, p. 14-19

The parts list on page 19 should be corrected to read

C7 = 2200µF 25V

When difficult to obtain, the BYW29/100 (D5) may be replaced by the BY229, which is rated at 6 A.

The text under the heading 'Calibration' should be replaced by:

4. Connect a multimeter between points G and H on the board, and adjust P4 until the measured voltage is 1 V lower than the voltage on the battery terminals.

MIDI program changer

April 1991, p. 14-17

The contents of the EPROM should be modified as follows:

address	data
OOBC	E5
00C7	80
00C8	CB
00C9	F5
00CA	7B
00CB	12
00CC	00
00CD	D2
00CE	C2
00CF	02
00D0	80
00D1	C2

Readers who have obtained the EPROM ready-programmed through the Readers Services may return it to obtain an update.

Electronic exposure timer

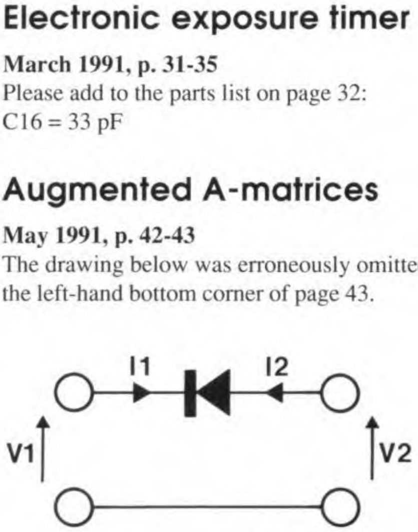
March 1991, p. 31-35

Please add to the parts list on page 32:
C16 = 33 pF

Augmented A-matrices

May 1991, p. 42-43

The drawing below was erroneously omitted in the left-hand bottom corner of page 43.



910058-14

UNIVERSAL I/O INTERFACE FOR IBM PCs

Those were the days when you could use your Commodore C64, Acorn Atom or ZX81 computer to control hardware intelligently. Model train systems, robots, greenhouse watering and temperature control systems — all within easy reach of the keen programmer with little or no knowledge of computer hardware. Alas, the coming of the IBM PC, the compatibles, the ATs and the 386-based systems, seems to have banished simple hardware interfacing, the PC being an expensive 'box' harnessing a lot of computing power, but restricted to use in an office environment. We do not agree that a PC is unsuitable for control applications: all it needs is the circuit described here: a low-cost fully buffered insertion card that forms a versatile, simple and safe link between the PC (whether an XT, AT or 386-based machine) and your own hardware.

A. Rigby

USING a computer and some home-brew software to control apparatus is sheer fun. A few lines of program code allow lamps to light, motors to start turning, and model trains to find their way on a complex track system. Sensors and other types of recorder device enable a computer to measure and store physical quantities from the 'real world' around us.

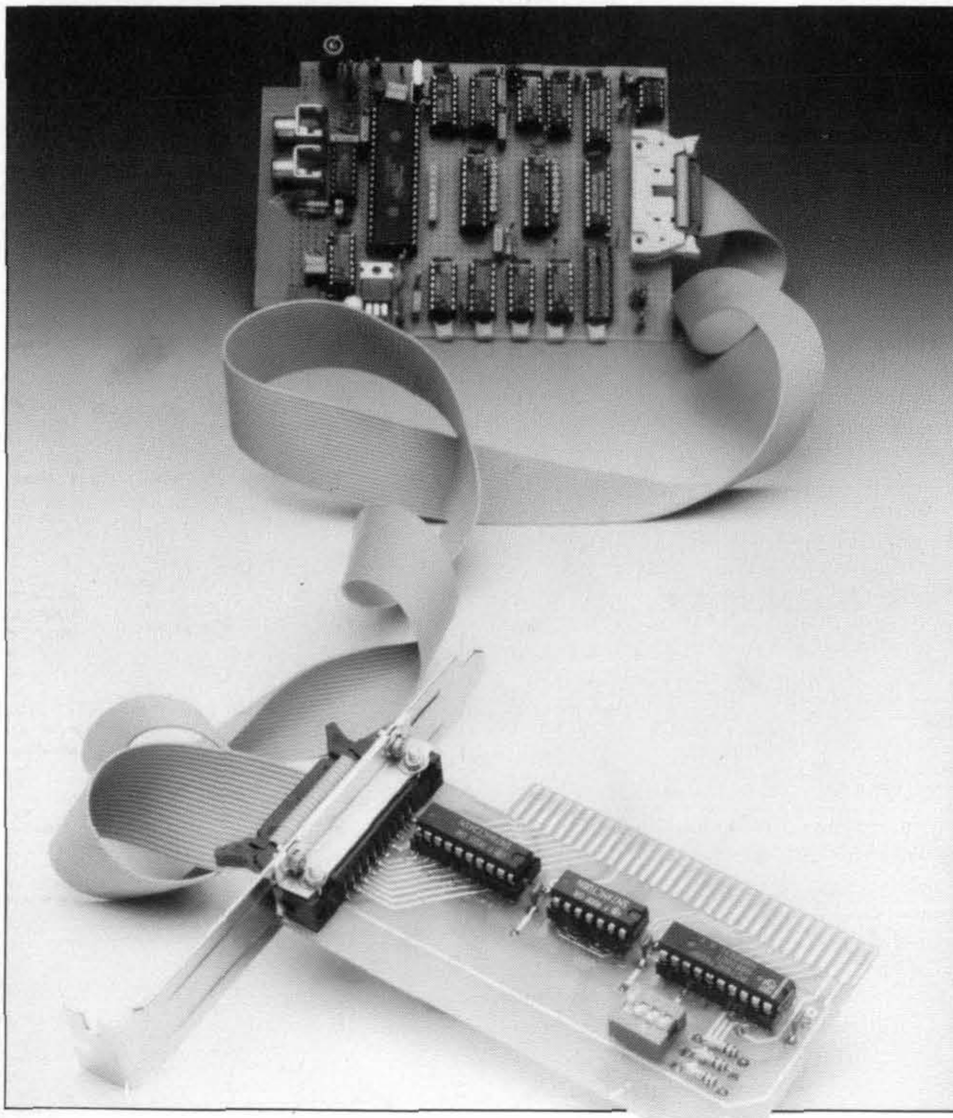
Unfortunately, an IBM PC or compatible appears to be less suitable for the above control applications as it is very much a closed system designed for office use. Does that imply that we have to say goodbye to the model train system and the computer-controlled hobby lathe? No! Many of you will have noted that PC interface cards are being offered for industrial control applications. Unfortunately, these cards are pretty expensive, so it's time for a low-cost solution.

Count the components

The circuit shown in Fig. 1 may well be the simplest PC I/O interface you have ever seen, having only three ICs, three resistors and three capacitors. The interface is suitable for all types of IBM PC and compatibles, i.e, XTs, ATs and 386-based systems.

The circuit acts as a buffer between the computer and the external hardware, and is set to operate at a unique address in a small area in the PC's I/O range.

As shown in the circuit diagram, the connector between the computer and the interface card is a type with 31 connections at each side. This is the well-known IBM PC expansion slot connector. A small number of signals available on this connector are used



for the present interface and/or the hardware controlled by it. The external hardware is connected to K1, a 20-way PCB header that provides the 8-bit wide databus, two address lines, an enable signal and the read and write signals. The computer's 5-V supply rail is also available on K1, allowing small (experimental) digital circuits to be powered without the need of an external supply.

The address decoding logic ensures that interface card occupies four addresses in the I/O range reserved for prototyping cards. The actual address setting is accomplished with three switches in DIP switch block S1. In all cases, a free base address must be used, i.e., the interface card must not share an I/O address with any other card in the PC. The four addresses in the block are selected individually by A0 and A1. For convenience the DIP switch settings are printed on the component overlay of the interface card (see the component mounting plan in Fig. 2b).

The three switches determine the logic level at inputs P0, P1, P2 and P3 of IC1. The pull-up resistors connected to these inputs provide a logic high level when a switch is opened. When a switch is closed, the relevant IC input is logic low. All other P inputs of IC1 are held at fixed logic levels.

Address lines A2 to A9 on the expansion bus are connected to inputs Q0 to Q6 of address comparator IC1. An AND gate, IC3d, combines address lines A8 and A9 at input Q6. This frees input Q7 for use by AEN, the address enable signal that indicates DMA (direct memory access) activity without an I/O address being decoded. Gate IC3c ensures that IC1 is enabled during a read or write operation only. When the binary code at the P inputs equals that at the Q inputs, output $\overline{P=Q}$ goes low. This signals the selection of the user hardware hooked up to connector K1. The $\overline{P=Q}$ output also actuates the $\overline{G}$ input of IC2, which enables the PC's databus to be connected to the databus on K1. The direction of the dataflow between the PC and the external hardware is determined by the level of the $\overline{RD}$ (read) signal. A low level means that data is transferred from K1 to the PC, while a high level means data transfer from the PC to K1. Both the $\overline{RD}$ and the $\overline{WR}$ (write) signal on connector K1 are provided by the PC and buffered by a gate, IC3a and IC3b respectively. The PC's databus is buffered by IC2. Since all PC signals are buffered, the risk of cross-effects between the PC and the external hardware is reduced to a minimum.

Construction

The compact printed-circuit board designed for the PC interface is shown in Fig. 2. Ready-made boards supplied through our Readers Services are provided with gold-plated bus contact fingers. Connector K1 is a so-called box header with right-angled PCB pins. It protrudes from a clearance cut in the support bracket attached to the rear side of the circuit board. This arrangement allows ready connection of a flatcable with an IDC connector.

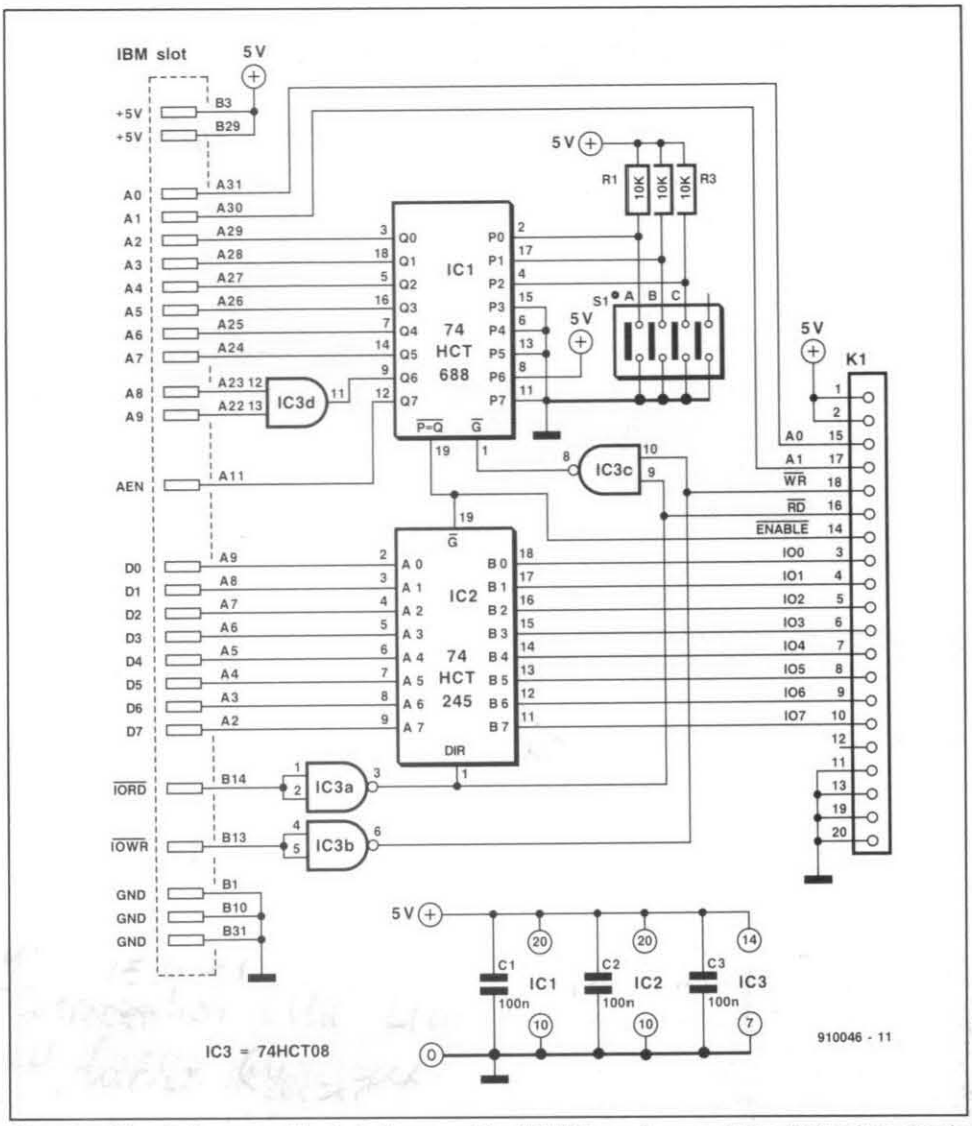


Fig. 1. Circuit diagram of the interface card for IBM PCs and compatibles. This ultra-simple circuit forms the ideal link between a PC and your own hardware developments.

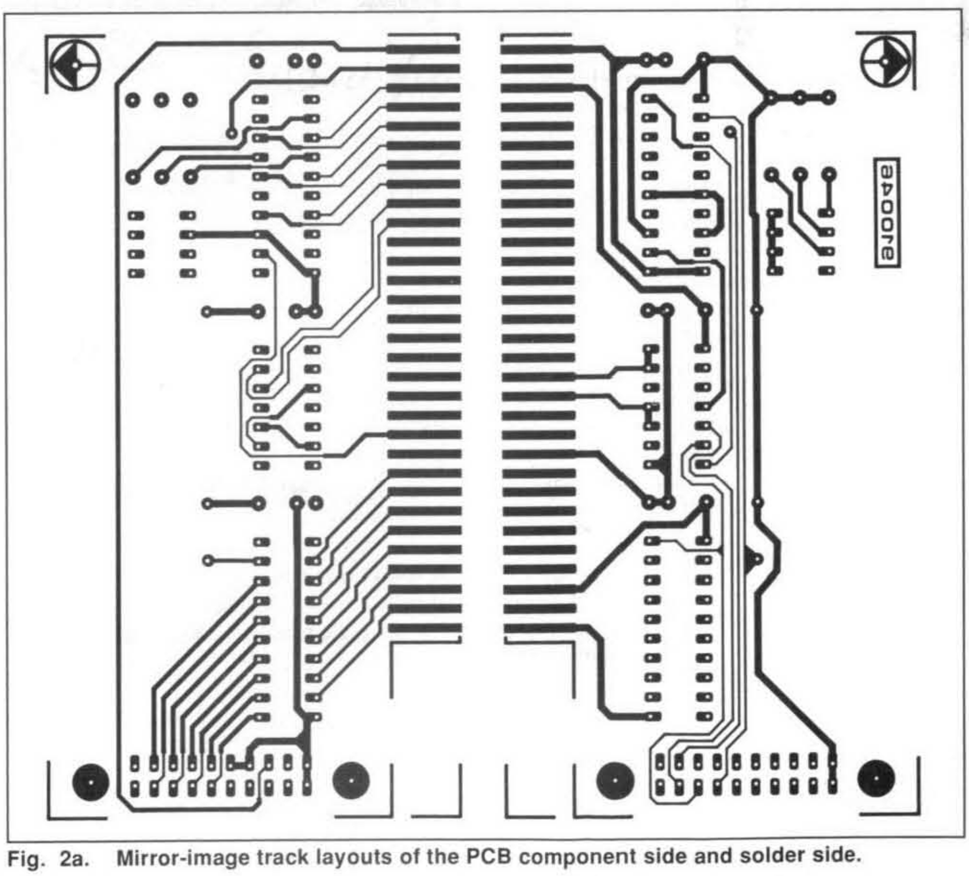
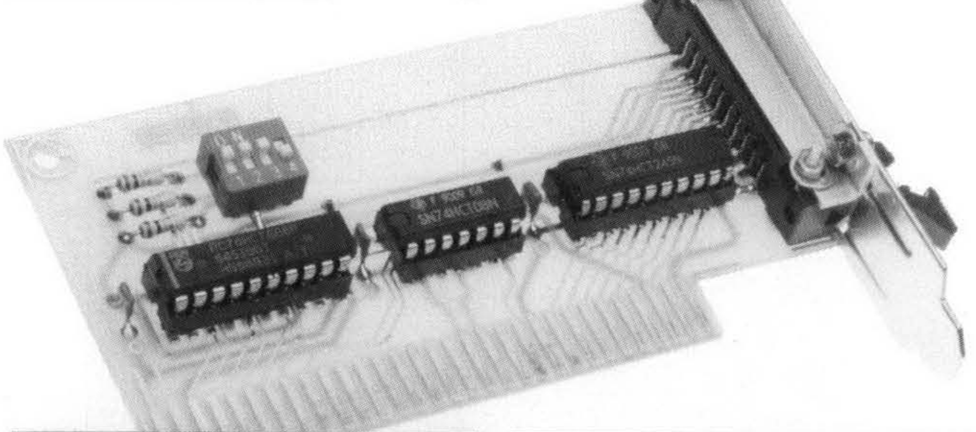


Fig. 2a. Mirror-image track layouts of the PCB component side and solder side.

After mounting all components, set the DIP switches to the desired I/O address. To avoid an address conflict, consult the manuals of other cards inserted in the PC to make sure the interface occupies a free address. Applications for the interface card will be described in future articles. ■



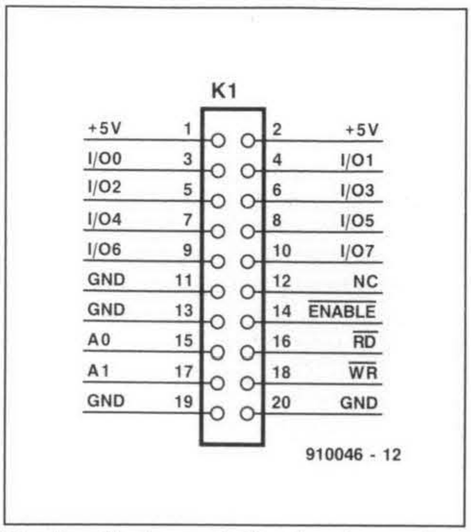


Fig. 3. Pinning and signal assignment on connector K1.

COMPONENTS LIST

Resistors:

3	10kΩ	R1;R2;R3

Capacitors:

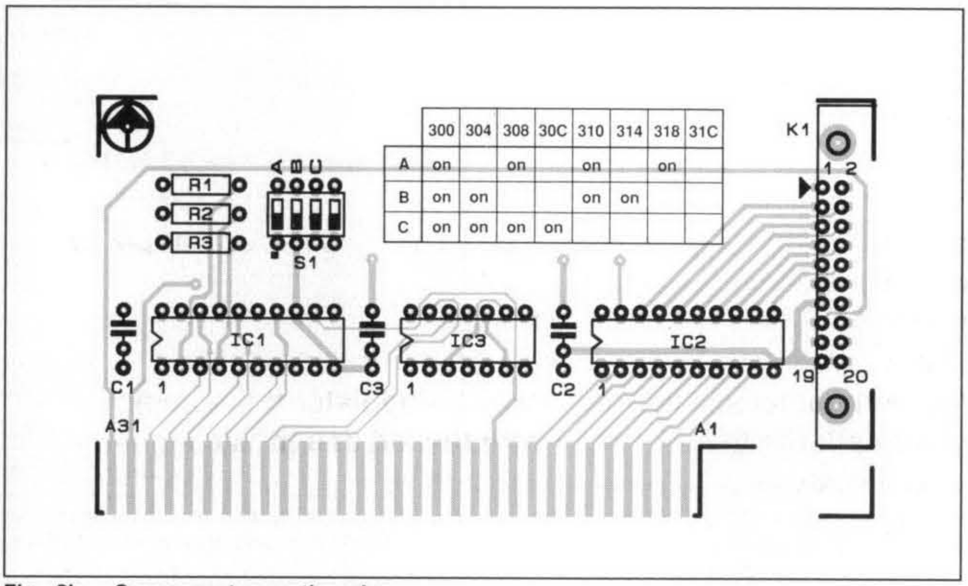
3	100nF	C1;C2;C3

Semiconductors:

1	74HCT688	IC1
1	74HCT245	IC2
1	74HCT08	IC3

Miscellaneous:

1	20-way header with eject handles and angled PCB pins	K1
1	4-way DIP switch	S1
1	printed-circuit board	910046



	300	304	308	30C	310	314	318	31C
A	on		on		on		on	
B	on	on			on	on		
C	on	on	on	on				

Fig. 2b. Component mounting plan.

NEW PRODUCTS

Floppy disk emulator for PCs

The recently introduced EDISK insertion card from DSS innovative electronics is capable of emulating a floppy disk drive in a PC. The maximum storage capacity of the EDISK is 4 MBytes with EPROMs fitted, or 1 MBytes with static RAMs fitted. The following EPROM types may be used: 27(C)512 (64 KBytes); 27(C)010 (128 kBytes); 27(C)020 (256 kBytes); 27(C)040 (512 kBytes). The card supports two 32 kByte SRAM types, the 62256 and the 621000.

The EDISK card enables you to replace the mechanical floppy disk drive A: by an all solid-state equivalent, which will be faster as well as more reliable, offering a speed of the order of a RAM disk. The main difference between the EDISK and a RAM disk, however, is that the contents can be stored in non-volatile EPROMs, or stored in static RAM. When the computer is switched off, the programs contained in the EDISK will be retained in static RAM by virtue of a back-up battery. Another difference is that the EDISK is automatically write-protected if EPROMs are used. The EPROMs are loaded with the aid of a programmer; their content can not be changed or erased by the PC user.

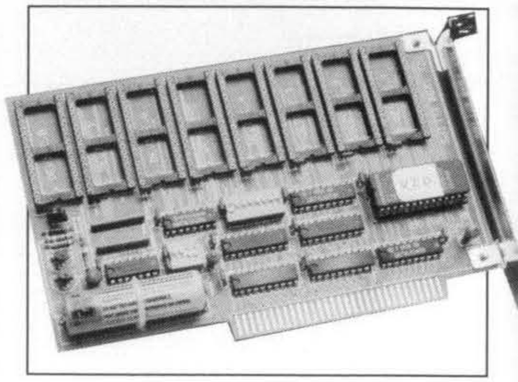
Since the EDISK is based on non-volatile memory, it is possible to boot the computer without a floppy disk, or even without a disk controller card.

The EDISK allows a combination of EPROMs and RAMs to be fitted, which effectively emulates two drives, drive A: containing EPROMs and drive B: SRAMs.

Applications of the EDISK include a 'stripped' XT computer that can be connected to a network as a diskless station. Alternatively, a motherboard fitted with an EDISK and a special I/O card can form a powerful control system. The use of SRAMs allow system parameters to be stored and altered as required.

For more information on the EDISK contact

DSS innovative electronics • Accustraat 25 • 3903 LX Veenendaal • Holland. Tel. +31 8385 41301. Fax: +31 8385 26751.



tion of the user hardware hooked up to connector K1. The $\overline{P=Q}$ output also actuates the $\overline{G}$ input of IC2, which enables the PC's databus to be connected to the databus on K1. The direction of the dataflow between the PC and the external hardware is determined by the level of the $\overline{RD}$ (read) signal. A low level means that data is transferred from K1 to the PC, while a high level means data transfer from the PC to K1. Both the $\overline{RD}$ and the $\overline{WR}$ (write) signal on connector K1 are provided by the PC and buffered by a gate, IC3a and IC3b respectively. The PC's databus is buffered by IC2. Since all PC signals are buffered, the risk of cross-effects between the PC and the external hardware is reduced to a minimum.

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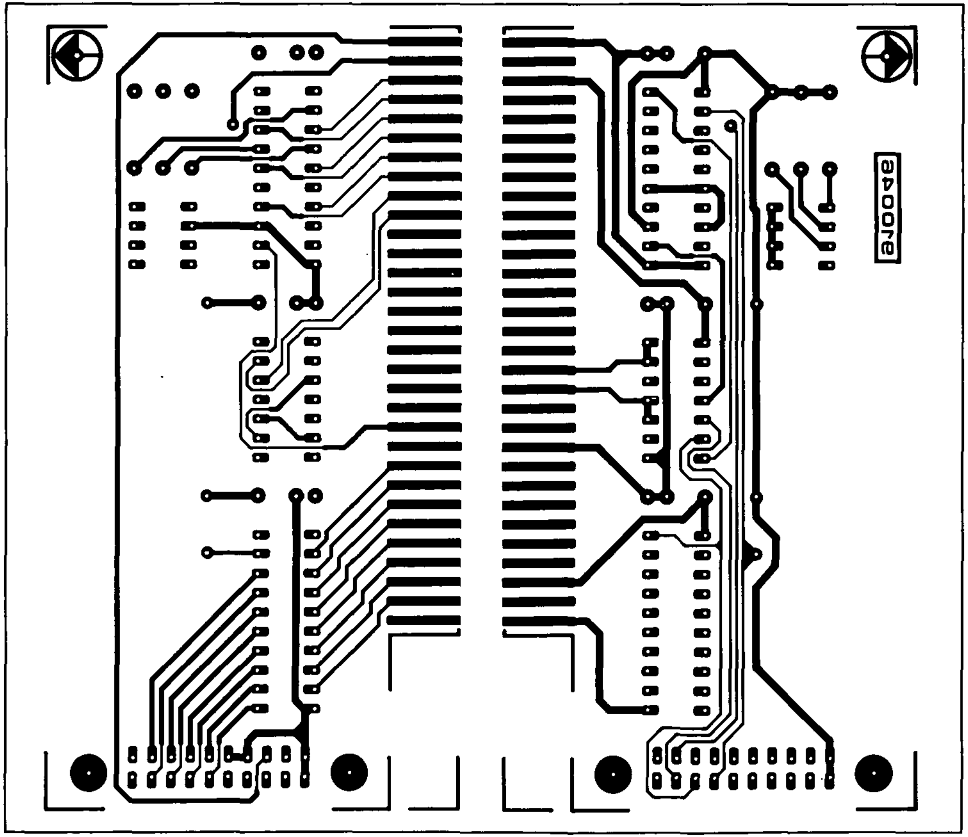
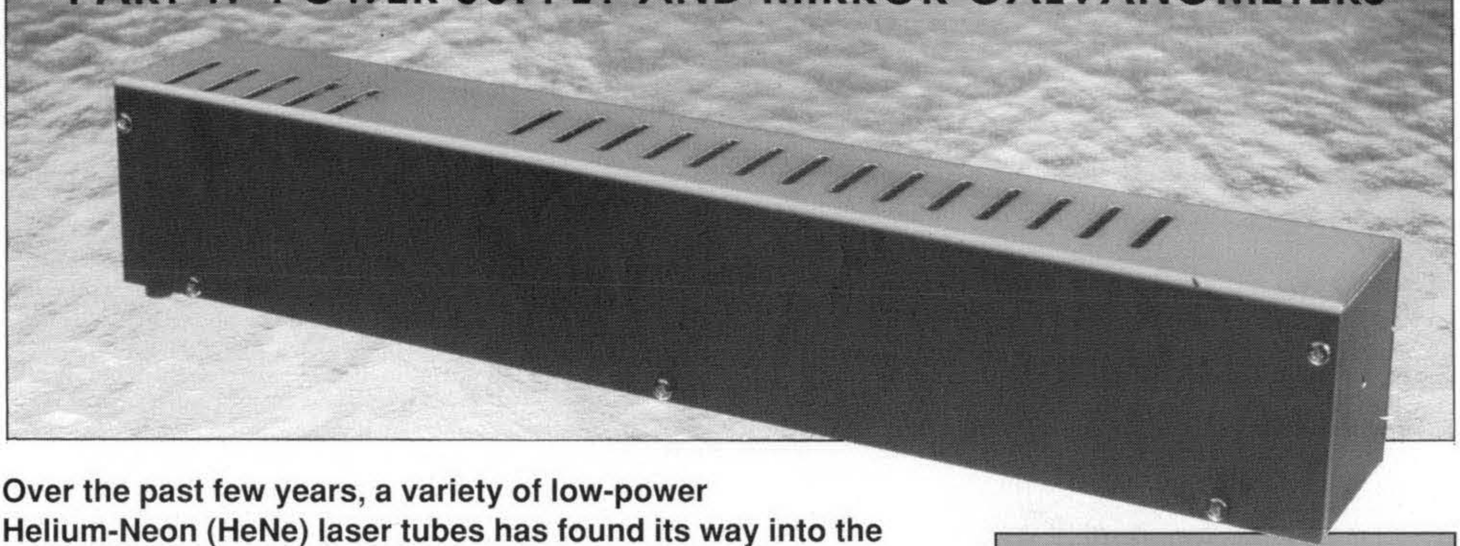


Fig. 2a. Mirror-image track layouts of the PCB component side and solder side.

909522-I

LASER

PART 1: POWER SUPPLY AND MIRROR GALVANOMETERS



Over the past few years, a variety of low-power Helium-Neon (HeNe) laser tubes has found its way into the electronics surplus trade circuit. Unfortunately, these laser tubes are nearly always sold without a suitable power supply, which, being a high-voltage unit, is not so simple to construct. In this two-instalment article we describe a power supply for the popular 2-mW class of laser tubes, a beam steering system based on mirror galvanometers and last but not least a state-of-the-art modulator that enables laser patterns to be created with the aid of a music signal. The circuits described are designed and marketed as kits by ELV.

MAIN SPECIFICATIONS

LASER EXCITER	
Type:	red Helium-Neon laser
Wavelength:	630 nm
Output power:	approx. 2 mW
Ignition voltage:	8,000 V
Operating voltage:	1,050 to 1,250 V
Laser current:	5 mA ±0.2 mA
Beam diameter at tube aperture:	0.75 mm
Beam divergence:	0.75 mm/m
Beam diameter at 10 m:	approx. 15 mm
POWER SUPPLY	
Input voltage:	10 to 16 V
Max. input power:	approx. 15 W
Max. output power:	10 W
Output curent (stabilized):	5.1 mA
Output voltage:	1.4 to 1.5 kV
Max. ignition voltage:	10 kV

MOST power supplies for laser exciters are fairly conventional circuits based on a cascade-type voltage multiplier. Unfortunately, these circuits exhibit very low efficiency and have the further disadvantage of being powered from the mains, which makes them dangerous units in many respect. Having a low-power laser tube complete with a suitable casing is one thing, actually making it work is quite another. The stumbling block is nearly always the power supply with its special transformer and high-voltage components.

The good news is that all owners of a two-milliwatt HeNe laser unit can now build a compact, high-efficiency switch-mode power supply: all the parts required to do so are contained in a kit supplied by ELV. The mirror galvanometers and associated control interface are also available in kit form. Many home-made lasers use conventional mirrors fitted on bulky loudspeaker drive units. The mirror galvanometer system discussed here achieves far better beam positioning accuracy, and is much smaller and more sensitive than any loudspeaker-based system. In addition, the mirror control system (to be described in Part 2) is an advanced circuit capable of absolute beam positioning, linearizing a part of the mirrors' frequency response, supplying a number of Lissajous patterns, and much more.

Mirror galvanometers

A mirror galvanometer is basically a measuring instrument based on a small mirror attached to a suspension system. The position of the mirror is determined by the level of electrical current sent through a coil in a permanent magnet assembly. A beam of light is reflected by the mirror and the spot of light moves, in principle, across a linear scale. In the present application, a miniature mirror galvanometer is fitted where the laser beam leaves the exciter tube, allowing the beam to be steered to give patterns. The mirror is positioned by a suitable drive signal, which may be supplied by an audio amplifier. By using two mirror galvanometers fitted at right angles, we are capable of deflecting the laser beam in two directions, horizontal (X) and vertical (Y).

From a constructional point of view, a mirror galvanometer is almost identical to a moving coil meter. The galvanometers used here, however, reverse the operating principle by using a coil as the stator and a suspended magnet instead of the other way around. The practical realization is shown in Fig. 1. The magnet is divided into two parts suspended from a glass carrier. The carrier is fixed to a flexible silicone pivot, and forms the base for about 40 deposited layers of glass that form a high-precision interference mirror. The irregularities on this surface are smaller than 60 nanometre, or about one tenth of the wavelength of the laser light. The reflection efficiency of the mirror is very high: 99.7% for light incident at an angle of 45°. This means that the mirror hardly contributes to the attenuation or divergence of the laser beam. Since the mirrors are precision instruments, the glass surface must never be touched — a fingerprint degrades the reflecting performance considerably.

The mirror is positioned with the aid of a coil wound around the moving parts. The

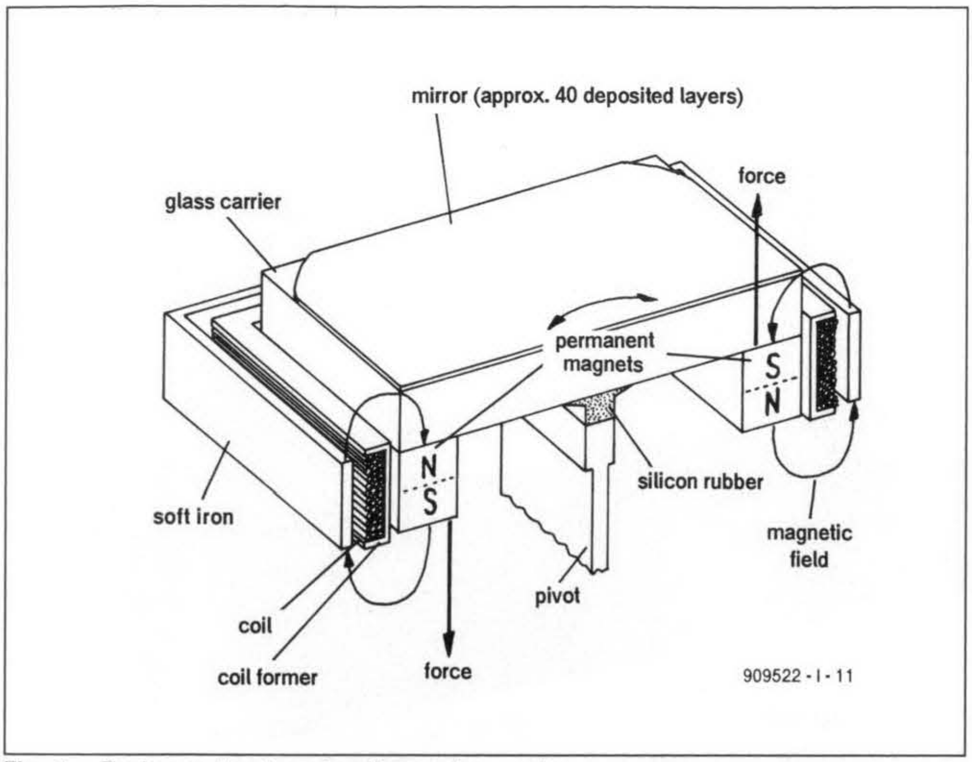


Fig. 1. Basic construction of a mirror galvanometer.

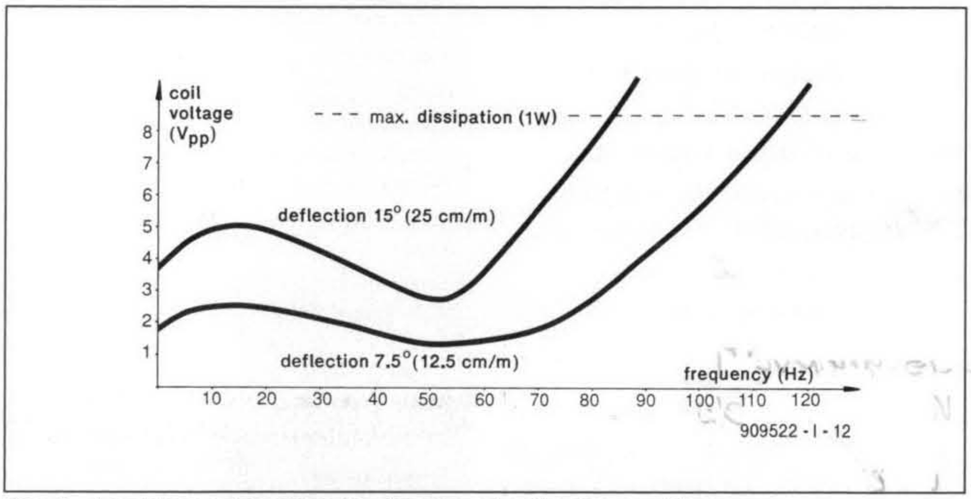
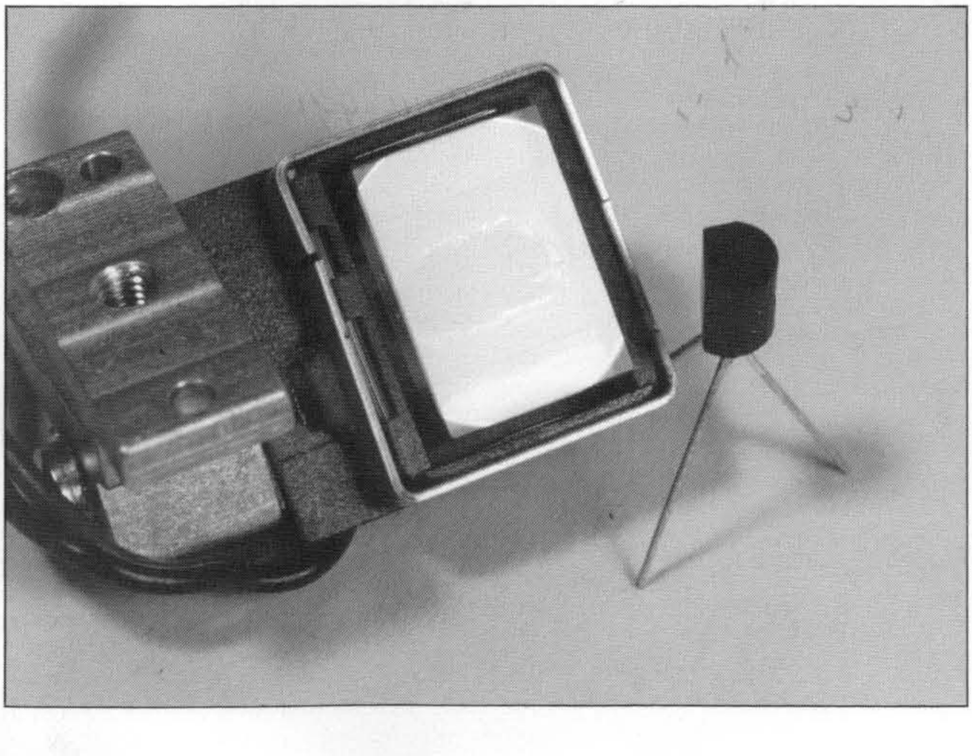


Fig. 2. Laser beam deflection as a function of coil voltage and frequency.



soft iron core around the coil concentrates the magnetic field in the direction of the magnets attached to the mirror. The resultant magnetic force makes the mirror tilt. The silicone suspension system provides a small counterforce that stops the mirror at a certain position, preventing it from tilting completely. The amount of deflection is determined by the magnetic field strength, the inertia of the mirror, and the counterforce of the pivot. Since the magnetic force depends on the level of current that flows through the coil, the deflection of the laser beam incident on the mirror can be controlled fairly accurately by controlling the current through the coil. In practice, however, it is easier to use a voltage rather than a current to control the mirror. Fortunately, this is not a problem since the coil represents a certain resistance. This resistance, albeit reactive, enables voltage drive of the coil.

The deflection of the mirror is not linear but dependent on the frequency of the signal applied to the coil — see Fig. 2. The curves show the correlation between the voltage and the frequency of the coil signal for mirror deflections of 7.5° and 15°. Fortunately, the actual response of the mirrors is a little better than suggested by the two curves. In practice, the deflection of the mirrors will be almost linear for single frequencies in the range from 0 to about 60 Hz. Serious deviations are, however, caused when more than one frequency is applied to the coil. This happens when the galvanometer coils are driven directly by a music signal. In most cases, however, accuracy is then not a point — what counts is an attractive pattern.

Electrically, the mirror system behaves not unlike an 8-Ω loudspeaker. This means that the deflection system can be connected to almost any AF power amplifier, provided the drive power is limited to about 1 W. Also, the frequency spectrum of the drive signal is limited to about 120 Hz as the mirrors are not capable of following 'faster' signals. When they are not removed or suppressed, signal components of 120 Hz and higher cause additional unnecessary heating of the mirror system.

High-voltage laser supply

As already mentioned, the present laser supply is somewhat different from most earlier designs because it is off the beaten track formed by mains transformers, cascades and series resistors. The high-voltage supply presented here has an input requirement of 10 to 16 V, and thus can be used independently of the mains. The current consumption is modest at about 1 A, allowing the unit to be powered from a car battery, a set of NiCd batteries, or a suitably rated mains adapter. The proposed PSU is a switch-mode DC-DC converter with current stabilisation to ensure that the laser tube supplies the maximum light intensity without having its lifetime degraded.

The circuit diagram of the power supply is given in Fig. 3. The input voltage is connected to reservoir capacitor C5 via a polarity

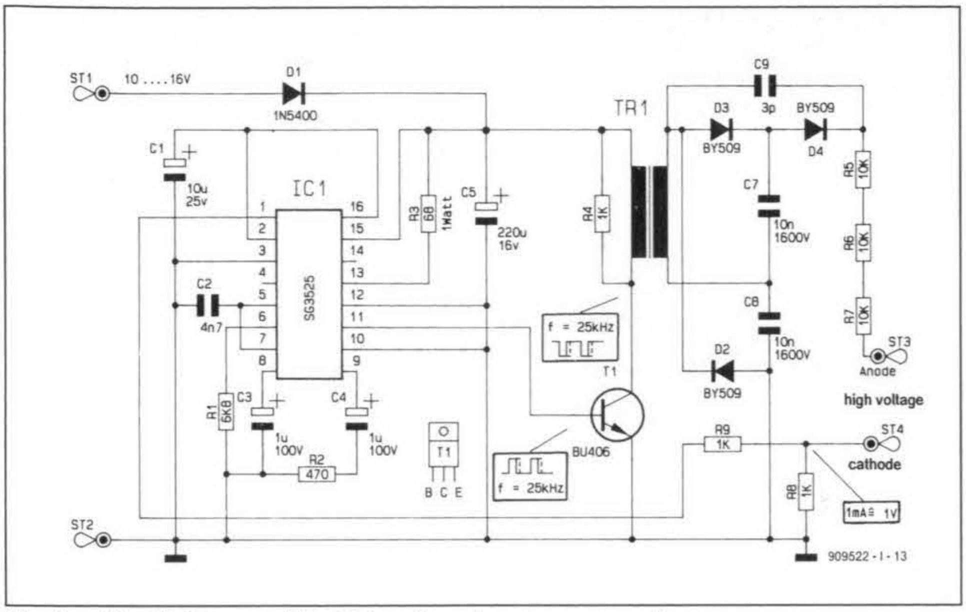


Fig. 3. Circuit diagram of the high-voltage laser power supply.

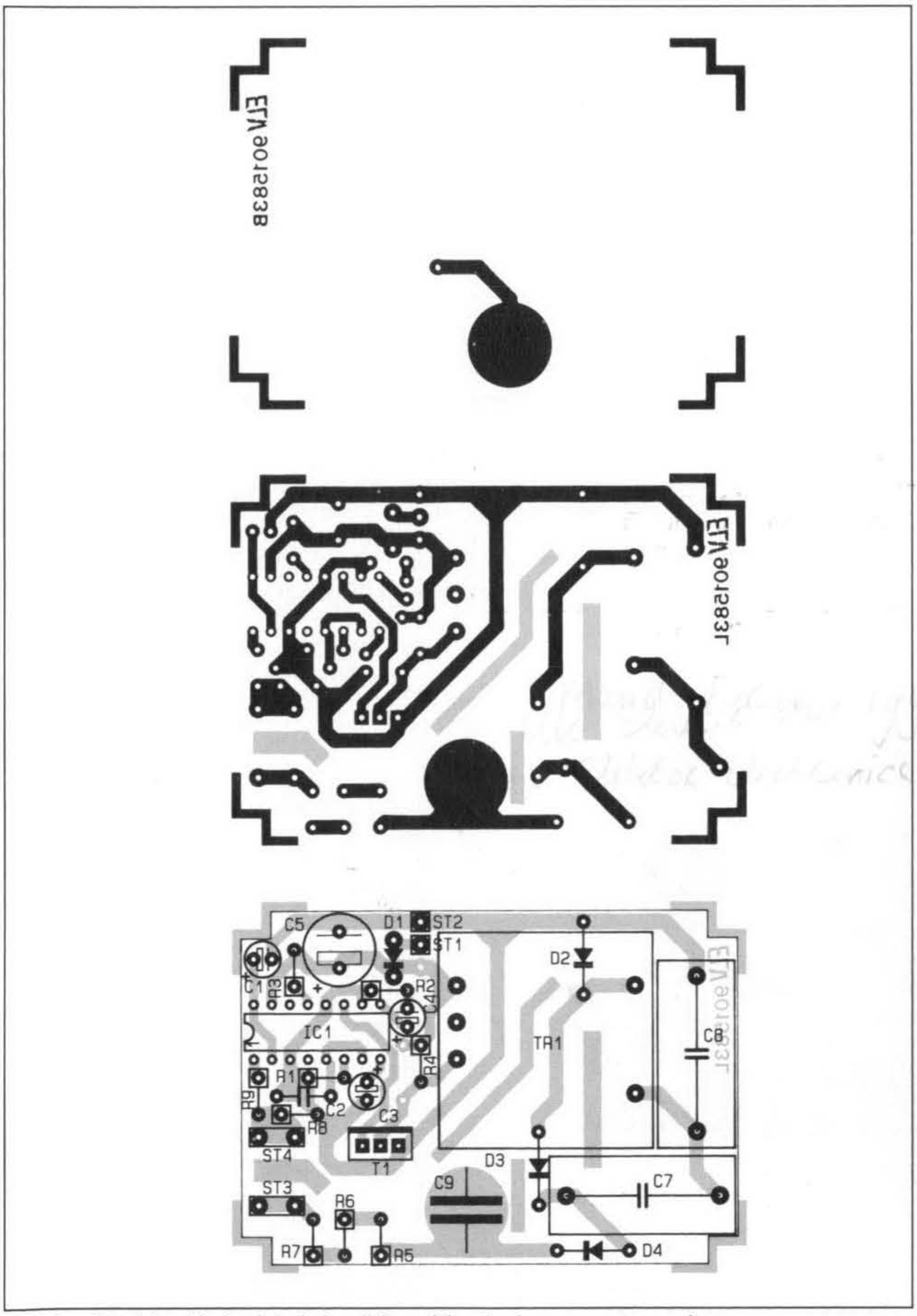


Fig. 4. Double-sided printed circuit board for the laser power supply.

reversal protection, D1. The positive terminal of C5 is connected to the primary winding of a ferrite-core transformer, Tr1. The current through this winding is switched by transistor T1. The transistor is driven by a pulse-width modulated (PWM) rectangular signal that allows the voltage induced in the secondary winding of Tr1 to be controlled. The transformer secondary is connected to a special rectifier/voltage multiplier. First, the secondary voltage is rectified by diodes D2 and D3 for the negative and positive half-cycles respectively. This means that the total voltage across reservoir capacitors C7 and C8 is about two times the peak value of the alternating voltage supplied by the secondary winding of Tr1. Since D4 conducts and C9 is relatively small, the high voltage is maintained even after ignition of the laser. The laser operating voltage is then 1,400 to 1,500 V. When ignition has not yet taken place, i.e., when the tube draws no current as yet, D4 and C9 plus D3 and C7 form a voltage doubler. This arrangement gives a total voltage multiplication of six times. Since the laser does not draw current at this stage, the voltage control system enables the maximum output voltage of about 10 kV to be achieved briefly. Immediately after ignition of the laser, the output voltage drops to the nominal operating value. This happens because of the small value of C9, and the action of the voltage control system.

Unfortunately, the voltage control circuit does not obviate a current limit resistor in the HV output line. However, the resistor used here has half the value of the one used in a conventional cascade-based laser supply, and introduces a power loss of only 0.75 W instead of the more usual 1.5 W or more.

COMPONENTS LIST

Resistors:

1	6kΩ8	R1
1	470Ω	R2
1	68Ω 1W	R3
3	1kΩ	R4;R8;R9
3	10kΩ	R5;R6;R7

Capacitors:

1	10µF 25V	C1
1	4nF7	C2
2	1µF 100V	C3;C4
1	220µF 16V	C5
1	1nF	C6
2	10nF 1,600V	C7;C8

Semiconductors:

1	1N5400	D1
3	BY509	D2;D3;D4
1	BU406	T1
1	SG3525	IC1

Miscellaneous:

1 ferrite-core HSP transformer Tr1
2 PCB terminal block
1 ABS enclosure
40cc epoxy resin (compound)

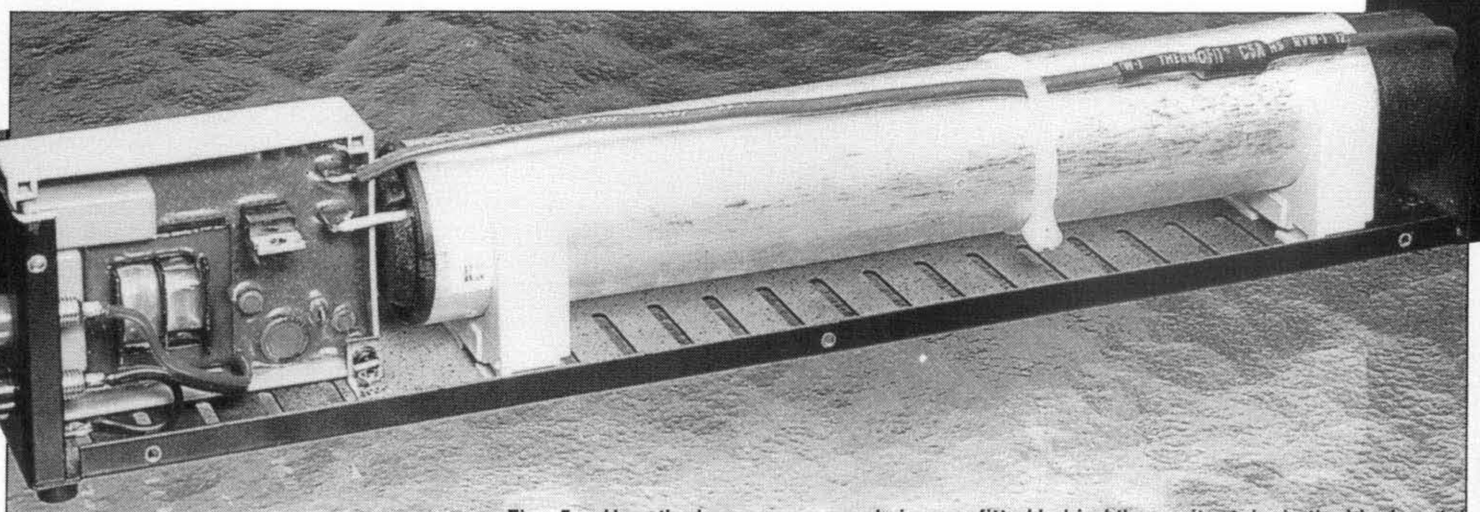


Fig. 5. Here the laser power supply is seen fitted behind the exciter tube in the black metal cabinet supplied with the kit.

The power supply has a stabilized output current of 5.1 mA to keep the laser beam intensity at the maximum level without shortening the tube life. The regulation circuit to achieve this is contained in IC1. The current consumption of the laser is measured by means of the voltage across R8. This voltage is passed to pin 1 of IC1 via R9. Pin 2 of the same IC is held at a constant voltage of 5.1 V. The two input voltages are compared, and the difference is used to control a rectangular wave generator with a variable duty factor (pulse/pause ratio). When R8 measures no current, i.e., when the laser has not yet ignited, the duty factor of the PWM signal is 1 (i.e., a square wave is produced). Hence, the transformer supplies the maximum secondary voltage. The frequency of the PWM signal is fixed at about 25 kHz, allowing a transformer to be used of a size much smaller than a mains transformer with an equal power rating.

It is well known that laser exciters have a very low efficiency. The laser tube used here is no exception: it requires about 10 watts of input power to supply about 2 milliwatt of laser light. That's an efficiency of 0.02%, and you will like to hear that the PSU used to power the laser is at least reasonably efficient.

Construction of the PSU

Although the laser PSU supplies only small currents, every possible care must be taken to prevent that any part of the circuit can be touched while it is operational. Remember: the maximum output voltage is about 10 kV, and the normal output voltage about 1,500 V. Both voltages can cause nasty and possibly harmful electrical shocks, so be careful.

Another source of danger is the light produced by the laser proper. **Never look direct into the laser beam; your eyes may be permanently damaged. Never point the laser beam at a person.** At some distance of the exciter, the beam is less harmful, but still, never look into it. During laser shows in discotheques and the like, the beam moves continuously and is unlikely to be harmful as the eye is 'hit' very briefly and at a considerable distance from the exciter.

The design of the printed-circuit board for the high-voltage PSU is given in Fig. 4. Remarkably, the board has only one, round copper area at the component side. This area forms one 'plate' of high-voltage capacitor C9, which is etched on the board. The single copper area at the component side is connected to the circuit by soldering the anode terminal of diode D3 at both sides of the PCB. The layout of the solder side of the PCB (Fig. 4) shows three grey areas that must be cut out with the aid of a jig-saw (fret-saw) to prevent arcing.

The construction of the PSU board is straightforward — simply follow the components list and the component overlay printed on the board. To prevent arcing, cut all soldered component terminals to a length of 1.5 mm or less. Do not cut its terminals and fit transistor T1 as high as possible above the PCB surface, but at the same time make sure its terminals are soldered securely. This leaves the metal tab of the transistor in contact with air after the supply is cast in a moulding compound.

The low-voltage connections, ST1 and ST2, are made either via solder terminals or via wires soldered direct to the copper pads. A wire length of 10 cm is sufficient when the PSU is to be built into the exciter cabinet. Use wire of a cross-sectional area of 0.4 mm^2 or greater. If you use stranded wire, make sure all individual wires pass through the PCB hole. One wire, however thin, may cause a short-circuit!

The last part to be fitted onto the PCB is the ferrite transformer (note: you must have fitted D2 and D3 beforehand, since they are located underneath the transformer). First, however, cut off the transformer terminals that are not connected to a winding. These terminals are easily identified.

Once the transformer has been fitted on the board, the PSU is ready for testing. Note, however, that this must never be done without a load connected. So, before switching on, connect either the laser exciter unit or a 250 kΩ resistor made from 25 series-connected 10 kΩ resistors with a minimum power rating of 0.33 W. When the laser ex-

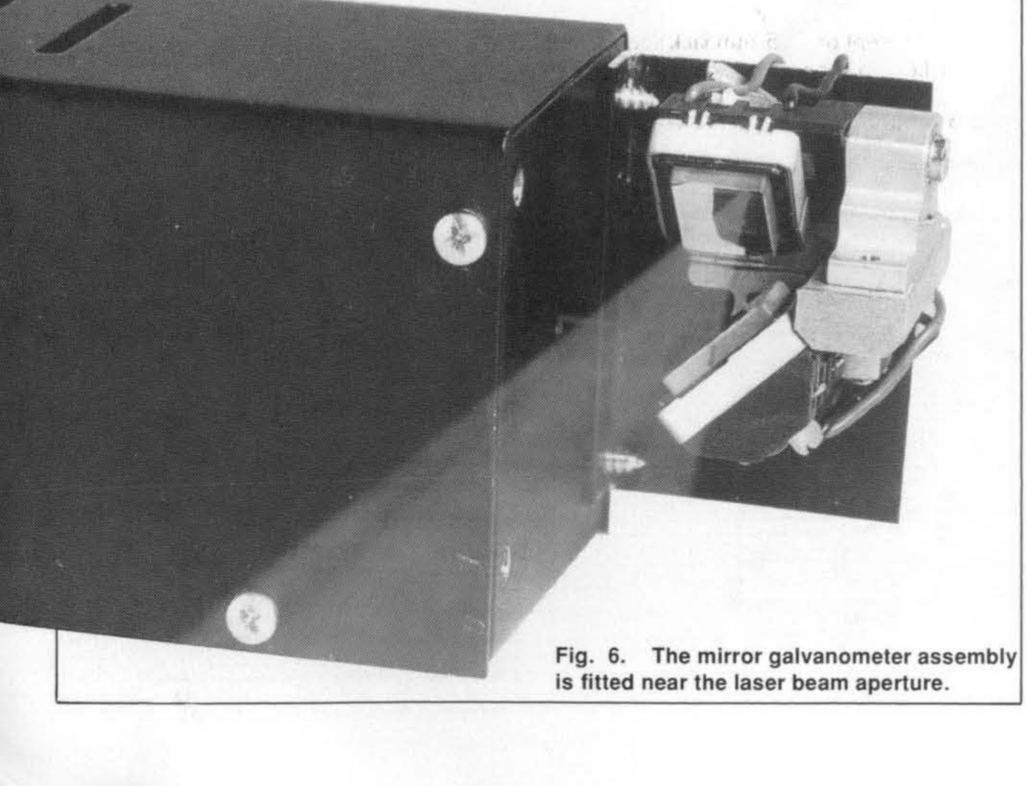


Fig. 6. The mirror galvanometer assembly is fitted near the laser beam aperture.

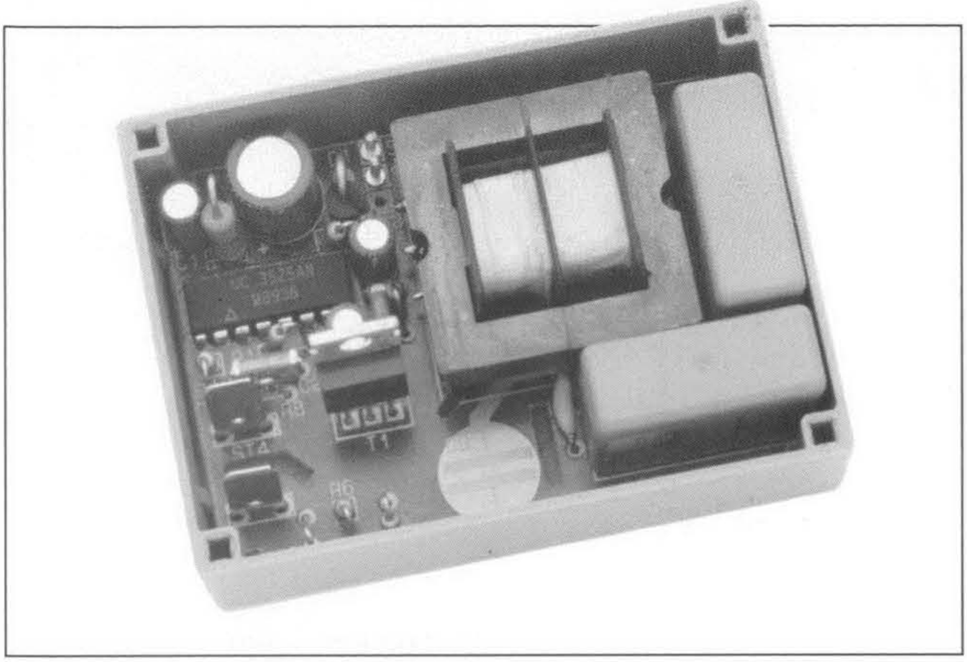


Fig. 7. Completed power supply board photographed before it was cast into a moulding compound to protect it against dust and moisture.

citer is used, connect the anode wire to PSU terminal ST3. The anode wire is black, and has a moulded in-line resistor. Connect the cathode of the laser to ST4. Switch on the PSU and measure the voltage across R_8. Switch off immediately and investigate the circuit for errors if this voltage is not between 5.0 and 5.2 V.

Although slightly cumbersome to make, the 'dummy' resistor will withstand unintended rough treatment far better than the laser exciter when it comes to testing the PSU. When the supply functions properly, its output voltage is between 1,300 V and 1,500 V. If your voltmeter can not measure such high voltages, connect it to a suitable point in the composite resistor and calculate the output voltage. Note, however, that the internal resistance of your voltmeter may affect the measurement considerably. If you are unable to measure the high voltage, rest assured that a voltage of 5.1 V ±1% across R_8 is a reliable indication that the current control will function all right when the laser is connected.

Connect the laser and check that it ignites almost instantly after the PSU is switched on. If not, and if the PSU is known to supply the correct (lower) operating voltage to the 250 kΩ dummy load, the minimum voltage to ignite the laser is probably not reached. Run a thorough check on the PSU circuit. Have you soldered the anode of D_3 at both sides of the PCB?

Once the power supply has been tested successfully, it is cast into a moulding compound. This is done to prevent arcing as a result of dust or humidity. Place the enclosure horizontally on a flat surface, and lower the PSU board into the box (the box and the moulding compound are contained in the kit supplied by ELV). Next, lift the PCB about 1 cm at the side of C_9 to allow the air to escape as the compound is poured into the box. The 2-component resin compound (approx. 40 cc) is mixed at a ratio of 1 part of hardener to 5 parts of resin, or at the ratio stated in the manufacturer's instructions. Mix the compound until a uniform colour is achieved. Next, pour the compound into the box, at the side of ST1 and ST2 (which have been connected to wires beforehand!). Allow the air to escape as indicated. Resistors R_5, R_6 and R_7 must be covered by the compound, but ST3 and ST4 must remain accessible. Allow the compound to set overnight.

Assembly

The photograph in Fig. 5 shows the high-voltage PSU fitted at the rear of the exciter cabinet. The PSU is secured to the rear plate of the cabinet with the aid of two small brackets. Fit a spring washer between the cabinet and each bracket. One of the screws used to secure the bracket also has a solder eye which is wired to the ground connector (the lower socket). The PSU must be fitted just above the bottom plate of the laser cabinet.

The rear plate of the laser enclosure is drilled to accept one 3.5-mm jack socket and two sockets for banana plugs. Connect the lower socket and the ring of the jack socket to ST2 (ground) on the PSU board. Connect the upper socket and the centre contact of the jack socket to ST1 (positive supply) on the PSU board.

Before fitting the plastic clamps for the laser tube, remove the protruding pieces from the sides. Fit the clamps with small screws and nuts, but do not tighten them as yet. Slide the tube into the clamps until its beam aperture is flush with the front side of the enclosure. Then push the tube about 0.5 mm deeper into the cabinet and tighten the clamps. The tube is preferably fitted to give horizontally polarized laser light. This is done to prevent reflections on smooth, non-metallic, horizontal surfaces (the so-called Brewster effect). The polarization is horizontal if the holes in the tube are positioned horizontally (i.e., to the left and the right of the tube).

Solder the black anode wire (the one with the internal resistor) to the upper pin on the PSU board (ST3), and the cathode wire to the lower pin (ST4). To ensure the greatest distance between the connections, solder the anode wire to the top side of ST3, and the cathode wire to the lower side of ST4.

The mirror galvanometer assembly is fitted to the laser cabinet with the aid of four screws that are inserted from the inside of the cabinet into the mirror unit. The photograph in Fig. 6 shows the mirror assembly attached to the laser cabinet, with the cover plate removed. The mirror assembly has two screws that enable the rest position of the mirrors to be adjusted.

Power supply and cooling

Any power supply with an output of 12 V d.c. at a nominal current of 1 A can in principle be used to power the laser. The safest and cheapest power supply is probably a ready-made mains adapter. In many cases, however, mains adapters do not meet the specifications as regards output current. The typical symptoms of an overloaded mains adapter are extreme heating and a slowly falling output voltage. Obviously, the mains adapter used to power the laser must be capable of supplying 1 A at 12 V for extensive periods. If you have doubts about the performance of your mains adapter, use a more powerful type, or build a separate 12 V supply.

Given its very low efficiency we can safely say that all of the input power of the laser (approx. 10 W) is dissipated as heat. To keep the operating temperature of the laser tube within safe limits, there should be ample airflow around the device. The laser cabinet shown in the photographs has slots in the side panels to prevent heat building up inside. ❑

Continued next month.

May 1991

A complete kit of parts for the laser is available from the designers' exclusive worldwide distributors:

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE

Telephone: +33 82837213
Facsimile: +33 82838180

The type code of the kit that contains the laser power supply, the laser exciter tube and the metal cabinet is **LPS12**. The type code of the kit that contains the mirror galvanometer assembly is **LA90**.

yet, D4 and C9 plus D3 and C7 form a voltage doubler. This arrangement gives a total voltage multiplication of six times. Since the laser does not draw current at this stage, the voltage control system enables the maximum output voltage of about 10 kV to be achieved briefly. Immediately after ignition of the laser, the output voltage drops to the nominal operating value. This happens because of the small value of C9, and the action of the voltage control system.

Unfortunately, the voltage control circuit does not obviate a current limit resistor in the HV output line. However, the resistor used here has half the value of the one used in a conventional cascade-based laser supply, and introduces a power loss of only 0.75 W instead of the more usual 1.5 W or more.

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Resistors:		
1	6kΩ8	R1
1	470Ω	R2
1	68Ω 1W	R3
3	1kΩ	R4;R8;R9
3	10kΩ	R5;R6;R7
Capacitors:		
1	10µF 25V	C1
1	4nF7	C2
2	1µF 100V	C3;C4
1	220µF 16V	C5
1	1nF	C6
2	10nF 1,600V	C7;C8
Semiconductors:		
1	1N5400	D1
3	BY509	D2;D3;D4
1	BU406	T1
1	SG3525	IC1
Miscellaneous:		
1	ferrite-core HSP transformer	Tr1
2	PCB terminal block	
1	ABS enclosure	
40cc	epoxy resin (compound)	

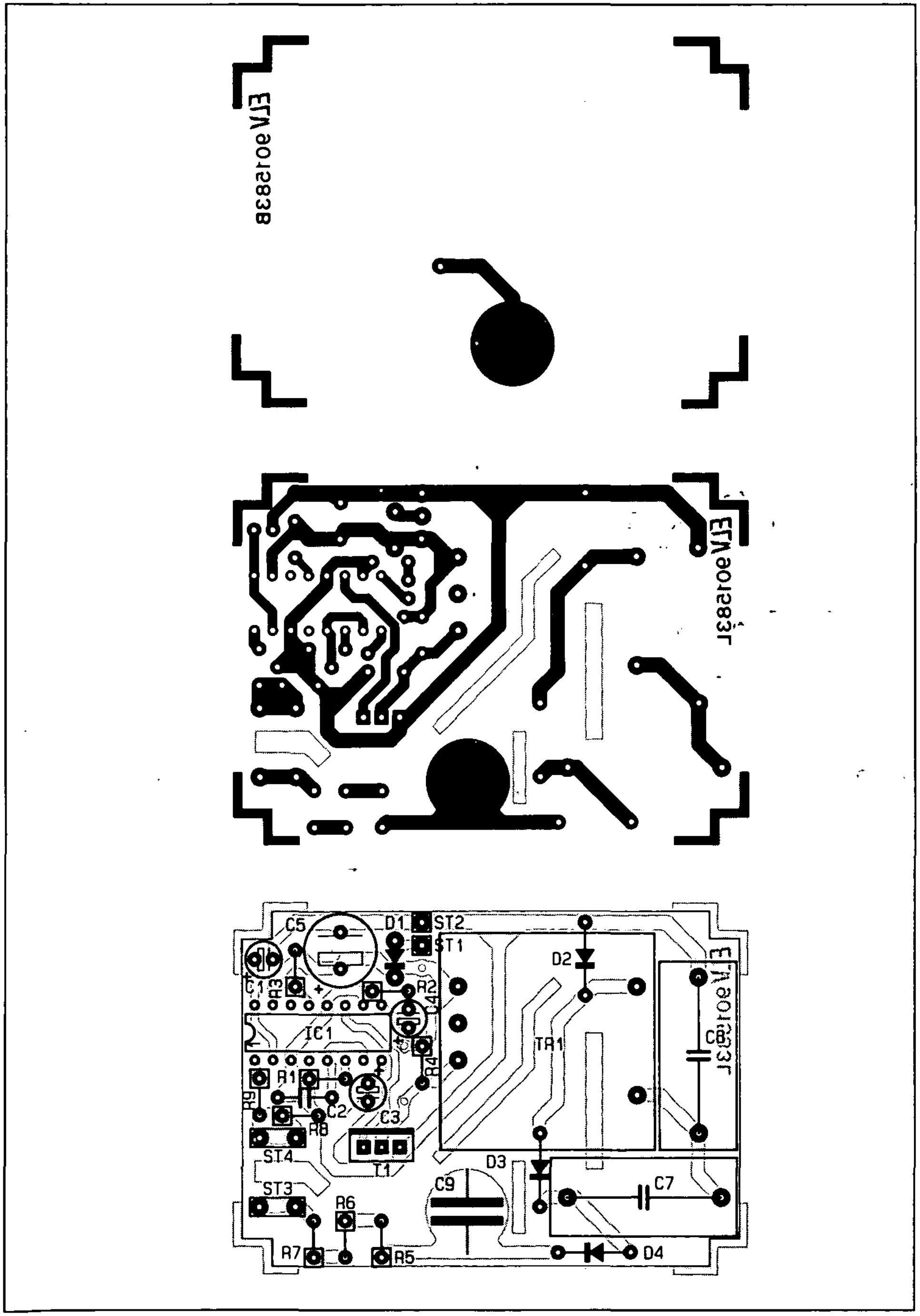


Fig. 4. Double-sided printed circuit board for the laser power supply.

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.

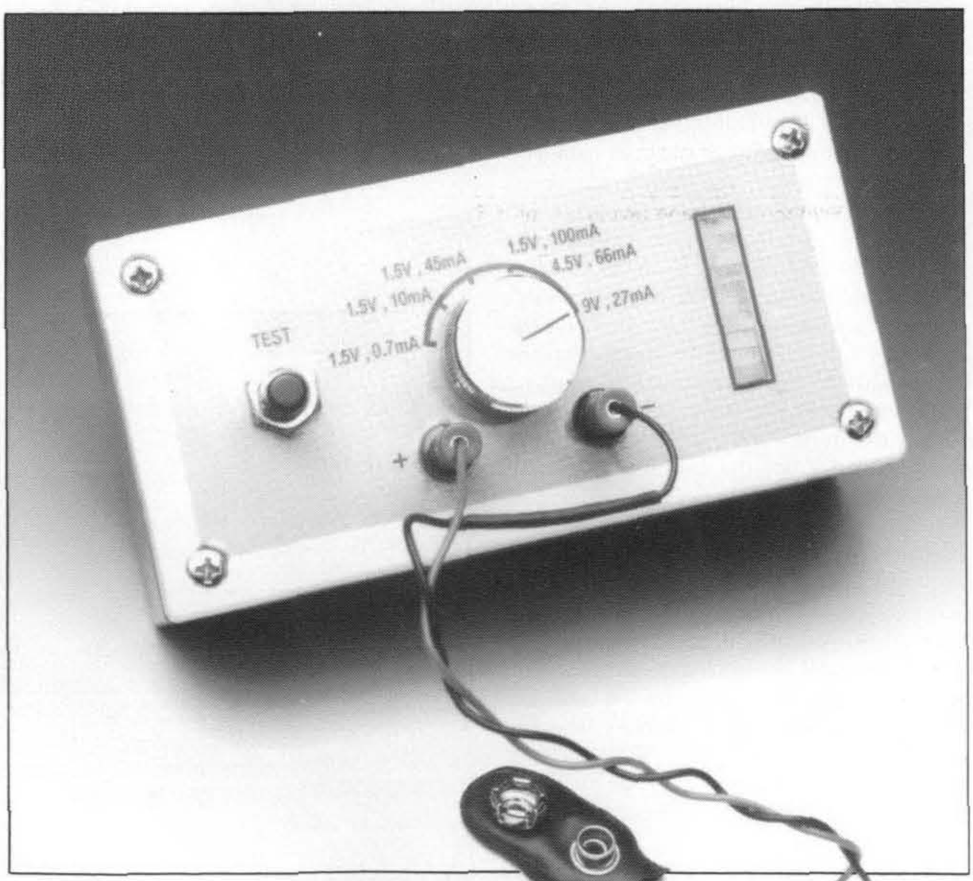
BATTERY TESTER

This month we describe a useful little instrument that gives a clear condition indication for the most popular types of dry battery. Based on one integrated circuit and a coloured LED bar that functions as a readout, the tester is both inexpensive and simple to build.

L. Lemon

RUNNING a quick condition check on the batteries before using portable equipment can help prevent a lot of frustration. The professional photographer, for instance, can not rely on his good fortune when he starts to use his flasher to make, say, a series of wedding photographs. He knows that there can be no excuses for mishaps: the flasher must work under all conditions, and must have fresh, tested, batteries. Similarly, if rechargeable batteries are used, they must be fully topped up and known to have their full capacity.

There are, however, instances where a partly exhausted battery can be used without problems. But then again, it is useful to have at least an indication of how long we can rely on the battery to supply its nominal voltage. For this purpose we can use the present tester. It has a number of ranges with different test currents for popular dry batteries of 1.5 V, 4.5 V and 9 V. On pressing the TEST button on the instrument, a VU-meter-like read-out gives an unambiguous full/usable/flat indication. No difficult-to-read scales or expensive moving coil meters here: just three colours: red for 'flat', orange for 'usable' and green for 'full'.



Loading, e.m.f. and measuring

If you have ever attempted to test a battery simply by measuring its voltage with a multimeter, you may have noted that the battery in question may be virtually exhausted even though its voltage is, say, less than 10% below the nominal value. The explanation for this is that degradation of battery capacitance is a matter of increasing internal resistance rather than decreasing cell voltage.

A multimeter with a very high internal resistance gives a corresponding indication of the e.m.f. (electromotive force) of the battery. We must hasten to add, however, that the term e.m.f. applies strictly to an unloaded source of electrical energy but is sometimes erroneously used as being equivalent to a potential difference. The e.m.f., E, of a battery will supply a current I to an external resistance R:

$$E = I\,(R + r) \qquad \text{[volt]}$$

where r is the internal resistance of the battery. From this equation it is simple to see

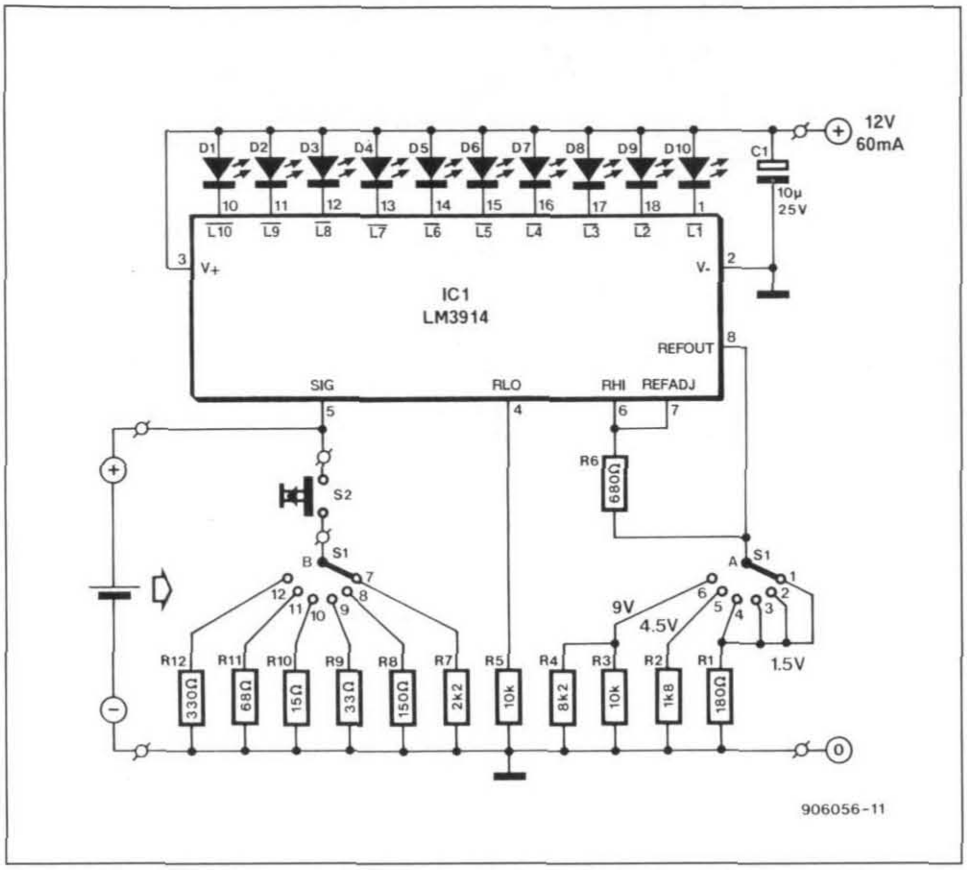


Fig. 1. The circuit diagram of the battery tester is basically a standard application of National Semiconductor's LM3914 LED bargraph driver IC.

that *E* may correspond closely to the nominal battery voltage as long as *r* is small with respect to *R*. Returning to the above battery test using a voltmeter, a digital multimeter will have a typical input resistance of several megohms, so *r* is bound to be small in any case, even it is, say, five times the value specified for a 'fresh' battery. The upshot is that the capacity (not the e.m.f.) of a dry battery can only be measured when the load resistance equals, say, ten times the internal resistance of the battery. If the battery is capable of supplying its nominal voltage at the resultant load current, its internal resistance is still relatively small. If the battery is exhausted, the internal resistance will have risen to a value that is no longer small with respect to the load resistance.

The circuit: a single-chip voltmeter

The circuit in Fig. 1 is a kind of VU (volume-unit) meter based on a single IC and LEDs D1-D10. The load resistors that set the battery test current, R7-R12, are selected with a rotary switch. The battery is effectively loaded when push-button S2 is pressed.

The dual-pole rotary switch in the circuit, S1, enables one of six load resistors to be selected. Each load resistor has a value geared to a particular type of battery, of which the capacity, as you probably know, depends on size, construction and chemical composition. The first four ranges are for 1.5 V batteries. The load currents are for button cells, penlight batteries (IEC-R6), IEC-R14 ('baby') batteries and IEC-20 ('mono') batteries. The selected load resistor is connected across the battery terminals when S2 is pressed. Assuming that the relevant battery is full, it supplies 0.7 mA (button cell), 10 mA ('penlight'), 45 mA ('baby') or 100 mA ('mono'). The test current for 4.5 V 'power pack' batteries is 66 mA, and 27 mA for 9-V (PP3 or IEC 6F22) batteries. Since the battery is connected permanently to the voltmeter, you can instantly see the difference between the open-circuit voltage (virtually the e.m.f.) and the loaded voltage, which is indicated when S2 is pressed. The larger the difference between the open-circuit voltage and the loaded voltage, the smaller the battery capacitance. Note that the term open-circuit voltage is strictly incorrect here since the battery is connected to the RLO and SIG pins of IC1. However, the LM3914 has such a high resistance between these pins that it is safe to say that the battery is not loaded. Hence, the test current flows only when S2 is pressed.

It should be noted that the tester can only be used with batteries or cells that exhibit a gradually falling voltage curve, as shown in Fig. 2. These batteries include carbon-zinc and alkali-manganese types, as well as nickel-cadmium types (which are rechargeable). The tester is not suitable for battery types whose voltage remains almost constant over the hours of service, and drops suddenly at the end. This type of voltage characteristic is illustrated in Fig. 2 for lithium, silver-oxide and mercury batteries.

Circuit description: introducing the LM3914

As shown in the block diagram in Fig. 3, the LM3914 from National Semiconductor has

Range	LED									
	D1	D2	D3	D4	D5	D6	D7	D8	D9	D10
	red	orange		green						
1.5 V	0.86	0.96	1.04	1.13	1.21	1.29	1.38	1.46	1.55	1.63
4.5 V	2.58	2.83	3.05	3.31	3.57	3.82	4.07	4.33	4.57	4.82
9.0 V	5.3	5.8	6.3	6.9	7.4	7.9	8.5	9.0	9.5	10.2

Table 1. Switch-on voltage levels for the LEDs on the battery tester.

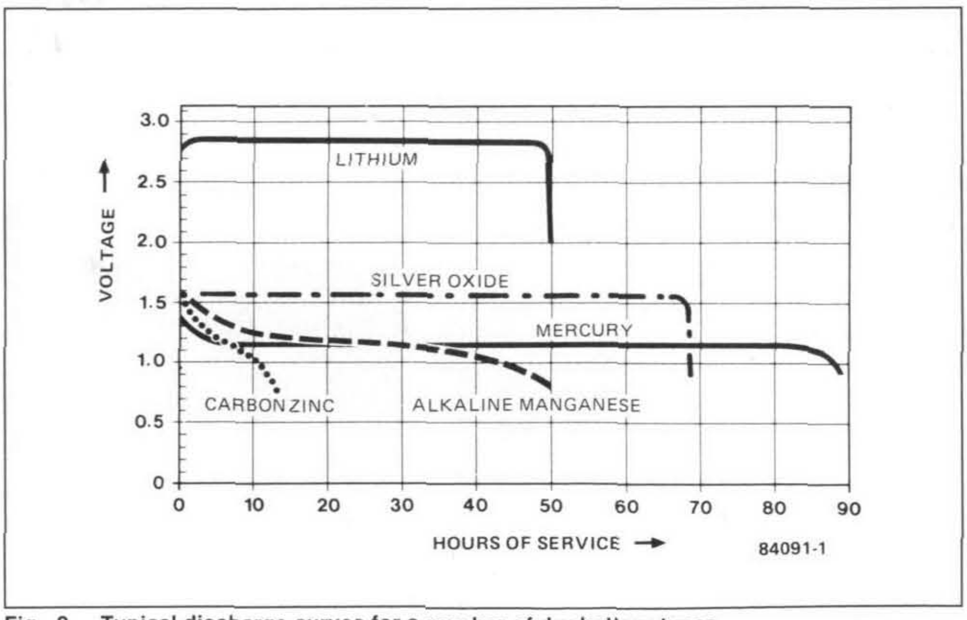


Fig. 2. Typical discharge curves for a number of dry battery types.

an internal network of series resistors connected between the RLO and RHI terminals (pins 4 and 6 of the IC). In the present application, the RLO terminal is connected to ground via R5, while the RHI input is held at a fixed voltage derived from the IC's on-board 1.25-V reference voltage source. The output of the reference is connected to IC pin 7, marked REF OUT. Hence, the internal resistor network is supplied with an accurately defined voltage. This, in turn, means that there are fixed voltages at the taps on the series network (i.e., at the junctions of the resistors that form the ladder network). There are 10 taps on the ladder, and each of these is internally connected to the +input of an associated comparator. The other inputs of the 10 comparators are connected to the output of an internal buffer opamp, whose input is connected to IC pin 5, marked SIG (for signal). The main function of this opamp is to protect the 10 comparators against reversed or too high input voltages.

Because of the ladder network in the LM1914, the comparator which is closest to the RLO input will toggle at the lowest input voltage. As the input voltage rises, more comparators toggle, until the 'highest' one, i.e., the one associated with the top of the ladder at the RHI input, changes state. The comparator outputs are connected to an internal decoder (not shown in the block diagram) with active-low outputs marked L1 to L10 in the circuit diagram. Here, the decoder is set to operate in the 'dot' mode, which means that only one LED lights at a time, depending on the input voltage applied to the tester.

The stabilized voltage at pin 7 of the LM3914 can be changed within certain limits by connecting pin 8 to a voltage divider. Note, however, that the voltage between pins 8 and 6-7 is fixed at 1.25 V. Thus, the indication range of the voltage meter can be changed by giving resistors R1 to R4 appropriate values. In the present circuit, this is achieved with the aid of a rotary switch which creates the three voltage ranges, 1.5 V, 4.5 V and 9 V. Table 1 shows the voltages at which the LEDs of the VU-meter light. The red LED lights when the battery is flat, one of the orange LEDs when the battery is usable, and one of the green LEDs when the battery is full, covering a range from usable to fully topped.

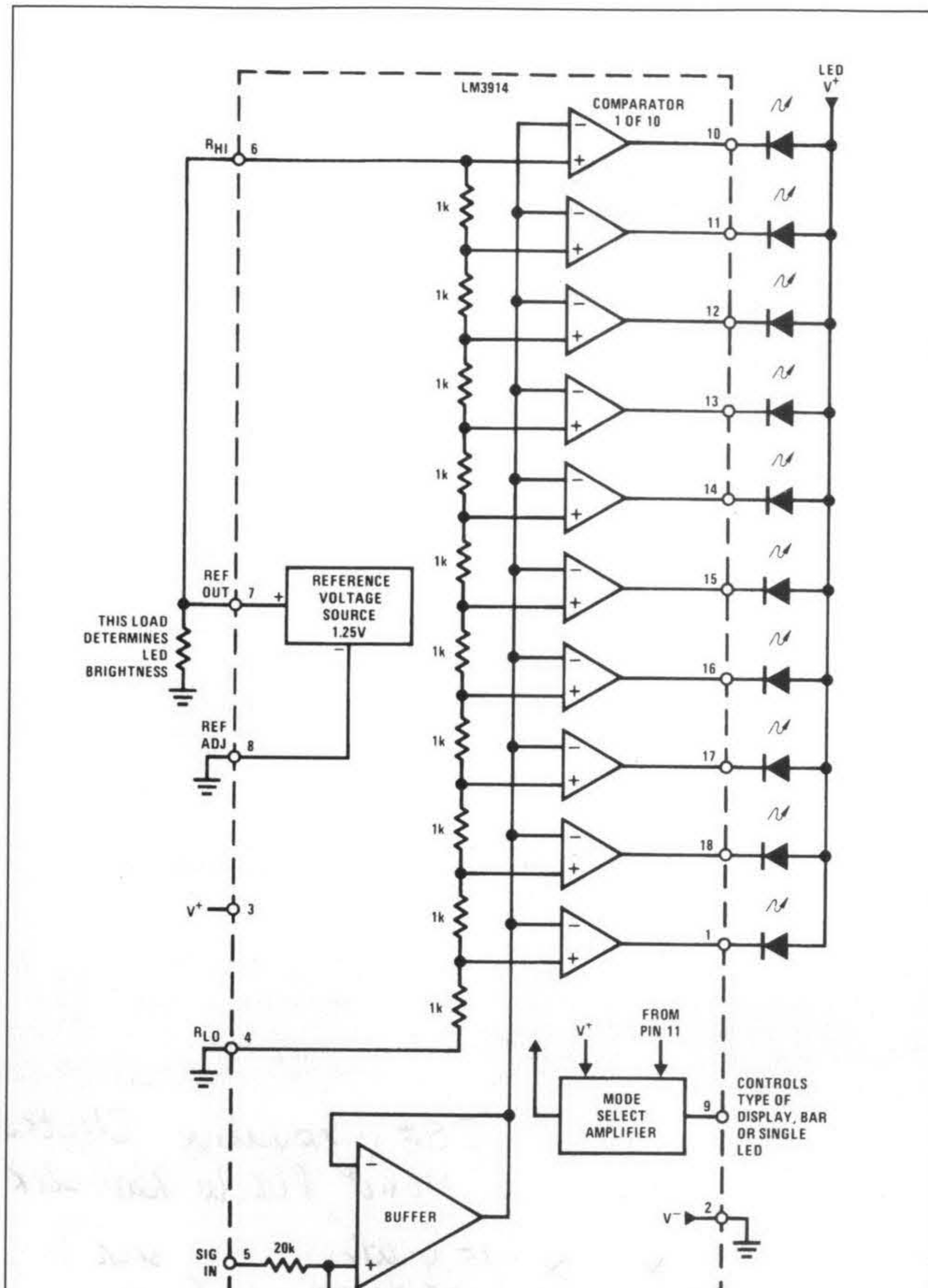
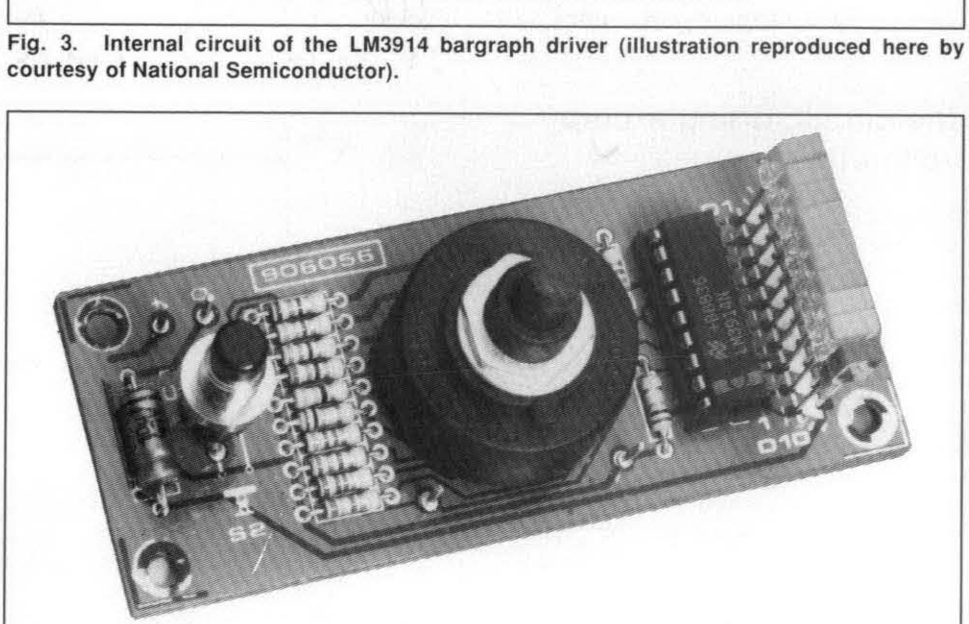


Fig. 3. Internal circuit of the LM3914 bargraph driver (illustration reproduced here by courtesy of National Semiconductor).

Building the battery tester

The track layout and the component mounting plan shown in Fig. 4 are used to produce a printed-circuit board for the battery tester (ready-made PCBs are unfortunately not available for this project). The internal construction of the prototype of the battery tester is shown in the Fig. 5. The 'TEST' button, S2, is connected to the board via two short wires before it is secured to the front panel of the ABS enclosure. Do not cut the terminals of the 10 LEDs before you solder them onto the board. The distance between the PCB and the front panel of the enclosure is determined by the mounting height of the LEDs. As shown in Fig. 5, the PCB is secured to the



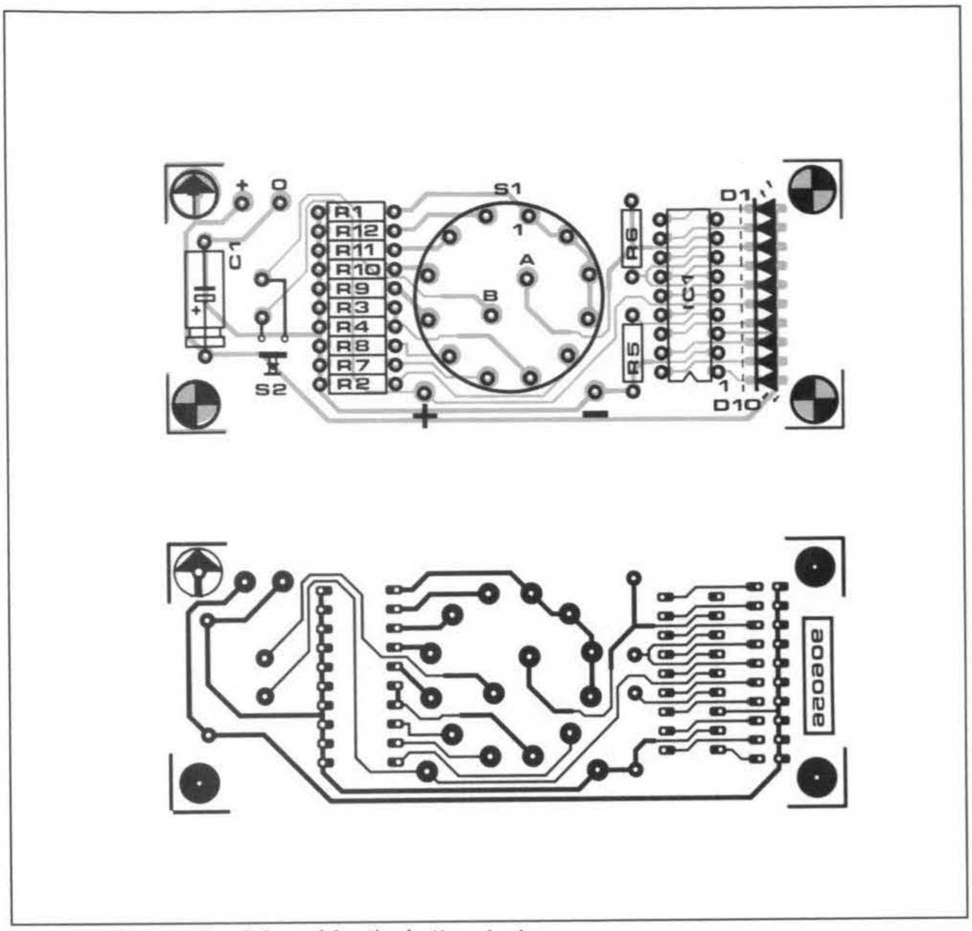


Fig. 4. Printed-circuit board for the battery tester.

COMPONENTS LIST

Resistors:

1	180Ω	R1
1	1kΩ8	R2
2	10kΩ	R3;R5
1	8kΩ2	R4
1	680Ω	R6
1	2kΩ2	R7
1	150Ω	R8
1	33Ω	R9
1	15Ω	R10
1	68Ω	R11
1	330Ω	R12

Capacitors:

1	10µF 25V	C1

Semiconductors:

7	green rectangular LED e.g., LGB480F	D1-D7
2	orange rectangular LED e.g., LYB480H	D8;D9
1	red rectangular LED e.g. LSB480H	D10
1	LM3914	IC1

Miscellaneous:

1	two-pole six-way rotery switch for PCB mounting	S1
1	push-to-make button	S2
1	ABS enclosure approx. dimensions: 110×45×55 mm	

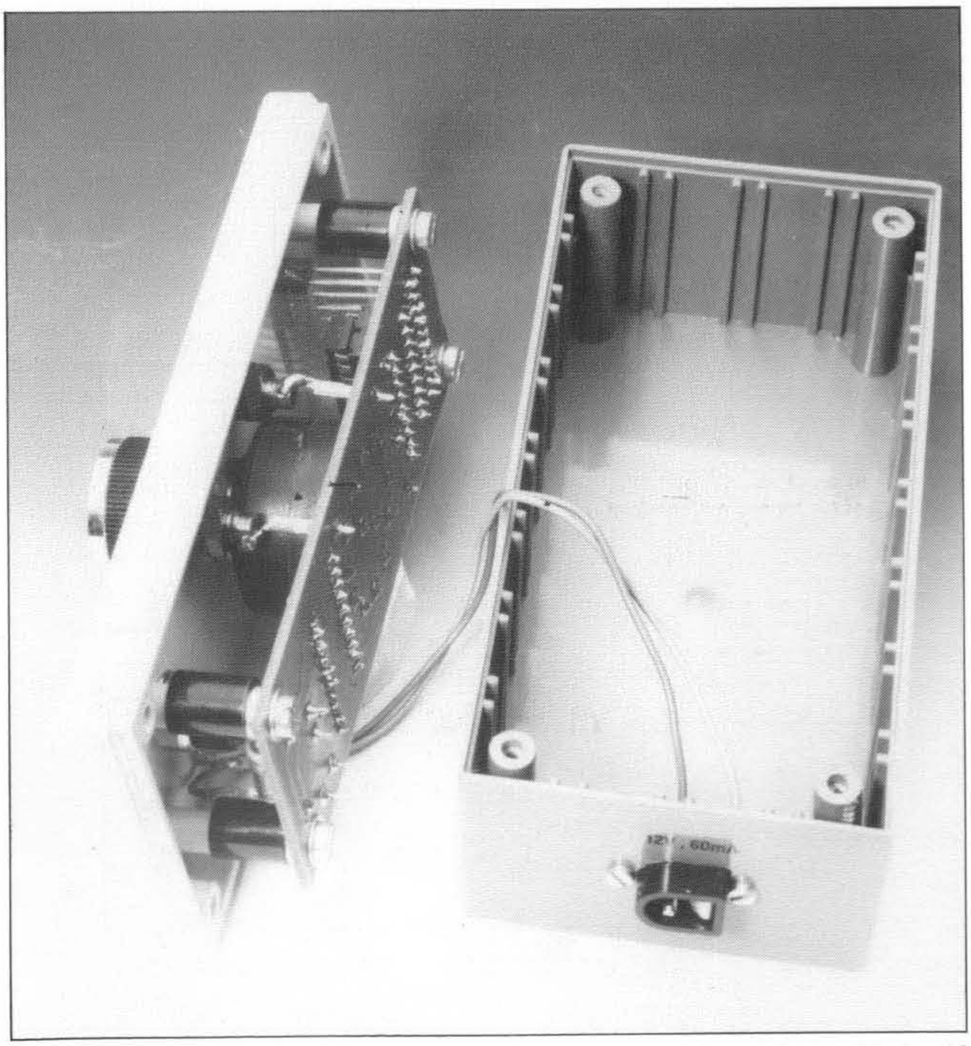


Fig. 5. As shown here, the PCB is secured to the front panel of the enclosure with the aid of four long M3 screws and plastic PCB spacers.

front panel of the box by four plastic spacers. Cut a slot in the front panel, and adjust the position of the LEDs such that their tops are just flush with the front panel surface.

The rotary switch used is a double-pole six-way type with solder pins for PCB mounting. The range and test current indications on the front panel are made with the aid of rub-down symbols as shown in Fig. 6. The tester is powered by a small mains adapter with a 12-V d.c. output which need not be regulated. This adapter is connected to the circuit via a small socket of the type used on portable cassette recorders and pocket calculators — see Fig. 5.

Finally, LED D1 lights when the tester is powered but not connected to a battery. It is, therefore, a perfect on/off indicator! ■

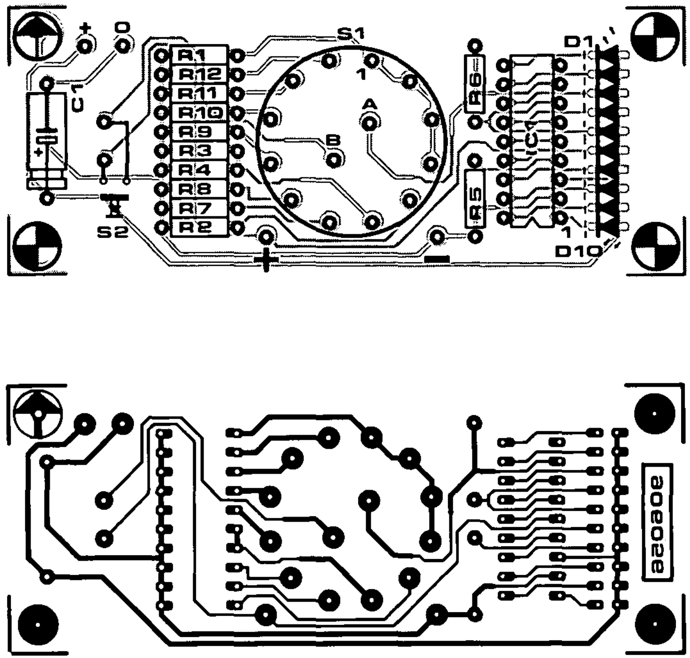


Fig. 4. Printed-circuit board for the battery tester.

COMPONENTS LIST

Resistors:

1	180Ω	R1
1	1kΩ8	R2
2	10kΩ	R3;R5
1	8kΩ2	R4
1	680Ω	R6
1	2kΩ2	R7
1	150Ω	R8
1	33Ω	R9
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Miscellaneous:

1	two-pole six-way rotery switch for PCB mounting	S1
1	push-to-make button	S2
1	ABS enclosure approx. dimensions: 110×45×55 mm	

VIDEO A-D/D-A CONVERTER

PART 1: INTRODUCING THE ICs

Although fast 8-bit video converters are available from a variety of manufacturers, their practical application has so far been over the head of the average electronics hobbyist with an interest in video signal processing. Fortunately, that situation has come to an end with the introduction of a number of simple to use A-D and D-A converter ICs from Philips Components. Two of these ICs, the TDA8708 ADC and the TDA8702 DAC, form the heart of an advanced video converter described in this two-instalment article. This month we discuss the basic operation of ADC and the DAC, followed next month by a constructional project intended to get you going with video encoding/decoding experiments, digital video processing, sync locking techniques, etc.

P. Godon (Philips Components, Paris)

TDA8708 analogue-to-digital converter

The TDA8708 contains more than just a fast ADC — in fact, it should be referred to as a 'video analogue input interface'. The internal diagram of the TDA8708, which is also available in a surface-mount assembly package (suffix -T), is given in Fig. 1. Clearly, the ADC is but one of a number of functional blocks contained in the IC. Each of the three analogue video inputs, VIN0, VIN1 and VIN2 may be driven by a CVBS (chrominance-video-blanking-synchronization) signal of an average level between 0.45 V and 1.6 V measured with respect to the analogue ground, AGND. The typical input capacitance of the VINx inputs is about 1 pF, while the input impedance lies between 10 kΩ and 20 kΩ (input capacitance and input impedance values without external components). The video input selection is effected by applying logic levels to the I0 and I1 inputs of the chip as shown below:

Table 1. Input channel selection

I1	I0	VINx
0	0	0
0	1	1
1	0	2
1	1	2

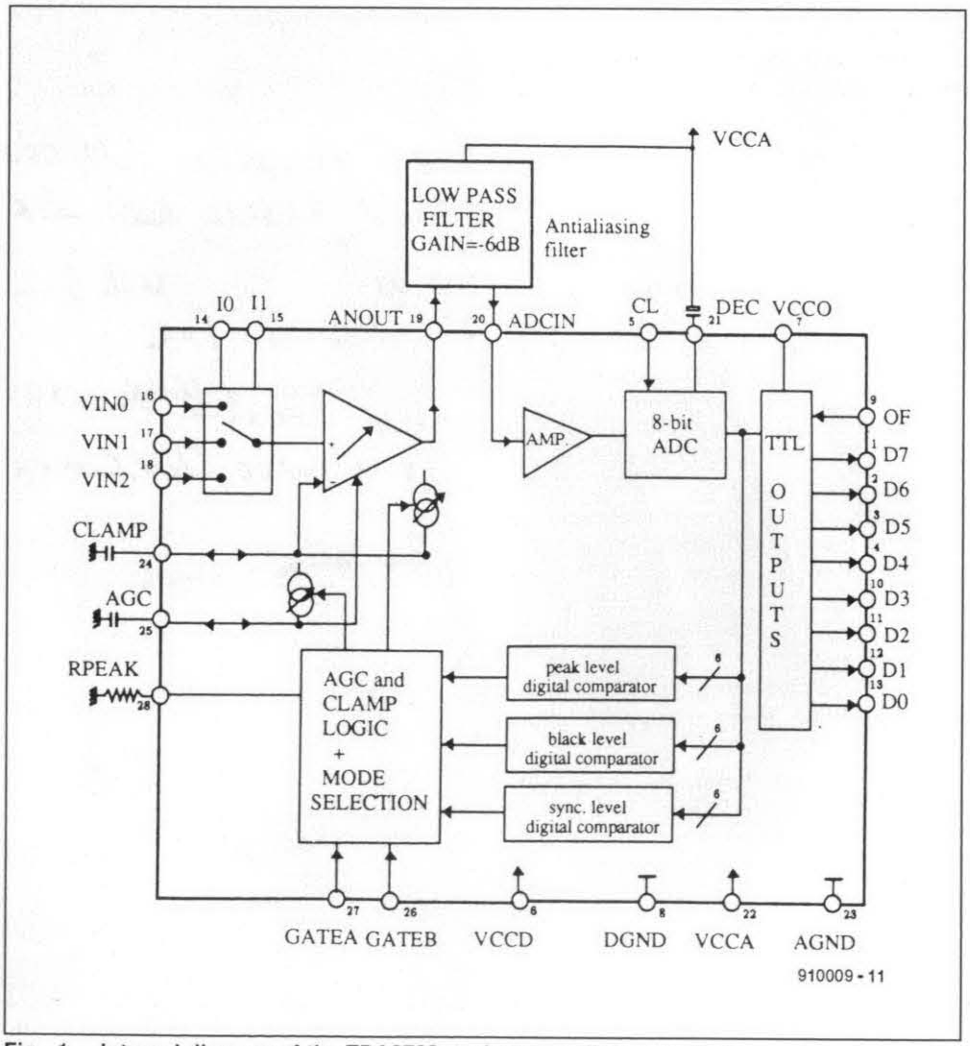


Fig. 1. Internal diagram of the TDA8708 analogue-to-digital converter.

The video amplifier features switchable AGC (automatic gain control) and clamping circuits. The AGC sets the upper video level (peak white); the clamping, the lower video level (base of sync pulse). As illustrated in Fig. 2, the peak white level produces the highest digital value, and the base of the sync pulse, the lowest digital value. The AGC and clamping sections receive information on the instantaneous drive level of the ADC via three comparators: one for the white level, one for the black level and one for the sync level. The AGC and the clamping logic each drive a pulsating direct current source that controls the video amplifier gain. The control current causes a voltage drop on the AGC capacitor connected to pin 25 of the TDA8708. The minimum and maximum video amplifier gains are 0.5 and 2.3 times respectively, corresponding to control voltage levels of 2.8 V and 4.0 V. As an option, the gain may be determined externally by omitting the AGC capacitor and applying a direct voltage instead. This, however, requires a fairly accurate temperature compensation circuit.

The gain of the internal amplifier is also controlled as a function of the lower video level: here, the voltage across the clamp capacitor at pin 24 is subtracted from the instantaneous video input voltage. A current limiting resistor, R_{PEAK} at pin 28, determines the level of the current diverted when the instantaneous level of the video signal exceeds the set limits. With $R_{PEAK} = 0\ \Omega$, the diverted current has a maximum level of 80 µA.

Table 2 lists the various possibilities of controlling the upper and lower video levels with the aid of logic levels applied to the GATE A and GATE B inputs of the chip.
In mode 1 (GATE A = GATE B = 1), the ADC output values are 255 for peak white and 0 for the base of the sync pulse. Mode 2 by contrast affords greater control over the conversion output values. As shown in Fig. 2, a positive pulse at the GATE A pin links the sync level to a value of 0. A positive pulse applied to the GATE B pin during the back porch sets the digital value of the top of the sync pulse to 64. The white level comparator is active all the time. Nominal levels of the video input signal produce a digital value of 213. This creates a safety margin of between 213 and 240. When the output value exceeds 240, the white level comparator reduces the gain of the video amplifier. This is done to reduce the risk of ADC overdrive to a minimum.

The analogue output, AN OUT (pin 19), of the video amplifier is connected to an external anti-aliasing low-pass filter. The maximum amplifier output current is limited to 2.5 mA, while the maximum output voltage is set to 1 V_{pp} at a nominal video input level of 1 V_{pp}. Table 3 lists a few additional characteristics relevant to the input circuit in the TDA8708.

The output signal of the external filter is fed back into the IC via the ADC IN pin. The ADC input may also be driven direct, provided the applied video signal has a voltage range of (V_{CCA}−1.6) V to (V_{CCA}−1.1) V. The input impedance of the ADC is about 50 MΩ,

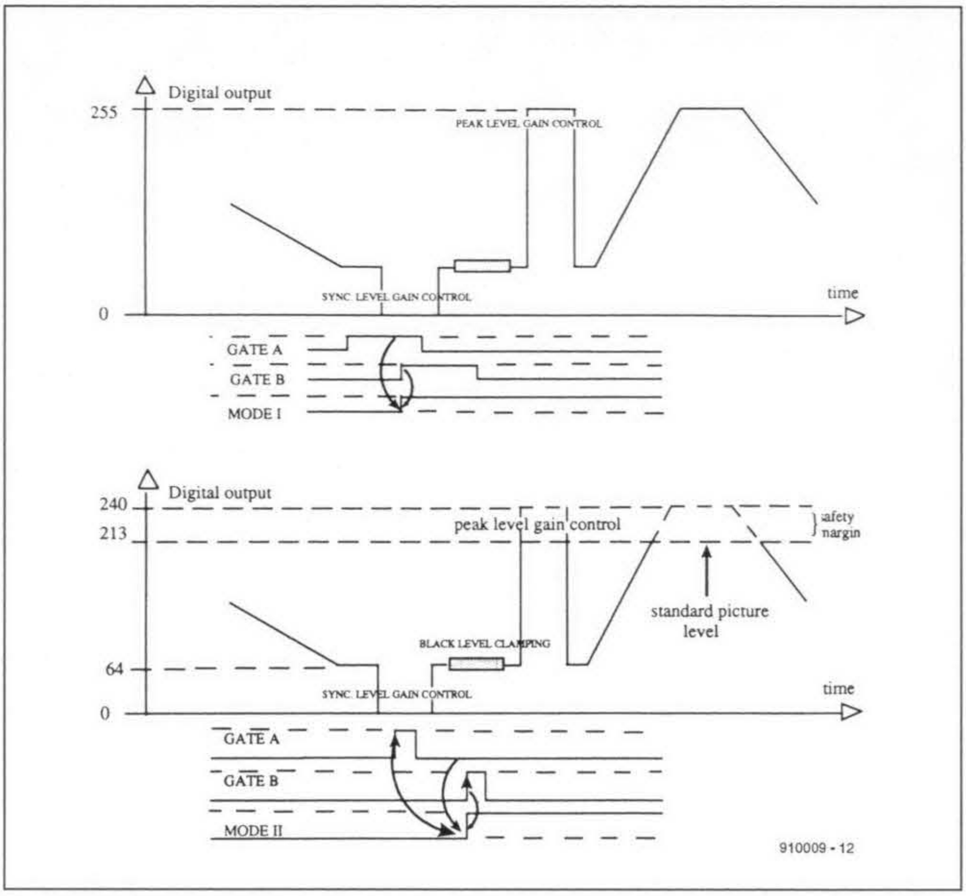


Fig. 2. Digital output value as a function of the analogue video input level in mode-1 (top drawing) and mode-2 (lower drawing).

Table 2. TDA8708 Mode Selection

GATE A B	Output	AGC	I_{CLAMP}	MODE
	<0	−2.5µA	I_{PEAK}	
1 1	0 - 255	−2.5µA	−2.5µA	1
	>255	I_{PEAK}	−2.5µA	
0 0	<240	0	0	
0 0	>240	0	I_{PEAK}	
1 0	<0	+2.5µA	0	
1 0	0 - 240	−2.5µA	0	2
1 0	>240	I_{PEAK}	0	
0 1	<64	+50µA	0	
0 1	64 - 240	−50µA	0	
0 1	>240	−50µA	I_{PEAK}	

Table 3. TDA8708 main technical characteristics

Symbol	Parameter	Min.	Typ.	Max.	Unit
V_{CCA}	analogue supply voltage	4.5	5.0	5.5	V
I_{CCA}	analogue supply current	—	37	45	mA
—	VINx crosstalk	—	−60	−55	dB
G_d	Gain error	—	2	—	%
Φ_d	phase error	—	2	—	°
B	−3dB bandwidth	12	—	—	MHz
S/N	noise distance	60	—	—	dB
SVRR	supply voltage rejection	—	45	—	dB
G	Gain	−4.5	—	6.0	dB

Table 4. TDA8708 pin function description

Pin	Symbol	Description
1-4	D7 (MSB) - D4	Digital outputs, bit 7 (MSB) through bit 4
5	CL	Clock input
6	VCCD	Digital supply voltage +5 V
7	VCCO	TTL output supply voltage +5 V
8	DGND	Digital ground
9	OF	Output format and chip enable (3-state)
10-13	D3 - D0 (LSB)	Digital outputs, bit 3 through bit 0 (LSB)
14,15	IO, I1	Video input channel selection
16-18	VIN0 - VIN2	Analogue video inputs 1, 2, 3
19	ANOUT	Analogue voltage output
20	ADCIN	ADC input
21	DEC	Decoupling capacitor for ADC
22	VCCA	Analogue supply voltage +5 V
23	AGND	Analogue ground
24	CLAMP	clamp capacitor
25	AGC	AGC capacitor
26,27	GATEB, GATEA	clamping and sync level control
28	RPEAK	current limit for AGC

and the input capacitance is as low as 1 pF.

The maximum clock rate of the ADC inside the TDA8708 is 30 MHz. The timing is illustrated in Fig. 4: the analogue input voltage is measured 2 ns (t_{su}) after the leading edge of the clock signal has exceeded the reference level of 1.5 V. The previous digital word remains on the outputs for another four to six nanoseconds, and is replaced by the new value after a delay t_d (16 to 20 ns).

The digital outputs of the ADC, D0 through D7, have a 20 μA current sinking or sourcing capability. The digital word is supplied either in binary form (with pin OF not connected) or as a two's complement value (with pin OF at logic 0). The ADC outputs are switched to high-impedance when the OF input is made logic high. Finally on the ADC, Table 4 provides the descriptions of the signals associated with the IC pins.

TDA8702 digital-to-analogue converter

This IC is of a much simpler layout than its counterpart — see the block diagram in Fig. 3. The pinning of the DAC is given in Table 5.

The digital data applied to the TDA8702 is accepted at the D0 through D7 inputs, which are TTL-compatible. The data are buffered by an input interface before they are loaded into a register. As shown in the block diagram, the register drives eight constant-current sources that translate the digital value into a corresponding analogue quantity, in this case, a current. Each digital output causes a current of about 85 μA to flow through the 75-Ω resistor. This in turn causes a voltage drop of about 6.3 mV with respect to the positive analogue supply voltage, V_{CCA}, provided the load impedance is relatively high. Table 6 provides information on the DAC output level for load impedances of 10 kΩ and 75 Ω.

While the output voltage (w.r.t. V_{CCA}) at the VOUT pin is proportional to the digital input value, the VOUTN pin supplies an inverted output signal. The signals at VOUT and VOUTN are typically coupled out via an electrolytic capacitor of 68 μF to 100 μF. Each of these introduces a voltage drop,

$$V_c = \frac{V_{OUT} \times R_L}{R_L + 75\Omega}$$

where R_L is the load impedance.

The circuit in Fig. 6 shows the two video outputs connected to a differential amplifier. This may be done to increase the suppression of hum and noise on the ADC supply voltage. A suggestion for a simple transistor-based buffer amplifier is given in Fig. 6.

Finally, a few notes on the timing of the TDA8702. As shown in Fig. 7, the DAC is transparent while CL (clock) is held logic low. Transparent means that the DAC follows the changes of the digital input signal. The

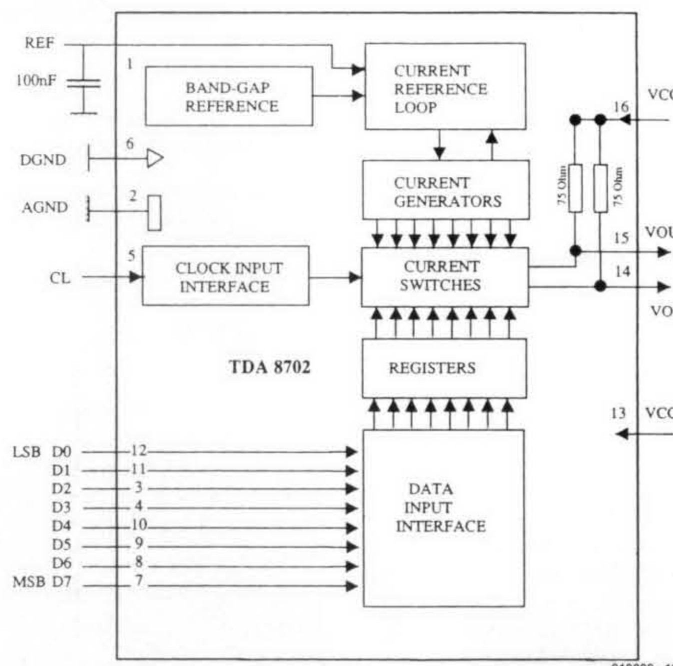


Fig. 3. Block diagram of the TDA8702 digital-to-analogue converter.

Table 5. TDA8702 pin function description

Pin	Symbol	Description
1	REF	Reference voltage source decoupling
2	AGND	Analogue ground
3,4	D2,D3	Digital video inputs bits 2, 3
5	CL	Clock input
6	DGND	Digital ground
7-10	D7-D4	Digital video inputs bit 7 (MSB) through 4
11,12	D1,D0	Digital inputs bit 1, bit 0 (LSB)
13	VCCD	Digital supply voltage +5 V
14	VOUT	Analogue video output
15	VOUTN	Inverted analogue video output
16	VCCA	Analogue supply voltage +5 V

Table 6. TDA 8702 output loading data

Code	Binary	Output voltage at Z_L = 10 kΩ V_{OUT} [V]	Z_L = 10 kΩ V_{OUTN} [V]	Z_L = 75 Ω V_{OUT} [V]	Z_L = 75 Ω V_{OUTN} [V]
000	000 000 00	0	−1.6	0	−0.8
001	000 000 01	−0.006	−1.594	−0.003	−0.797
...					
128	100 000 00	−0.8	−0.8	−0.4	−0.4
...					
254	111 111 10	−1.594	−0.006	−0.797	−0.003
255	111 111 11	−1.6	0	−0.8	0

TDA8702 latches the current digital value when the clock signal reaches a value of 1.3 V on the leading edge. To ensure that the output voltage is stable until the trailing edge of the clock signal, the input data must be stable at least 0.3 ns before, and 2 ns after, the threshold is reached. Table 7 lists the main technical characteristics of the TDA8702. ❑

Next month's final instalment of this article will describe a construction project based on the TDA8708 and TDA8702.

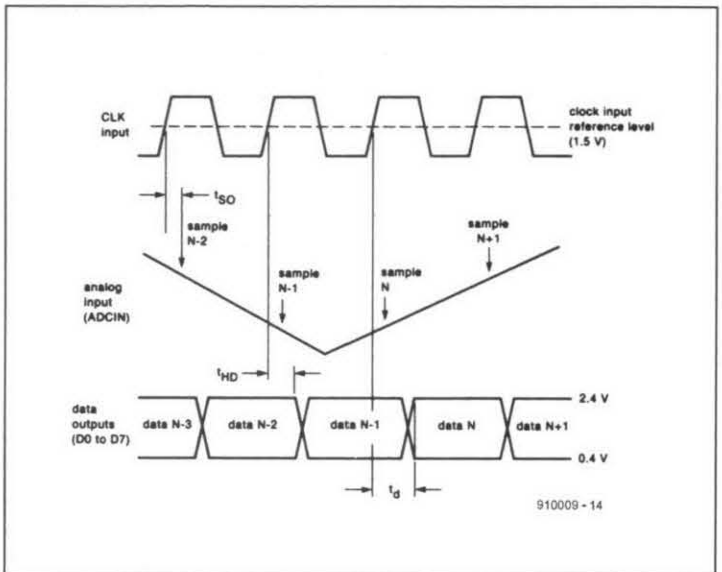


Fig. 4. TDA8708 data conversion timing.

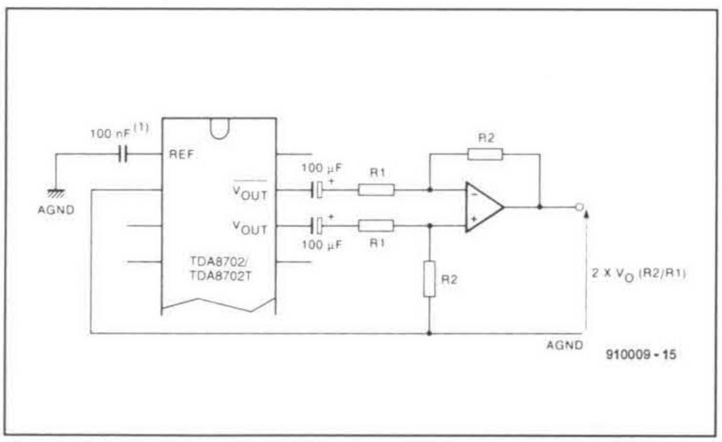


Fig. 5. Video output buffer based on a differential amplifier.

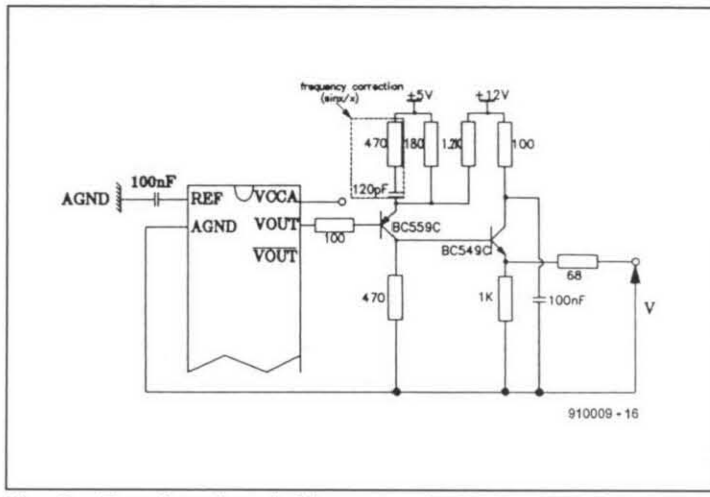


Fig. 6. Transistor-based video output buffer for the TDA8702.

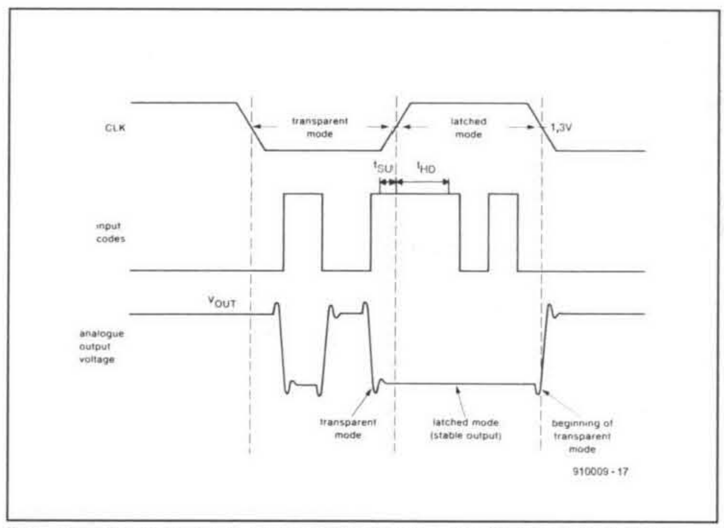


Fig. 7. Transparent and latching modes of the TDA8702 DAC.

Table 7. TDA8702 main technical characteristics

Symbol	Parameter	Min.	Typ.	Max.	Unit
V_{CCA}	Analogue supply voltage	4.5	5.0	5.5	V
I_{CCA}	Analogue supply current	-	26	32	mA
V_{CCD}	Digital supply voltage	4.5	5.0	5.5	V
I_{CCD}	Digital supply current	-	23	30	mA
$V_{OUT(N)}$	Output voltage (Z_L=10kΩ)	−1.45	−1.6	−1.75	V
	(Z_L=75Ω)	−0.72	−0.8	−0.88	V
V_{OFF}	Output offset	-	−3	−25	mV
$\Delta V_{OUT(N)}/\Delta T$	Temperature co-eff. V_{OUT}			200	µV/K
$\Delta V_{OFF}/\Delta T$	Temperature co-eff. V_{OFF}			20	µV/K
B_{3dB}	3-dB bandwidth of analogue output		150		MHz

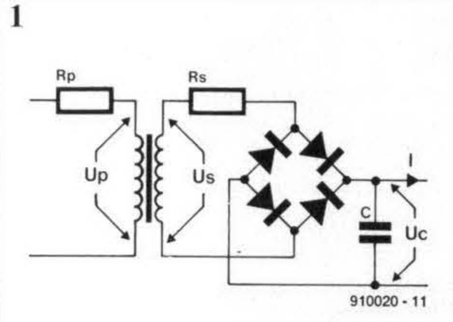
Symbol	Parameter	Min.	Typ.	Max.	Unit
$Z_{OUT(N)}$	Output impedance		75		Ω
DLE	Differential linearity error			±0.5	LSB
ILE	Internal linearity error			±0.5	LSB
t_{su}	Data setup time	0.3			ns
t_{HD}	Data hold time	2.0			ns
t_{pd}	Propagation delay			1.0	ns
t_{s1}	Settling time 10-90%		1.1	1.5	ns
t_{s2}	Settling time ±1 LSB		6.5	8.0	ns

RECTIFIER CALCULATIONS

What is the ripple voltage across the buffer capacitor? What is the level of the peak current through the rectifier diodes? The values of these and many other quantities are often obtained by rule of thumb, but what is the basis of that? This article aims to answer that question.

A TYPICAL, fundamental mains power supply is shown in Fig. 1. In this, U_p is the primary voltage; R_p represents the total losses at the primary side; U_s is the secondary voltage; R_s represents the total losses at the secondary side; U_C is the open-circuit output voltage (e.m.f.).

1



It is advantageous in most calculations to combine R_p and R_s into a single loss representation at the secondary side and call this R. The value of R (in Ω) is given by

$$R = (U_s/U_p)^2 R_p + R_s. \qquad [1]$$

This means, in effect, that the transformer need no longer be a part of any calculation; the circuit is fed by a source that provides a voltage U_s, has an internal resistance R and, for nominal loads, has negligible inductance.

Unfortunately, R can not be calculated readily owing to lack of data, and measuring it is also not possible. An ohmmeter will not do, because that does not take into account the losses caused by stray magnetism. Similarly, the resistance offered by the diodes can not be measured realistically.

For a proper measurement to be made, a variac is needed at the primary side of the transformer (begin with $U_p = 0$). The buffer capacitor, C, is short-circuited by an ammeter (whence the variac).

Start by adjusting the variac until the secondary winding provides the nominal level of current specified by the manufacturer. The primary voltage needed for this depends on the variac, the transformer, the rectifier diode(s), and the ammeter. The current indicated by the ammeter is the r.m.s. value of the short-circuit current, I_{sc}. Now, R (in Ω) may be calculated from

$$R = U_s / I_{sc}. \qquad [2]$$

Draw first, calculate later

Figure 2 shows a few voltages that are involved in the calculations. If rectifiers were ideal components, the waveform of U_s would indeed be as drawn. Since, however, ideal components do not exist, the voltage that is available to charge the buffer capacitor is lower than U_s by the knee voltage of the diodes, U_d. The level of that knee voltage depends on the number of diodes and on the forward voltage of each diode. The bridge rectifier in Fig. 1 will lower U_s by about 2 V. The voltage, U, after the rectifiers is thus

$$U = U_s - U_d. \qquad [3]$$

In most textbooks, the waveshape of that voltage is shown slightly differently: roughly as the dashed line in Fig. 2 immediately below the peak of U. In reality, the voltage will have a shape somewhere between the dashed and solid curves of U_C, depending on the ratio charging current : discharge current.

The shape of the (solid) curve of U_C is explained as follows. From the moment that U becomes larger than U_C, current flows from the voltage source to the buffer capacitor via resistor R. The level of the current is determined by the difference between U and U_C, which is U_R, and the value of R across which that difference voltage exists.

At the onset, U_R is tiny and the charging current, I_{ch}, is smaller than the discharge current, I_{dis}: U_c will then drop to U_{min} (the minimum value of U_C). From there, U_C rises and may go on rising after U has begun to drop again (because I_{ch} is then still larger than I_{dis}). However, at a given instant, U becomes too small and the capacitor starts to discharge. How much U_C will drop depends on the current, I_L, that the capacitor must supply to the load and on the duration, t_{dis}, of the discharge current.

The level of load current is known from the specification of the power supply. The duration of the discharge current is equal to the period of the rectified voltage minus the time necessary to charge the capacitor. The charging period is (as yet) unknown, but in general it is much smaller, in fact, negligible, compared with the discharge period. That means that the discharge time is equal to the period, T_r, of the rectified voltage. Note that T_r depends on the frequency and the method of rectification. For instance, in full-wave rectification, the capacitor is charged twice as fast as in half-wave rectification. That means that in half-wave rectification $T_r = 1/f$ and in full-wave rectification $T_r = 1/2f$.

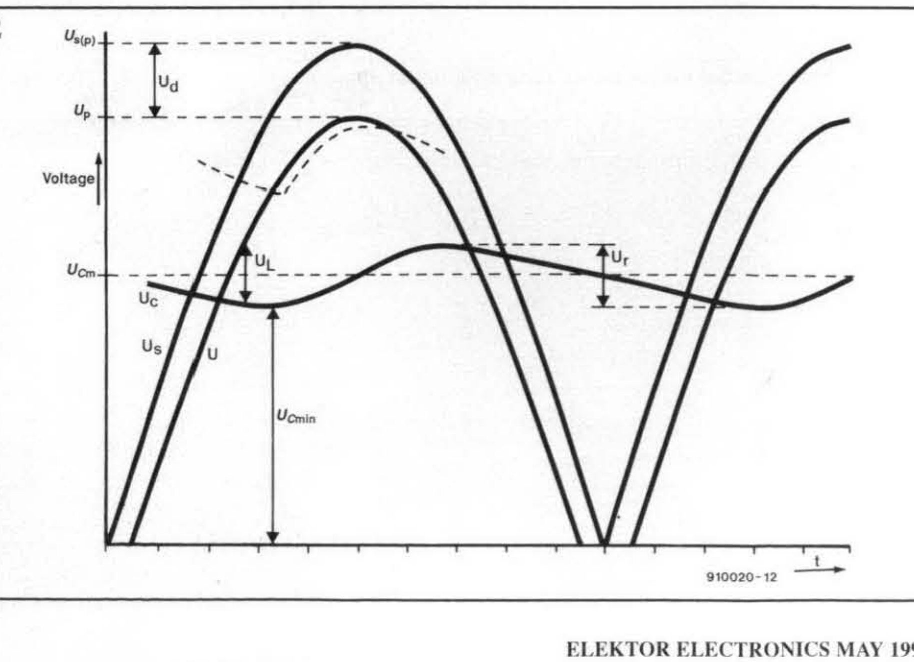
The charge, Q, that the capacitor can supply in that time is given by

$$Q = IT_r. \qquad [4]$$

The ripple voltage, U_r, may be calculated from

$$U_r = Q/C = I/2fC. \qquad [5]$$

2



Apart from the ripple voltage, the mean capacitor voltage, U_{Cm}, or, if voltage regulators are used, the minimum capacitor voltage, U_{Cmin}, is important. The exact computation of these voltages is, unfortunately, fairly complex and, therefore, in this article approximations are used, based on the data that have been found so far. Which of the two approximations given must be used for the mean capacitor voltage depends primarily on the ratio $R:R_L$, where R represents the total resistance of the transformer and the rectifier and R_L is the load through which current I_L flows. If R_L is large with respect to R (the usual case), I_L is small with respect to the maximum current with which the buffer capacitor can be charged. This means that the capacitor can be charged to the peak value, U_p, of the available voltage. The mean capacitor voltage is then

$$U_{Cm} = U_p - {}^1\!/_2 U_r \qquad [6]$$

and the minimum capacitor voltage is

$$U_{Cmin} = U_p - U_r. \qquad [7]$$

If R_L is not much larger than R, the capacitor will not charge to U_p and [6] and [7] are no longer valid. However, U_{Cm} may then be calculated with the aid of Fig. 3.

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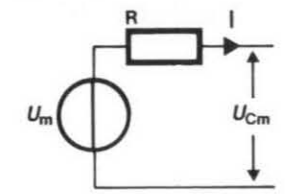

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The voltage source, which represents the transformer and rectifier, provides a direct voltage that is equal to the mean value of the non-smoothed direct e.m.f., U, supplied by the transformer and rectifier. The internal resistance of the source is represented by R. If a current I is drawn from the circuit, that is, the circuit is loaded, the output voltage, U_C, will reduce by the voltage drop across R, that is

$$U_C = U - IR. \qquad [8]$$

The level of U is dependent on the method of rectification: with full-wave rectification it is equal to $2U_p/\pi$, and with half-wave rectification it is equal to U_p/π.

The maximum current through the rectifier diode(s), $I_{d(p)}$ is

$$I_{d(p)} = U_p/R \qquad [9]$$

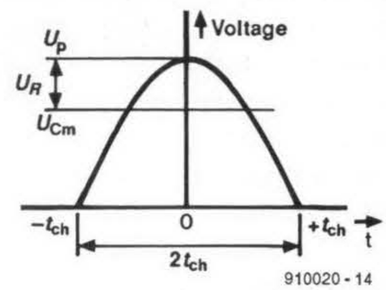
This level of current flows normally only when the power supply is switched on, since the capacitor voltage is then zero. It can, however, also flow in conditions of very heavy (over-) loads when the minimum of the ripple voltage is zero. Normally, however, the capacitor voltage is appreciably higher and the current that then flows is equal to the maximum voltage drop across R (roughly $U - U_C$) divided by R, that is

$$I_d = (U - U_C)/R \qquad [10]$$

The rectifier diodes must be able to cope continuously with this level of current, and should also be able to withstand, for short periods, the maximum current $I_{d(p)}$.

The considerations in this article allow the practical design and calculation of the rectifier section of a power supply. The formulas given are not one hundred per cent accurate for all theoretical considerations; for instance, no account has been taken of the time constant, τ, presented by resistance R and buffer capacitor C. ■

Some additional mathematics



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The buffer capacitor is charged via resistor R for a period $-t_L$ to $+t_L$. The charge, Q, stored in the capacitor in that period depends on the average charging current, I_{ch} and the length of the period, $t = 2t_{ch}$:

$$Q = I_{ch}t. \qquad [11]$$

The average current depends on the resistance, R, in the charging circuit and the average voltage, U_R, across that resistance:

$$Q = U_R 2t_{ch}/R. \qquad [12]$$

So that

$$U_R 2t_{ch} = 2U_p \int_0^{t_{ch}} \cos(\omega t)\,\mathrm{d}t =$$

$$= 2U_p \sin(\omega t_{ch}) \qquad [13]$$

and

$$Q = \frac{2U_p \sin t_{ch}}{\omega R} = \frac{2I_p \sin t_{ch}}{\omega} \qquad [14]$$

where I_p is the peak value of the current that the supply can deliver for short periods, that is, U_p/R.

To keep the mean value of the capacitor voltage constant, the charge removed from the capacitor must equal the input charge, that is,

$$I_{ch}\,2t_{ch} = I_{dis}\,T_r, \qquad [15]$$

where I_{dis} is the discharge current and T_r is the period of the rectified voltage (as explained in the text, $T_r = 1/f$ in half-wave rectification and $= 1/2f$ in full-wave rectification).

Since the incoming and outgoing charges are equal,

$$\sin(\omega t_{ch}) = \omega I_{dis} T_r / 2I_p, \qquad [16]$$

so that,

$$t_{ch} = \frac{1}{\omega} \arcsin\left(\frac{\omega I_{dis}\, T_r}{2 I_p}\right) \qquad [17]$$

Depending on the method of rectification, $\omega T_r = \pi$ (full-wave rectification) or $\omega T_r = 2\pi$ (half-wave rectification). If it is assumed that I_p is several times (or even many times) larger than I_{dis}, the following approximations are obtained:

$$t_{ch} = \frac{\pi I_{ch}}{2\,\omega I_p} = \frac{I_{ch}}{4 f I_p} \qquad [18]$$

or

$$t_{ch} = \frac{2\,\pi I_{ch}}{2\,\omega I_p} = \frac{I_{ch}}{2 f I_p} \qquad [19]$$

where it is assumed that

$$\arcsin(x) \approx x \text{ if } x \le 0.5.$$

The level of the ripple voltage, U_r, is determined by the amount of charge removed during the period that the capacitor is not being charged. The length of that period is $T_r - 2t_{ch}$, so that

$$U_r = Q/C = I_{dis}\,(T_r - 2t_{ch})/C. \qquad [20]$$

In a well-designed power supply, the capacitor is charged rapidly and $t_{ch} \ll T_r$, so that a simplification may be made:

$$U_r = I_{dis}\, T_r/C \qquad [21]$$

or

$$U_r = I_{dis}/2fC \text{ (full-wave rectification)} \qquad [22]$$

or

$$U_r = I_{dis}/fC \text{ (half-wave rectification).} \qquad [23]$$

SCIENCE & TECHNOLOGY

Augmented *A*-matrices:

a new circuit technique suitable for use with home computers

by Michael Soper, MA

Many articles have been written on the use of two-by-two matrices for the representation of two-port or four-terminal networks. Computer design now makes such techniques seem laborious when CAD facilities are available. But many designers still do not have access to such advanced facilities and need powerful techniques to help them analyse their proposed circuits. The advantage of the techniques outlined in this paper is that, unlike standard matrix techniques, they may represent forward-biased diodes as well as linear active or passive circuits.

To describe the system, we start off with the standard circuit where a shunt impedance Z has these equations:

$$V_1 = V_2$$

and

$$I_1 = V_2/Z - I_2$$

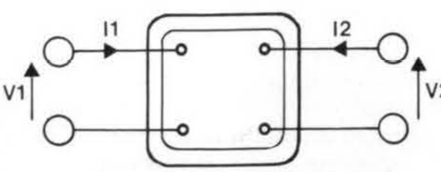
and is represented by the matrix

$$\begin{bmatrix} 1 & 0 \\ 1/Z & 1 \end{bmatrix}$$

The whole standard system is summarized as follows:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = [A]\begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}$$

represents this circuit



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where

$$A = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix}$$

and the circuit equations are:

$$V_1 = a_{11} V_2 - a_{12} I_2$$

$$I_1 = a_{21} V_2 - a_{22} I_2$$

The system in use here is a three-by-three matrix system which represents these equations:

$$V_1 = a_{11}V_2 - a_{12}I_2 - u.1$$

$$V_2 = a_{12}V_2 - a_{22}I_2 - y.1$$

$$1 = 1$$

The third equation is purely formal and thus the three-by-three matrix has this form:

$$\begin{bmatrix} a_{11} & a_{12} & u \\ a_{12} & a_{22} & y \\ 0 & 0 & 1 \end{bmatrix}$$

Another notation may be used for this:

$$\begin{bmatrix} a & b & u \\ c & d & y \\ 0 & 0 & 1 \end{bmatrix}$$

The vectors to use are:

$$\begin{bmatrix} V \\ I \\ 1 \end{bmatrix}$$

This is the standard voltage/current vector used in transmission matrices augmented by the extra column:

$$\begin{bmatrix} u \\ y \\ 1 \end{bmatrix}$$

All this will be explained more fully in the following.

A drawback of A-matrix technique is that although linear passive circuits can be represented adequately for design purposes by two-by-two matrices (called A-matrices), these are not adequate for the d.c. design of active circuits. This is because of the necessity of bias components which can not be calculated by what is essentially an incremental method. For example, there is unfortunately no matrix for a forward biased silicon diode.

This restriction has its origin in the mathematical fact that two-by-two matrices can not be used to move the origin to which the

$$\begin{bmatrix} V \\ I \end{bmatrix}$$

vectors of the model are referred. This restriction may be removed by an artifice: replace

$$\begin{bmatrix} V \\ I \end{bmatrix} \text{ by } \begin{bmatrix} V \\ I \\ 1 \end{bmatrix}$$

and use three-by-three matrices. These may be termed 'augmented' A-matrices and they have the form:

$$\begin{bmatrix} a & b & u \\ c & d & z \\ 0 & 0 & 1 \end{bmatrix} \qquad [1]$$

where u, z are complex numbers as are, of course, a, b, c, d.

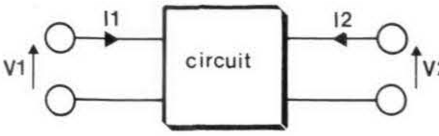
Standard *A*-matrix technique

We will summarize the standard A-matrix technique and then show how the augmented matrices are an improvement.

A-matrix technique has its origin in the equations:

$$V_1 = aV_2 - bV_2 \qquad [2a]$$

$$I_1 = cV_2 - dV_2 \qquad [2b]$$



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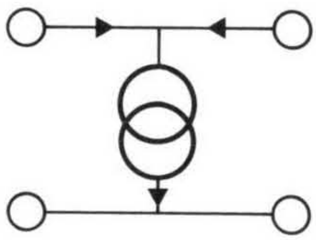
where V_1, I_1 are the input voltage and current, and V_2, I_2 are the output voltage and current. Equations [2a] and [2b] may be summarized neatly in matrix form:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = [A] \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \qquad [3]$$

The simple series impedance, Z, and the parallel admittance, Y, have therefore the corresponding A-matrices:

$$\begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix} \quad \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \qquad [4]$$

A formulation like this is particularly suitable for use in computer programs, but it has a serious limitation: V and I are incremental, not absolute, as may be seen from the fact that a shunted current-source can not be represented by an A-matrix.



$$\begin{bmatrix} ? \\ ? \end{bmatrix}$$

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New system

The liberation of the standard technique from this restriction is relatively easy, since the restriction has its root in the fact that two-by-two complex matrices can not be used to represent translations in the (V, I) plane. The solution is to replace each vector

$$\begin{bmatrix} V \\ I \end{bmatrix} \quad \text{by} \quad \begin{bmatrix} V \\ I \\ 1 \end{bmatrix}$$

and each matrix A by:

$$\begin{bmatrix} A & \begin{matrix} u \\ Z \end{matrix} \\ 0\ 0 & 1 \end{bmatrix} \qquad [5]$$

To give an example: suppose the circuit is

in which the diode is forward biased and has zero forward impedance. The matrix for a silicon diode then summarizes the situation:

$$\begin{bmatrix} 1 & 0 & 0.7 \\ 0 & +1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

and

$$V_1 = V_2 + 0.7$$
$$I_1 = -I_2 \qquad [6]$$
$$1 = 1$$

Current source

Let us now consider the matrix of a current source

$$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & -i \\ 0 & 0 & 1 \end{bmatrix} \qquad [7]$$

which provides these equations:

$$V_1 = V_2$$
$$I_1 = -I_2 - i.$$

The source has infinite impedance and supplies a current i. These matrices may be multiplied just as the unaugmented matrices: the result represents two circuits in cascade. The advantage of this system is that it is absolute and can represent absolute levels of voltage and current.

To model a transistor, we must include the knee voltage of the emitter-base junction and an emitter-collector working voltage. A small-signal silicon n-p-n transistor with a 0.7 V e-b voltage in a common-emitter configuration might have a matrix

$$\begin{bmatrix} -7\times 10\text{E}-4 & -20 & 0.7 \\ -10\text{E}-6 & -0.24 & 10\text{E}-6 \\ 0 & 0 & 1 \end{bmatrix}$$

When it is known what d.c. voltage offsets are required at input and output, the matrix may be computed as follows:

$$M = \begin{bmatrix} 1 & 0 & V_{\text{knee}} \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} A & \begin{matrix} 0 \\ 0 \end{matrix} \\ 0\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 & -V_{\text{w}} \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

If the matrix is to represent a transistor, and A is the usual A-matrix for, say, a common-emitter configuration, A may be found from the h parameters, where h is the determinant of the hybrid matrix by choosing

$$a = -h/h_{21}$$
$$b = -h_{11}/h_{21}$$
$$c = -h_{22}/h_{21}$$
$$d = -1/h_{21}.$$

Empirical tests may be the best way to obtain optimum parameters for the augmented A-matrix.

A related question is: 'Given the augmented A-matrix of a transistor, how can the surrounding circuit values be designed?'

In a common-emitter configuration, the emitter resistor, the collector resistor, and the two base bias resistors are the minimum requirement.

This is best shown by an example as follows.

Design procedure

In the design of a small-signal, single-stage n-p-n bipolar silicon transistor amplifier in a common-emitter configuration, let A be the augmented matrix of the transistor. For convenience, we write the column vectors in rows.

1. Solve $A(Z, 0, 1)^T = (V_k, 0, 1)^T$

for V_k, where Z is adjusted for zero input current.

Then, let the supply voltage be V and the transistor d.c. current, I.

$$A \begin{bmatrix} 1 & R_L & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} [V, -I, 1]^T = [0.7, I/\alpha h_{22}, 1]^T$$

2. Solvefor R_L.

3. Let j be the current in the potential divider and dv half the peak-to-peak signal value; then:

$$\begin{bmatrix} 1 + R_2/R_1 & R_2 & 0 \\ 1/R_1 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

$$[0.7 + \text{d}v, -I/\alpha h_{22}, 1] = [V, j, 1]$$

Solve for R_2 and R_1.

System use

The use of this system has been outlined in the foregoing; it allows, with care, circuits containing forward- or reverse-biased diodes and transistors, as well as passive components, to be easily designed. ■

References.

"Matrix Algebra - 2" by G.H. Olsen, *Wireless World*, April 1965.

"Useful matrix theorem with applications to electronic circuit theory" by M.C. Soper, *Electronics Letters* (IEE), 18 July 1985, Vol. 21 No. 15.

"Electromagnetism and Time" by M.C. Soper, *Awareness*, Vol. 12 No. 2, 1983-84.

Introductory circuit theory by Guillemin, Wiley 1953.

"Metalogic" by M.C. Soper, *Awareness*, Vol. 16, No. 3, 1989.

CORRECTIONS

Wattmeter

April 1991, p. 32-35

With reference the circuit diagram, Fig. 1, the right-hand terminal of the lower section of switch S2 should be connected to the circuit ground. This point is indicated by a dot.

In the adjustment procedure given on page 35, the references to presets P4 and P5 have been transposed. Contrary to what is stated, P4 sets the VY offset, and P5 the VX offset. The functions of the presets are shown correctly in the circuit diagram, Fig. 1.

To improve the accuracy of the instrument, connect R5 direct to the circuit ground instead of junction R6-R7. Finally, all circuit board tracks carrying mains current must be strengthened with 2.5-mm^2 cross-sectional area solid copper wire if currents higher than about 5 A are measured.

80C32/8052 Single-board computer

May 1991, p. 17-23

When a CPU type 8031 or 8052AH-BASIC is used, IC1, IC2, IC3, and IC8-IC12 must be 74HC**T** types. Jumper B is erroneously reffered to as Br2 in the text under "On-board EPROM programmer". Contrary to what is stated, this jumper must be fitted only when an EPROM is to be programmed — for all other use of the SBC, it must be removed. Also note that jumper B may only be fitted when the programming LED is out.

Sequential control

July/August 1991, p. 61

Motor M should be a d.c. type, not an a.c. type as shown in the circuit diagram.

Digital phase meter

June 1991, p. 32-39

In Fig. 5, the switch between input 'A' and IC1 should be identified 'S1', and that between input 'B' and IC2 'S2'. Switch S4 is an on/off type, not a push-button as shown in the diagram. Capacitors C3 and C6 are shown with the wrong polarity. The component overlay of the relevant printed-circuit board (Fig. 8) is all right.

Universal NiCd battery charger

June 1991, p. 14-19

The parts list on page 19 should be corrected to read

C7 = 2200µF 25V

When difficult to obtain, the BYW29/100 (D5) may be replaced by the BY229, which is rated at 6 A.

The text under the heading 'Calibration' should be replaced by:

4. Connect a multimeter between points G and H on the board, and adjust P4 until the measured voltage is 1 V lower than the voltage on the battery terminals.

MIDI program changer

April 1991, p. 14-17

The contents of the EPROM should be modified as follows:

address	data
OOBC	E5
00C7	80
00C8	CB
00C9	F5
00CA	7B
00CB	12
00CC	00
00CD	D2
00CE	C2
00CF	02
00D0	80
00D1	C2

Readers who have obtained the EPROM ready-programmed through the Readers Services may return it to obtain an update.

Electronic exposure timer

March 1991, p. 31-35

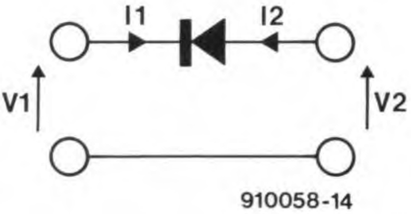
Please add to the parts list on page 32:

C16 = 33 pF

Augmented A-matrices

May 1991, p. 42-43

The drawing below was erroneously omitted in the left-hand bottom corner of page 43.



APPLICATION NOTES

The contents of this column are based on information received from manufacturers in the electrical and electronics industries and do not imply practical experience by *Elektor Electronics* or its consultants.

D.C.-TO-D.C. CONVERTER

(SGS-THOMSON MICROELECTRONICS)

DC-TO-DC converters are also known as switching regulators, which, in the opinion of many, is a more appropriate name. The converter described in this article is based on SGS-Thomson's new UCxx84x family of regulators.

The average output voltage of the converter is controlled by varying the pulse-width of a regulator, IC_1 in Fig. 2.

The regulator described is a step-up type that converts a direct voltage input of 12–16 V to one of 18 V. The maximum output current is 3 A at 18 V. The efficiency is 73%.

The new chips all operate according to the pulse-width control technique. In contrast to similar chips, for instance, those from National Semiconductor, the SGS-Thomson devices do not have the power transistor on board. Although this increases the number of external components required, it offers greater freedom in the design of the regulator circuit. The internal circuit of the regulator chip is shown in Fig. 1.

The various members of the UCxx84x family differ in ambient temperature range, the difference between on and off voltage, the accuracy of the internal reference voltage source and the maximum duty ratio. The housing is indicated by a single letter: J for ceramic mini-DIP; B and D for DIP-14 and SO-14 respectively. The circuit described here uses the consumer version UC3843N.

The principle of operation is that the counter-e.m.f. generated in inductor L_1 is added to the input voltage. When T_1 is switched on for a time t_1, the current through L_1 rises and energy is stored in the inductor. When T_1 is switched off for a time t_2, the energy stored in the inductor is transferred to the load via diode D_1 and the inductor current drops.

When T_1 is on, the voltage across L_1 is

$$U_L = L(\mathrm{d}I/\mathrm{d}t) \qquad [1]$$

and this gives the peak-to-peak ripple current in the inductor as

$$\Delta I = (U_i/L)t_1, \qquad [2]$$

where U_i is the input voltage.

The instantaneous output voltage, U_o, is

$$U_o = U_i + L(\Delta I/t_2) = U_i(1+t_1/t_2) = U_i(\Delta I/[1-k]). \qquad [3]$$

The large capacitor, C_5, across the load ensures that the output voltage is continuous.

Note from Eq. [3] that the output voltage

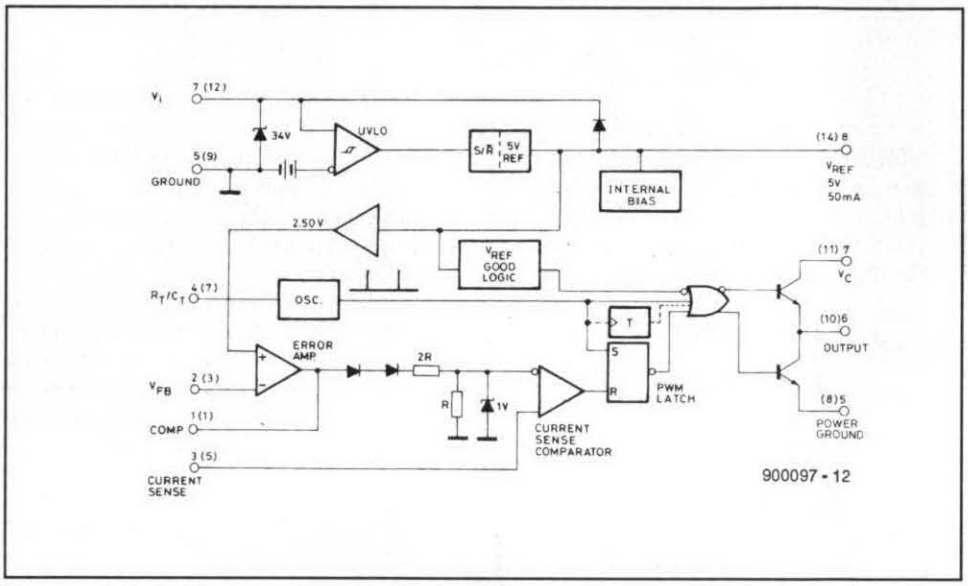


Fig. 1. Diagram of the internal circuit of the UCxx84x regulator.

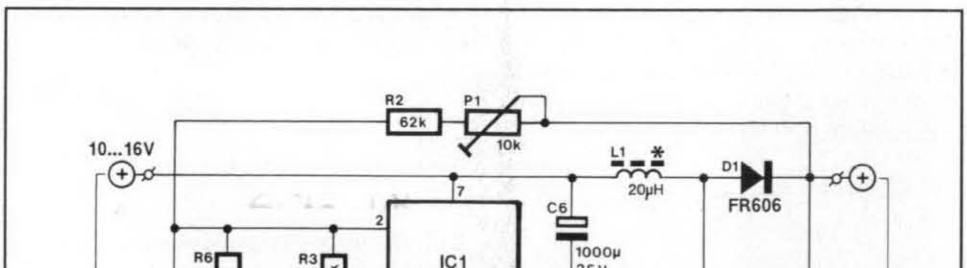


Fig. 2. Circuit diagram of the step-up switching regulator controlled by the UCxx84x.

depends on the duty factor, k, that is, the ratio of the on and off times, t_1/t_2, of T_1. The minimum output voltage occurs when $k = 0$. However, T_1 can not be switched on continuously such that $k = 1$. For values of k tending to unity, U_o becomes large and very sensitive to changes in k.

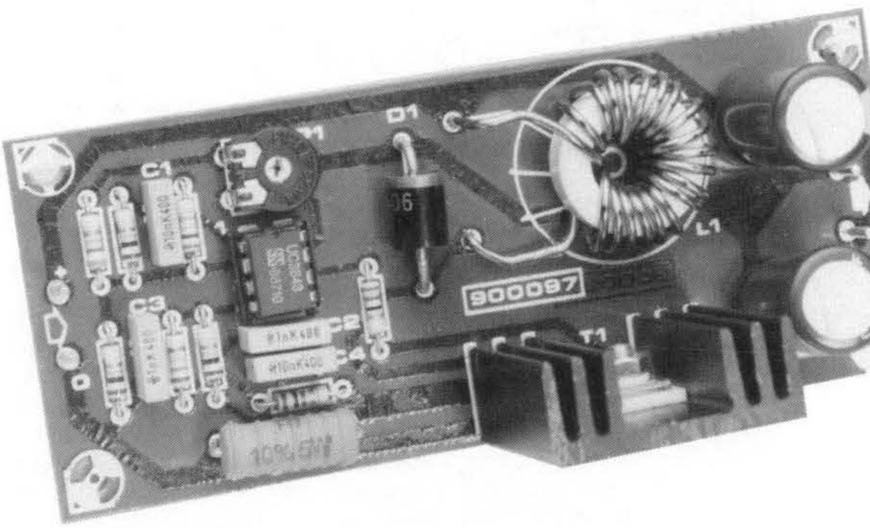
Circuit description

The circuit shown in Fig. 2 is based on regulator IC_1. This device controls the on and off times of T_1 and thus the on-off switching of the current through L_1.

As explained earlier, the voltage across C_5 depends on the ratio of the on and off times of T_1. The regulator chip (pin 2: error amp) measures the output voltage with the aid of potential divider P_1-R_2-R_5 and compares the part of it that exists across R_5 with the internal 2.5 V reference voltage. On the basis of this, it adjusts the duty factor of the pulse at its output (pin 6), which switches T_1, in a manner that ensures the required level of output voltage.

The pulse rate is constant and kept at about 50 kHz by the oscillator on board the UC3843N. The duty factor may vary from 0% to 50% (x844/5) or from 0% to 100% (x842/3).

A peculiarity of the regulator chip is the way the current in the transistor circuit is measured: this is done indirectly by the voltage drop across source resistor R_9. This voltage is integrated by R_7-C_2 and then applied directly to the input of the current sense comparator on board IC_1. This method improves the speed of control appreciably.



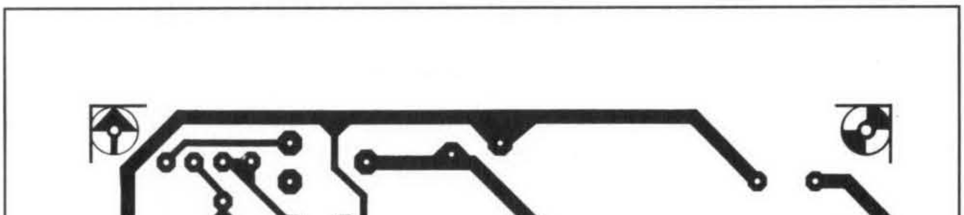


Fig. 3. Printed-circuit board for the d.c.-to-d.c. converter.

Construction

If the converter is constructed on the PCB shown in Fig. 3 (not available through our Readers' services), no difficulties should be encountered. If the board is mounted on to the metal rear panel of an existing power unit, T_1 normally does not require an additional heat sink. It must, however, be insulated with the aid of mica washers and heat conducting paste, since its metal base is NOT at earth potential.

Since D_1 can get pretty warm, it is advisable to mount it 2–3 mm above the board.

Inductor L_1 consists of 19 turns 1.5 mm dia. enamelled copper wire on a Siemens Type R20/7 ring core. The turns should be distributed more or less evenly across the core, but this is not very critical.

The efficiency of the converter depends greatly on the quality of the components used. Therefore, a fast FET instead of a bipolar transistor is used for T_1. Also, a fast diode is used for D_1; the circuit would operate with a simple silicon diode, but the transit times would not be short enough.

The converter is calibrated by applying an input of 10–16 V provided by an accurate power supply and setting the output voltage to 16–20 V as required with P_1. ■

References

SGS Application: *Industrial and Computer Peripheral ICs.*

SGS *Power Supply Application Manual.*

"Switch-mode power supplies", *Elektor Electronics*, October 1987.

COMPONENTS LIST

Resistors:
R1 =22 Ω
R2 = 62 kΩ
R3 = 100 kΩ
R4–R6 = 10 kΩ
R7 = 1 kΩ
R8 = 15 kΩ
R9 = 0.1 Ω; 5 W
P1 = 10 kΩ preset

Capacitors:
C1, C4 = 10 nF
C2, C3 = 1 nF
C5, C6 = 1000 μF; 35 V; upright

Semiconductors:
IC1 = UC3843N
I1 = BUZ11
D1 = FR606 (Taiwan Semiconductors)

Miscellaneous:
L1 = 20 μH (see text)
Ring core R20/7, K1 (Siemens)
Heat sink for T1 (SK09)
Enamelled copper wire 1.5 mm dia.

Electronics, October 1987.

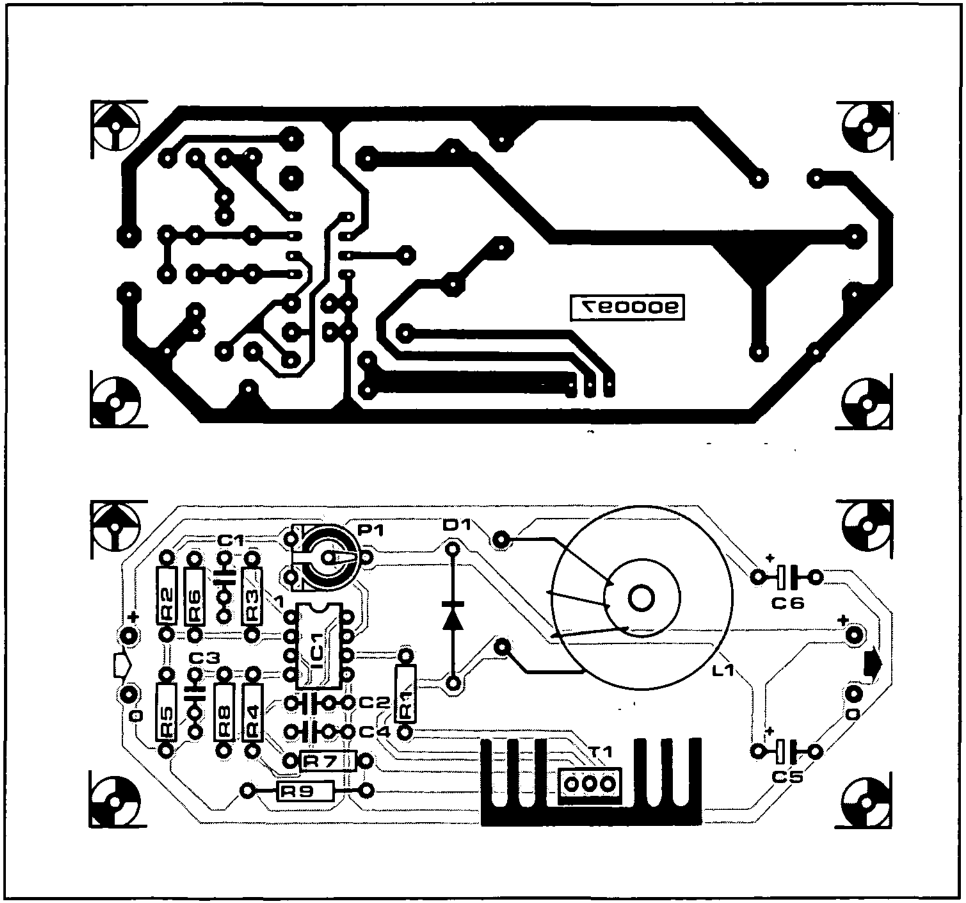


Fig. 3. Printed-circuit board for the d.c.-to-d.c. converter.

COMPONENTS LIST

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Ring core R20/7, K1 (Siemens)
Heat sink for T1 (SK09)
Enamelled copper wire 1.5 mm dia.

CONDUCTANCE METER

from an idea by J. Vaessen

Conductance, the inverse or reciprocal of resistance, is not a quantity that many of us measure daily: we normally deal with resistance. However, in some instances, for example, in liquids, it is easier to measure conductance than resistance.

GENERALLY speaking, the measurement of conductance is not any harder than measuring resistance. In fact, in case of an analogue ohmmeter it is sufficient to state the inverse of the printed figure to convert the instrument to a conductance meter. However, most ohmmeters operate from a direct voltage or current. If that were used in a liquid, it would invariably give rise to some sort of electrolysis that would distort the measurement.

To counter the electrolysis, the meter must operate with an alternating voltage or current of a sufficiently high frequency. The principle of such a meter is shown in Fig. 1a. Here, a rectangular-wave generator provides a voltage that alternates between 0 and a reference voltage, U_{ref}. That potential is applied to one end of the unknown conductance, G_x; the other side is kept at $\frac{1}{2}U_{ref}$. This means that the potential across G_x is a rectangular voltage that swings between $\pm{}^1/_2U_{ref}$, which is a true alternating voltage. To determine the conductance, the potential across one of the resistors R_{ref} needs to be measured.

The relationship between G_x and the output voltage, U_o, follows from Thevenin's Theorem and the Superposition Theorem. For convenience, in the following the direct voltage setting of ${}^1/_2U_{ref}$ will be ignored. That being the case, Fig. 1b is the a.c. equivalent of Fig. 1a. From this,

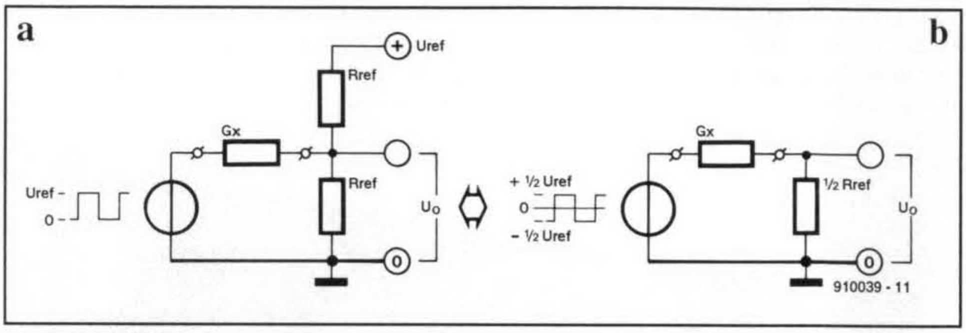


Fig. 1. The principle of measurement is similar to that used in an ohmmeter.

$$U_o = \frac{1}{2}U_{ref}\frac{\frac{1}{2}R_{ref}}{\frac{1}{G_x}+\frac{1}{2}R_{ref}}$$

If we make the term ${}^1/_2U_{ref}$ in the denominator much smaller than $1/G_x$, the formula may be rewritten as

$$U_o = {}^1/_4U_{ref}\,R_{ref}\,G_x.$$

However, this has advantages as well as drawbacks. The advantages are that the meter will have a linear scale and that R_{ref} will be small in spite of the fact that conductance measurements are normally used in case of high impedances. After all, R_{ref} would be impractically high for some meter ranges. It is also possible to draw a new scale for the meter, which allows for the assumption. The drawbacks are that the assumption causes some degradation of accuracy and that in this method of measurement the greatest accuracy is obtained when $1/G_x = R_{ref}$.

The circuit

The circuit diagram of the conductance meter is given in Fig. 2.

The rectangular-wave oscillator is based

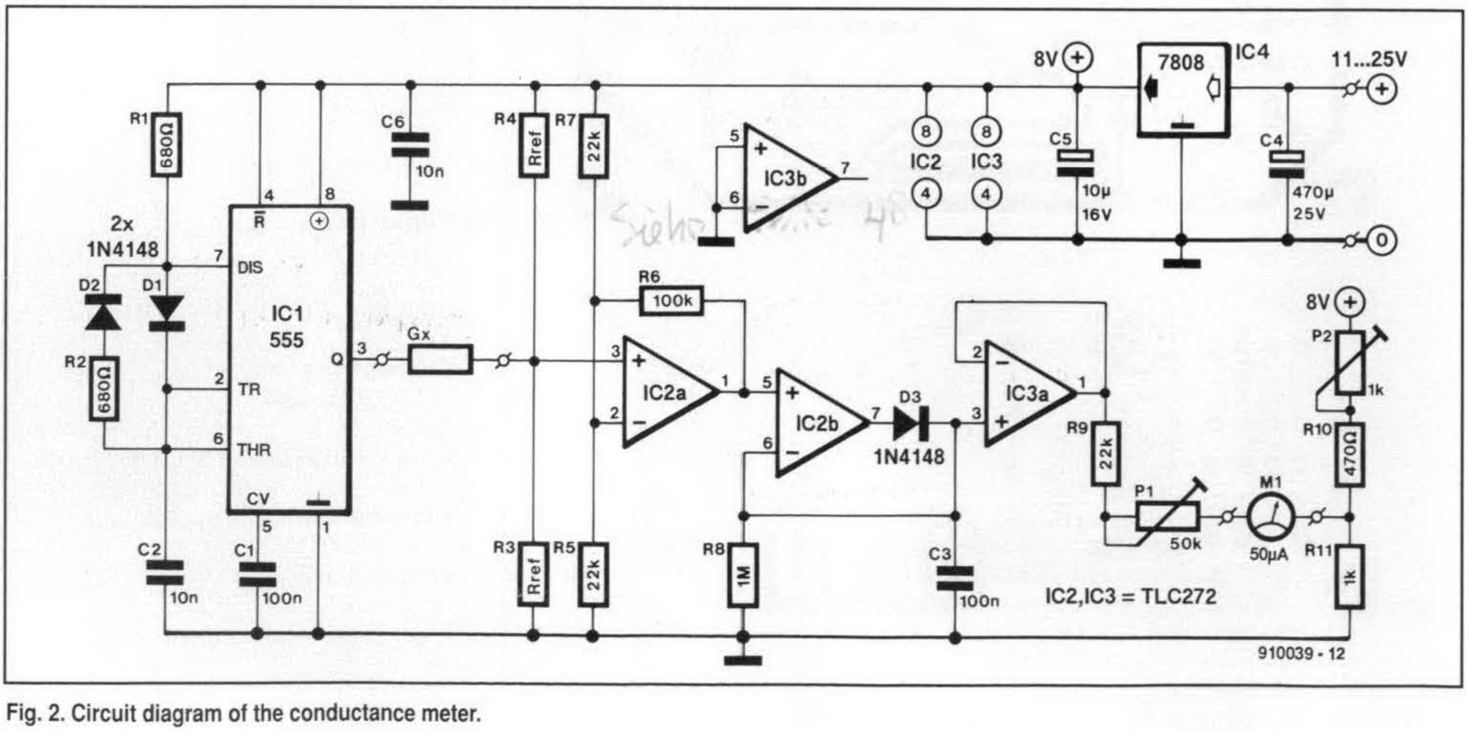


Fig. 2. Circuit diagram of the conductance meter.

on the well-known Type 555 (IC_1). Its frequency is 10 kHz, which is high enough to prevent electrolysis in a liquid.The output of the oscillator is applied to the measuring circuit as indicated in Fig. 1 (G_x, R_3, R_4).

The output of the measuring circuit is buffered by opamp IC_{2a} and amplified about ×10. This figure comes about because both R_3 and R_4 must have a value of $0.2/G_{max}$, where G_{max} is the value of the conductance at full-scale deflection—f.s.d.

The amplifier is followed by a high-quality rectifier circuit based on IC_{2b}. Buffer capacitor C_3 is charged via D_3 regularly, fast and accurately to the peak value of the voltage at the non-inverting input of the opamp, while it discharges slowly via R_8.

The potential across C_3 is applied to the non-inverting input of buffer IC_{3a}, which drives the meter. Network R_9-P_1 serves to set the full-scale value of the meter. Potential divider R_{10}-R_{11}-P_2 serves to zero the meter.

The circuit is powered via voltage regulatr IC_4. The regulator ensures that the supply voltage may be used to derive the reference potential.

A probe for measuring in liquids is made readily from a print header. Two pins of the header—preferably gold-plated—form an excellent probe with fixed distance between the electrodes. The link between the electrodes and the connecting wires should be made waterproof with the aid of two-component epoxy resin or potting compound.

Calibration

Depending on the application, the meter may be calibrated in two ways. If it is to be used for measuring the conductance of solids only, a high-value resistor may be used as the calibrating conductance. The meter is adjusted with P_1 and P_2 as described later.

If the meter is intended for measurements in liquids, the shape of the probe, that is, the distance between the two electrodes, plays a role. The measurand is expressed in $\mu S\ cm^{-1}$, that is, micro-siemens per centimetre. The 'per centimetre' emphasizes that the distance between the electrodes is a factor.

To calibrate the meter including probe accurately, a calibrating liquid is required. A saturated calcium-sulphate (anhydrite—$CaSO_4$) solution is eminently suitable for this purpose: at 20 °C, this has a conductance of 1976 $\mu S\ cm^{-1}$. The solution is prepared by mixing calcium-sulphate in distilled water until the liquid accepts no more of the $CaSO_4$. Note, however, that calcium-sulphate is only slowly soluble in water.

By adding an identical volume of distilled water to the solution, the conductance will be halved. Doubling the volume again by adding distilled water will halve the conductance once more. While diluting the solution, make sure that it remains at 20 °C, because the conductance varies with temperature.

Start the calibration before the power is switched on by manually adjusting the moving-coil meter to zero.

Next, switch on the power and let the instrument warm up for a few minutes, after which the microammeter is zeroed by adjusting P_2 as appropriate. There should be nothing connected to the measuring electrodes or terminals.

Then, connect the calibrating resistor or immerse the probe in to the calibrating liquid and adjust P_1 till the microammeter indicates the correct value. Since this affects the setting of P_2 slightly, the adjustments of the two potentiometers should be repeated a couple of times. ■

DESIGN IDEAS

The contents of this column are based solely on information supplied by the author and do not imply practical experience by *Elektor Electronics*

Versatile pulse-width modulator

by Dr U. Kunz

THE CIRCUIT shown here generates a pulse train in which the width of each individual pulse is determined by a control voltage. This property makes the circuit suitable for use as a multiplex-decoder in, for instance, a remote control system, but other applications are, of course, also possible.

If the input signals applied to IC_1 originate in analogue sources, they can be transmitted in digital form over a two-wire system or via a wireless system. In other words, the circuit can also serve as an analogue-to-digital converter with eight channels. Its use in that function is readily accomplished with the aid of a computer.

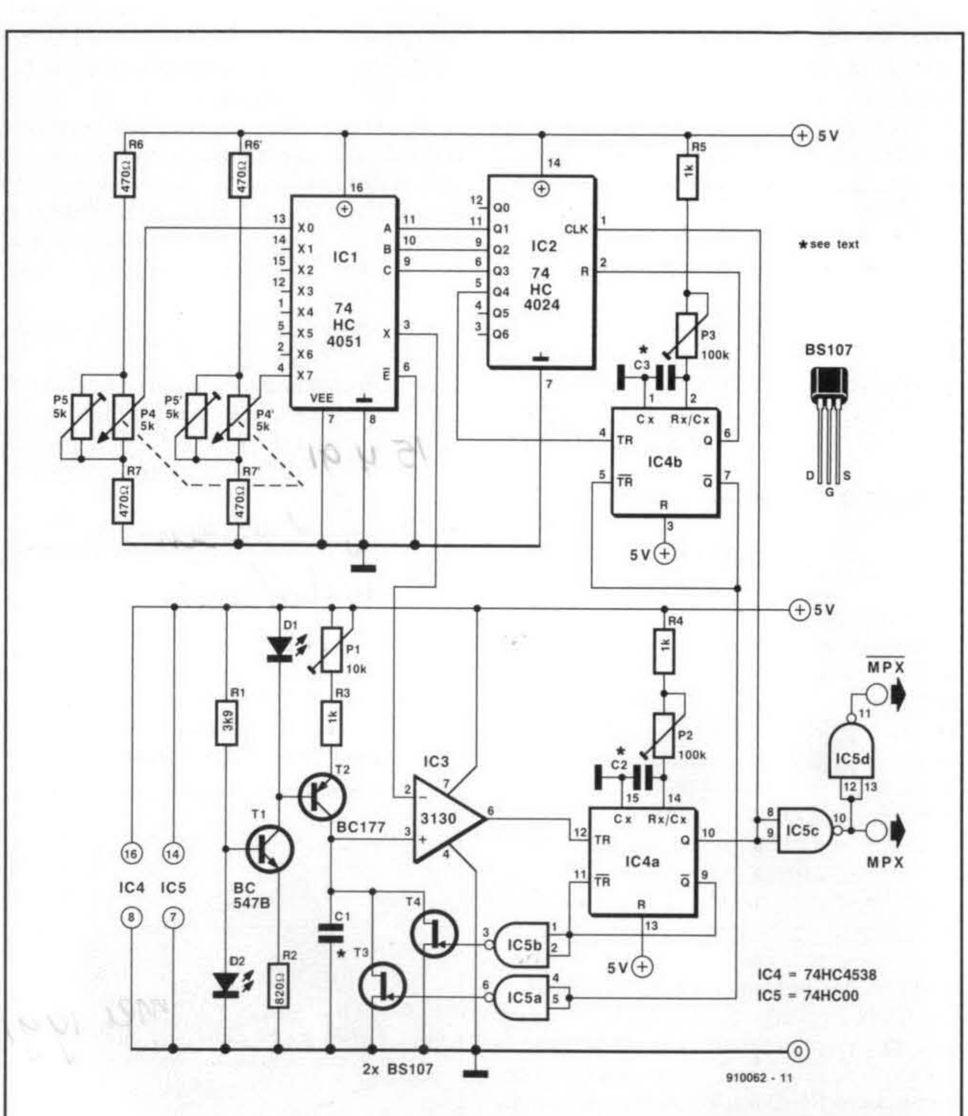
The constant-current source, based on T_1, T_2 and associated components charges capacitor C_1 linearly. When the rising potential at the non-inverting input (pin 3) of IC_3 reaches the level of the voltage at the inverting input (pin 2), the output of the opamp becomes $+U_b$.

The consequent leading edge at pin 12 of IC_4 starts monoflop IC_{4a}, and this results in the output going high. At the same time, C_1 discharges via IC_{5b} and T_4.

When IC_{4a} toggles, binary counter IC_2 advances one step. Since outputs Q1, Q2 and Q3 of the counter are connected to the digital inputs of multiplexer IC_1, this IC applies a different voltage, which is presettable with P_4-P_4', to the input of IC_3.

Since C_1 discharges via T_4, the process just described repeats itself, but with a different input voltage. This ensures that the time lapsed before comparator IC_3 toggles is different from that in the previous cycle. The setting of the potentiometers or, if none is used, the voltage level at the inputs of IC_3 influences the width of the pulse at pin 10 of IC_4. In other words, the input voltages are translated into pulse widths.

At the eighth cycle, pin Q4 of IC_2 goes high and this starts monoflop IC_{4b}. The time constant of this stage is much longer than that of IC_{4a}. The leading edge at pin 6 of IC_{4b} is used to reset the counter. Gate IC_{5a} and T_3 prevent C_1 from charging until IC_{4b} toggles. The long pulse so generated, which appears at the outputs of IC5c and IC_{5d}, may be used for time synchronization (for instance, to reset a counter) in the receiver unit. When IC_{4b} toggles, the whole process starts afresh.



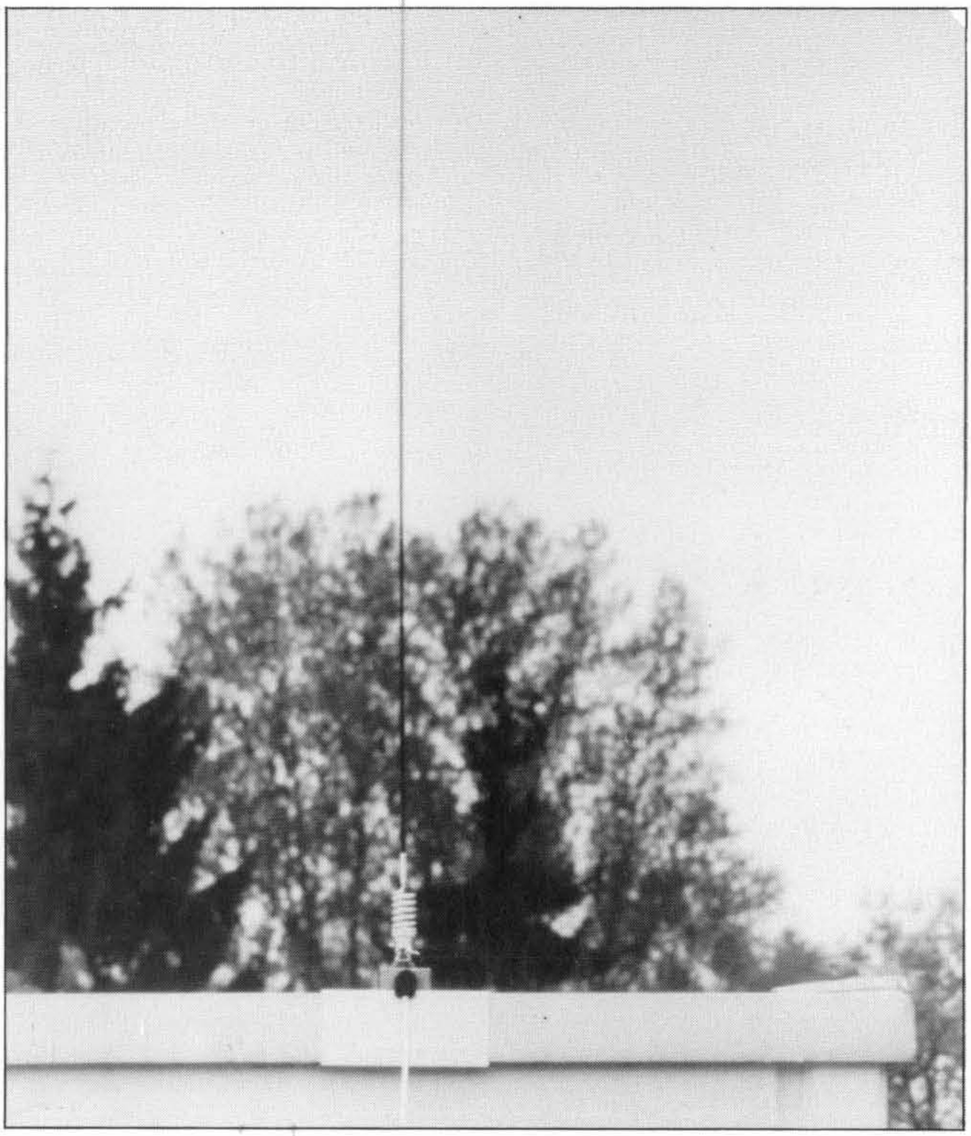
The values of C_1, C_2 and C_3, which determine the time constants of the pulse generation, must be chosen to fit the particular application.

Applications

The circuit was designed as a multiplex encoder for a remote control system, in which potentiometers P_4-P_4' are the control joysticks. P_5-P_5' serve to restrict the control range of P_4-P_4'. This is advantageous in applications where great sensitivity of control is needed. Since the time constants for the various cycles can be set independently of one another, and the time lapses are controlled not only by potentiometers, but also by voltage sources, the field of applications is very wide.

The possibility of using the circuit as an A-to-D converter has already been mentioned. Other uses include a weather station, where the sensor output voltages for air pressure, temperature, wind speed, humidity, and so on, are applied to the inputs of IC_1. ■

WIDEBAND ACTIVE ROD ANTENNA



The starting point for this design is the well-known 5⁄8-lambda whip antenna with base coil as used for mobile communications in the 2-m VHF band. Interestingly, this type of antenna is readily tuned to 6-metres. Add a wideband low-noise preamplifier and you have a wideband rod antenna for a large frequency range (20 kHz to 150 MHz).

Dipl.-Ing. Jo Becker, DJ8IL

A familiar sight on cars fitted with a VHF band mobile radio, the 5⁄8-λ whip aerial offers a gain of 2 to 3 dB over the ¼-λ rod, known as the Marconi antenna. The coil seen at the base of the 5⁄8-λ antenna serves to lengthen it electrically and match it to a 50-Ω coax cable. Most 5⁄8-λ antennas for the 2-m amateur radio band have a length of about 1.25 m, which is about 0.22 λ in the 6-m band. The additional inductance formed by the base coil gives the antenna an electrical length of a little over 1/4 λ, which allows ready connection to a 50-Ω coax cable.

Active antenna

Long established for maritime radio communications, the active antenna is gaining popularity with users of general coverage receivers (100 kHz to 30 MHz). Compared with the familiar long wire, the active antenna is unobtrusive, small and simple to install (although its final location will have to be given some thought in view of interference). Where a 'full-size' vertical antenna for general coverage reception has a minimum size of about 6 m and a weight of more than 5 kg, the active version presented here weighs a modest 400 g and has a length of only 1.3 m.

In principle, an antenna for the frequency range below 20 MHz or so can be shortened to about 1 m without degrading reception. This is so because the level of man-made and natural noise is then still higher than the noise level of the receiver. However, the problem with such a short antenna is that it has a relatively high radiation resistance. As a rule of thumb, short antennas for the SW frequency range have a capacitive base impedance, C_A, of about 10 pF/m. This means that they must be fitted quite close to an amplifying impedance transformer with a 50-Ω output for the coax cable to the receiver input. This assembly of a short rod and a wideband RF amplifier fitted at its base is called an active antenna. The amplifier is provided with its supply voltage via the coax cable and a simple *L-C* decoupling network at the receiver input.

Transmit/receive operation in the 2-m and 6-m band is enabled by a relay that disconnects the amplifier and connects the antenna to the coax cable when transmitting.

Table 1. Main technical data

- **Passive operation; transmitting or receiving:**
  - VSWR = 1.4 - 1.6 in 2-m amateur radio band
  - VSWR = 1.2 - 1.7 in 6-m amateur radio band
  - Loss introduced by relay and switching circuit: 0.1 dB (2-m band)
  - Permissible transmit power: >50 watt
- **Active operation at U_b = 11 to 15 V (13.5 V typ.)**
  - Current consumption: 60 mA
  - Field strength conversion constant $k_A = U_0 / E = 0.5$ m
  - Ripple of k_A: ±1 dB between 150 kHz and 65 MHz
  - Noise level $P_{N0} / \Delta f = -155$ dBm / Hz ±1.5 dB between 1 and 60 MHz
    $U_{N0} / \sqrt{\Delta f} \cong 4$ nV / √Hz
  - Equivalent noise level: 8 nV / m × 1 / √Hz
  - Large-signal behaviour: 2nd-order intercept point IP_2 = 48 dBm; 3rd-order intercept point IP_3 = 30 dBm

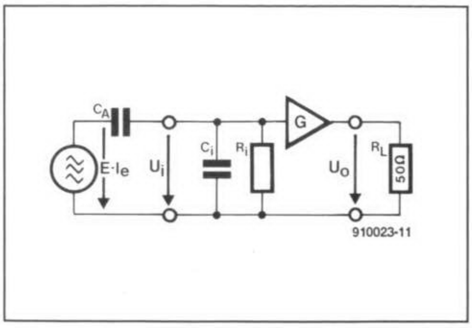


Fig. 1. Basic schematic and parameters of an active antenna.

The relay is controlled remotely from the receiver or transceiver.

As illustrated in Fig. 2, the active rod antenna supplies a virtually constant voltage relative to the field strength of signals in the range from VLF to VHF. It is precisely this characteristic that makes the active rod suited to relative field strength measurement. Fitted on a car roof, it allows the in-band and out-of-band signal levels of SW stations to be monitored. The antenna is also suitable for mobile long-distance SW reception, provided you take the trouble to drive your car as far as possible from large cities, broadcast transmitter sites and industrial sites. At home, the antenna is best fitted to the metal protection parts of the roof (see the introductory photograph). An alternative location is the balcony railing.

Given its small size, the performance of the active antenna is quite impressive, as shown by the measurement data collected in Table 1.

Background noise and background theory

The field strength conversion factor, k_A, of an antenna of effective length $l_e \cong l/2$ (for values of l smaller than $\lambda/8$) is determined with the aid of the basic circuit shown in Fig. 1. As an example, the factor is calculated at a frequency of 10 MHz and at the associated gain, G, measured as 1.91 (see Fig. 2):

$$k_A = \frac{U_o}{E} = \frac{l}{2} \cdot \frac{C_A}{C_A + C_i} \cdot G$$

$$= \frac{1.25\text{m}}{2} \cdot 10^{-2/20} = 0.5 \text{ m}$$

Next, the intermodulation-free drive margin of the amplifier, P_{max}, and the dynamic range, DR, are calculated with the aid of the second-order and third-order intercept points, IP_2 and IP_3. The purpose of this calculation is merely to compare the performance of one active antenna with that of another. For the practical construction, it has no significance.

$$P_{max} = \frac{1}{3}\,(P_{N0} + 2\,IP_3) \qquad \text{[dBm]}$$

$$DR = P_{max} - P_{N0} = \frac{2}{3}\,(IP_3 - P_{N0}) \quad \text{[dBm]}$$

All power levels in the above equations are entered in decibel-milliwatts (dBm) normalized at 50 Ω for RF; 0 dBm equals 1 mW. For SSB in a bandwidth of 2.5 kHz:

$P_{No} = -155$ dBm + 34 dB = −121 dBm
hence $P_{max} = -20$ dBm, and $DR = 101$ dB

For CW in a bandwidth of 500 Hz:

$P_{No} = -155$ dBm + 27 dB = −128 dBm
hence $P_{max} = -23$ dBm, and $DR = 105$ dB

To many of you, voltage levels may be more familiar than the above dBm values. Assuming an antenna impedance of 50 Ω, and remembering that

$$U = \sqrt{P \cdot R} \qquad \text{[V]}$$

the resultant figures may have more meaning. The result for SSB is a noise voltage, U_{N0}, of 0.21 μV, a maximum antenna voltage, U_{max}, of 22 mV, and a maximum field strength, E_{max}, of 44 mV/m.

For CW we obtain the following values: U_{N0} = 0.09 μV, U_{max} = 16 mV, and E_{max} = 32 mV/m.

In practice, when the field strength of a received station exceeds E_{max}, the intermodulation and interference signals produced by the amplifier will exceed its own noise. The effects of intermodulation cause a number of spurious signals which may be heard in the receiver if they are strong enough to top the level of the background noise produced by man-made, atmospheric and other natural sources.

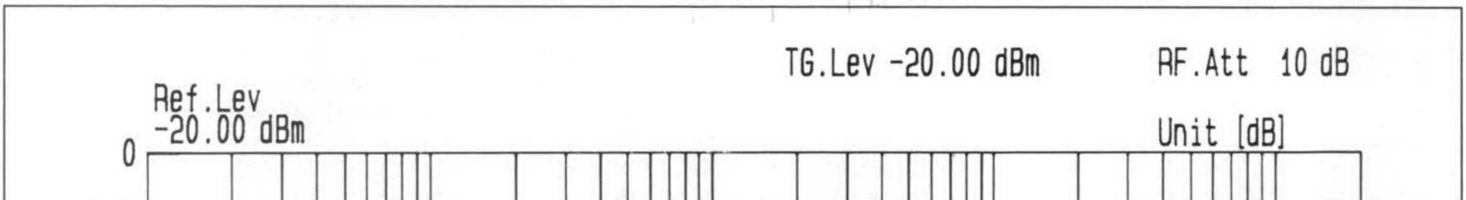
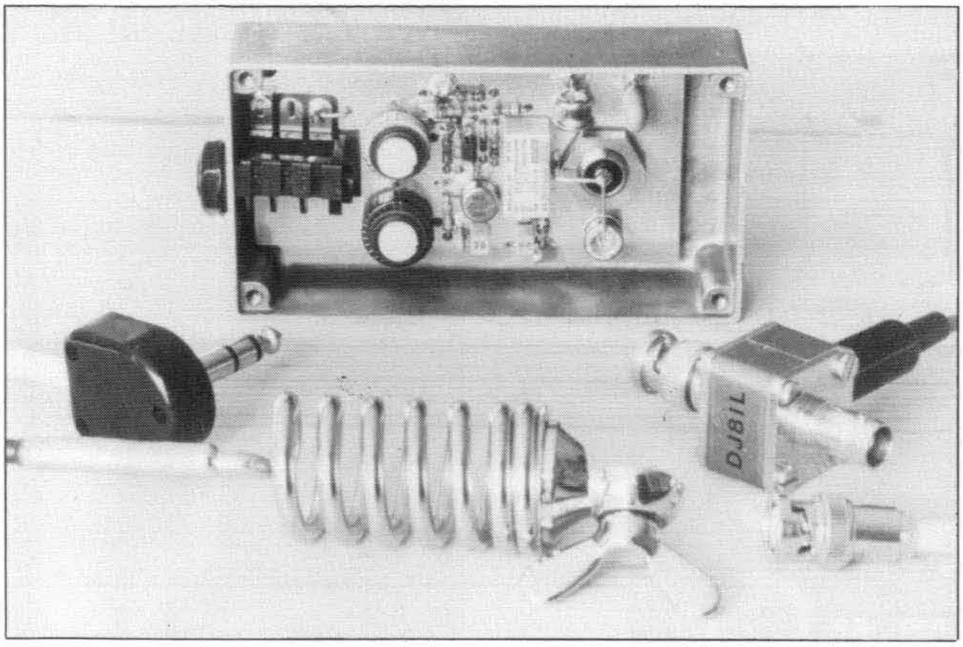


Fig. 2. Frequency response of the active antenna. The rod was simulated by a capacitor of 15pF and a resistor of 25Ω.



The active antenna discussed here was tested with a Yeasu FT-757GXII SW transceiver. The noise produced by the active antenna was audible only above 10 MHz. The dynamic range of the receiver with the active antenna switched on is smaller than that of the receiver proper, while the dynamic range of the receiver alone is greater than that of the preamplifier alone.

The achievable signal-to-noise ratio can only be improved by means of a directional antenna, which, as most radio amateurs know, have a size of at least $\lambda/2$, and offer a relatively small bandwidth.

From Fig. 2 we could be lead to believe that the active antenna could be used up to nearly 200 MHz. However, when the electrical length of the rod is $\lambda/4$ or greater, the antenna starts to form a low impedance. This means that the preamplifier (which is essentially an impedance converter) can be omitted, and the rod connected direct to the input of the VHF receiver. Consequently, the VHF receiver will produce a low noise voltage relative to its own noise level. For an impedance of 50 Ω, the equivalent input noise voltage of the amplifier is 1 nV/√Hz.

Circuit concept

As to the wideband amplifier, if you are after low amplifier noise rather than a high dynamic range, consider the following alternatives:

- a single-stage VMOS-FET based amplifier utilizing, for instance, the VN0808M or the VN66AK from Siliconix;
- a two-stage amplifier with a source follower at the input and an emitter follower with strong feedback at the output.

The second alternative has certain advantages: the input capacitance C_i is small; the

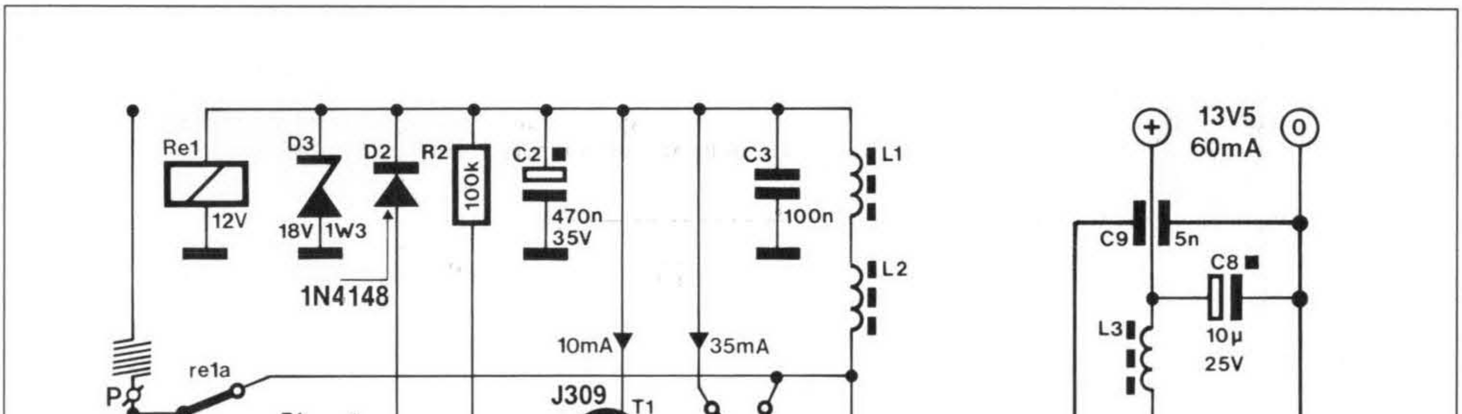


Fig. 3. Circuit diagram of the active antenna, consisting of a ⅝-lambda whip with base coil, a wideband RFamplifier and a supply voltage coupling circuit. Relay Re1 is actuated in receive mode only when the amplifier is functional also. During transmit operation, the supply voltage to the amplifier is switched off, so that the whip antenna is connected direct to the transceiver.

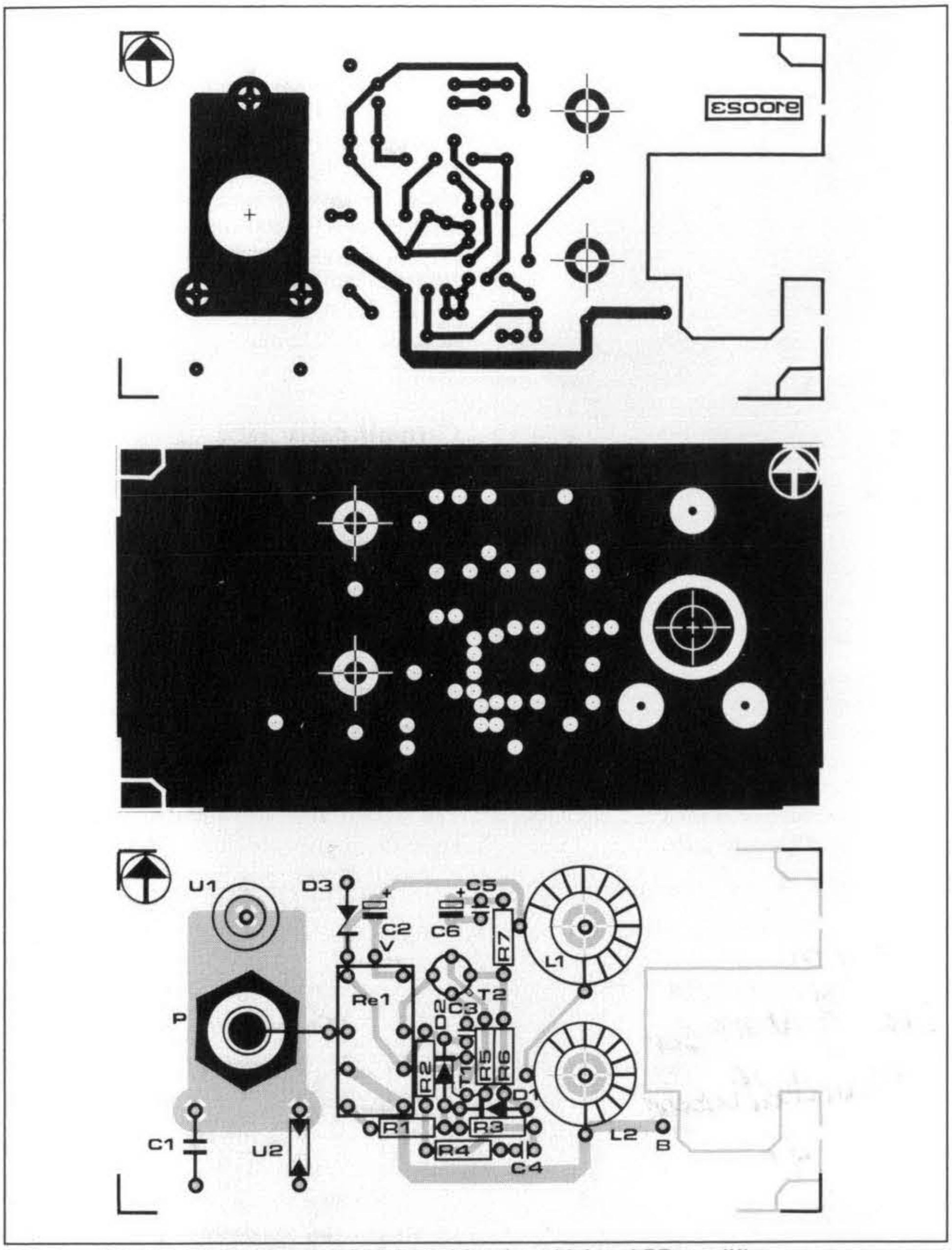


Fig. 4. Double-sided printed circuit board for the wideband RF amplifier.

COMPONENTS LIST

Resistors:

1	27Ω	R7
2	47Ω 0.3W	R1,R8
1	82Ω	R5
1	120Ω 0.3W	R6
2	100kΩ	R2,R4
1	2MΩ2	R3

Capacitors:

1	470pF 3kV ceramic	C1
1	5nF feedthrough	C9
1	100nF 10 V ceramic	C5
2	100nF 35V ceramic	C3,C4
1	470nF ceramic	C7
1	470nF 35V tantalum	C2
1	4μF7 10V tantalum	C6
1	10μF 25V tantalum	C8

Semiconductors:

2	1N4148	D1,D2
1	18V 1.3W zener diode	D3
1	J309 (Siliconix)	T1
1	2N5109 (Motorola)*	T2

Miscellaneous:

1	12V relay with 2 c/o contacts (SDS** type DS2E-12V)	Re1
1	noble-gas surge arrester 145V @ 5,000A	U1,U2

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noise level U_N is low; the current consumption is modest and, importantly, the amplifier has no adjustments.

The circuit shown in Fig. 3 is a source follower based on a Type J309 n-channel junction FET (T_1) from Siliconix or National Semiconductor. This transistor offers a good large signal performance by virtue of the relatively high and constant drain current. To stabilize the drain current, the source resistor, R_5, is not connected to ground but in parallel with the base-emitter junction of the output amplifier transistor, T_2. This is a Type 2N5109 medium-power wideband transistor designed for use in CATV head-end stations. The transistor (manufactured by Motorola) is marked by low noise and excellent linearity. If you can not get hold of the 2N5109, you may use the 2N5943, 2SC1252 or 2SC1253 as a near equivalent.

The linearity of the output stage is ensured by a relatively high degree of feedback. The gain of the stage is set to about 1.9, allowing k_A factors of between 0.2 and 0.5 to be achieved. The output impedance of the amplifier is 50 Ω to allow conventional coax to be used. The output signal is fed to the receiver via a diplexer consisting of R_8, L_3, C_7 and the input impedance of the receiver.

The supply voltage arrives at the amplifier and the transmit/receive relay via chokes L_1 and L_2. The chokes must be capable of withstanding the transmit power while not introducing significant losses or resonance at any frequency within the range covered by the active antenna.

Two small gas-filled surge arrester tubes are provided to protect the amplifier against voltage peaks caused by lightning and 'electrostatic rain'. In addition, two diodes, D_1 and D_2, and a series resistor, R_1, protect the

Table 2. Inductor winding data

L1: 22 turns of 0.2 mm dia. copper enamelled wire on a ferrite ring core of o.d./i.d.=16 mm/9.6 mm; h = 6.3 mm; type B64290-K45-X830 from Siemens.
L2: 20 turns of 0.2 mm dia. copper enamelled wire on a ferrite ring core of o.d./i.d. = 14 mm/9 mm;h = 5 mm; type 43220209718 from Philips Components. Alternative type: FT50A-61 (Micrometals/Amidon).
L3: 21 turns of 0.2 mm dia. copper enamelled wire on a ferrite ring core of o.d./i.d. = 16 mm/9.6 mm; h = 6.3 mm; type B64290-K45-X830 from Siemens.

Main electrical data of the ferrite cores:
L1, L3: material N30, μ_i=4,300; A_L=2.77 μH. L1=1.3 mH; L3=1.2 mH.
L2: material 4C6; μ_i=120; A_L=53 nH; L2=21 μH.

Siemens distributor in the UK is ElectroValue Limited • 28 St Judes Road • Englefield Green • Egham • Surrey TW20 0HB. Telephone: (0784) 33603. Telex: 264475. Fax: (0784) 35216. Northern branch: 680 Burnage Lane • Manchester M19 1NA. Telephone: (061 432) 4945.

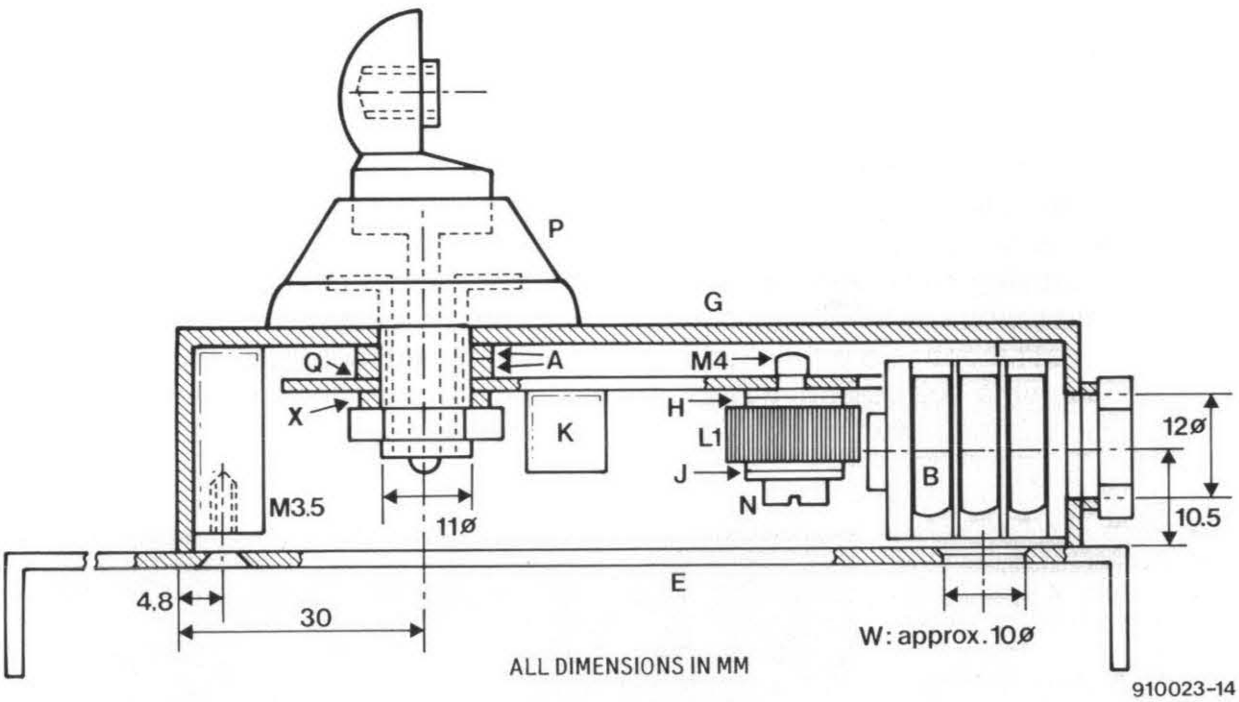


Fig. 5. Mechanical outline of the active antenna. The whip antenna is bolted to the top part of the base, P.

gate of the J309. The zener diode, D_3, has three functions: first, it limits the maximum supply voltage; second, it prevents back-e.m.f. from the relay coil; and third, it acts as a protection against polarity reversal of the supply voltage.

Construction and earthing

Figure 4 shows the double-sided, not through-plated, printed circuit board for the amplifier. The component side is largely unetched to allow it to function as a ground plane. All component terminals that are connected to ground are soldered at both sides of the PCB. Note that all component terminals, whether grounded or not, must be kept as short as possible.

The construction of the board is largely apparent from the photograph of the prototype. Note the orientation of the tab on the case of T_2. If you use a transistor with a terminal connected to the case (the near equivalents of the 2N5109), this terminal is soldered to ground.

The FET, T_1, is fitted with its flat side facing D_2, R_2 and the relay. The centre terminal of the FET (source) is not indicated on the component overlay. The outer two terminals are the drain (near C_3) and the gate (near D_1).

The frequency response of the amplifier will be a little smoother than shown in Fig. 2 when toroid core L_3 is allowed more space than in the 'cramped' prototype shown in the photograph. The supply/coax connection should be sealed to prevent rain or dew entering the coax cable or the amplifier enclosure. An N-type plug and socket may be used because they are waterproof. The disadvantage however is the relatively high cost. The author used a cheaper alternative in the form of a jack socket and a mating plug. After 1½ years of continuous use on the roof no traces of corrosion could be detected inside the amplifier enclosure.

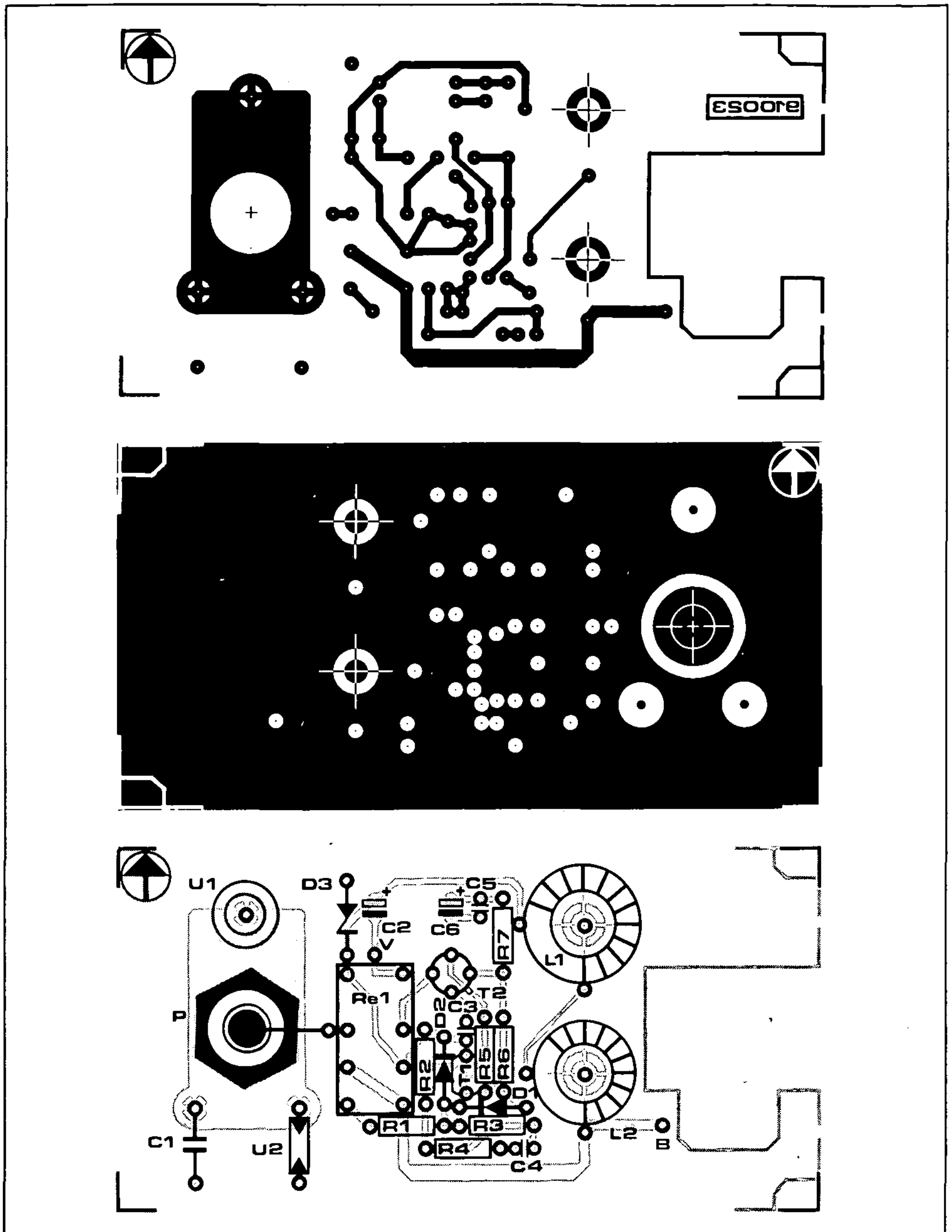
Table 3. Mechanical parts

o.d. = outside diameter
i.d. = inside diameter

A: Metal washer o.d./i.d.=16/11 mm; 4 mm thick.
B: 6.3 mm dia. jack socket.
E: chrome-plated aluminium support plate.
H,J: Isolating washer o.d./i.d.=12/4 mm; 2 mm thick.
K: relay.
N: nylon screw M4×15 mm.
Q: stainless steel spring washer o.d./i.d.=16/11 mm; 0.2 mm thick.
S: 6.3 mm dia. angled jack plug.
X: isolating washer o.d./i.d.=16/11 mm; 2 mm thick.
W: water exhaust
G: diecast case 111×60×27 mm (Eddystone Radio 27134P).
P: antenna base.

In the introductory photograph the antenna is seen attached to the zinc rim (of width *w*) on the edge of a flat roof. The antenna is secured to a 2-mm thick U-shaped chrome-plated aluminium piece (E) that is clamped on to the metal rim of the roof. The size of the aluminium plate is 200 (L) × (100+*w*+100 mm) (W) mm. The zinc rim on the roof is connected to the lightning conductor system. It should be noted that there may exist a up to a few 100 mV of hum between the lightning conductor system (which is earthed) and the protective earth on the mains system, to which the RF equipment is connected. To prevent a stray current flowing via the relay coil and L_1, L_2 and L_3 (as a result of the potential difference, which is essentially hum), the two earthing systems are capacitively coupled via C_1 The value of C_1 is 470 pF, which is much greater than C_A. The capacitor effectively prevents harmonics of the hum voltage occurring in the VLF range of the receiver. Finally, use insulating spacers between the PCB and the metal enclosure, as shown in Fig. 5. This is done to ensure that the two earth systems are not interconnected. ■

Note:
This article was translated and edited by the staff of *Elektor Electronics* from an originally German version that appeared in the 2-1990 issue of CQ-DL magazine. We thank the editors of CQ-DL for their permission to reproduce Mr. Becker's article.



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fier and the transmit/receive relay via chokes L1 and L2. The chokes must be capable of withstanding the transmit power while not introducing significant losses or resonance at any frequency within the range covered by the active antenna.

PREAMPLIFIER FOR MOVING-MAGNET PICK-UP

by T. Giffard

The basic function of a pick-up element is to translate the motion of the stylus into an electrical signal. The most popular cartridges in use are the moving magnet (fixed coil and tiny magnet) and the moving coil (fixed magnet and tiny coil). In both designs the vibrations of the stylus cause fluctuations in a magnetic field. Last month we published a preamplifier for advocates of the moving coil design; this month we turn our attention to those who favour a moving magnet element.

MOVING magnet carrtridges have a relatively high output and require a load of some 47 kΩ in parallel with a specified capacitance, normally 200–500 pF, for optimum performance.

The output voltage of the cartridge increases with frequency as shown in Fig. 1 following the recording characteristic of LP records in accordance with the IEC recording standard. This is a 1976 adaptation of the well-known RIAA (Recording Industry Association of America) recording standard. The preamplifier is therefore required to have a playback characteristic as shown in Fig. 1 with de-emphasis time constants of 8 ms, 3180 µs, 318 µs and 75 µs,corresponding to roll-off points of 20 Hz, 50 Hz, 500 Hz and 2120 Hz respectively. The output level is 0.8–2.0 mV cm^{-1} s^{-1} of groove modulation velocity, resulting in a mean output of some 3–10 mV at 1 kHz.

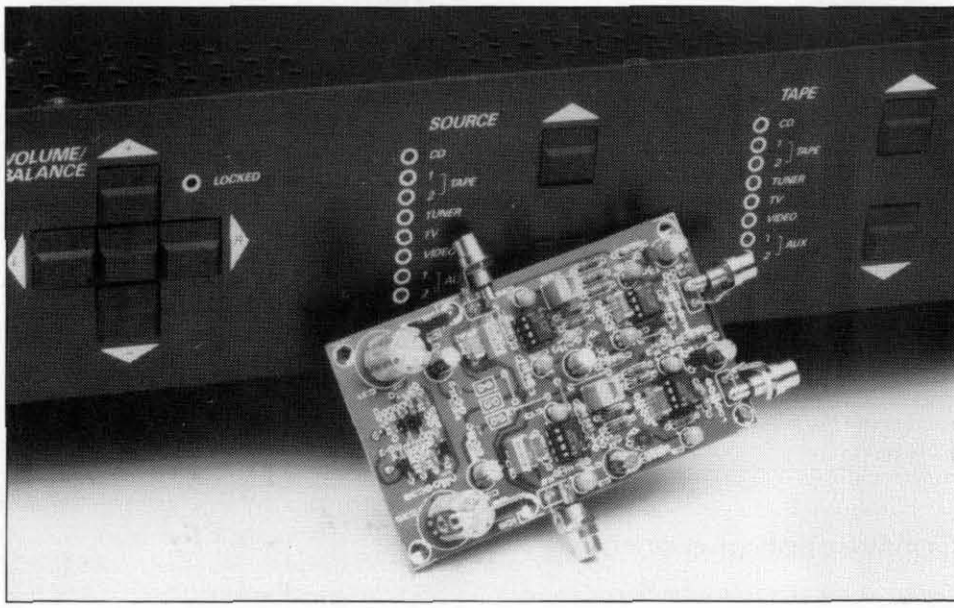
The RIAA recording characteristics did not specify the 8 ms time constant which, particularly in the present preamplifier, is of importance since it attenuates all kinds of subsonic sound (at 2 Hz by as much as 20 dB compared with the RIAA characteristic). The use of the IEC rather than the RIAA characteristic means that the amplifier reproduces records cut according to the RIAA standard with a greater attenuation of any rumble, while those cut in conformity with the IEC standard are reproduced correctly. This assumes, of course, that the components in the correction networks are close tolerance types.

Circuit description

This is the first quality preamplifier designed around opamps that we have ever published. The opamps used have the lowest noise figure currently commercially available—more about this later.

The circuit diagram of the preamplifier is shown in Fig. 2; it will be discussed on the basis of the left-hand channel.

The signal from the pick-up element is applied to the non-inverting input of IC_1. The input impedance is formed by the parallel network R_1-C_1. The correct value of these components is normally given by the manu-



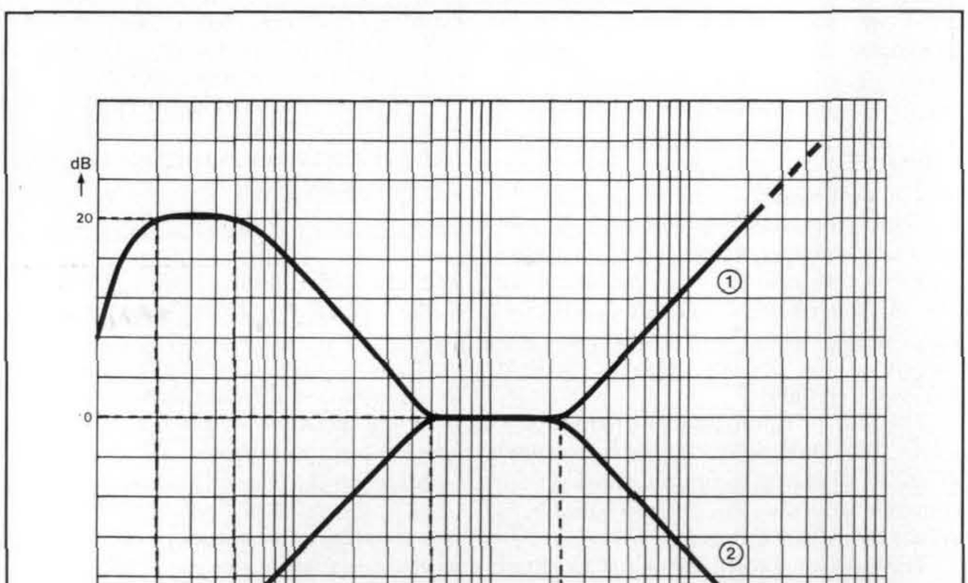


Fig. 1. Recording (1) and playback (2) characteristics according to the 1976 IEC standard.

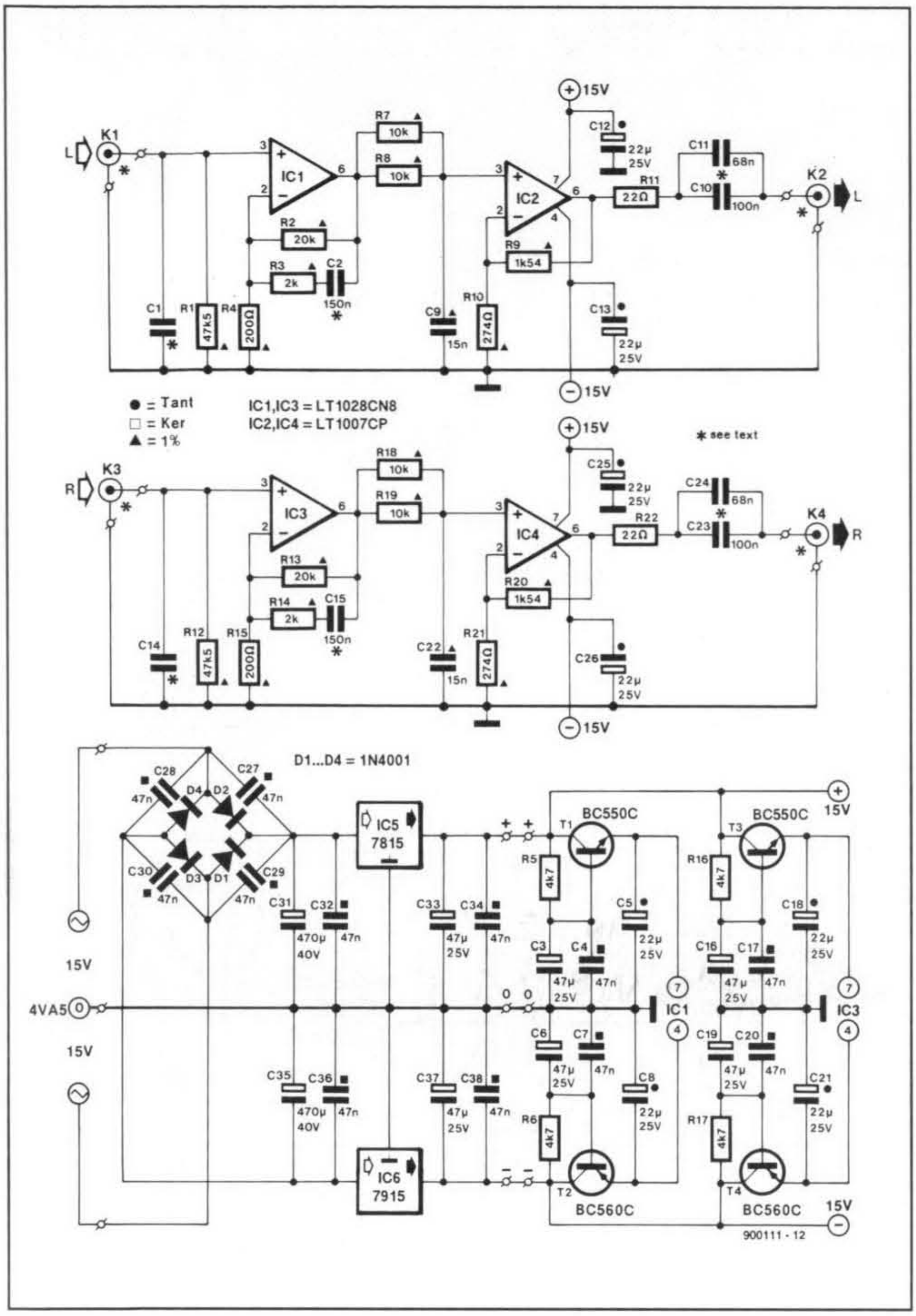


Fig. 2. Circuit diagram of the stereo preamplifier and associated power supply.

facturer of the cartridge, although that of the resistor is usually taken as 47 kΩ. The value of the capacitor varies between 200 pF and 500 pF. Note that C_1 is shunted by the capacitance of the cable, which is normally 100–200 pF and this must, of course, be deducted from the capacitor value specified by the manufacturer. If no manufacturer specification is available, make R_1 = 47 kΩ and C_1 = 100 pF.

The feedback of low-noise opamp IC_1 has been arranged to ensure that the opamp does not only provide an amplification of ×10, but also the required correction of the low frequency part of the signal. The high-frequency part is corrected by R_7-R_8-C_9. Apart from the 20 Hz roll-off point, that completes the frequency response requirement according to the IEC standard.

The output of IC_1 is applied to IC_2, whose characteristics are virtually identical with those of the well-known OP27. The amplification of the stage is ×6.5.

Although R_{11} at first sight appears to have no function, it does, in fact, stabilize the opamp if the load is highly capacitive.

The 20 Hz filter is formed by the parallel combination of C_{10} and C_{11} and the input impedance of the power amplifier connected to K_2. The value of the capacitors may be calculated from $C_{10}+C_{11} = 1/40\pi R_i$, where R_i is the input resistance (47 kΩ). The old RIAA characteristics may be retained by giving both C_{10} and C_{11} a value of 470 nF.

The power supply for the stereo preamplifier contains no fewer than six regulator stages. The transformer rating is about 4.5 VA.

Any spurious noise signals produced by the rectifier diodes are suppressed by capacitors shunting the diodes.

The ±15 V output of regulators IC_5 and IC_6 is suitable for powering IC_2 and IC_4, but for the low-noise opamps it needs further smoothing in T_1–T_4. Strictly speaking, since these transistors are not saturated, they function not so much as regulators but more as filters. Their bases are driven via *RC* networks with a very low cut-off point (0.7 Hz). This arrangement ensures effective suppression of any residual hum and other noise extant on the output voltage.

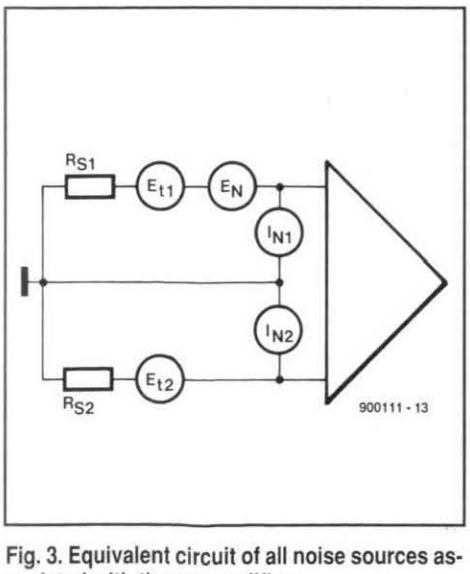


Fig. 3. Equivalent circuit of all noise sources associated with the preamplifier.

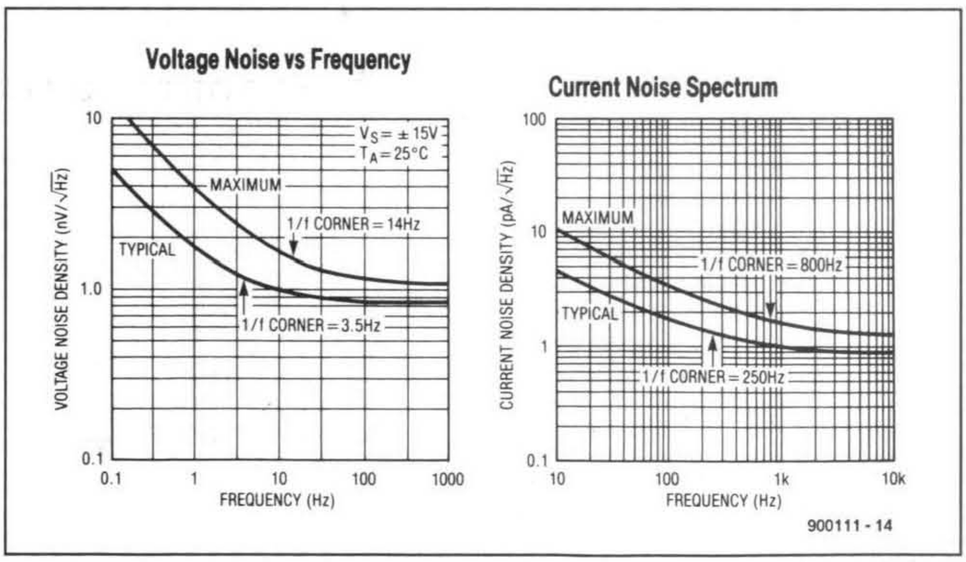


Fig. 4. Voltage noise and current noise as a function of frequency (Courtesy Linear Technology).

Noise

Noise figures do not tell the whole story of the noise that an amplifier will produce. The level of noise is determined primarily by the input circuit(s). The equivalent circuit in Fig. 3 shows where the sources of noise are to be found. Sources E_N, I_{N1} and I_{N2} represent all possible causes of noise in the opamp. Sources E_{t1} and E_{t2} represent the thermal noise level of the source impedances (R_{s1} and R_{s2}) that are connected to the opamp. The noise level at 25 °C is about $0.13\sqrt{R}$ (nV $\sqrt{\text{Hz}^{-1}}$) where R is expressed in Ω. Although this noise is not under the direct control of the designer, it does have an effect on the amplifier. The extent of its effect can only be assessed once we know more about the noise caused by the opamp.

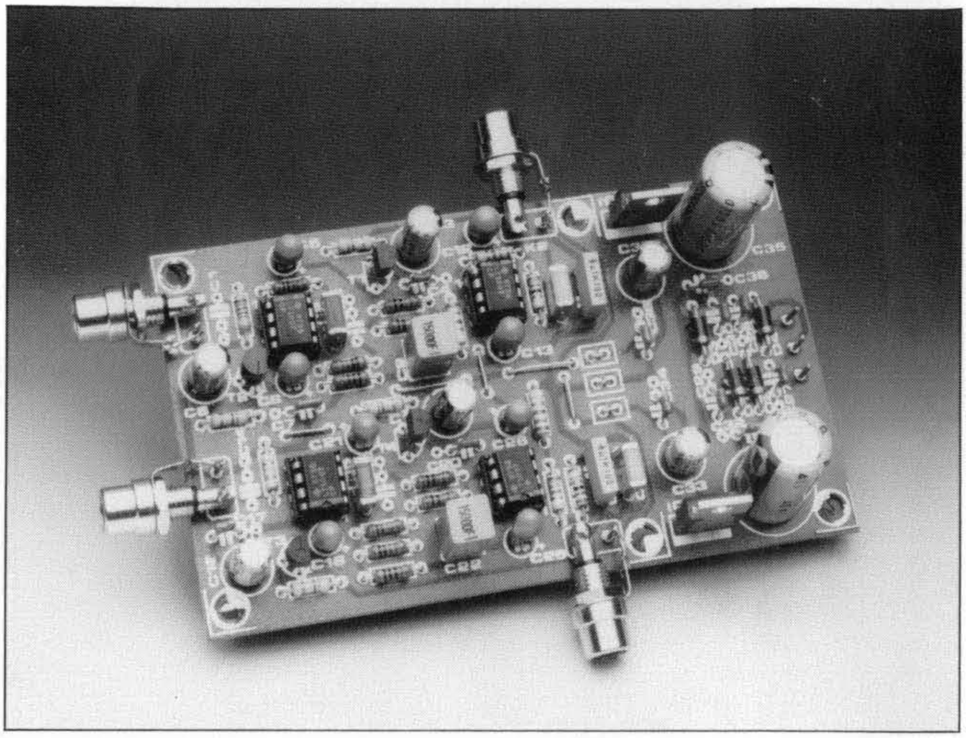
The noise voltage, E_N, and noise current, I_N, as function of frequency are shown in Fig. 4. In both, the characteristic may be divided into two: a flat part and a part where the amplitude rises with decreasing frequency. This comes about because, basically, both voltage noise and current noise consist of two types of noise: white noise and $1/f$ noise.

White noise is characterized by by its level being the same at all frequencies, whereas the level of $1/f$ noise is inversely proportional to the frequency. The characteristic curve is therefore a descending line ($1/f$ component greater) that slowly becomes horizontal (white noise greater). The frequency at which the levels of the two components are equal is usually called the rolloff frequency.

The effects of these noise sources may be limited by three golden rules:

- The source impedance must be kept as low as possible to keep thermal noise low. This in turn will keep down the level of the noise voltage caused by noise currents I_{N1} and I_{N2} across the source impedances. Note, however, that with the LT1028 the effects of the noise current are already exceeded by those of the thermal noise when the source impedance is smaller than 20 kΩ. Nevertheless, to ensure minimum noise in the LT1028, the source impedance must be less than 400 Ω to ensure that both the noise current and the thermal noise may be ignored: only the noise voltage then still plays a role. In their data book, Linear Technology therefore state that the use of the LT1028 is sensible only when the source impedance <400 Ω.
- The $1/f$ rolloff frequency must be chosen as low as necessary (and possible). This ensures that the amplifier has a good signal-to-noise ratio at even low frequencies. It does not make sense to use a $1/f$ rolloff frequency that lies beyond the bandwidth of the amplifier.
- Match the bandwidth of the amplifier to that of the signal. Noise that lies outside the bandwidth of the amplifier is amplified to a lesser degree than noise within it. In other words, the output of an amplifier with a correct bandwidth will contain less noise than that of an amplifier with too large a bandwidth.

With the LT1028 a big step has been taken



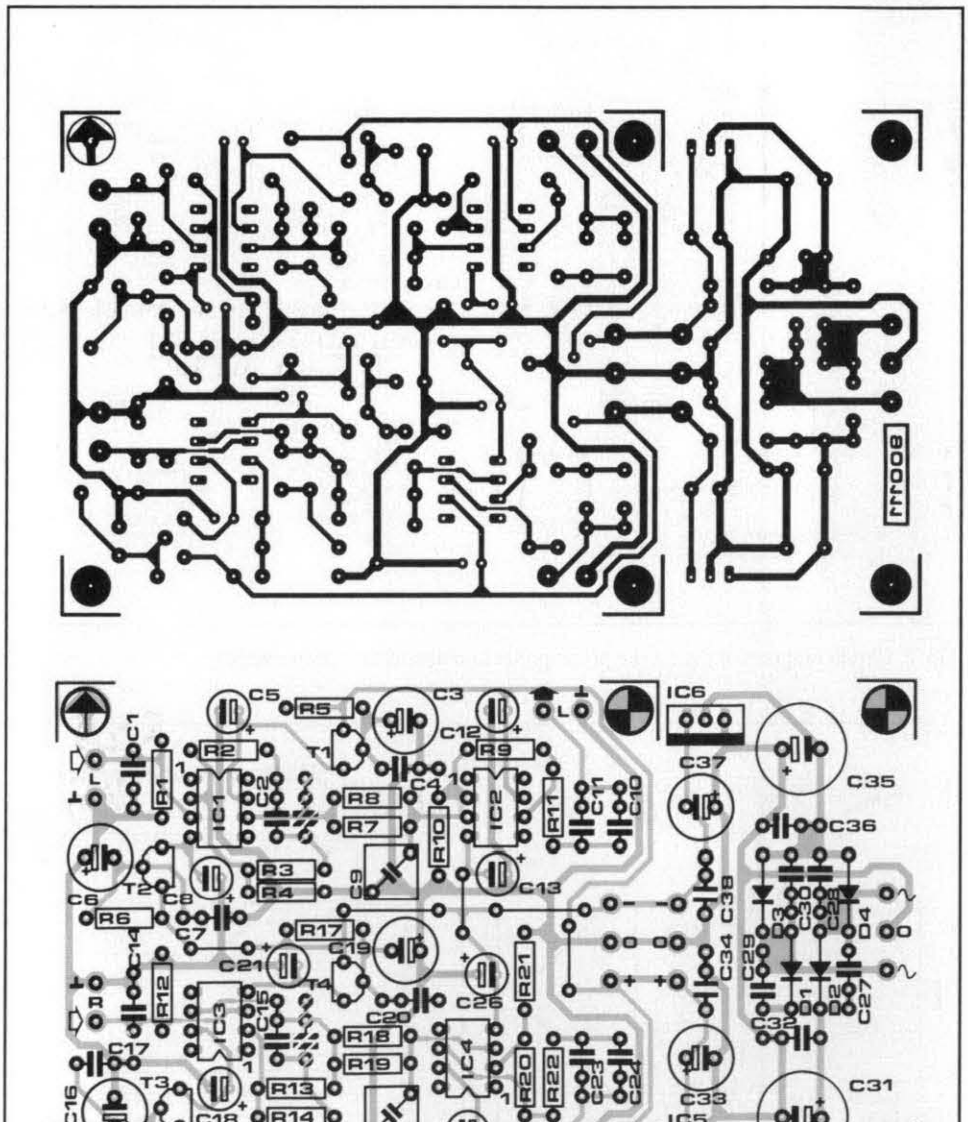


Fig. 5. The printed-circuit board for the preamplifier is available through our Readers' services.

towards a low-noise design, but the designer has little control over the source impedance. A moving-magnet cartridge usually has an impedance of 1–2 kΩ and that means that not the opamp but the pick-up element produces most noise. In other words, the amplifier is too good for a moving-magnet element.

There are, however, high-output moving-coil elements with an impedance of about 200 Ω; when these are used, it is the noise voltage of the opamp that will determine the signal-to-noise ratio.

In the latter case, the $1/f$ frequency also becomes important. For the LT1028 that lies at about 10 Hz. That is well below the 20 Hz rolloff frequency designed into the preamplifier: no problem here. In case of a moving-magnet element, the frequency at which the noise of the cartridge is exceeded by the $1/f$ noise is lower still, so no problem here either.

The question is, however, what are we going to do with the bandwidth? After all, the correcing networks are affected by the bandwidth. To keep the bandwidth as narrow as consistent with the amplifier requirements, the entire IEC correction network should be incorporated in the feedback loop. Only then, no more noise than absolutely unavoidable will be amplified. There is, however, also the requirement of keeping the source impedances small. This means that the value of R_4 (R_{15}) must be kept low (200 Ω in the prototype).

Since the consequent 40 μF capacitor would be very large (it has to be a film type, because an electrolytic one would not do), the 20 Hz high-pass filter can not be accommodated in the feedback loop.

Similarly, the 2122 Hz low-pass filter can not be included in the feedback loop, because an opamp that is used as a non-inverting amplifier has an amplification of not less than ×1. The required attenuation of the highest frequencies can not, therefore, be achieved by the feedback loop.

Construction

As always, a high-quality amplifier as described here should preferably be constructed on the printed-circuit board shown in Fig. 5. The most conspicuous aspect of this is the earth track that divides the left- and right-hand channels.

It is also possible to separate the power supply from the the remainder of the board, at least as far as its part without transistors T_1–T_4 is concerned. Those transistors are located as close as possible to IC_1 and IC_3 for reasons explained earlier.

This arrangement makes it possible for a range of a.c. input voltages to be used: the amplifier works readily from ±7.5–20 V. In all cases, it is essential to keep the mains transformer or adapter well away from the board. If, for instance, the amplifier is built as a stand-alone unit (in a metal enclosure), it is advisable to house the transformer in its own (metal) enclosure away from the amplifier.

The connecting cables between the amplifier and the pick-up element must be as short as feasible. It is not a bad idea to build the amplifier within the record player.

Populating the board is fairly straightforward, although it requires absolute first-class soldering. It is necessary to tin the earth track again before any other work is begun. The only components that require extra care are:

R_1, C_1, R_{12}, C_{14}, which ensure correct termination of the pick-up element. Their value is as indicated in the documentation of the element. Note that the cable capacitance must be deducted from the specified capacitance to arrive at the correct values for C_1 and C_{14}. If no data on terminating the element are available, use the values given in the parts list.

C_2, C_{15} should preferably have been 1% polystyrene types, but those are so large that the size of the PCB would have to be increased by almost 40%. It was therefore decided to use MKT—metallized polyester (PEPT)—types. These must, however, be carefully selected to ensure accuracy of the circuit. Apply soldering heat for a short time only, because the value of MKT capacitors changes when they are overheated.

C_{11}, C_{10}, C_{23}, C_{24} form, together with the input impedance of the amplifier connected to K2 and K4, the 20 Hz high-pass filter. The correct value is calculated from

$$C_{10} + C_{11} = C_{23} + C_{24} = 1/40\pi R_i.$$

Note that the board allows the use of two capacitors in parallel.

It is important that the capacitors just discussed are of exactly the same value in the lefty- and right-hand channels. This means that two capacitors for the C_2 and C_{15} positions that have the same value are preferred over two values that are not equal but are closer to 150 nF. ■

PARTS LIST

Resistors:

All resistors are 1% metal film types unless otherwise stated

R1, R12 = 47.5 kΩ[1]
R2, R13 = 20 kΩ
R3, R14 = 2 kΩ
R4, R15 = 200 Ω
R5, R6, R16, R17 = 4.7 kΩ[2]
R7, R8, R18, R19 = 10 kΩ
R9, R20 = 1.54 kΩ
R10, R21 = 274 Ω
R11, R22 = 22 Ω[2]

Capacitors:

C1, C14 = 100 pF[1]
C2, C15 = 150 nF MKT (match)
C3, C6, C16, C19, C33, C37 = = 47 μF; 25 V; radial
C4, C7, C17, C20, C27–C30, C32, C34, C36, C38 = 47 nF, ceramic
C5, C8, C12, C13, C18, C21, C25, C26 = 22 μF; 25 V; tantalum
C9, C22 = 15 nF; 1%; polystyrene
C10, C23 = 100 nF; MKT
C11, C24 = 68 nF; MKT[1]
C31, C35 = 470 μF; 40 V; radial

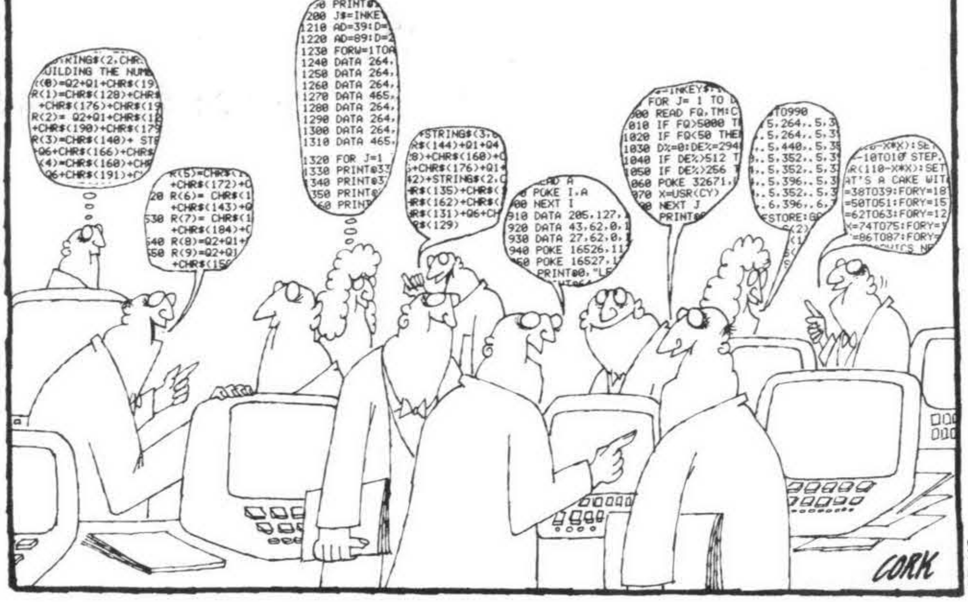
Semiconductors:

D1–D4 = 1N4001
T1, T3 = BC550C
T2, T4 = BC560C
IC1, IC3 = LT1028CN8[3]
IC2, IC4 = LT1007CP[3]

Miscellaneous:

K1–K4 = gold-plated phono sockets
Mains adapter or transformer, 4.5 VA, with 2×15 V secondary
PCB Type 900111

[1] See text
[2] carbon film type
[3] Linear Technology



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types of noise: white noise and $1/f$ noise.

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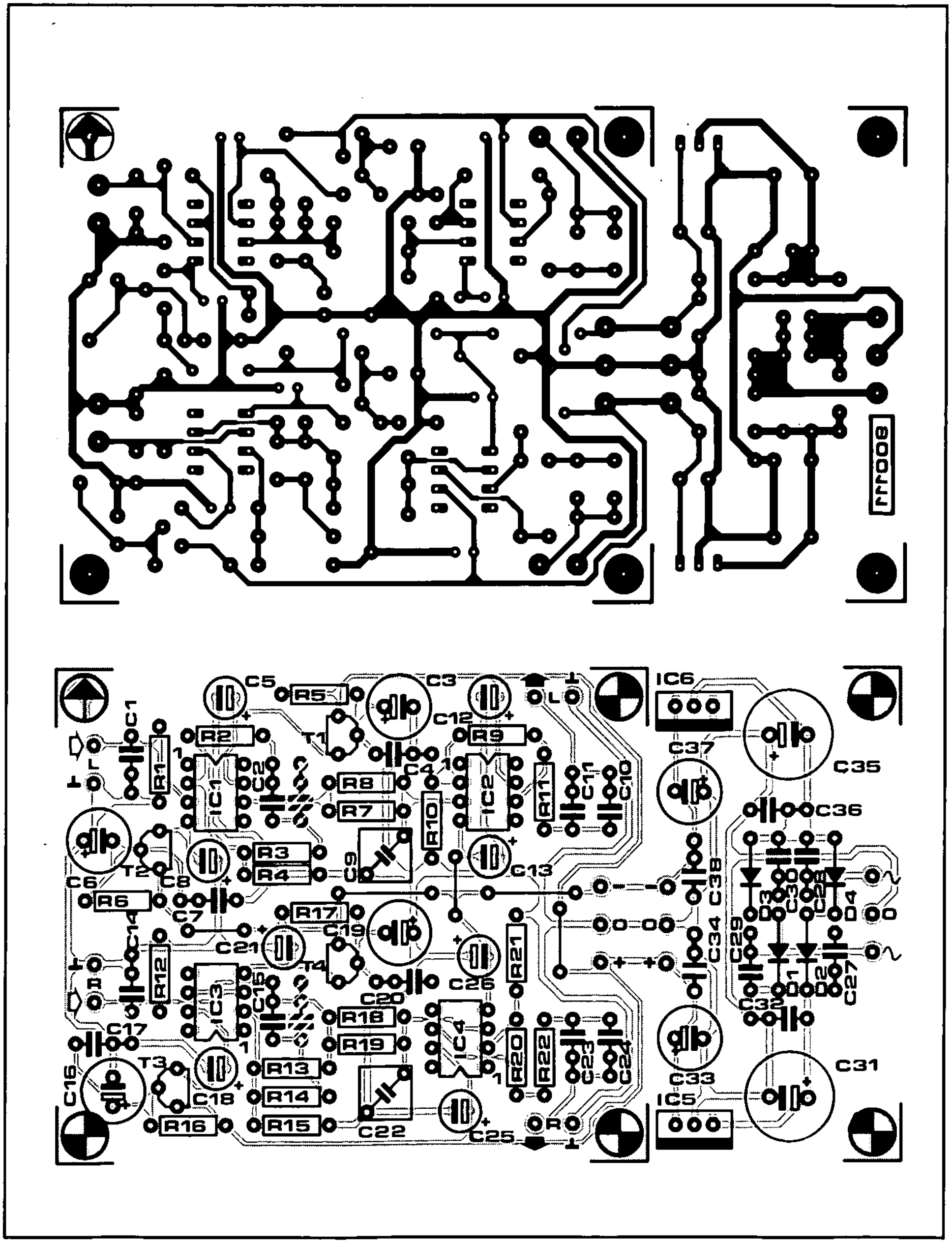

Fig. 5. The printed-circuit board for the preamplifier is available through our Readers' services.