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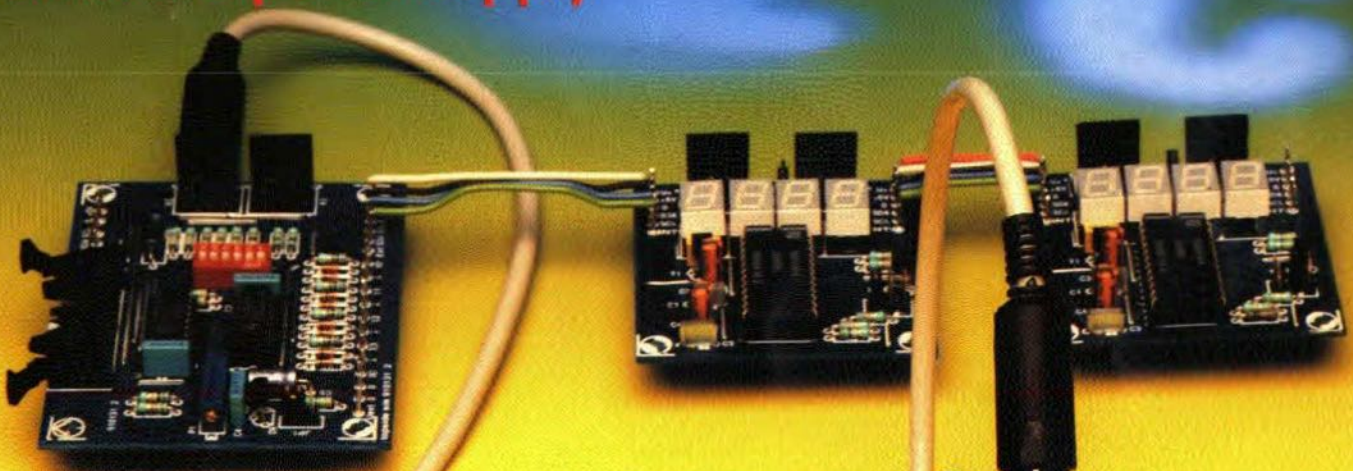
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ELEKTOR ELECTRONICS

I²C interface for PCs

Audio/video switching unit

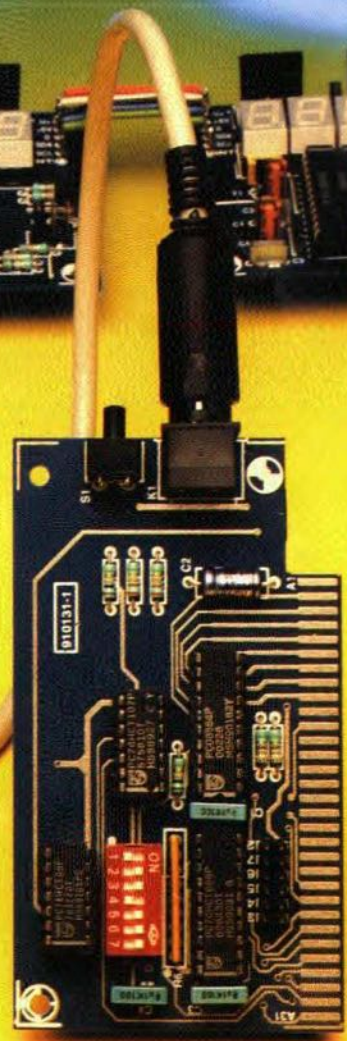
Switch mode power supply



Mini square-wave generator

Direct digital synthesis-DDS

PWM temperature controller



Correction:

'Alexander Graham Bell'
(December 1991)

The last line of this article has unfortunately been omitted; the entire last sentence of the article should read: "In this regard, therefore, Alexander Graham Bell is largely responsible for beginning the evolution of a 'global village' culture".

In next month's issue:

among others:

- AD/DA/IO for I²C
- 8051/8032 assembler course
- Measurements on power supplies
- FM tuner – Part 1
- LC meter
- AF drive indicator
- The flat-top 80 antenna
- Simple timer

Front cover

The I²C interface (Inter-IC Communication) is a Philips invention that has been in use for many years to enable ICs to communicate with each other in complex electronic equipment such as radios, video recorders and television sets. The insertion card shown here (and described on pages 36–40) puts your PC in control of up to ten I²C compatible ICs.

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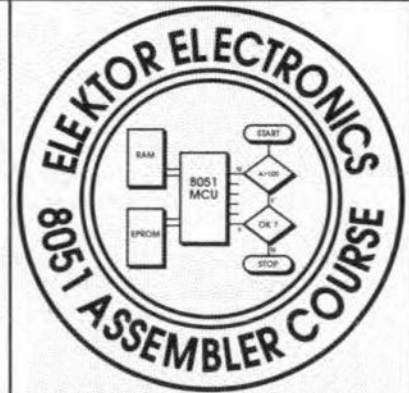
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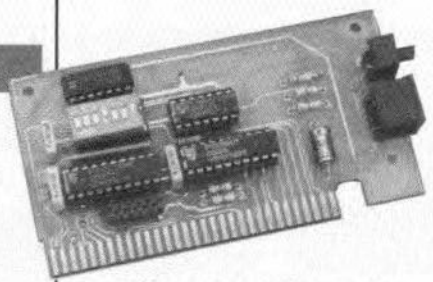
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IMPROVING PORTABLE RADIO PERFORMANCE

The performance of the average portable radio (not including the tiny plastic one and the huge ghetto blaster) is usually quite adequate for general domestic and portable listening. Unfortunately, for the more serious DX enthusiast, various limitations appear: however good the audio amplification may be, it can not improve an inadequate signal at the RF input, or a signal which is subject to heavy interference. All is not lost, and the 'RF pick-up' performance can be improved with some simple low-cost DIY construction. Simple as the suggestions given here may appear, the results can be quite dramatic!

by Richard Q. Marris G2BZQ

MOST transistor portable radios are battery-operated, and can also be plugged into the a.c. mains supply for home use. These radios typically operate on LW, MW, SW and VHF FM. In some cases, one or more of these bands may be eliminated. The average size will probably be around 28x15x6.5 cm (11x6x2.5 inch).

For the experiments to be discussed, the following radios were used, and results later

cross-checked with other makes:

- A Grundig Music Boy (30x18x5 cm). Coverage: LW, MW, SW (5.9 MHz to 18 MHz) and VHF FM;
- A Matsui MR4099 (28x15x5 cm). Coverage: continuous from 150 KHz to 30 MHz, plus VHF FM. Modes: AM, FM, SSB, CW, mono/stereo.

The answer to the previously mentioned

problems for the DX listener is to improve the antenna system. However, separate different treatments are required for the VHF, SW, MW and LW bands.

The short-wave (SW) bands

Most portables have a short telescopic rod antenna. Although this may seem adequate at first sight, a large number of stations are

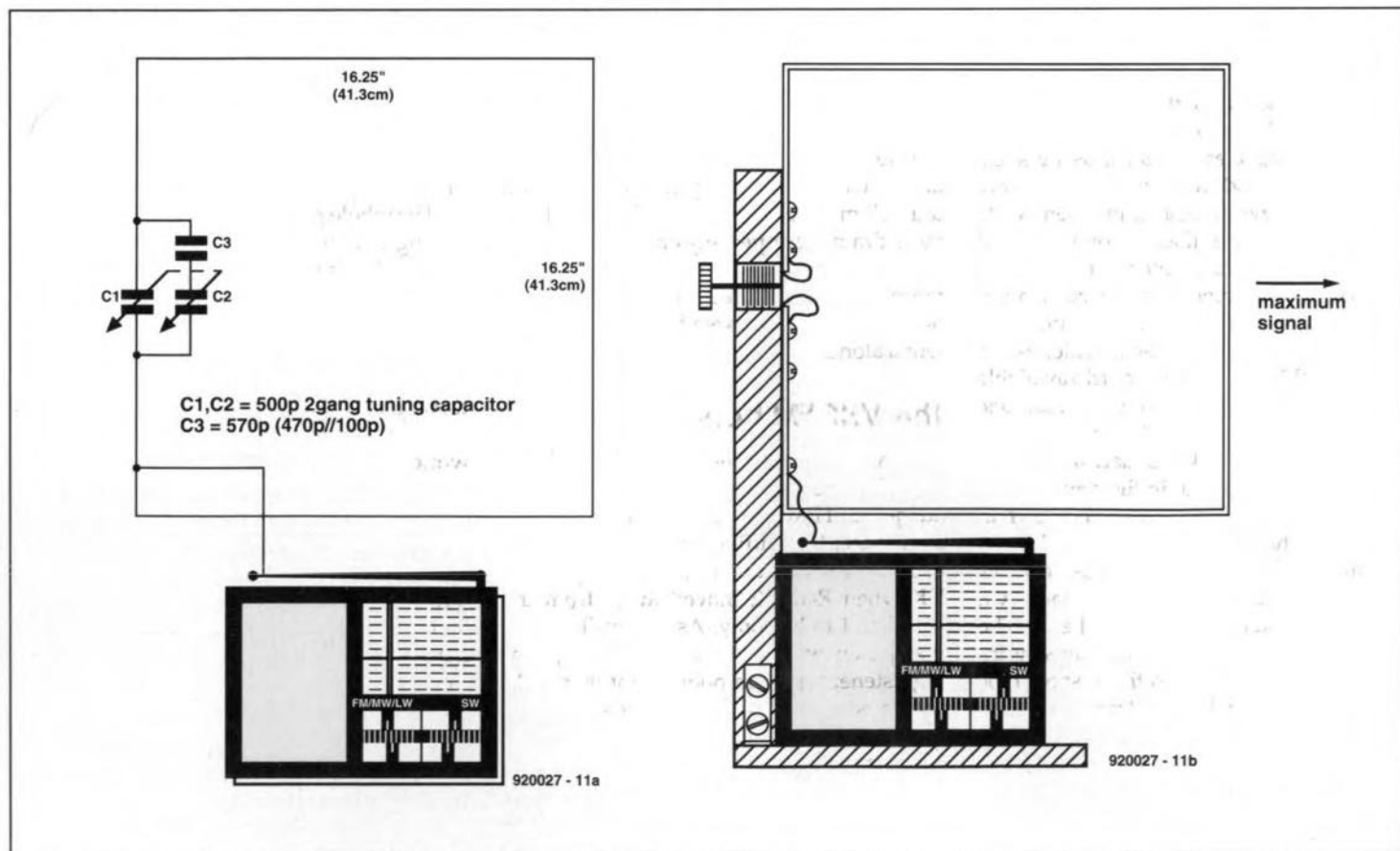


Fig. 1. Circuit (left) and basic mechanical layout of the antenna for the short-wave bands.

being missed completely or obliterated by chaotic interference. If it is possible to fit an external wire antenna, more stations become available, but the interference increases equally. Interference is often directional and consists of other stations, atmospherics, and static electric storms. In addition we have a multitude of man-made electrical interference received from thermostats, TV sets, home computers, vacuum cleaners, electric drills, and motorized vehicles. Legally, of course, all these are fully 'suppressed' but in practice much suppression leaves a lot to be desired, and just does not exist in many cases.

The antenna shown in Figs. 1a and 1b covers the frequency range from 6 MHz to 26 MHz. It consists of a 41.3x41.3 cm (16.5x16.5") square loop made of aluminium strip, and is resonated with a 2-gang 500+500 pF variable capacitor, C1-C2, with C3 in series with one section. A tap, with crocodile clip and lead, is taken from the loop to the tip of the retracted telescopic antenna on the portable radio. The tap on the prototype is about 2 cm (3/4") 'up' from the left-hand corner, giving the best mid-frequency match between the telescopic antenna rod and the larger loop—see Fig. 1a.

The drawing in Fig. 1b shows the physical layout, consisting of a wood base on which the radio stands, plus a vertical member to support the square loop and the resonating capacitors. The whole assembly is unidirectional, as indicated in Fig. 1b, and can stand on a simple turntable to rotate in order to achieve maximum signal and minimum interference.

The construction of the SW antenna is illustrated in Fig. 2. The base should be sufficiently large to accommodate the radio, and heavy enough to prevent the loop tipping over. The vertical member is screwed and bracketed to the rear of the base.

The vertical member should ideally be a good insulating material such as perspex. Failing this, wood was used, thoroughly dried in the airing cupboard, and then, while still warm, given a liberal application of polyurethane varnish, which soaks in as the wood cools. This operation was carried out of doors because of possible toxic fumes.

The loop is made of 2-cm wide, 3-mm thick (0.75x0.125") aluminium strip available in DIY stores. The length used had one round side, and was later bent into a 41.3x41.3 cm (16 1/4x16 1/4") square in a vice. A 3-cm (1.25") gap was cut in the centre of the left-hand side, and holes drilled as shown in Fig. 2. Next, the loop was fastened to the vertical member with brass woodscrews as shown, with the connections made via solder tags fitted under the screw heads. The tuning capacitor, C1+C2, was mounted as indicated, with C3 underneath. A short croc lead was clipped on to the retracted antenna tip, after putting the radio on the base (see Fig. 1b).

Operation is simply a matter of tuning the radio to the desired frequency, and resonating the loop to that frequency. Resonance manifests itself by an increase in signal.

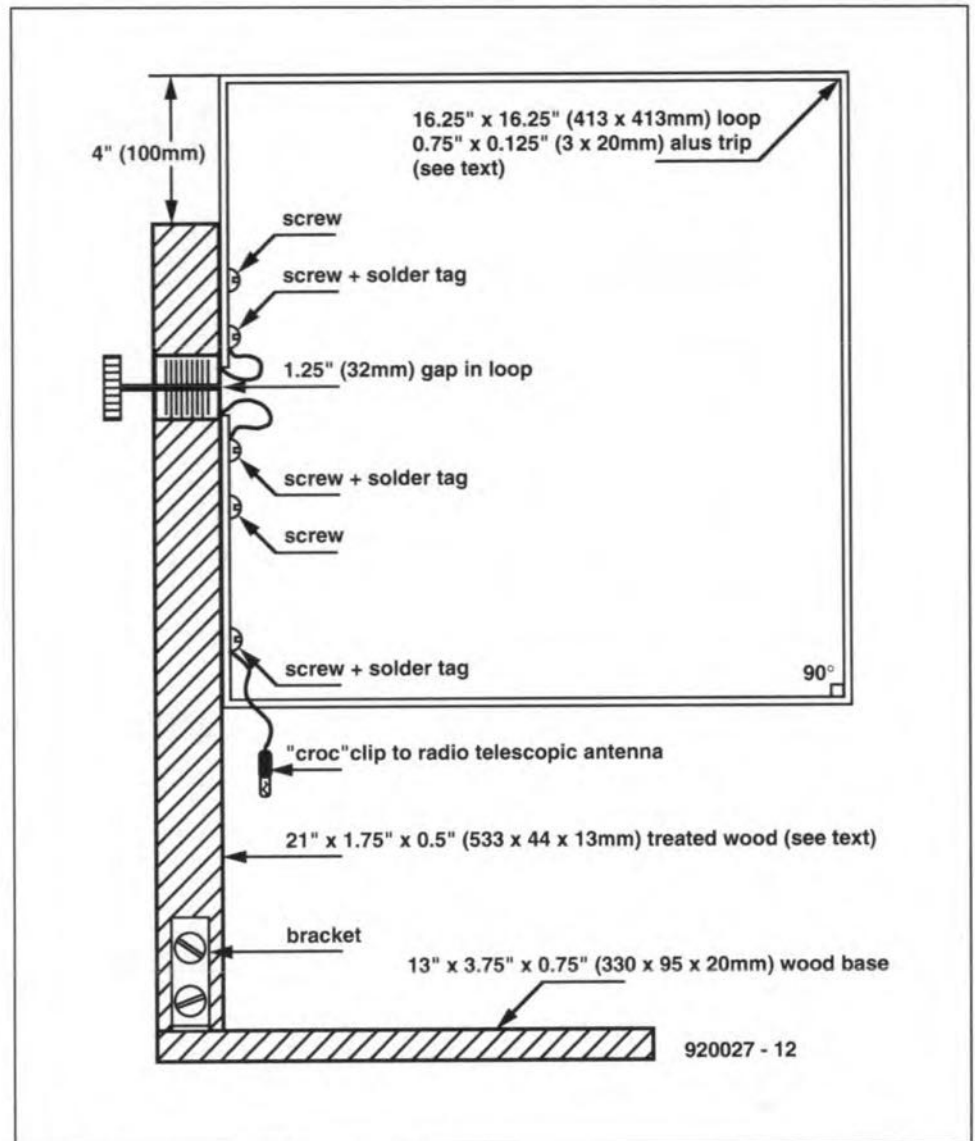


Fig. 2. Construction of the short-wave antenna.

Rotation of the loop and radio will increase the signal further, and greatly reduce interference and background noise. The results are often quite dramatic: a clean signal is obtained where previously it was weak and unintelligible owing to interference. Even more dramatic, especially on the 25-m, 19-m and 16-m broadcast bands, clean long-distance stations have been heard that just did not exist with the extended telescopic antenna alone.

The VHF FM band

In an area where strong VHF FM signals are available, the telescopic antenna may well be adequate. However, in remoter, or hilly, areas, reception difficulties are often experienced. This was found out by many in the UK when Radio 2 moved away from the MW band to FM only. As one well-known disk-jockey was heard to say to a complaining listener who had poor reception on FM with a portable: "Well try moving the radio around the room and waggle the telescopic aerial about".

Actually, this non-technical talk hit the nail on the head. The difficulty was well-illustrated to the author, using both the Matsui and the Grundig, when frequently

visiting a fishing village on the West Dorset coast. There, FM reception is not too good, and Radio 2 was 'difficult', except in just one spot in the room, which was standing the radio on top of the heating radiator!

The whole problem was solved by locally buying a ribbon dipole FM antenna for about £2, hanging it up in the room, and coupling it as shown in Fig. 3. The result was good reception on all BBC FM channels, plus a number of continental stations.

The 300-Ω ribbon feedline from the antenna is terminated with a 4-turn coupling loop, made from PVC-covered hook-up wire wound around the retracted telescopic antenna. The coupling can be optimized by sliding the coil, right or left, along the telescopic rod. It must be stressed that this was only a temporary expedient which 'worked', and might prove a project for someone to look into more seriously. A similar arrangement should be possible using co-axial feedline from an antenna.

The medium-wave (MW) band

As many of you will know, it is possible, with a good receiver and a good antenna, to

receive trans-Atlantic stations on MW during the night time hours. Many enthusiastic listeners concentrate on MW DX. Using the Matsui, with in-built ferrite rod antenna, North American stations have been heard on the West Dorset coast, but can not be heard at the author's home address in Berkshire, about 100 miles (165 km away). However, with the antenna arrangement shown in Figs. 4 and 5, quite dramatic DX results can be achieved, and also remove much of the night time MW interference and noise.

The antenna consists of a large tuned loop coupled inductively to the in-built rod antenna in the portable. The relative polar diagrams of the large loop and the ferrite loop are shown in Fig. 4. While the maximum signal received is on the end of the large loop, it is on the long side in the case of the ferrite rod.

The large outer loop consists of seven closewound turns of wire on a 75 cm (29.5") former, which has to be fabricated. The wire type is PVC-covered single-strand with a 0.6-mm conductor, and an outside diameter of 1.2 mm. This loop, tuned with a 500-pF variable capacitor (see Fig. 4b), covers the frequency range from 510 kHz to 1600 kHz on the prototype, i.e., 588 m to 180 m.

The 75-cm diameter loop former was made from two standard white Formica strips each 20x5 cm (8x2") wide. One Formica strip, with the white surface facing outward, was bent round, and the ends glued together with a 7.5-cm (3") overlap, producing a rather flexible circle. EvoStik wood adhesive was used. The second Formica strip was glued around the inside of the circle, with the white surface facing inward, producing a laminate, 75-cm diameter, 5-cm wide (29.5x2"), white loop former.

The seven wire turns were closewound at the centre (see Fig. 5) of the large loop former. The whole was mounted on a plastic box containing the 500-pF variable capacitor mounted on one side.

The box should be about 23x24x9 cm

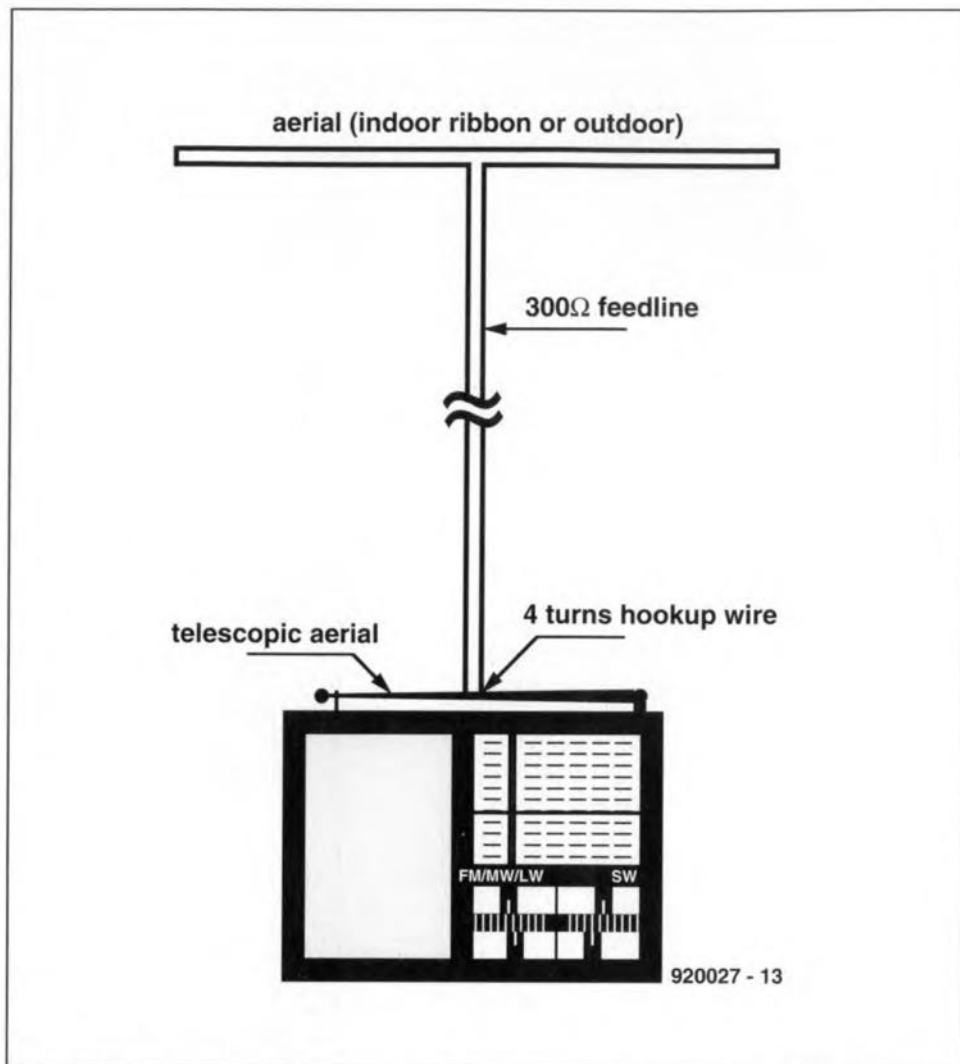


Fig. 3. The VHF antenna is made from an inexpensive indoor ribbon dipole, and is coupled inductively to the radio.

(wxdxh) (9x9.5x3.5"), and can be plastic or made of wood. On the prototype, a Peak Freans coloured plastic biscuit box was used, with the labels removed. Being made of semi-rigid plastic, it was reinforced inside with a couple of wood strips (see below). The sides are cut away just enough to accom-

modate the loop so that the box lid can be snapped on (see Fig. 5; side view). The variable capacitor is mounted in the centre of one side. Adhesive and wood struts were used to hold the loop firmly in position.

Operation is quite simple. The portable radio is stood on the box top as shown in

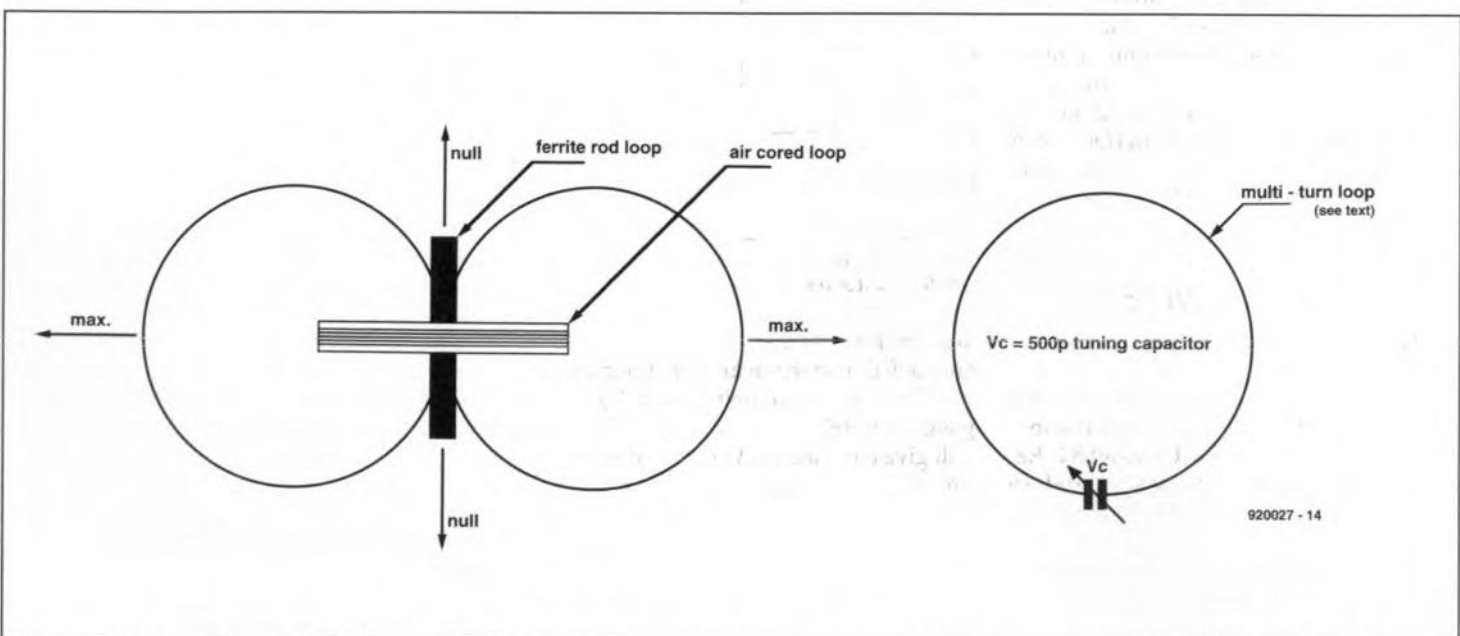


Fig. 4. Polar diagram (left) and circuit of the medium-wave antenna. Note that this works in conjunction with the ferrite loop in the radio.

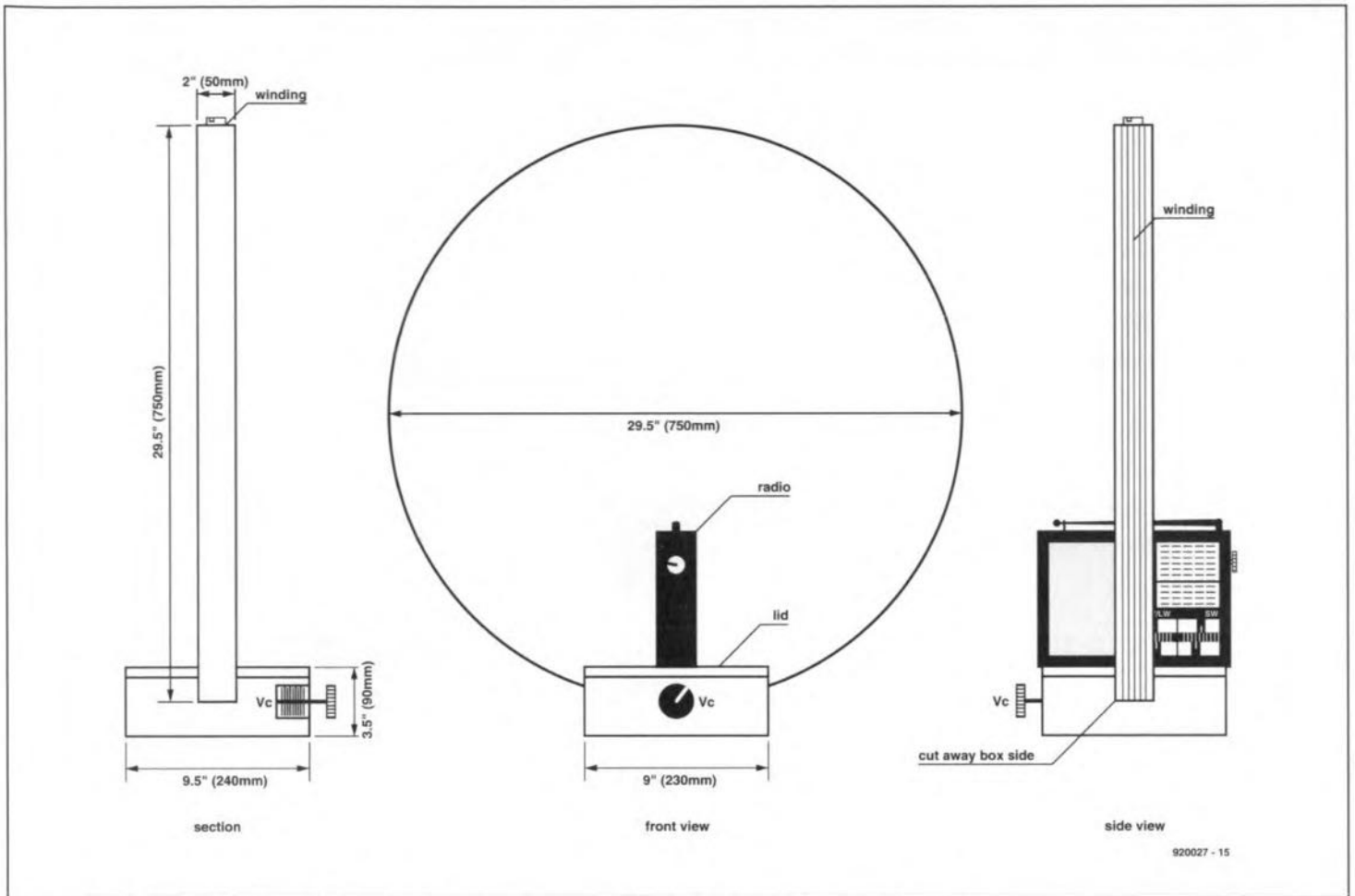


Fig. 5. Construction of the MW (or LW) antenna: section (left); front view (centre); side view (right).

Fig. 5, and a weak station selected. Bringing the loop variable capacitor to resonance results in a very substantial increase in signal strength, and reduction in interference. Note that on strong local stations the receiver may be overloaded.

In use, the whole assembly should be rotated for maximum signal and minimum adjacent channel interference, which is prevalent on the MW band at night. A simple turntable is useful for this operation.

In practice it has been found that the loop antenna enables both the Grundig and the Matsui portable to receive DX stations with adequate signal strength and minimum interference. Previously, these DX stations, including trans-Atlantic ones, could not be heard, or were obliterated by interference. A simple idea to enable the DX enthusiast to use an average MW portable for DX reception!

The long-wave (LW) or MW/LW bands

A long-wave version of the 75-cm diameter loop can be made by increasing the number of turns on the MW version by about 3.7. Reducing the number of turns would produce a loop to listen to the spectrum between the MW and LW bands.

A suggested two-band (MW/LW) version is shown in Fig. 6. It uses a MW winding, as previously, and alongside it a separate LW winding consisting of 3.7 times

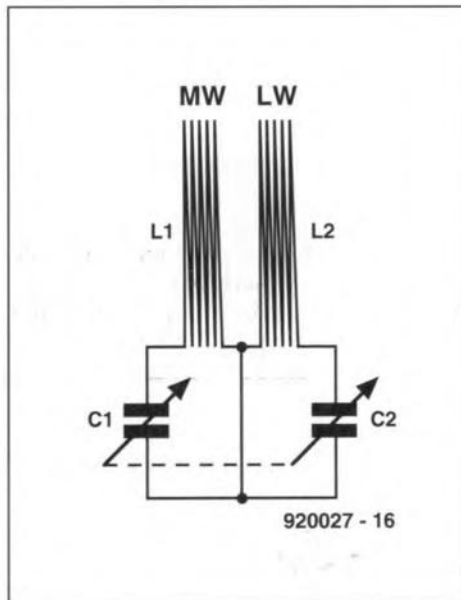


Fig. 6. Suggested tuning circuit for the MW/LW antenna.

the number of turns of 32SWG (0.3 mm dia.) enamelled copper wire (or thinner). The windings are separately tuned by a two-gang 500+500 pF variable capacitor. This will give combined MW/LW reception on a portable radio without antenna switching.

Conclusion

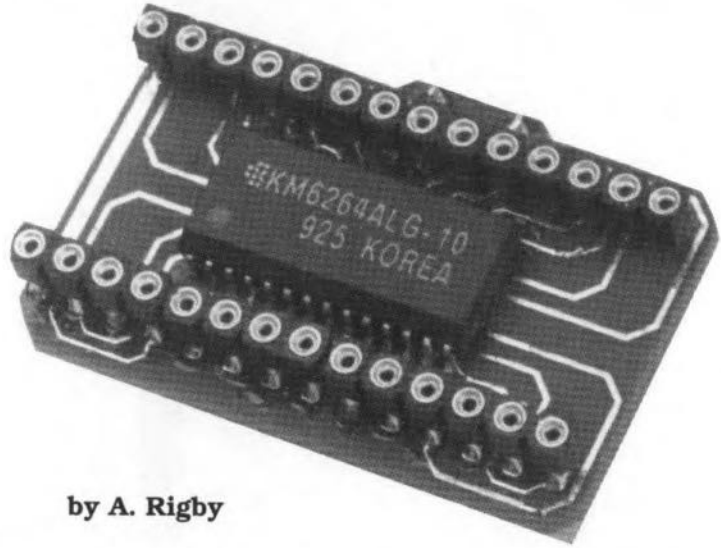
The above is an amalgam of separate experiments carried out over the last few years. There is no doubt room for further experi-

ments by enthusiasts, especially in the case of the VHF FM band, where a quick *ad hoc* arrangement was found to work, and left at that. The MW (and LW) loop could be improved by anyone who can dream up a loop former 50% or 100% larger in diameter, and, of course, have somewhere to house it when it is finished, such as an attic.

Happy listening!

RAM EXTENSION FOR MINI Z80 SYSTEM

The Z80 card described last month will be perfectly suitable for many applications. There are cases, however, in which RAM is a must, and that is why we now present a simple extension circuit for 8, 16 or 24 Kbyte of static RAM.



by A. Rigby

CHALLENGING, this Z80 programming without RAM, but the amount of data that may have to be stored at one time during the execution of a machine code program soon exceeds the capacity of the registers contained in the Z80. In that case, RAM (ran-

dom access memory) must be used, although it is possible to go a long way without it.

Some applications, for instance, a data logger, simply can not work without RAM because it is required to store amounts of data that can not possibly be 'crammed' into

the registers, even if the Z80 sports so many.

RAM in an EPROM socket

The RAM extension developed for the mini Z80 system (Ref. 1) is suitable for other

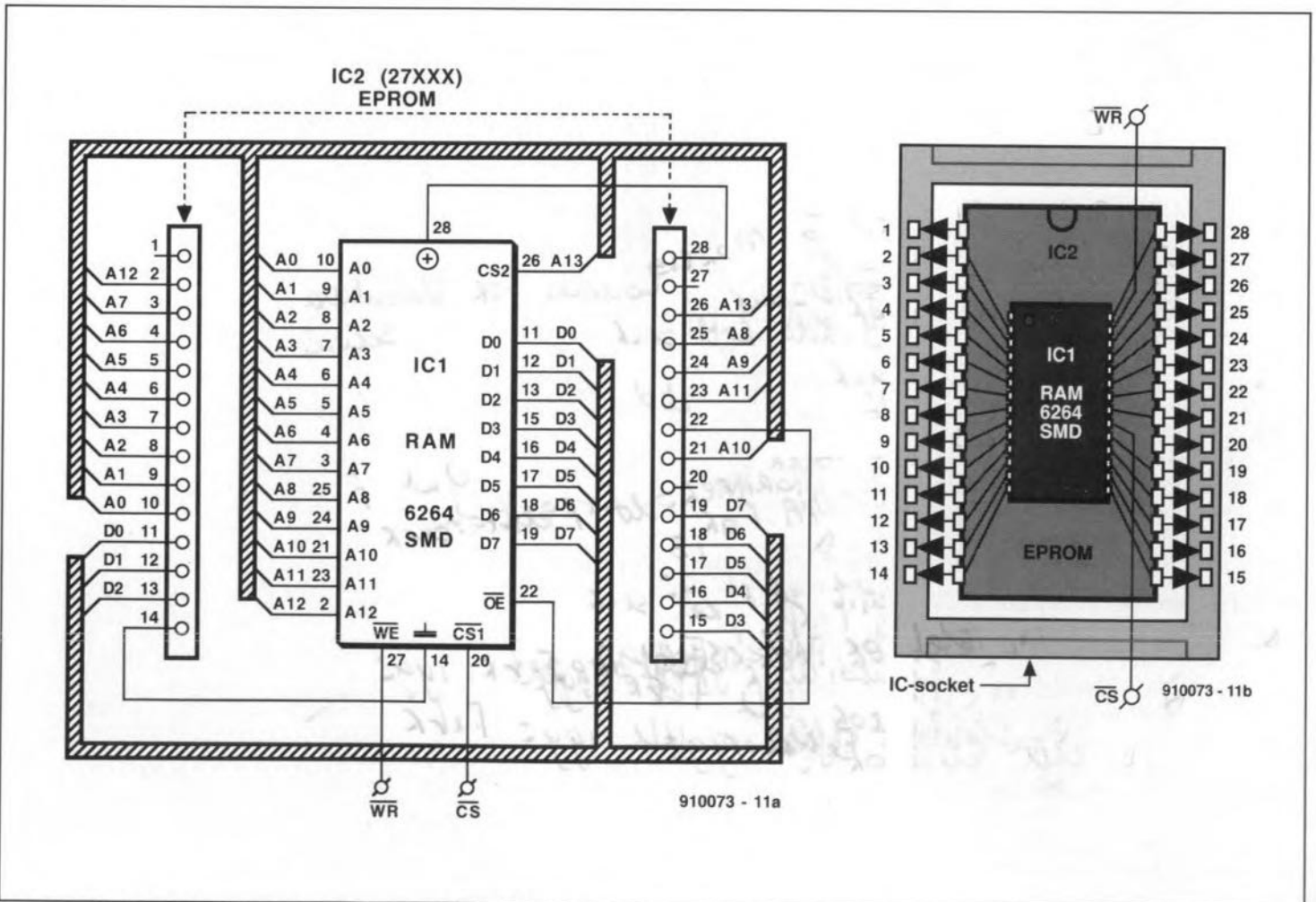


Fig. 1. Circuit diagram and wiring diagram of the RAM card.

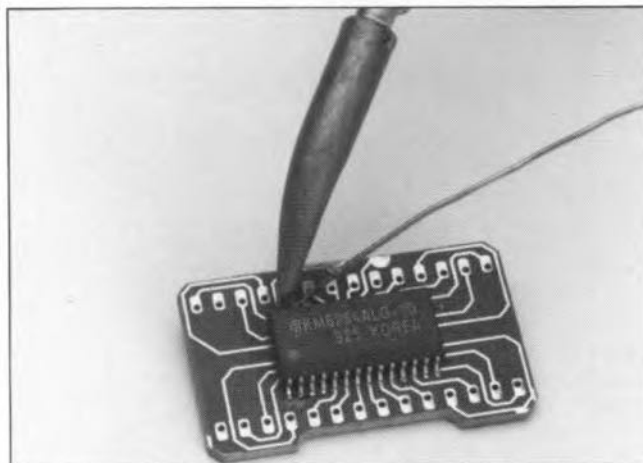
microprocessor systems also. The RAM unit is hardly larger than a 27128 EPROM, and has the same pinning. It is simply plugged into the socket provided for the EPROM on the mini Z80 board. To reduce the size of the computer even further, it is possible to do without the socket altogether, and fit the RAM extension direct on to the Z80 board.

Figure 1 shows the circuit diagram of the RAM extension, as well as a kind of wiring diagram that indicates (approximately) what the connections look like in practice. The PCB accommodates two ICs: a RAM and an EPROM. The socket for the EPROM is very important in this plug-on assembly because it functions as the connector between the Z80 board and the RAM board. The EPROM originally fitted on the Z80 board is removed and relocated to the 'first floor', i.e., on the RAM extension board. All 28-pin EPROMs in the 27xxx series can be used, provided they are electrically suitable, and 'fit' in the available memory space.

Unfortunately, two signals are required by the RAM that are not available on the EPROM socket on the Z80 board. These signals are WR (write) and CS (chip select). The first is never required by an EPROM, and the



The terminals of the RAM IC must be bent as close as possible to the enclosure to enable the chip to be fitted in between the two pin rows.



The RAM is secured by first soldering two diagonally opposite pins to the copper spots. Next, align the remaining pins with the copper spots underneath, and solder carefully using thin solder wire and a low-power solder bit with a small tip.

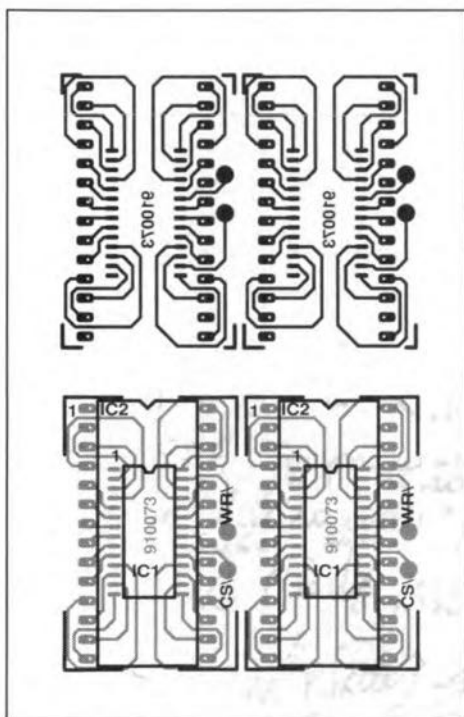
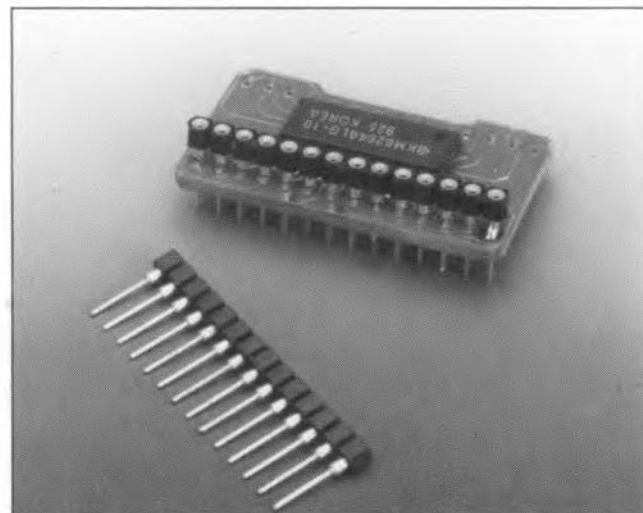
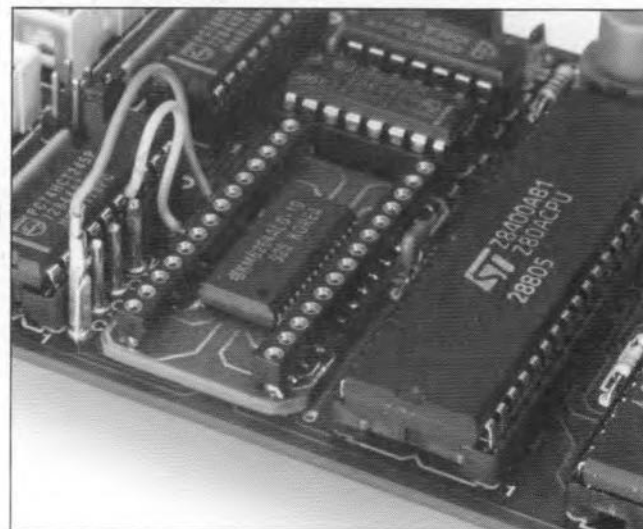


Fig. 2. Track layout (mirror image) of the printed circuit board. Note that the components are fitted at the track side.



The IC socket/connector is made from two 14-way strips with extra long pins (approx. 1 cm).



The RAM extension may be inserted into an IC socket, or fitted direct on to a computer board. If 8 KBytes are not sufficient, further modules are simply stacked.

COMPONENTS LIST

Semiconductors:

- 1 6264 RAM (SMA) IC1
- 1 2764/27128 EPROM (see text)

Miscellaneous:

- 2 14-way pin strip with 1-cm long pins K1;K2
- 1 Printed circuit board (for two RAM modules) 910073

second can not be used simply because it selects the EPROM. Fortunately, the connection for the \overline{CS} signal required for the RAM already exists on the Z80 board.

The circuit diagram indicates that the second chip select signal of the RAM, CS2, is connected to address line A13. This means that the RAM is selected at all addresses with A13 at '1', and has the advantage of avoiding 'image' areas where the RAM contents are duplicated. Remember, the \overline{CS} signal available on the Z80 card selects blocks of 16 KBytes, and, when used on its own, would cause the RAM contents to appear double in the memory addressed by the Z80. If you use the RAM extension in another system, and you are bothered by the fixed link between A13 and the CS input, simply break this connection between the two ICs, and connect pin 26 of IC1 to pin 28 (+5 V). If necessary, this modification can be carried out on a completed RAM unit also.

Construction

The RAM is built on a single-sided printed circuit board shown in Fig. 2. The PCB is supplied double, so you have to cut it in two.

First, make the PCB as narrow as you can. To keep the Z80 computer as compact as possible, reduce the size of the RAM board such that there is virtually no PCB material outside the outer copper tracks. The board may also be made narrower between pins 4 and 11 with the aid of a jig-saw. At the other side, it may be made narrower locally, leaving the spots for \overline{CS} and \overline{WR} to protrude.

The components are mounted on to the board as illustrated in Fig. 3. The RAM module so made may be mounted on to any circuit board as if it were a 28-pin memory IC. If it is soldered, it can be fitted flush on to the PCB surface because there are no copper tracks at the underside.

When the RAM extension is to be plugged into a socket, pins are used which are slightly longer than IC socket pins. These pins are supplied on 36-way or 72-way strips, and are inserted into the respective holes from the component side. Solder at the component side.

When it is intended to fit the RAM board direct on to a motherboard, the longer pins are not required, and push-through pins with a standard length may be used. The length of the pins should enable soldering at the component side of the RAM board, and make the pins protrude 2-3 mm at the underside of the board for soldering to the motherboard. Figure 3 shows a RAM card made in this way.

RAM cards may be stacked via the EPROM socket. The system EPROM, if used, is always fitted in the EPROM socket on the RAM board at the top of the stack.

Mini Z80 card with RAM

The connections for the \overline{WR} and \overline{CS} signals are available on the Z80 board. There are, in fact, three chip select signals, so that a maximum of three RAM cards may be connected.

```

;*****
;   SIMPLE RAM TEST PROGRAM FOR MINI Z80 BOARD
;*****
;   IC5 and IC6 jumpers set for input only
;   Be careful: a write can be destructive
;*****

output1: equ 0           ;address for K1
output2: equ 1           ;address for K2
input1:  equ 2           ;address for K3
input2:  equ 3           ;address for K4

                org 00h

;*****
;   NON DESTRUCTIVE RAMTEST
;   CHECKS THE ADDRESS AREAS 0E000H...0FFFFH,
;   0A000H...0BFFFH AND 06000H...07FFFH FOR RAM
;*****
ramcheck: ld  hl,0e000h   ;first RAM address of CS3\
           ld  de,02000h   ;size of one RAM module
           ld  c,2
next$:    ld  a,(hl)      ;load contents of RAM address
           ld  b,(hl)
           cpl a          ;complement the contents
           ld  (hl),a     ;write to RAM
           ld  a,(hl)     ;reload contents
           and b          ;compare with old value
           ld  (hl),b     ;restore contents
           jr  nz,noram$  ;if not zero then no RAM
                           ;location
                           ;increment address counter
                           ;decrement counter
           inc hl
           dec e
           jr  nz,next$
           dec d
           jr  nz,next$
           jr  endram$
noram$:   ld  hl,0a000h   ;first RAM address of CS2\
           ld  de,02000h   ;size of one RAM module
           dec c
           cp  1
           jr  z,next$
           ld  hl,06000h   ;first RAM address of CS1\
           cp  0
           jr  z,next$
           jr  ramcheck

endram$:  dec hl
           ld  sp,hl      ;load stack pointer with
                           ;highest RAM location
again$:  call begin      ;initialization
           call test     ;copy input inverted to output
           jr  again$

;*****
;   OUTPUT INITIALIZATION ROUTINE
;*****
begin:    ld  a,0
           out (output1),a
           out (output2),a
           ret

;*****
;   READ INPUT AND OUTPUT INVERTED
;*****
test:    in  a,(input3)
           cpl a
           out (output1),a
           in  a,(input4)
           cpl a
           out (output2),a
           ret

```

910073-12

Fig. 4. This program for the mini Z80 system can be used to test the freshly installed RAM extension.

Although this gives a total of 24 Kbytes of RAM, 8 Kbytes will be sufficient in most cases. The address ranges available for the RAM modules are:

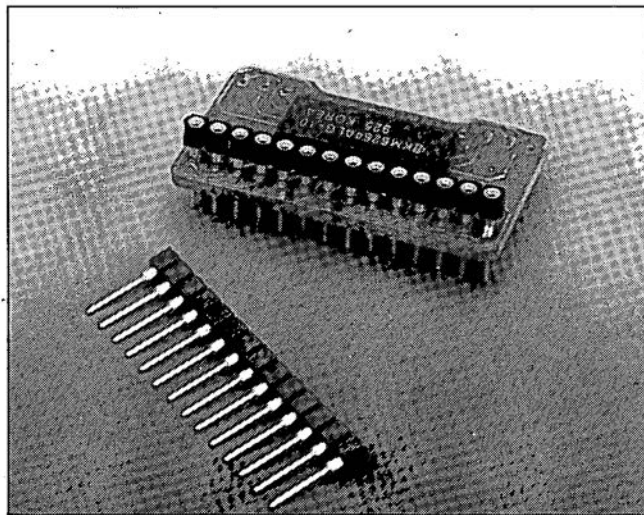
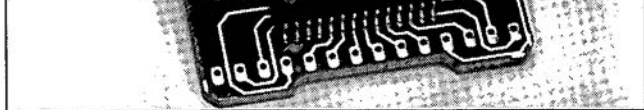
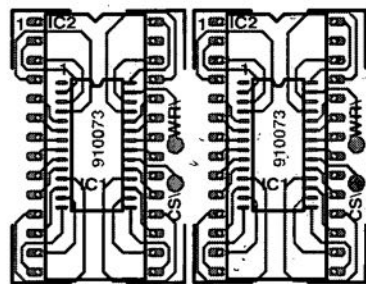
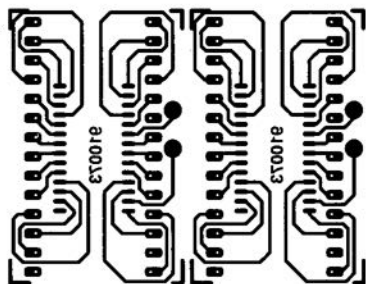
- CS1: 6000_H to 6FFF_H
- CS2: A000_H to BFFF_H
- CS3: E000_H to FFFF_H

The presence of RAM allows all Z80 instruc-

tions to be used, i.e., it does away with the restrictions listed in Table 2 in Ref. 1. To help you get started, Fig. 4 shows a little program that tests the RAM, and makes the stack pointer point to the highest RAM address. ■

Reference:

1. "Mini Z80 system", *Elektor Electronics* January 1992.



per spots underneath, and solder carefully using thin solder wire and a low-power solder bit with a small tip.

The IC socket/connector is made from two 14-way strips with extra long pins (approx. 1 cm).



Fig. 2. Track layout (mirror image) of the printed circuit board. Note that the components are fitted at the track side.

MEASUREMENT AMPLIFIER

by J. Ruffell

Measurement amplifiers are used to magnify the output of all kinds of sensor so that the signal can be processed as appropriate in, among others, a test instrument, a computer, or a control loop.

THE measurement amplifier has been designed as a stand-alone unit to make it universally usable. It can, therefore, be operated manually or via a computer.

It is seen from the block diagram of the amplifier in Fig. 1 that, apart from the power supply, it is designed as a twin unit: the two sections (channels) are absolutely identical. If the amplifier is to feed an analogue-to-digital (A-D) converter, it is important that the level of the input signal to the amplifier matches the input level of the converter. That means that large signals must be attenuated, and small ones be amplified, in a frequency-independent manner. The passive attenuator at the input of the amplifier enables input signals between ± 200 mV and ± 200 V to be processed. The subsequent amplifier stages ensure that the output level is suitable for a variety of converters and test instruments.

Furthermore, it is often necessary that the output signal has an appropriate offset, that is, is superimposed on a certain direct voltage, since several instruments and A-D converters can handle only positive inputs. To that end, the amplifier has provision for an offset voltage that can be varied between -5 V and $+5$ V.

Moreover, the amplifier must also be able to function as an impedance converter, since most sensors have a high-impedance output, whereas the amplifier output must have a low impedance to enable it to be linked to a wide variety of instruments.

Finally, the amplifier must be able to deliver relatively high output currents to make it suitable for driving the majority of test instruments.

The circuit

Since the two amplifier channels are identical, the following description will refer to channel A only.

The attenuator and associated frequency compensating network consist of four series-connected impedances that have identical time constants. Provided the variable capacitors have been adjusted correctly, the division ratio is constant over the frequency range 0–300 kHz. Since the range selection occurs at the output of the attenuator, the impedance that loads the signal source at the input is the same in all ranges. It consists of a resistance of 1.11 M Ω shunted by a capacitance of 10 pF, and is, therefore, identical to the input impedance of most commercial multimeters and oscilloscopes. Finally, it is important that



the attenuator is terminated into a high impedance to match the FET input of IC₃.

The Type TLC2201 in the IC₃ position can handle rectangular signals with a frequency of up to 125 kHz and sinusoidal ones with a frequency of up to 500 kHz. The use of this opamp makes the amplifier sufficiently stable for a resolution of 12 bits. When that sort of accuracy is not required, an LF356 may be used instead of the TLC2201. That opamp

can handle rectangular signals up to 300 kHz and sinusoidal ones up to 500 kHz, but its temperature behaviour is fairly poor. To achieve a resolution of 12 bits, the ambient temperature must be kept to within ± 5 °C, compared with ± 46 °C in the case of the TLC2201. Another opamp that may be used is the OPA602 which gives a resolution of 10 bits for a temperature change of ± 54 °C and of 12 bits for ± 13 °C.

The input of IC₃ is protected against too

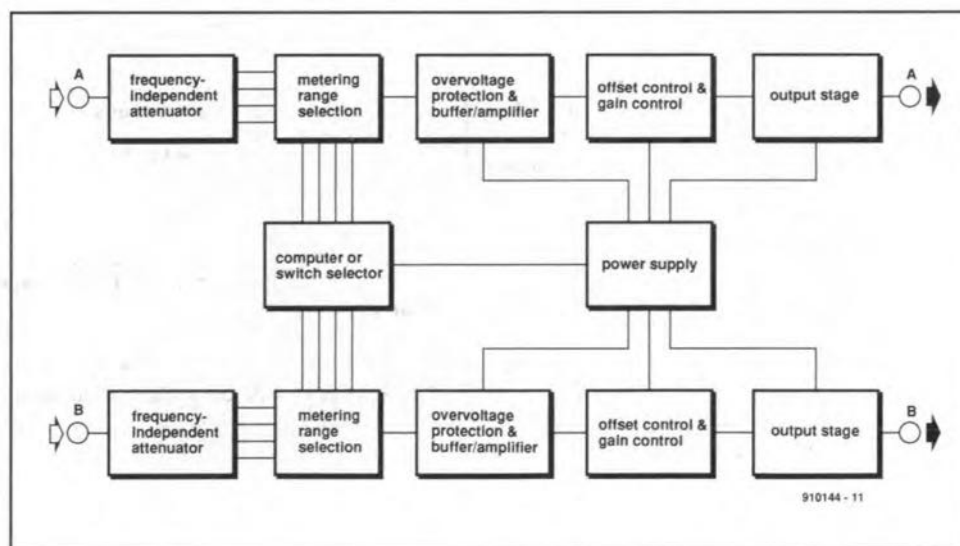


Fig. 1. Block diagram of the measurement amplifier.

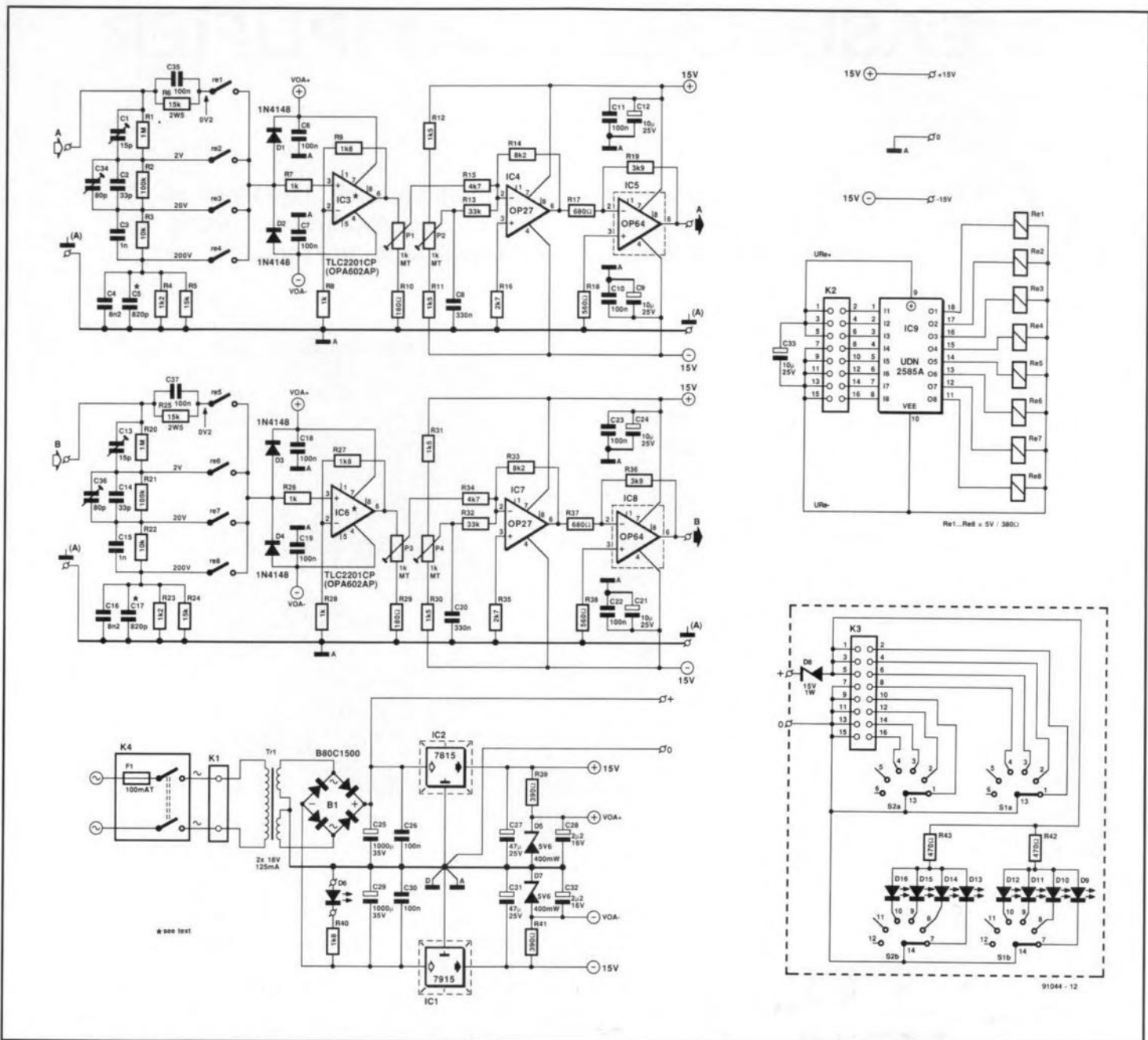


Fig. 2. Circuit diagram of the complete measurement amplifier.

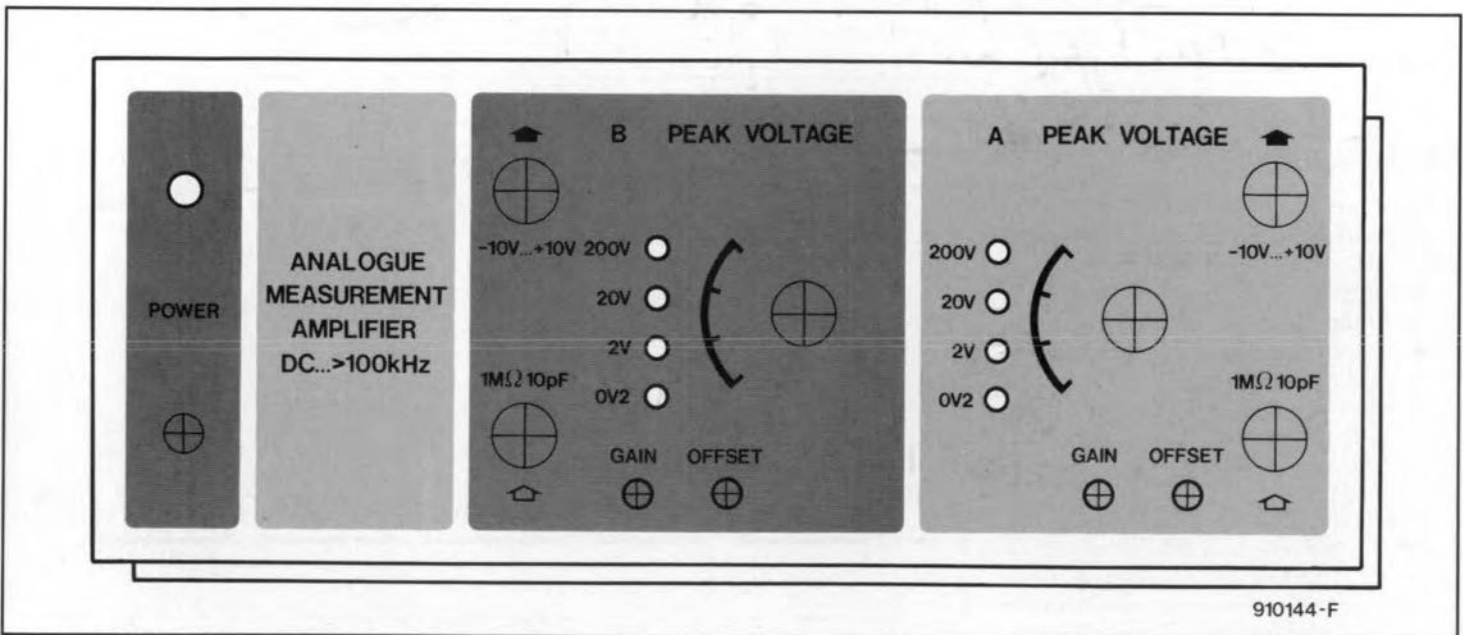


Fig. 3. The front panel foil for the measurement amplifier is available through our Readers' services.

high an input voltage by diodes D₁ and D₂, which short-circuit signals $>\pm 6.2$ V to the supply lines. The resistors in the attenuator (total resistance in the 2 V, 20 V and 200 V ranges >1 M Ω) ensure that the currents through these diodes cannot become too large. In the 200 mV range, the protection is not so good, although even then, R₆ limits the current to a large extent. None the less, some care should be exercised when connecting input signals (start at the 200 V range).

In the next stage, based on IC₄, an offset voltage between -3.75 V and $+3.75$ V, derived

from potential divider R₁₁-R₁₂-P₂, is added to the measurand (signal being measured). The OP27 functions here as an adder with an amplification of 1.4, so that the offset voltage at its output ranges from -5.3 V to $+5.3$ V. The amplification is preset with P₁. The low-noise OP27 is excellent value for money: it is inexpensive, fast, exhibits little drift and has excellent common-mode rejection.

The output stage is based on a Type OP64 opamp, IC₅. Like IC₄, this stage also inverts the signal, so that its output is in phase with the input of the amplifier. The OP64 is de-

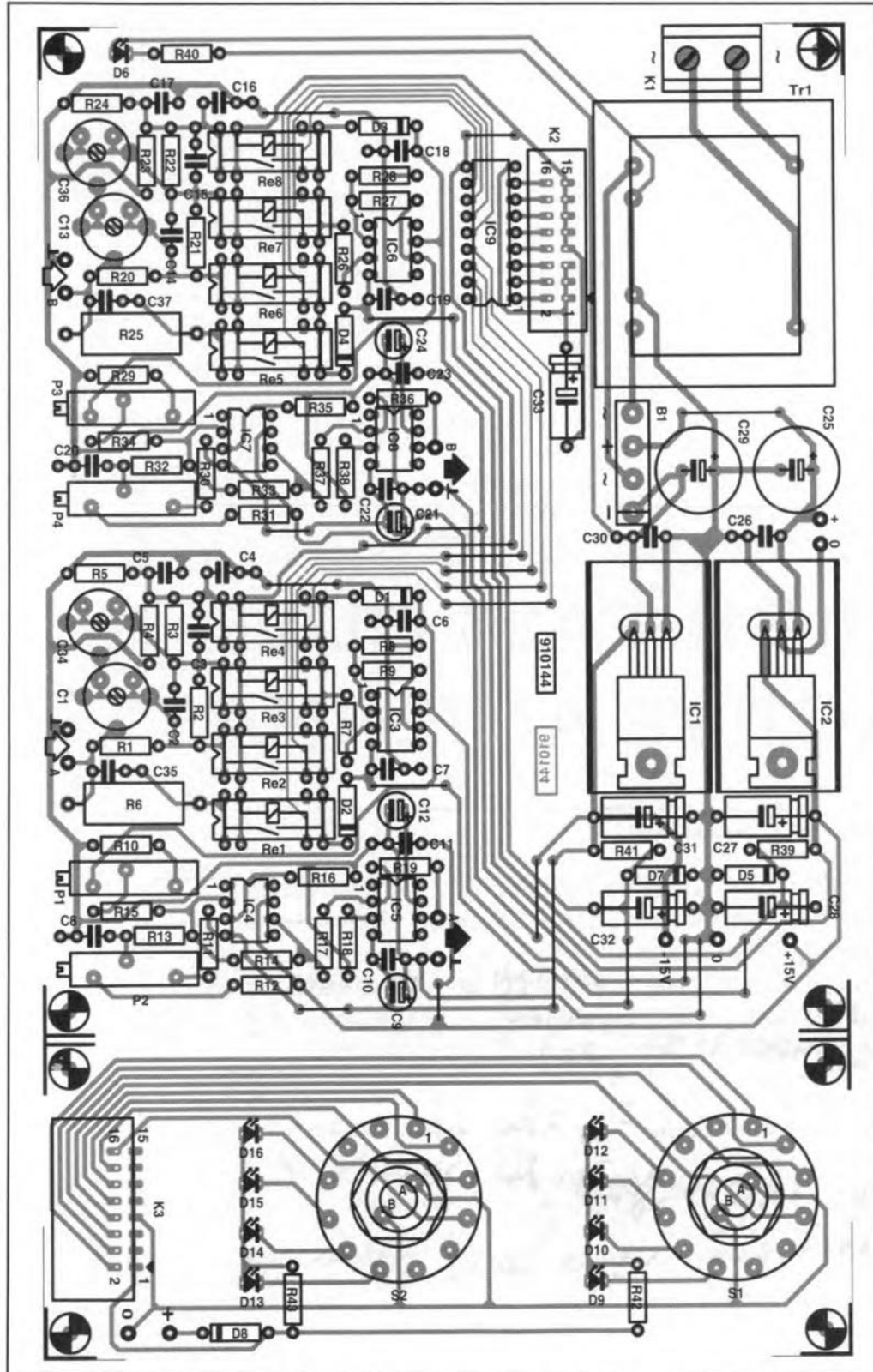


Fig. 4. The printed-circuit board for the measurement amplifier must be cut into two before any construction work is done on it.

PARTS LIST

Resistors:

- R1, R20 = 1 M Ω
- R2, R21 = 100 k Ω
- R3, R22 = 10 k Ω
- R4, R23 = 1.2 k Ω
- R5, R24 = 15 k Ω
- R6, R25 = 15 k Ω , 2.5 W
- R7, R8, R26, R28 = 1 k Ω
- R9, R27, R40 = 1.8 k Ω
- R10, R29 = 180 Ω
- R11, R12, R30, R31 = 1.5 k Ω
- R13, R32 = 33 k Ω
- R14, R33 = 33 k Ω
- R15, R34 = 4.7 k Ω
- R16, R35 = 2.7 k Ω
- R17, R37 = 680 Ω
- R18, R38 = 560 Ω
- R19, R36 = 3.9 k Ω
- R39, R41 = 390 Ω
- R42, R43 = 470 Ω
- P1-P4 = 1 k Ω multiturn preset

Capacitors:

- C1, C13 = 15 pF trimmer
- C2, C14 = 33 pF
- C3, C15 = 1 nF
- C4, C16 = 8.2 nF
- C5, C17 = 820 pF
- C6, C7, C18, C19, C26, C30, C35, C37 = 100 nF
- C8, C20 = 330 nF
- C9, C12, C21, C24 = 10 μ F, 25 V
- C10, C11, C22, C23 = 100 nF ceramic
- C25, C29 = 1000 μ F, 25 V radial
- C27, C31 = 47 μ F, 25 V
- C28, C32 = 2.2 μ F, 16 V
- C33 = 10 μ F, 25 V
- C34, C36 = 80 pF trimmer

Semiconductors:

- D1-D4 = 1N4148
- D5, D7 = 5.6 V, 400 mW zener
- D6 = 5 mm LED, red
- D8 = 15 V, 1 W zener
- D9-D16 = 3 mm LED, green
- B1 = B80C1500
- IC1 = 7915
- IC2 = 7815
- IC3, IC6 = TLC2201CP (see text)
- IC4, IC7 = OP27
- IC5, IC8 = OP64
- IC9 = UDN2585A

Miscellaneous:

- K1 = 2-way PCB terminal block, pitch 7.5 mm
- K2 = 16-way male box header
- K3 = 16-way right-angle male box header
- K4 = mains entry with integral DPST switch and fuse holder
- F1 = fuse, 200 mA, delayed action
- S1, S2 = 2-pole, 6-position rotary switch
- Tr1 = mains transformer, 2 \times 18 V, 125 mA
- Re1-Re8 = DIL relay, 5 V, 380 Ω Case 80 \times 200 \times 180 mm (3 $\frac{1}{8}$ \times 7 $\frac{7}{8}$ \times 7 $\frac{1}{16}$ in)
- 2 heat sinks 83 K W⁻¹
- 2 IC heat sinks 29 K W⁻¹
- PCB 910144
- Front panel foil 910144-F

signed for driving low-impedance (down to 150 Ω) loads; its (continuous) output current can be as high as 80 mA. It is not truly proof against short circuits, but, according to the data sheet, it can withstand short circuits for up to 10 seconds. A drawback is its dissipation, which makes a heat sink essential. Also, its amplification should not be set below 5 to prevent instability and a tendency to oscillation.

The measuring ranges are selected by relays Re₁–Re₄. These are energized by IC₉, whose control inputs are TTL compatible and can thus be driven by a computer. The relays make the amplifier slightly more expensive, but their use provides electrical insulation between analogue and digital circuits. Integrated switches would have been less expensive, but these suffer from too narrow a voltage range, a high on-resistance and imperfect channel separation.

To operate the relays manually, two two-pole, four-position rotary switches are needed. One section of the switches actuates the range indicator LEDs, and the other is connected to the control inputs of IC₉.

Note that the interface to enable the amplifier to be controlled by a computer is still in development: it is hoped to publish its design within a couple of months.

The power supply is a conventional design providing ± 15 V and, with the aid of zener diodes D₅ and D₇, ± 5.6 V for the TLC2201s. Note also that the supply voltage for IC₉ is taken direct from the rectifier, that is, before

the regulator; it is stabilized by zener diode D₈. This arrangement ensures that any noise pulses caused by the change-over of relay contacts are not transmitted to the amplifier.

Construction

If the amplifier is constructed on the PCB shown in Fig. 4, no real difficulties should be encountered. Note, however, that before any construction is begun, the board must be cut into two. One section is for the rotary switches and LEDs, and the other for the power supply, attenuators and amplifiers.

The smaller section must be mounted directly behind the front panel. Since the rotary switches have six positions (four-position ones are difficult to obtain), their travel must be blocked at position 4 with the aid of the small washer provided.

Owing to its heat dissipation, D₈ should be mounted a few millimetres above the board. Circuits IC₅ and IC₈ must be fitted on heat sinks.

The two boards are interlinked by a short length of flatcable.

The input and output sockets on the front panel should be good-quality BNC types. Any connections to and from them should be in screened cable.

The completed amplifier should be installed in a suitable metal case: a proposed front panel layout is shown in Fig. 3 (the foil for this is available through our Readers' services). The mains entry, preferably with in-

tegrated fuse holder, should be fitted at the back of the housing.

Calibration

Do not compensate the offset of the individual opamps, since the direct voltage component of the output signal is a sum set by P₂ (P₄). This preset needs to be readjusted only when the amplification has been changed with P₁ (P₃).

Before commencing the calibration, let the amplifier warm up for about five minutes.

Connect a function generator to the input and the probe of an oscilloscope to the junction D₁–D₂ (D₃–D₄). Select the 200 V range and set C₁ (C₁₃) and C₃₄ (C₃₆) to the centre of their travel. Inject a rectangular, 1 kHz signal at a level of 10 V. Although the value of C₅ (C₁₇) is shown as 820 pF, its correct value, owing to the tolerance of the other circuit elements, may lie between 200 pF and 2.2 nF. The signal on the screen of the oscilloscope should be rectangular: if it has overshoot or rounded corners, the value of C₅ (C₁₇) must be lowered or increased respectively. Once the correct value of these capacitors has been found (and corresponding capacitors fitted), reduce the output of the function generator to 1 V, and adjust C₁ (C₁₃) with the 20 V range selected till the trace on the oscilloscope is rectangular again. Finally, adjust C₃₄ (C₃₆) with the 2 V range selected to reobtain a rectangular signal on the oscilloscope. ■

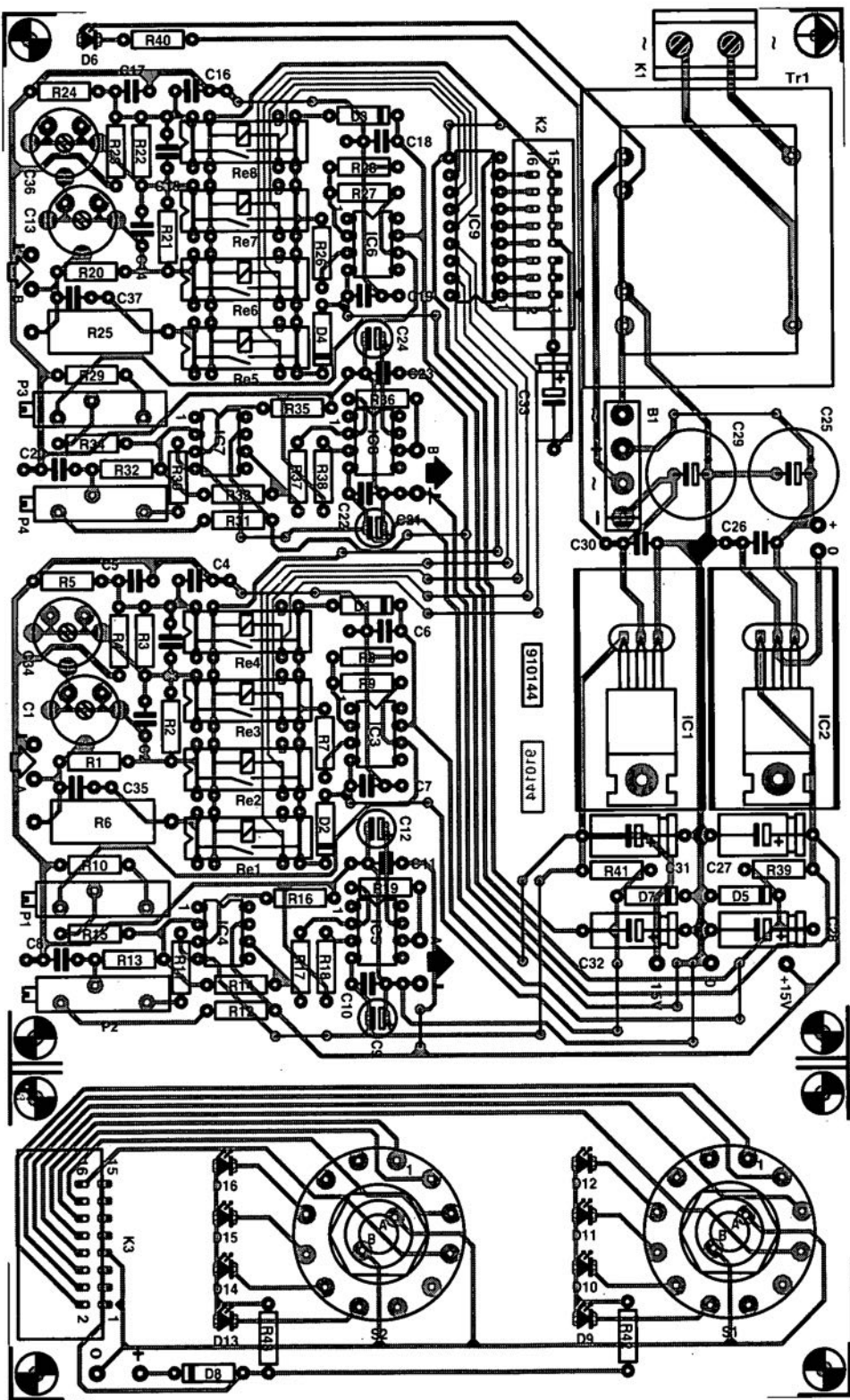


Fig. 4. The printed-circuit board for the measurement amplifier must be cut into two before any construction work is done on it.

CORRECTIONS & TIPS

Low-frequency counter

(January 1992, p. 44)

The parts list on p. 44 erroneously states that capacitors C_{11} and C_{12} are tantalum types. Since the polarity of the voltage across these capacitors may be inverted, the capacitors should be bipolar aluminium types.

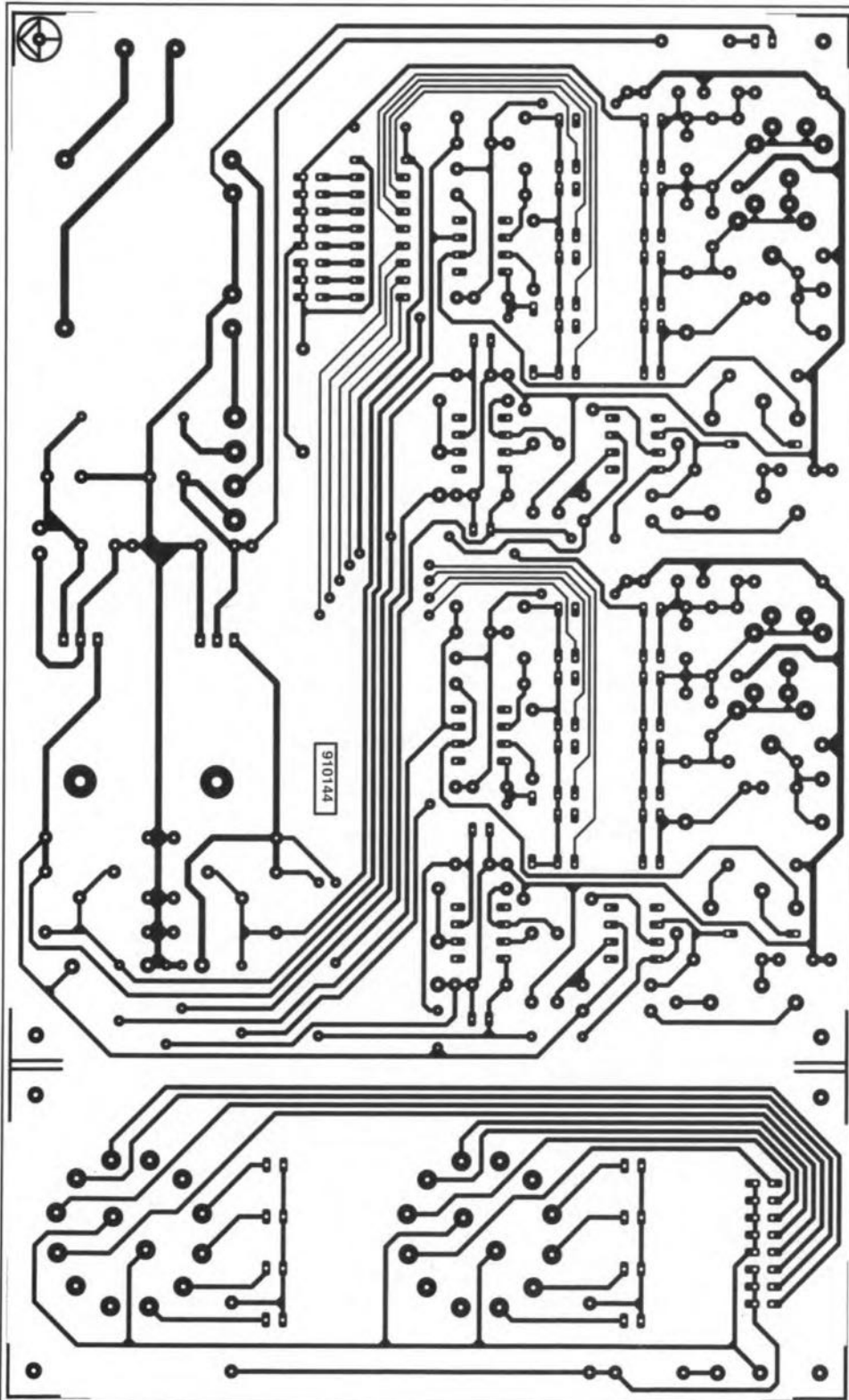
Under 'Construction' on p.45, it is stated in the penultimate paragraph that 'the connection between the input socket and C_{10} must be single screened cable'. In fact, the connection is so short that screened cable is not

necessary.

Measurement amplifier

(February 1992)

Owing to a misunderstanding, the track side of the printed circuit board (p.23) was not included with the article in our February 1992 issue. Our apologies for this oversight. The missing drawing is shown below.



Automatic cycle lights

(July/August 1991, p. 49)

Sir—In the construction of 'Automatic cycle lights', I have encountered three problems.

1. Triggering of IC_{1b} at input $-T$ (pin 11) and R (pin 13). A trailing edge at $-T$ triggers the IC if R is high. It is, however, possible that R is still low or is just changing state. A (not very elegant) solution to this is to connect the line from \bar{Q} (pin 7) to R to junction R_2-D_1 (+ battery) via a $1\text{ k}\Omega$ resistor.
2. A short pulse caused by the switching on of the battery triggered input $+T$ of IC_{1b} (pin 4), which switched the battery off again. This was cured by connecting a 470 nF capacitor between $+T$ and earth.
3. Triggering at $+T$ of IC_{1b} was so sensitive that even a tiny movement of the bicycle causes the battery to be switched off. In other words, if you don't hold the bicycle absolutely still, its lights will flash on and off. The sensitivity can be made variable by replacing resistor R_4 by a $100\text{ k}\Omega$ preset

Helge Bergmann, Hannover

Mini square-wave generator

(February 1992, p. 60)

Sir—On page 61 of 'Mini square-wave generator', you refer to the 7805 regulator as a "low-drop regulator, which ensures low power dissipation". I would disagree with that description: in my books that regulator is definitely not a low-drop type.

P. Thompson, Bristol

You are right: we apologize for that error.
Editor

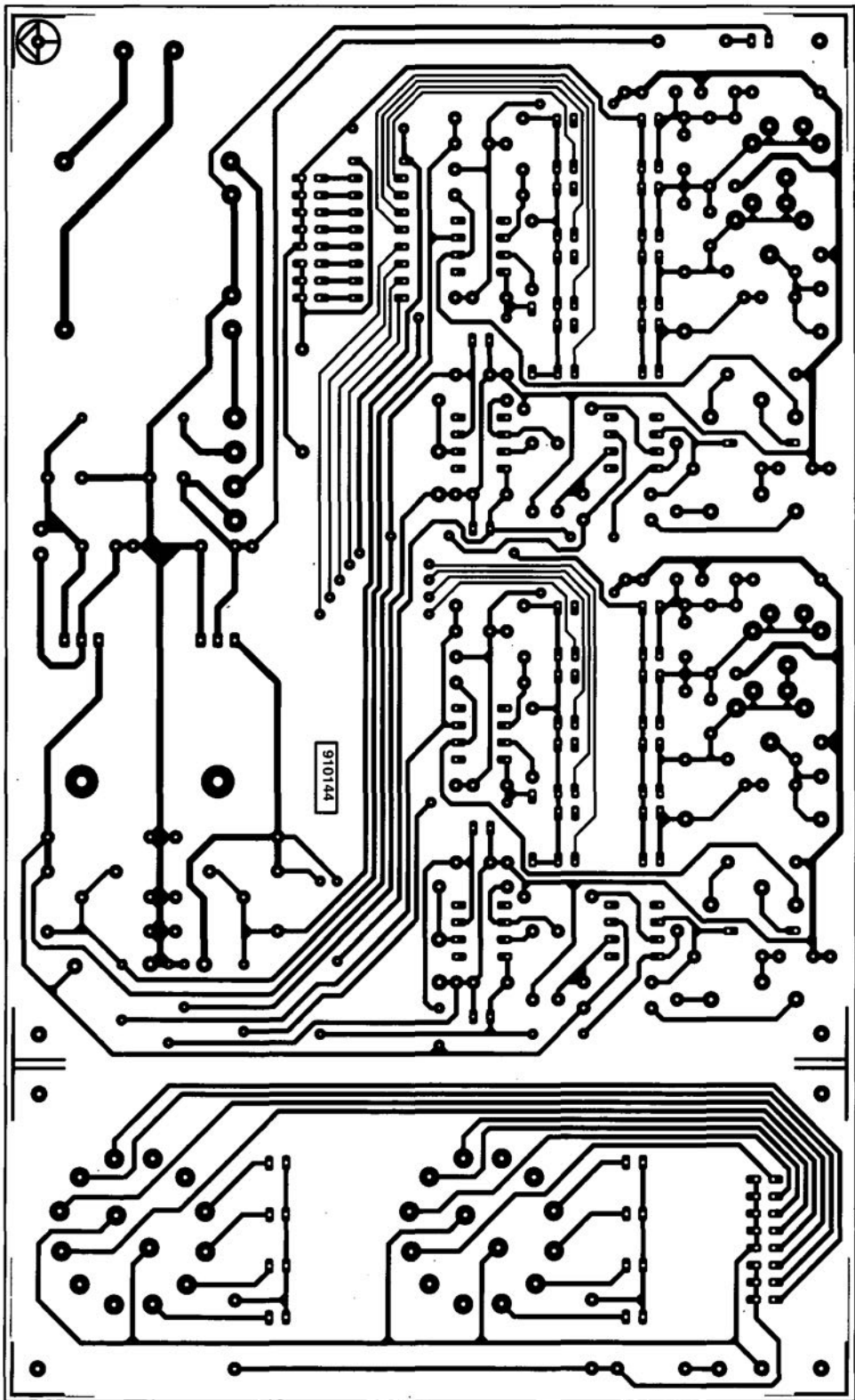
Build a compact-disc player

(January 1992, p. 36)

Sir—I think that your contributor, T. Giffard, in his article 'Build a compact-disc player' must be a lot more wealthy than I am if he considers £249 "an affordable price" (available from only one advertiser in your January issue!). Especially as Philips' own personal compact disc player complete and ready made costs only £149 (from Argos).

J. Easton, Watchet, Somerset

The price mentioned in the advertisement is HFL (Dutch guilders) 249 (equivalent to about £80.00).
Editor



910144

RC HIGH-PASS FILTER FOR ACTIVE ANTENNAS

From an idea by J. Becker

Powerful transmitters operating on medium and long waves may cause considerable interference in the reception of short-wave signals. A high-pass filter to suppress the interfering signals is, of course, the solution. Since, however, the active antennas to which such a filter must be added are normally out in the open, it is desirable to be able to control the filter remotely.

POWERFUL medium-wave or long-wave transmitters operating not far from an active short-wave antenna can cause serious interference in the reception of short-wave signals. The interfering signals, which are generally harmonics and sum frequencies of the transmitter signal, can be suppressed effectively by a critically damped high-pass filter at the input of the active antenna described in Ref. 1.

The transfer characteristic of the active antenna with high-pass filter at its input is given in Fig. 1. Curve 1 shows that signals

at frequencies above 4 MHz are passed unhindered, whereas those in the medium-wave and long-wave bands are suppressed effectively (>35 dB). When interference is caused by a short-wave transmitter, a filter with a characteristic like curve 2 can give some relief from the interference from about 14 MHz; signals below roughly 7 MHz are suppressed effectively.

The circuit diagram of the filter is given in Fig. 2. The filter proper is formed by C_1 , L_1 and damping resistor R_1 . The remainder of the circuit serves to operate relay Re_1 ,

which switches the filter on and off.

Terminals **a** and **c** are connected, respectively, to the anode and cathode of diode D_2 on the active antenna. Capacitor C_1 is then connected directly to the gate of T_1 on the active antenna.

The remote control circuit is given in Fig. 3. This is a modified section of the power supply of the active antenna—see Fig. 3 in Ref. 1, which is installed near the input of the receiver. The two push-button switches enable the filter to be switched on and off without the need of an additional cable.

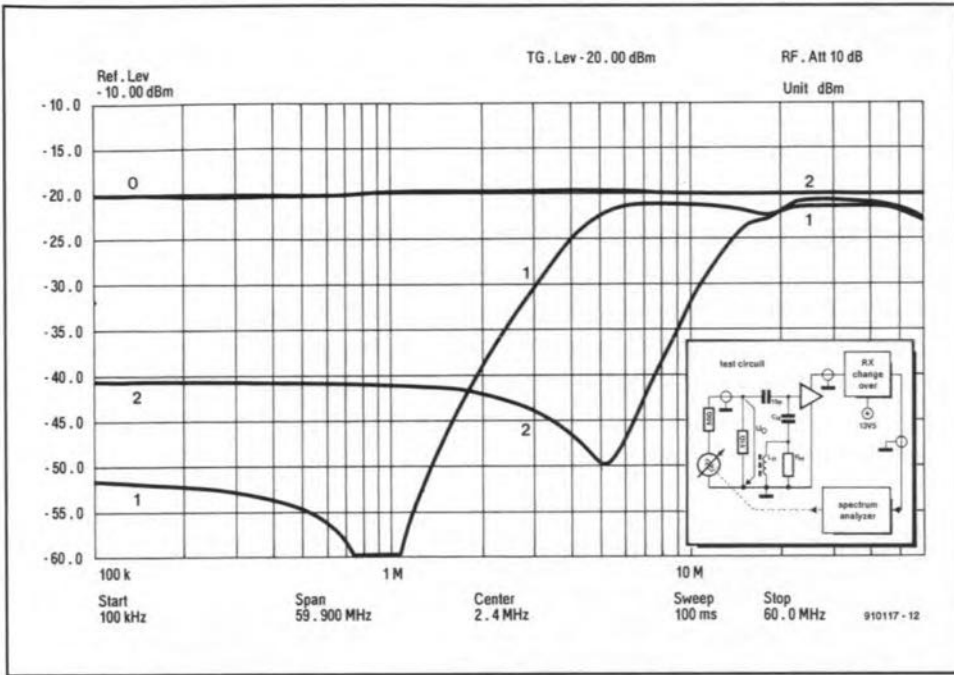


Fig. 1. The output signal of the antenna amplifier with the high-pass filter switched on. Characteristic 1 is correct for the suppression of interfering signals in the medium-wave and long-wave bands. Characteristic 2 is necessary when interfering signals up to 7 MHz must be suppressed. Either of the two curves can be obtained by the correct choice of certain components as indicated in the components list.

Normally, when the supply is switched on, the filter is not actuated since both T_1 and T_2 are cut off. When S_1 is pressed, the supply voltage is increased briefly (because resistor R_w in Fig. 3 is short-circuited), so that a current flows into the base of T_2 via D_1 . The relay is then energized and the filter is actuated. As soon as T_2 is switched on, T_1 conducts also and from then on maintains the base current to T_2 . That means that even when S_1 is released, T_2 , and thus the filter, remains on. When then S_2 is pressed, the supply voltage is interrupted, the relay is deenergized, both T_1 and T_2 are cut off, and the filter is inactive.

Preset P_1 serves to set the change-over voltage of the relay to 12.1 V: this is exactly midway between the supply voltage of 10.7 V required by the antenna and the control voltage of 13.5 V derived from the receiver.

The filter is best built on the small PCB shown in Fig. 4. When this board has been completed, it is small enough to be installed in the case housing the active antenna. As already stated, the filter is connected to the antenna via pins a and c.

Reference:
'Wideband active rod antenna', *Elektronika*, May 1991

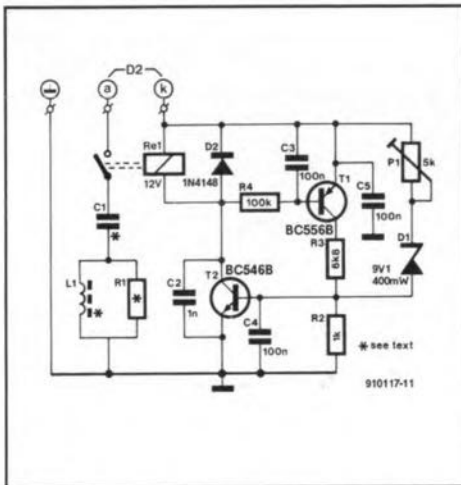


Fig. 2. Circuit diagram of the high-pass filter.

PARTS LIST

- Resistors:**
 R1 = 1 k Ω (curve 1) or 330 Ω (curve 2)
 R2 = 1 k Ω
 R3 = 6.8 k Ω
 R4 = 100 k Ω
 P1 = 4.7 k Ω preset

- Capacitors:**
 C1 = 1 nF (curve 1) or 270 pF (curve 2)
 C2 = 1 nF

C3-C5 = 100 nF

- Inductors:**
 L1 = 33 μ H (curve 1) or 3.3 μ H (curve 2)

- Semiconductors:**
 D1 = zener 9.1 V, 400 mW
 D2 = 1N4148
 T1 = BC556B
 T2 = BC546B

- Miscellaneous:**
 Re1 = reed relay, 12 V, coil = 1 k Ω

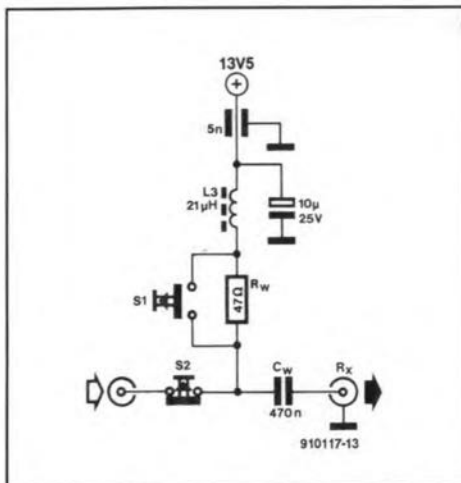


Fig. 3. Modified power supply of the active antenna.

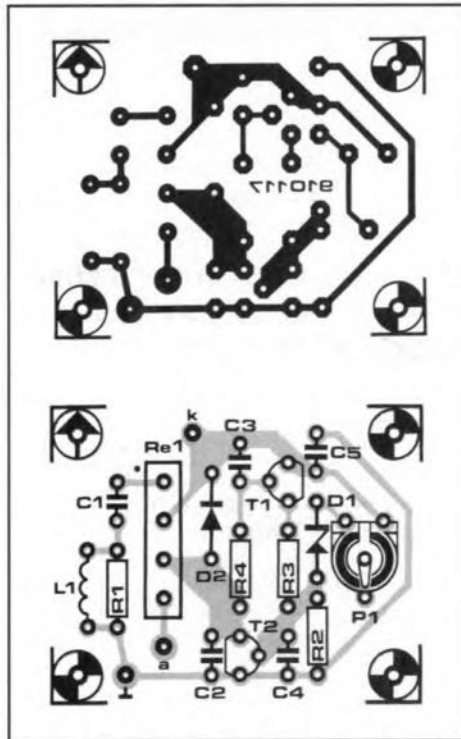
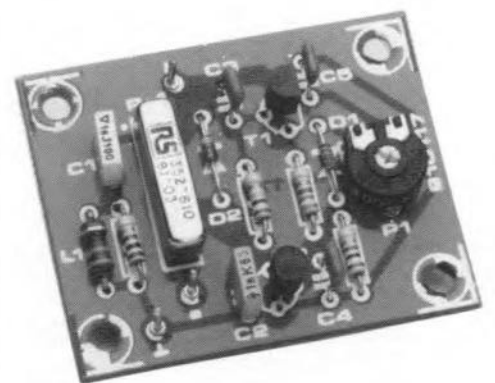


Fig. 4. Printed-circuit board for the high-pass filter.



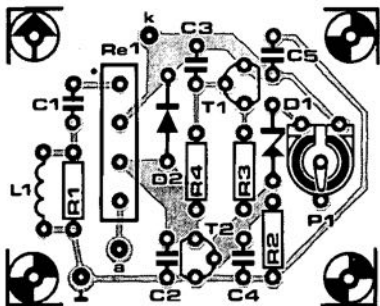
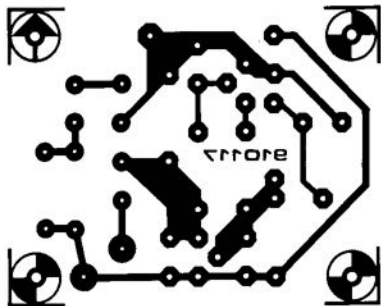


Fig. 4. Printed-circuit board for the high-pass filter.

CAPACITORS IN AF CIRCUITS

by H. Baggott

It is well known that coupling capacitors in hi-fi audio circuits can adversely affect the tone quality of the circuits. Unfortunately, there are frequently good reasons that make their use unavoidable. But how do you decide which type to use in a high-quality amplifier?

There are so many different types and makes on the market that this is not an easy decision. This article is intended to help in choosing a suitable capacitor.

CONVENTIONAL capacitors are made of two thin metal foils separated by a thin insulator or *dielectric*, such as mica or a man-made fibre. This sandwich is rolled or folded into a compact size and covered with an insulating coating. A wire terminal is attached to each foil. To increase the capacitance, the dielectric should be as thin as possible. This can only be done at the expense of limiting the maximum voltage that can be applied before the insulator ruptures because of the intense electric field. Another important factor is the resistivity of the dielectric. Thin, large-area shapes increase the *leakage* resistance between the foils and thus degrade the capacitor.

In ceramic and plastic-film capacitors, the metal-film plates are deposited directly on to the dielectric. Plastic dielectrics have very high resistivity so that the leakage resistance is very small.

Electrolytic capacitors are made of an oxidized metal foil in a conducting paste (dry) or solution (wet). The thin oxide film is the dielectric between the metal foil and the paste or solution. Since the film is very thin, the capacitance is large. The metal foil is normally made of aluminium or tantalum.

The capacitance, C , of a capacitor is determined by the dimensions of the foils and the thickness and relative permittivity, ϵ_r , of the insulator:

$$C = \epsilon_r A / d \times 8.85 \times 10^{-12} \text{ [farad]},$$

where A is the surface area of the foils in m^2 and d is the distance between the foils in m . The ϵ_r of polyester is about 3, while that of tantalum oxide is around 11.

The thickness and type of material of the dielectric determine the breakdown voltage of the capacitor: therefore, a high-voltage type is larger than a low-voltage one.

A capacitor is a non-linear electrical component, which makes it very useful in a number of applications. Its specific characteristic is the frequency-dependent reactance, X_C , which, for an ideal capacitor, is

$$X_C = 1 / 2\pi f C \text{ } [\Omega].$$

This would appear to indicate that the reactance characteristic of a capacitor is a con-



stantly dropping curve (on a logarithmic scale). This is, of course, not so, because the reactance would then really become 0Ω .

This is not the only non-ideal aspect of a capacitor. Apart from internal resistance (which is, of course, unwanted), a capacitor also has self-inductance, the magnitude of which is determined by its construction, the manner in which the terminals are connected to the foils and the length and shape of the terminals.

Furthermore, no dielectric is a perfect insulator; therefore, leakage currents will occur and these play an increasing role as the voltage across the capacitor becomes higher.

Figure 1 shows the equivalent circuit of a practical capacitor: C is the real capacitance and this is shunted by the insulating resistance, R_p of the dielectric. In series with that combination is a resistance, R_s , which represents the minimum transfer resistance of the capacitor from one terminal to the other. In

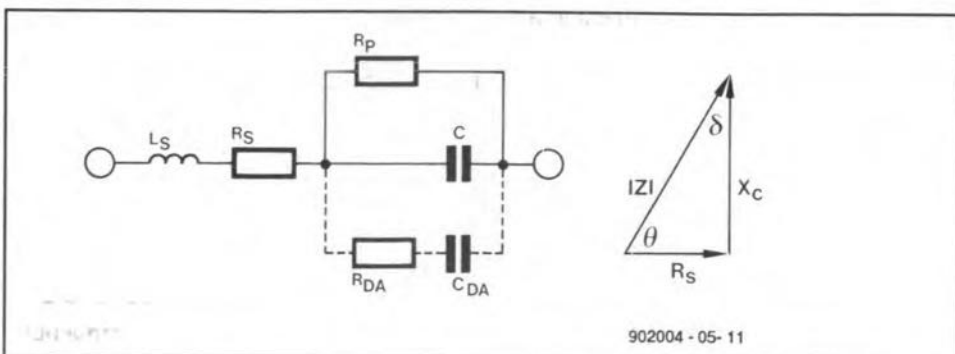


Fig. 1. The equivalent circuit and vector diagram of a capacitor.

series with that network is the self-inductance, L_s , of the capacitor. Furthermore, in parallel with C and R_p is a series network consisting of R_{DA} and C_{DA} , which represents the dielectric absorption of the capacitor. This is a less well-known property of capacitors. The dielectric absorption, DA, is a charge displacement phenomenon in the dielectric that causes a sort of memory lapse (a delayed transfer of acquired energy). Very few manufacturers quote the DA in their datasheets. This phenomenon, which affects the sound quality of the circuit in which the capacitor is used, will be reverted to later on in the article.

The impedance vs frequency characteristic, obtained from measurements on a 2.2 μF capacitor, is shown in Fig. 2. Up to about 200 kHz, the capacitor behaves almost ideally: the impedance diminishes linearly. At around 900 kHz, resonance is brought about by C and L_s (see Fig. 1). The minimum impedance at that point is virtually equal to R_s . At frequencies above 2 MHz, the capacitor behaves as a pure inductance (L_s).

Specifications

Parameters to look for in manufacturers's specifications and data sheets are given below.

- The *dissipation factor*, $\cot\theta$ or $\tan\delta$, indicates the losses caused by R_s , and should thus have a low value (θ is the phase angle; δ is the loss angle). A low value is particularly important if the capacitor is to be used in a cross-over filter. Note that $\tan\delta$ is frequency-dependent and is approximately equal to $2\pi fCR_s$. Some manufacturers give the value of R_s separately for large-value electrolytic capacitors.
- Only a few manufacturers give the value of the *dielectric absorption* (DA) and then only in the case of film capacitors. Here again, the lower this value, the better the capacitor. The DA of electrolytic capacitors is so large that it is never quoted.
- The *insulation resistance*, R_p , is normally of the order of hundreds of megohms and seldom plays a role in audio applications.
- The dissipation may be indicated by the *power factor*, which is equal to the ratio $R_s/Z (= \sin\delta)$.
- The *temperature behaviour* is usually given for a certain dielectric—see Fig. 3.
- The *capacitance* and associated *tolerance* of most capacitors, but not HF types, is normally given at a frequency of 1 kHz.
- The *working voltage* must, of course, be higher than the maximum voltage that will occur across the capacitor. Note that the manufacturer may state this as a direct voltage or as an alternating voltage.

What kind of capacitor?

Since this article deals with capacitors for audio applications, we will restrict us to the kinds of capacitor that are available in relatively high values: ceramic and mica capacitors are, therefore, not considered. That leaves film, electrolytic and paper capacitors.

Paper capacitors are hardly seen these days, although they are found in some equipment

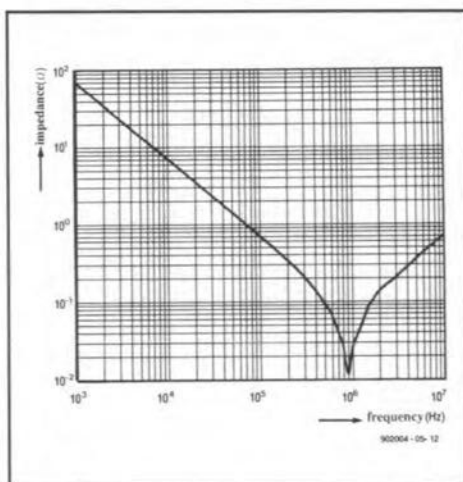


Fig. 2. Impedance vs frequency characteristic of a metal-plated polypropylene (MKP) film capacitor.

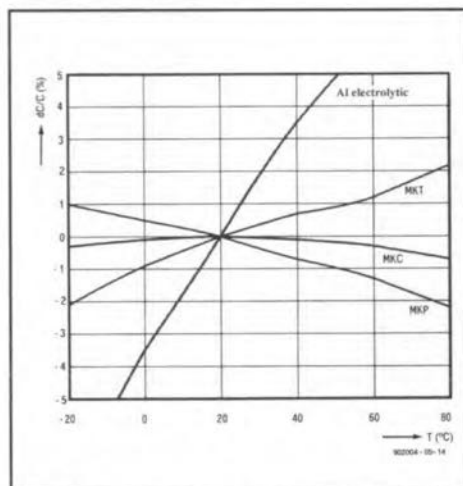


Fig. 3. Temperature-dependent behaviour of four different kinds of capacitor.

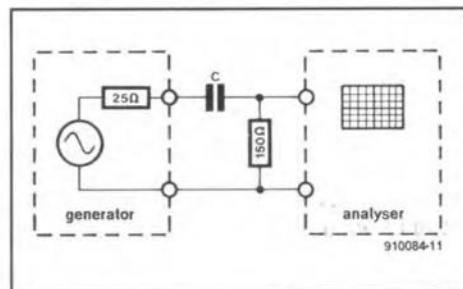


Fig. 4. Setup for measuring the harmonic distortion of various capacitors.

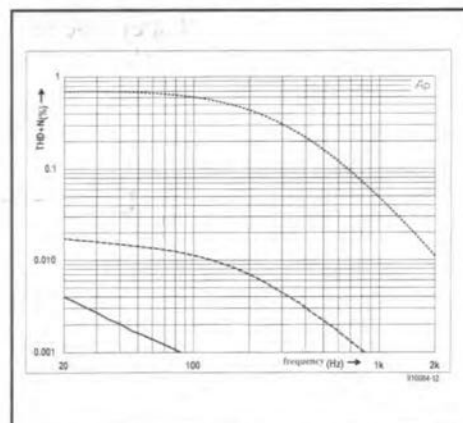


Fig. 5. Harmonic distortion vs frequency curves for tantalum (top), electrolytic and MKT (bottom) capacitors.

from the Soviet Union. Their quality is in general very good.

Of the film capacitors, *polyester* (Mylar™) types form the largest and least expensive group. Their quality is fairly good and their dimensions are reasonable.

Polycarbonate (MKC) capacitors have slightly better properties than polyester types. They are the right choice for circuits where good temperature behaviour is important.

Polypropylene (MKP) types are better still, but are generally rather larger than the previous two types.

Polystyrene (MKS or Styroflex™) capacitors are definitely the best for use in audio circuits, but they are fairly large and generally not available in values above 0.5 μF .

Electrolytic capacitors are decidedly inferior to film capacitors. Their tolerance is fairly large and this makes them unsuitable for use in filters.

A special version of the traditional wet electrolytic capacitor is the *bipolar* type that is used primarily in cross-over filters, although they can be useful in other audio circuits.

Finally, *tantalum* capacitors are not really suitable for processing audio signals, because, owing to their construction, they exhibit semiconductor effects.

Measurement results

A large number of measurements on a variety of capacitors from different manufacturers showed the following results.

The measured value of capacitance deviated from the stated value by not more than 2% in the case of *polypropylene* capacitors; not more than 4% in most of the tested *polyterephthalate* (MKT) types; and up to 20% in the case of *electrolytic* capacitors.

The measured *dissipation factor*, $\tan\delta$, of all tested capacitors was low.

The *third harmonic distortion* (THD) of capacitors used in a high-pass filter with a load resistance of 100 Ω , measured at 250 Hz, was <0.001% for all film types, and varied between 0.011% and 0.025% with electrolytic types.

The *dielectric absorption*, DA, was measured by charging the capacitors for 5 minutes at a direct voltage of 1.5 V, then short-circuiting them for 3 seconds and finally measuring the residual voltage with a voltmeter with 50 M Ω input impedance. This deviates somewhat from the MIL-C-19978-D test, but it is felt to give a better insight into the relation between C_{DA} and R_{DA} . The DA for all MKP types was 0.01% or smaller; varied between 0.05% and 0.11% with MKT types; and varied between 0.63% and 3.3% with electrolytic types.

It is clear from these measurements that differences between capacitors with the same dielectric are small. This has, no doubt, a lot to do with the fact that capacitor manufacturers buy the foil from a small number of producers. The tests threw up a few bad results even with the more expensive types. In other words, even when you buy an expensive capacitor, you have a (very) small chance of getting a rogue.

The poor DA figures of electrolytic capacitors are probably the reason that these components often adversely affect the sound quality of audio circuits, which is not at all evident from their THD figures. Note that the DA and the THD have no direct relationship.

The self-inductance of the capacitors tested was negligibly small: 50 nH in the case of $2.2\text{ }\mu\text{F}</math> capacitors. Modern production methods appear to result in minimal self-inductance: most of this is formed by the terminals (length and shape).$

As an aside: when procuring the many capacitors for the tests, it was found that the larger values are normally stocked by loud-

speaker DIY dealers, but not by many general electronics retailers.

In the audio circuit

Where quality is paramount, leave out anything from the signal path that is not strictly necessary is good advice. But, be careful, because poor components in the feedback loop of an opamp or power amplifier do, of course, also adversely affect the quality of the signal. Furthermore, the power supply also plays a role: it is advisable to shunt its large electrolytic capacitors with film types of not less than $0.47\text{--}1.0\text{ }\mu\text{F}</math> to improve the circuit's performance at higher frequencies.$

Fig. 9. Capacitors are made of two thin metal foils separated by a thin insulator or dielectric, such as mica or a man-made fibre. In plastic-film capacitors, the metal plates are deposited directly on to the dielectric. This 'sandwich' is rolled or folded into a compact size and covered with an insulating coating. An axial wire lead is attached to each foil. Shown here are the various production stages of a metal-plated polyterephthalate film (MKT) capacitor.

Electrolytic capacitors are made of an oxidized metal (usually aluminium or tantalum) foil in a conducting paste ('dry electrolytic') or solution ('wet electrolytic'). The thin oxide film is the dielectric between the metal foil and the solution or paste. Since that film is very thin, the capacitance is large: values from $1\text{ }\mu\text{F}</math> to $10\,000\text{ }\mu\text{F}</math> are available. The largest values can only be used in circuits where the applied voltage is low to avoid breakdown of the dielectric.$$

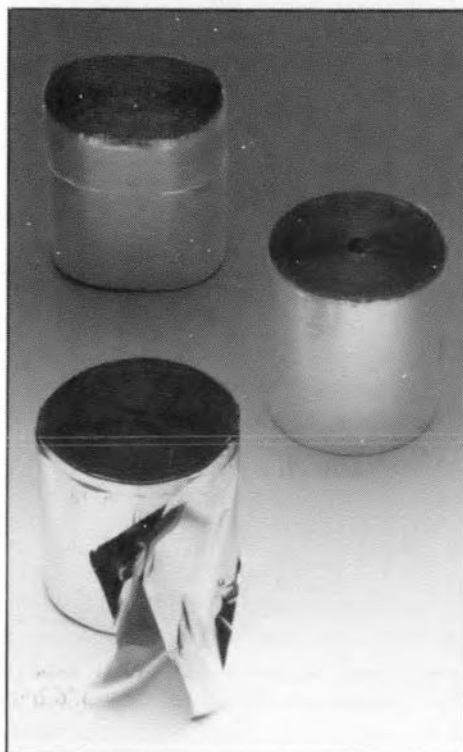
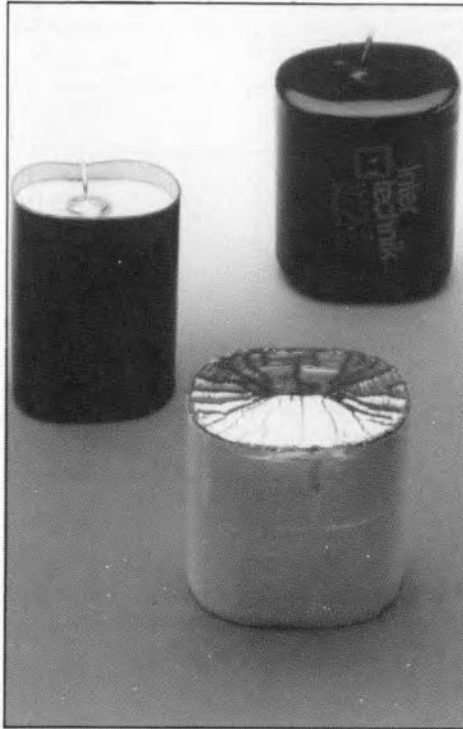


Figure 4 shows an interesting setup for investigating the kinds of irregularity capacitors produce. It is a high-pass filter with

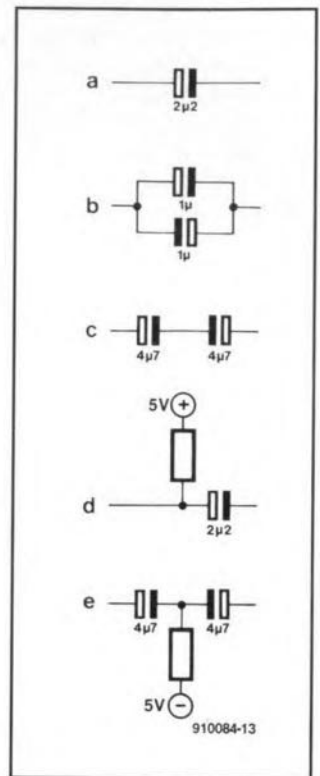


Fig. 6. Possible configurations of electrolytic capacitors.

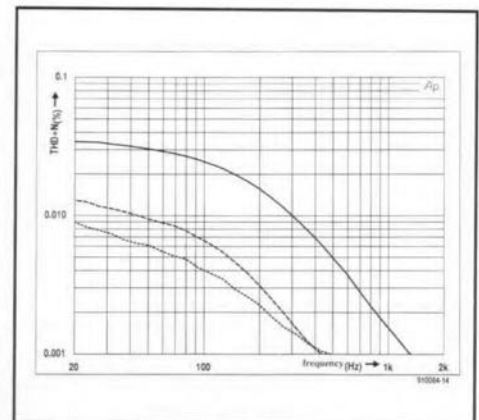


Fig. 7. Harmonic distortion vs frequency characteristics of configurations in Fig. 6a (top); 6b, and 6c (bottom).

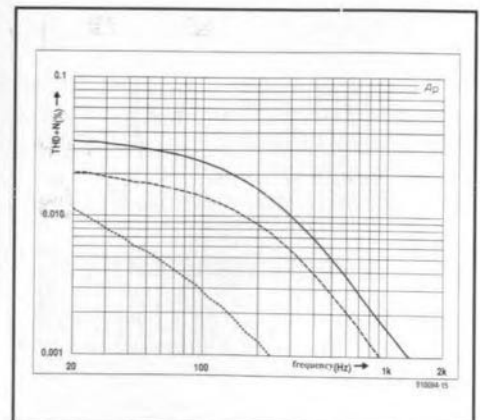


Fig. 8. Harmonic distortion vs frequency characteristics of configurations in Fig. 6a (top); 6d, and 6e (bottom).

cut-off frequency at around 400 Hz. A relatively low load is used to better show up any deficiencies (high loads improve the distortion factor). A frequency of a few hundred hertz is necessary to show how the capacitor behaves below the cut-off point.

Figure 5 shows the harmonic distortion of three types of capacitor: MKT, wet elec-

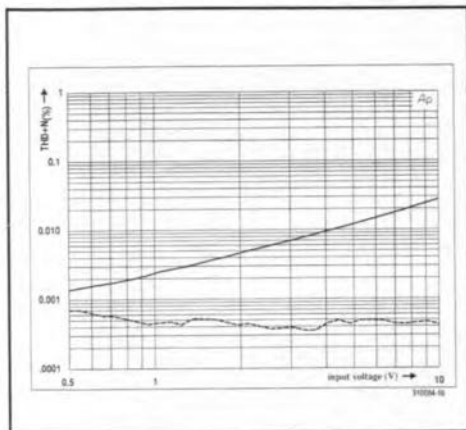


Fig. 10. Distortion vs applied voltage characteristic of an electrolytic capacitor (top) and a metal-plated polypropylene type; the test frequency was 500 Hz.

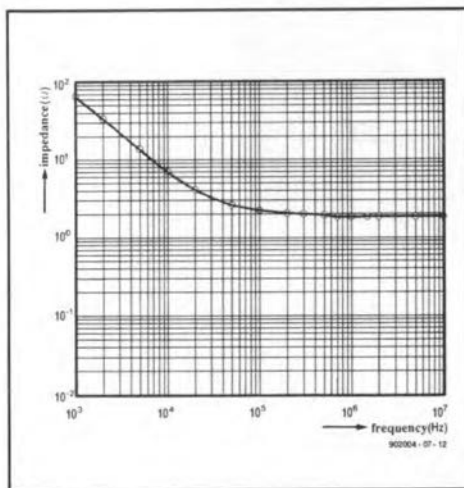


Fig. 11. Impedance vs frequency characteristic of an electrolytic capacitor shows that above 10 kHz the component does no longer behave like a capacitor.

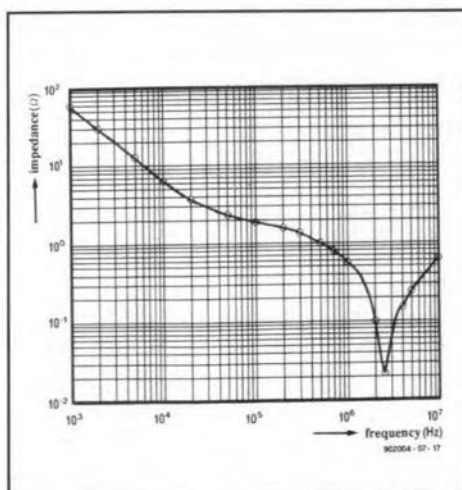


Fig. 12. The impedance vs frequency characteristic of an electrolytic capacitor shunted by a film type is still far from ideal.

trolytic and tantalum. The tantalum is poor, the electrolytic is reasonable and the MKT produces virtually no distortion (ignore its distortion at very low frequencies because that is caused by the measurement setup).

It appears, therefore, that the distortion increases below the cut-off point of the RC combination, that is, when the voltage across the capacitor increases: the same condition that a coupling capacitor experiences. It could be concluded that it would be advantageous to give the coupling capacitor a much larger value than necessary, that is, to choose a cut-off point of 1 Hz instead of 10 Hz. In principle, the area of distortion would then also be shifted downwards and largely fall outside the audio range. A measurement with a 100 μF capacitor in the setup of Fig. 4 showed that the distortion did, indeed, shift down relative to that with a 2.2 μF capacitor, but it also showed that the distortion characteristic became much steeper upwards. In other words, large-value electrolytic capacitors produce a relatively much larger distortion than smaller ones.

It is undoubtedly best to use MKP or MKT types for coupling capacitors. Unfortunately, that is not always possible owing to non-availability or lack of space, and electrolytic types must then be used. To find out how to keep the distortion caused by these components as low as possible, measurements were carried out on several configurations of electrolytic capacitors as shown in Fig. 6; the resulting distortions are shown in Fig. 7 and Fig. 8. The level of the input signal was 2 V r.m.s.

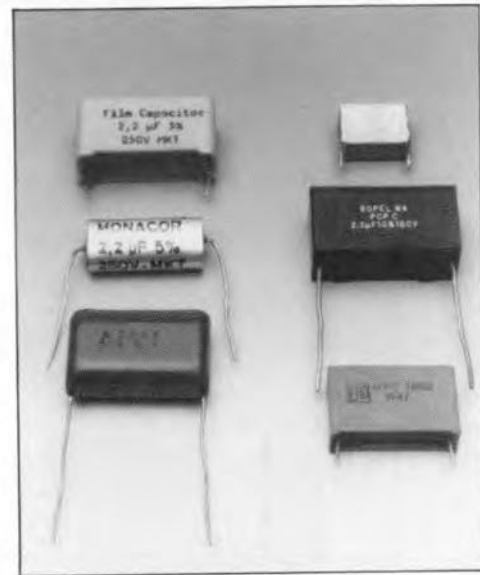
The distortion of a single electrolytic capacitor (Fig. 6 a) is fairly high as seen in Fig. 7. The distortion of an anti-parallel network (Fig. 6 b) is much smaller, but this configuration can handle alternating voltages of a few volts only; moreover, the direct voltage across it must be virtually zero. The series network (Fig. 6 c) is better still; it behaves, in principle, as a bipolar electrolytic capacitor.

In a practical circuit, there will be a direct potential across the coupling capacitor, which can have beneficial results. For instance, with a direct voltage of 5 V across a single electrolytic capacitor (Fig. 6 d), the distortion, as shown in Fig. 8, is noticeably smaller. If the same voltage is applied to a series combination (Fig. 6 e), a small improvement in distortion compared with that of the setup in Fig. 6 c results. This is easily realized with the aid of a resistor of, say, 100 k Ω , to the negative supply line.

Figure 10 shows the distortion measured on an electrolytic capacitor and an MKP type at a frequency of 500 Hz and input signals of 0.5–10 V r.m.s. It is clear that the distortion caused by the electrolytic type is voltage-dependent. When, therefore, large signal levels (as in valve amplifiers) are processed, the quality of the coupling capacitor is even more important than with small signals.

In these tests, only harmonic distortion was measured. In the case of electrolytic capacitors, this is caused primarily by odd harmonics, which are particularly offensive to the human ear. There are, however, other types

of distortion, such as that caused by DA. This causes irregularities in the dynamic behaviour of a capacitor and also muffles the sound at low frequencies. Figure 11 shows the impedance characteristic of a typical 2.2 μF electrolytic capacitor, which, compared with the curve of Fig. 2, is poor. At 20 kHz, the curve is already well away from the ideal line, while above that frequency, the impedance sticks at 2 Ω . In preamplifiers, this does not matter all that much, because the terminating impedances there are of the order of a some thousands of ohms. In low-impedance circuits, however, it does. To retain the



proper functioning at higher frequencies, the electrolytic capacitor is often shunted by a film type. This is, however, effective only if the value of the film capacitor is not too small compared with that of the electrolytic type. Figure 12 shows what happens when a 2.2 μF electrolytic capacitor is shunted by a 0.22 μF MKT type: the characteristic improves but only at fairly high frequencies. For good results, the value of the film capacitor must be not less than one third of that of the electrolytic type.

The shunting has no effect on the DA, because the worst component in the parallel network determines the DA. Only when the values of the film capacitor and the electrolytic capacitor are about the same is the DA of the combination reduced to about half that of the electrolytic capacitor.

Figure 12 also shows that shunting the electrolytic capacitors in a power supply with film types makes real sense. The electrolytic capacitors alone hardly decouple the supply lines at higher frequencies, whereas the film capacitors ensure that signals up to a few MHz are suppressed effectively so that they cannot cause interference in the audio circuits.

AUDIO/VIDEO SWITCHING UNIT



As more and more audiovisual equipment is cheerfully stacked up in your living room, connecting it all is bound to become a problem sooner or later. What do we want? A number of audio and video signals have to be switched between different equipment: for instance, we want the pictures produced by the hi-fi video recorder to be visible on two TV sets at the same time, while the recorder should still allow us to choose between recording TV sound or a signal from 'the stereo'. All this is possible with the electronic switch discussed here, which can be given as many inputs as you think necessary.

by T. Giffard

GONE are the days when every home had one TV set and one radio, and many of you will have grown accustomed to the presence of a stereo set, a video recorder, several TV sets, a camcorder, a portable CD player, or a DAT recorder in the living room and elsewhere in the home. Unfortunately, linking all this wonderful equipment is never easy, since low-frequency (audio) as well as high-frequency (video) signals have

to be switched and routed without losses and cross-talk. Apparently, a kind of 'switchbox' is in order.

The audio/video switching unit described in this article has the function of a versatile signal router that allows two or more devices to be connected to a single video input on a TV set or a video monitor. This feature is particularly useful with older TV sets.

Switching and routing video signals is not as easy as audio signals because the signal bandwidth is much greater (approx. 6 MHz instead of 20 kHz). Since an ordinary rotary switch is not suitable for this function, Philips Components have developed an integrated circuit capable of switching two (stereo) audio and two video signals simultaneously, and electronically. Fig. 1 shows the block diagram of this IC, the TDA8440,

which we introduced in Ref. 1. Although this IC has been designed primarily for connection to an I²C bus, it can also be controlled by ordinary logic levels. This is made possible by the double function of the 'select' inputs on the TDA8440, S₀, S₁ and S₂. Normally, these three inputs are used to locate the IC at one of seven possible addresses in an I²C system, allowing up to seven TDA8440s to be connected in parallel to an I²C bus. However, when all three 'select' inputs are made high simultaneously, the IC is switched to non-I²C mode, in which it can be controlled with ordinary logic signals. The input channel selection is then effected by appropriate control of the I²C bus inputs, SCL and SDA. In non-I²C mode, the SDA input may be used to switch between audio/video source '1' (SDA=+12 V) or audio/video source '2' (SDA=0 V). The SCL input is used to set the gain of the video amplifier. The gain is 2 when SCL=+12 V, or unity when SCL=0 V. When a number of TDA8440s are connected in parallel, the OFF input may be used to switch off 'unused' ICs. The TDA8440 is actuated by applying a logic low level to the OFF input. A high level disables the IC, when all outputs are switched to high-impedance.

The switchbox

With flexibility in mind, the circuit of the audio/video switch (and the associated printed circuit board) is designed such that it can be controlled via push-buttons as well as via the I²C bus. Furthermore, the switching unit has a modular structure, which enables inputs to be added as necessary. In practice, this means that a fair number of video sources (with accompanying audio signals) can be linked to a SCART connector on, say, a TV set.

Figure 2 shows the circuit diagram of the switching unit, which contains two identical modules. The circuit description below refers to the upper module in Fig. 2. Each module has two video inputs and two stereo audio inputs. The video and audio sources connected to these inputs can be switched on to the common outputs. The input and output impedance of the video channels is set to about 75 Ω with the aid of terminating resistors.

Thanks to the high-impedance 'off' mode of the TDA8440s, the outputs of the modules can be connected in parallel without problems. This makes it possible to use any number of modules together, so that it should always be possible to realize the required number of inputs. Note, however, that the capacity of the power supply may be limiting factor. The supply as drawn allows up to 8 modules (16 video inputs) to be used simultaneously. Although there exist 1.5-A versions of the 7812 voltage regulator (7812CV and others) that would allow up to 12 modules to be powered, the resultant power dissipation may well cause overheating of the regulator as the heat-sink is relatively small.

Returning to the circuit diagram, it will

MAIN SPECIFICATIONS (AUDIO CHANNELS)

Harmonic distortion (20 Hz to 20 kHz):	<0.004%
Dynamic IM distortion (square wave 3.15 kHz; sine wave 15 kHz):	<0.002%
Audio channel separation:	70 dB
Audio crosstalk (1 - 2):	-80 dB
Audio/video crosstalk suppression:	>90 dB
Amplitude error (20 Hz to 20 kHz):	+0 / -0.2 dB
Audio amplification (1 kHz):	-0.26 dB

Specifications measured at an audio bandwidth of 20 Hz to 20 kHz, and a source impedance of 600 Ω.

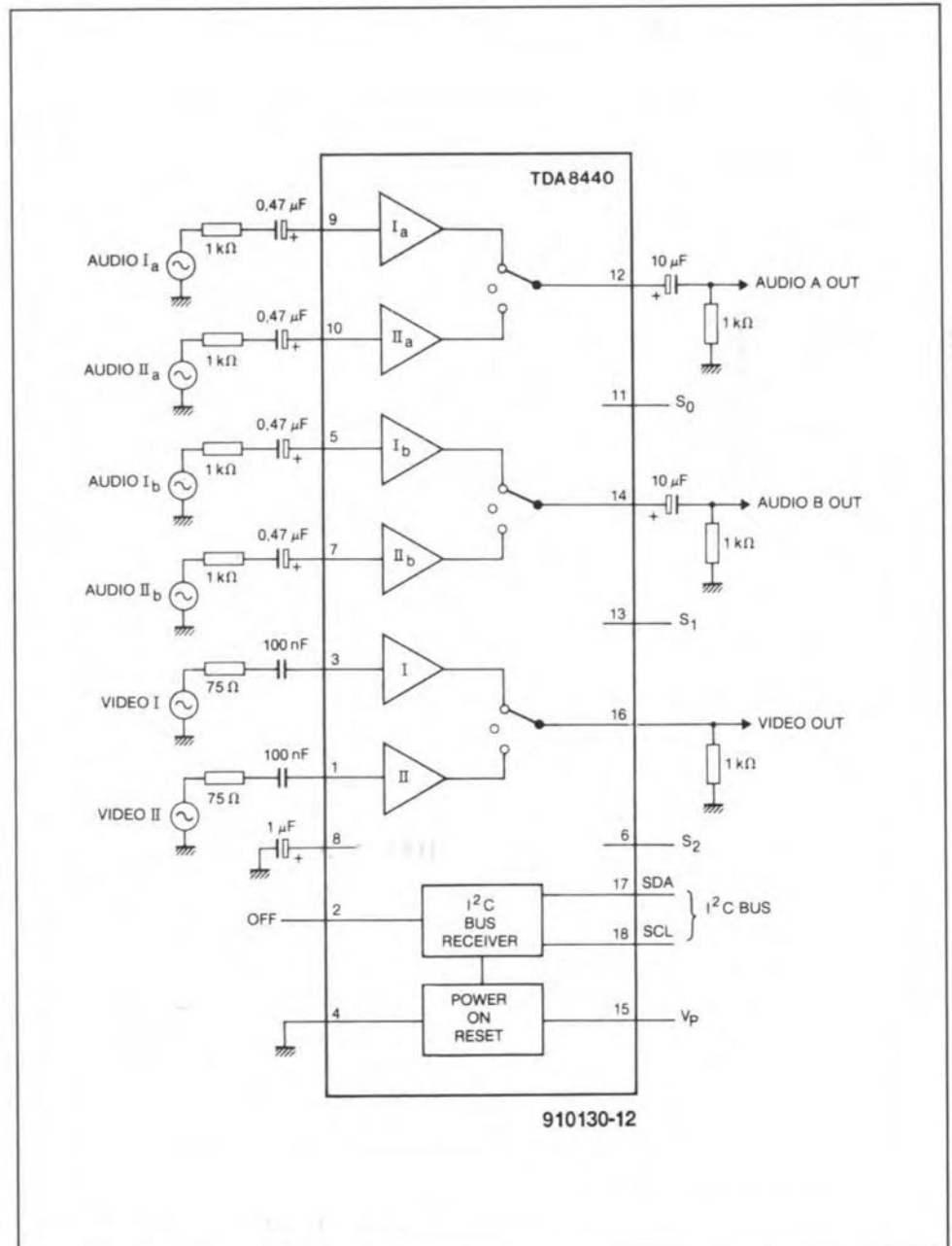


Fig. 1. Block diagram of the TDA8440.

be noted that a fair number of components is required to perform a basically simple function. This is caused partly by the need of correct termination on all video inputs. Also, two bistables are called for to be able to use the TDA8440 in non-I²C mode.

The IC is switched to non-I²C mode by

setting jumpers JP2-JP5 to the positions shown in the circuit diagram. Jumper JP1 is set to the position corresponding to the required gain of the video amplifier.

When one of the keys is pressed (on one of the modules shown in the circuit diagram or any other connected, identical, module),

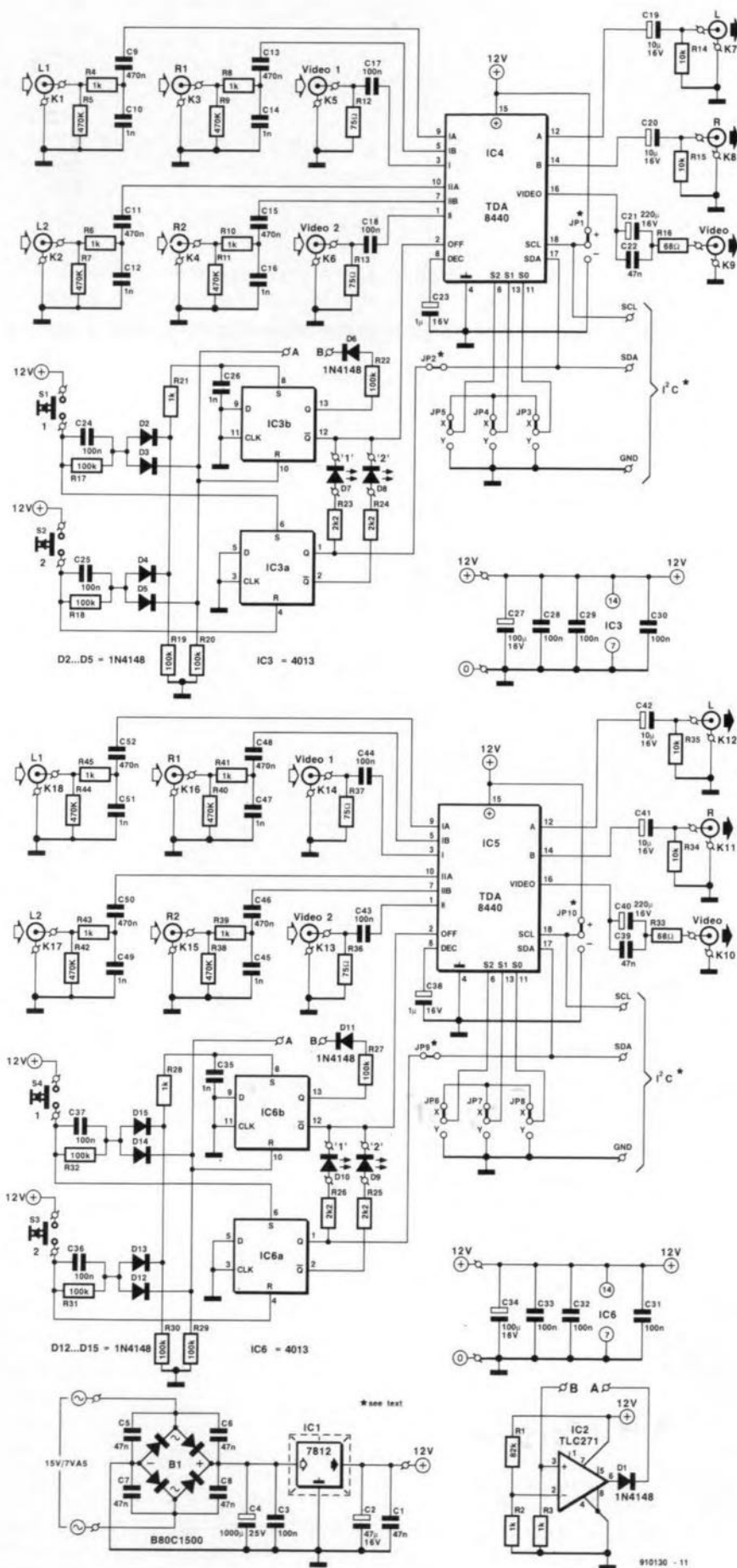


Fig. 2. Circuit diagram of the audio/video switching unit.

COMPONENTS LIST

Resistors:

2	68Ω	R16;R33
4	75Ω	R12;R13;R36;R37
12	1kΩ	R2;R3;R4;R6;R8; R10;R21;R28;R39; R41;R43;R45
4	2kΩ	R23-R26
4	10kΩ	R14;R15;R34;R35
1	82kΩ	R1
10	100kΩ	R17-R20;R22; R27;R29-R32
8	470kΩ	R5;R7;R9;R11; R38;R40;R42;R44

Capacitors:

10	1nF	C10;C12;C14; C16;C26;C35; C45;C47;C49;C51
7	47nF ceramic	C1;C5-C8;C22;C39
15	100nF	C3;C17;C18;C24; C25;C28-C33; C36;C37;C43;C44
8	470nF	C9;C11;C13;C15; C46;C48;C50;C52
2	1μF 16V radial	C23;C38
4	10μF 16V	C19;C20;C41;C42
1	47μF 16V	C2
2	100μF 16V radial	C27;C34
2	220μF 25V	C21;C40
1	1000μF 25V	C4

Semiconductors:

1	B80C1500	B1
4	LED high efficiency	D7-10
11	1N4148	D1-D6;D11-D15
1	7812	IC1
2	TDA8440	IC4;IC5
1	TLC271	IC2
2	4013	IC3;IC6

Miscellaneous:

8	3-way pin header	JP1;JP3-JP8;JP10
2	2-way pin header	JP2;JP9
18	phono (RCA) socket	K1-K18
4	SPST push button	S1-S4
1	Printed circuit board	910130

the common reset line, 'A' is actuated via a discrete OR gate, D3-D5-R20. All bistables connected to the reset line are cleared, so that their Q outputs go high. Another thing happens in the module that contains the pressed key: when key S1 or S2 is pressed, the associated bistable, IC3b, receives a set pulse in addition to the reset pulse. This set pulse is much longer than the (common) reset pulse, so that the bistable will eventually be set. Since the Q output of this bistable is connected to the OFF input of IC4, the electronic switch is actuated. This means that only the pressed key is enabled. All TDA8440s are disabled at power on, so that none of the video sources is connected to the output of the circuit.

Depending on whether S1 or S2 is pressed, IC3a is set or reset respectively. This bistable determines whether audio/video source '1' or '2' is connected to the output.

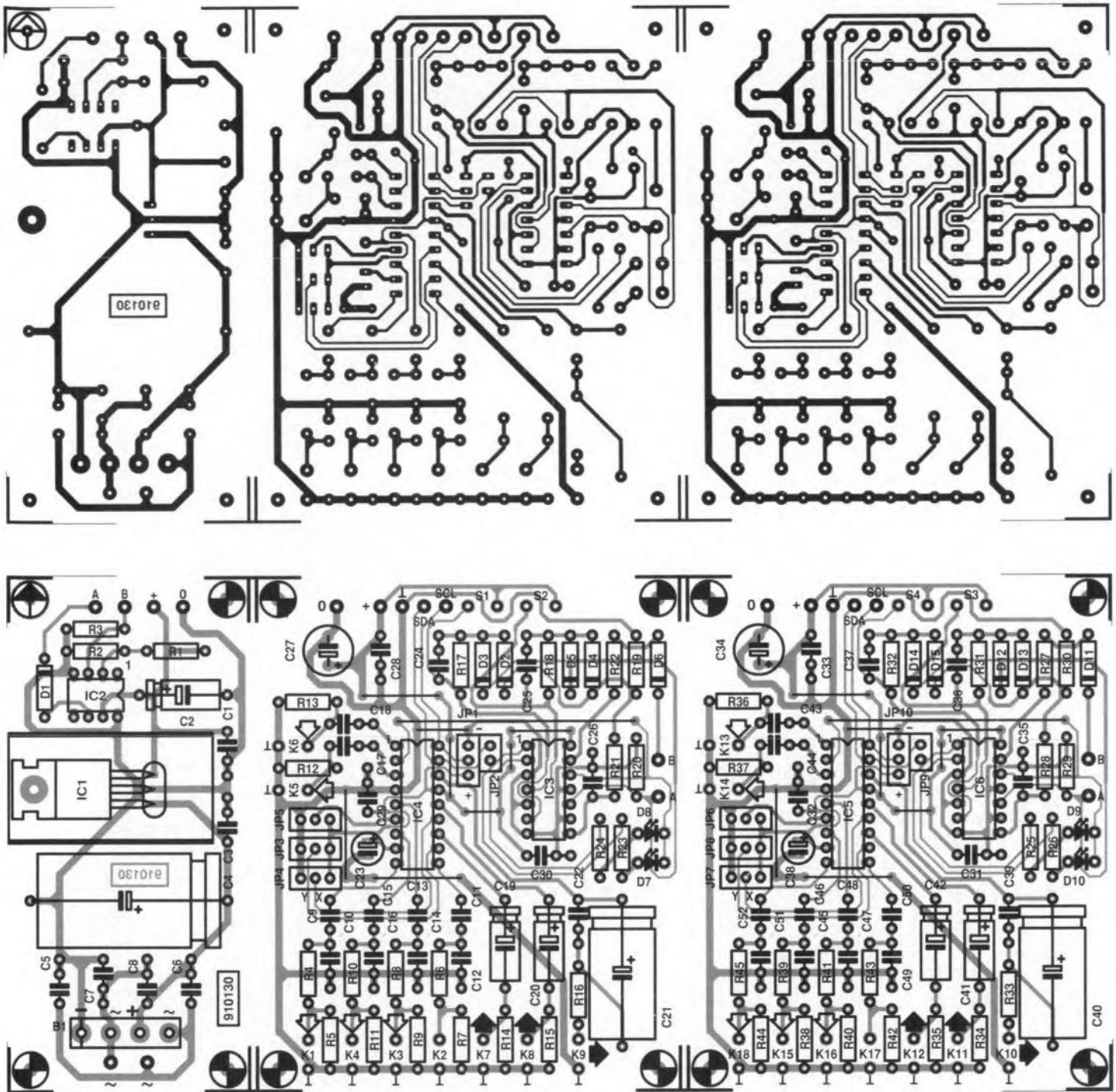
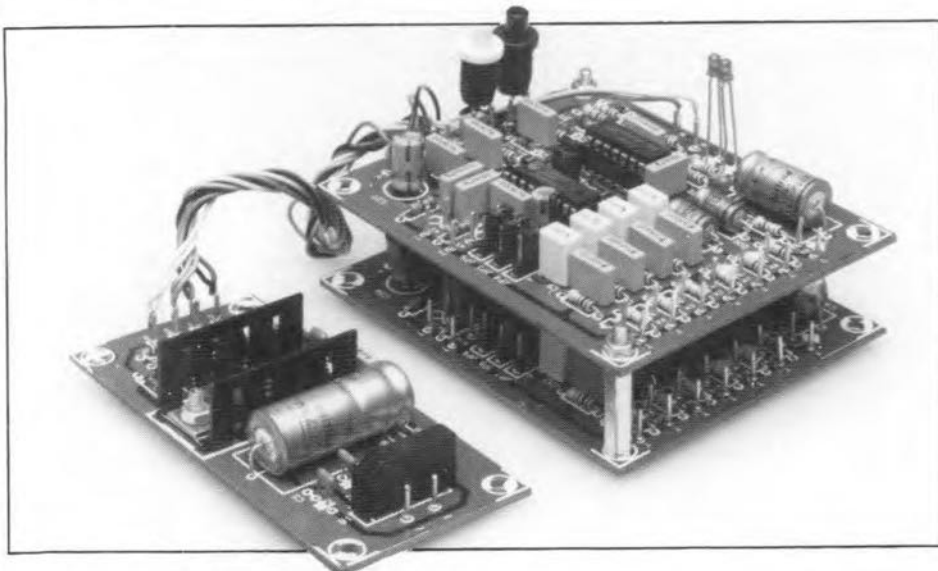


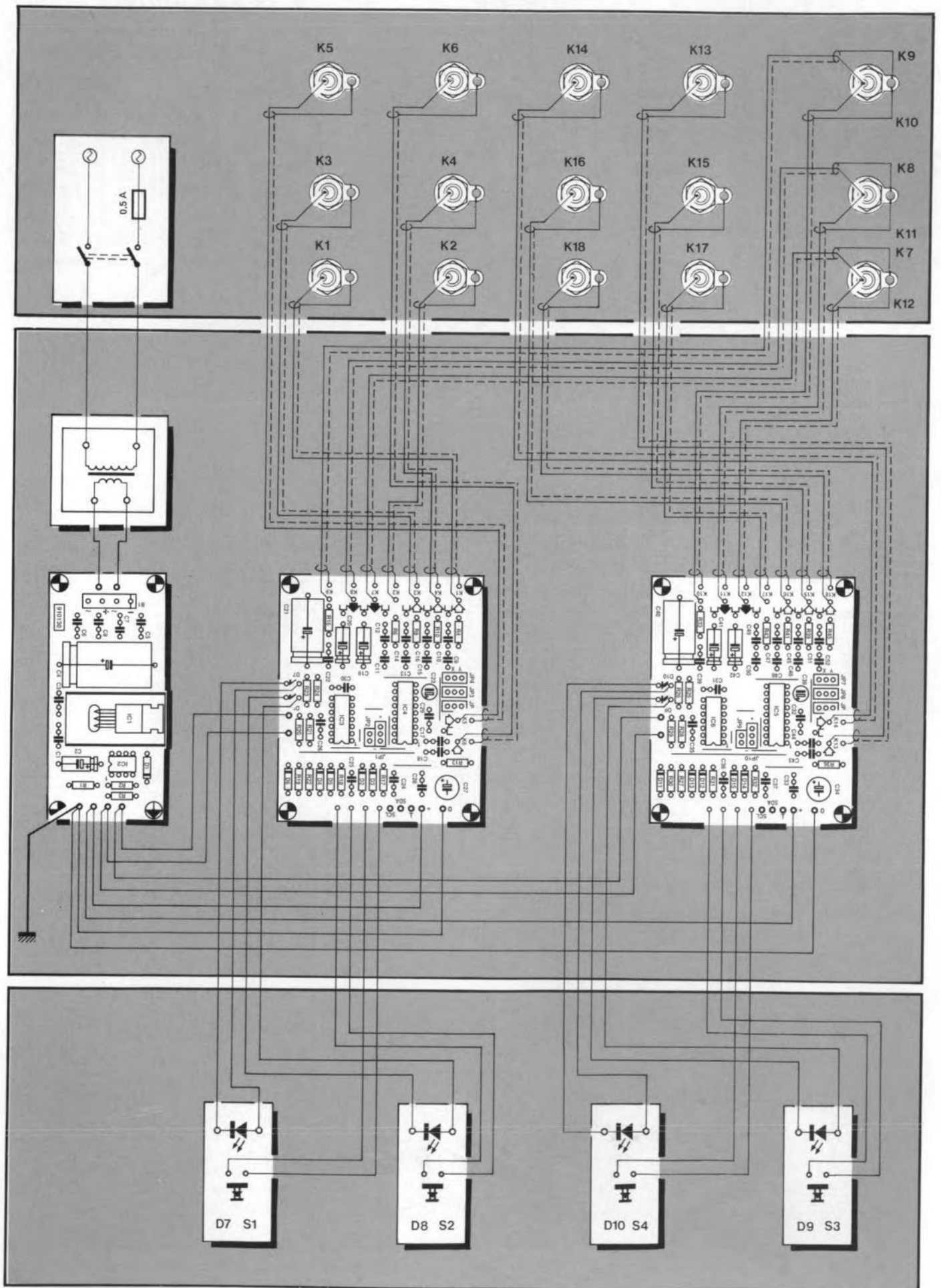
Fig. 3. Printed circuit board design.



Two LEDs connected between the outputs of the bistables indicate the source selection.

When the circuit is controlled via an I²C bus, jumpers JP1 and JP2 are omitted, and the desired address is set with the aid of jumpers JP3, JP4 and JP5. In this way, you can connect up to seven modules to an I²C bus system. All control logic, which includes the bistables, the switches and the associated discrete components, may then be omitted. Information on control commands and the bus protocol used with an I²C system may be found in the article 'I²C interface for PCs' elsewhere in this issue. In addition to this information, you may want to consult handbooks IC12a and IC12b published by Philips Components.

Note that the circuit diagram shows two identical modules which are built on separ-



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Fig. 4. Wiring diagram for a two-module version of the audio/video switching unit.

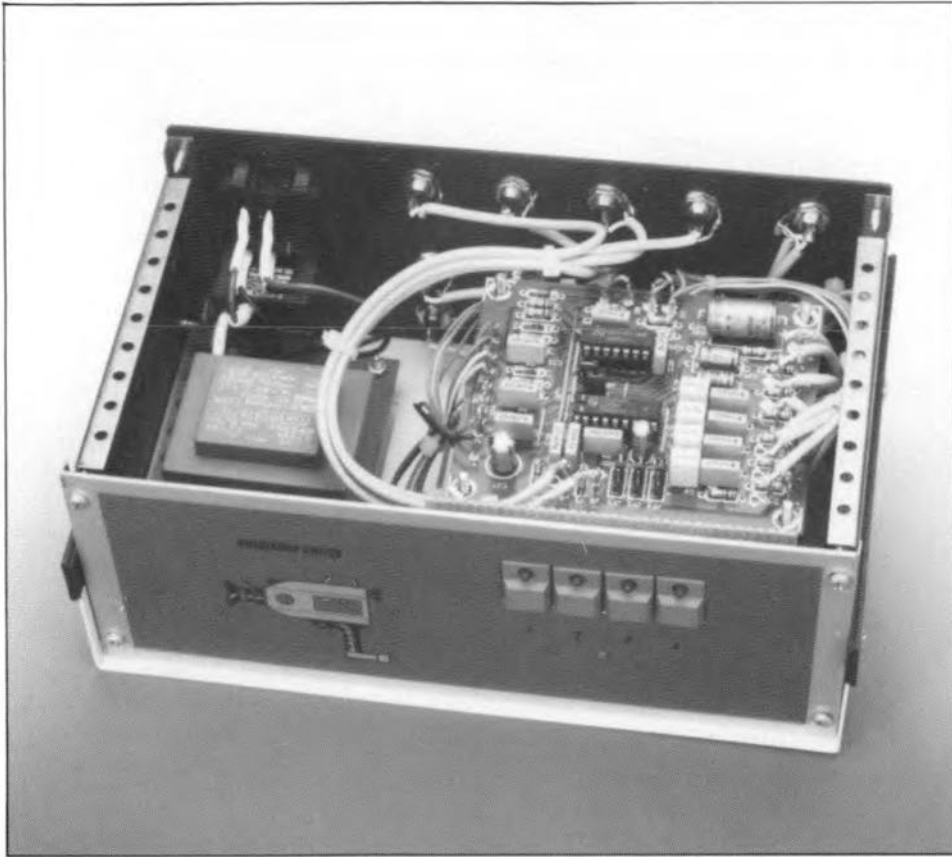


Fig. 5. Prototype of the A/V switching unit built into a Type LC850 enclosure from Telet.

ate PCBs. The PCB shown in Fig. 3 allow you to build two modules, which share a common power supply and a reset interface.

Construction

To prevent cross-talk between channels, the circuit is best built on the printed circuit board of which the layout is shown in Fig. 3. Before this PCB is populated, it must be cut

into three to separate the two modules and a section that contains opamp IC2 and the common power supply. When more than two modules are used (four inputs), only the module sections of the PCB are used. The supply board is left over, and put aside for other purposes as it is fairly universal.

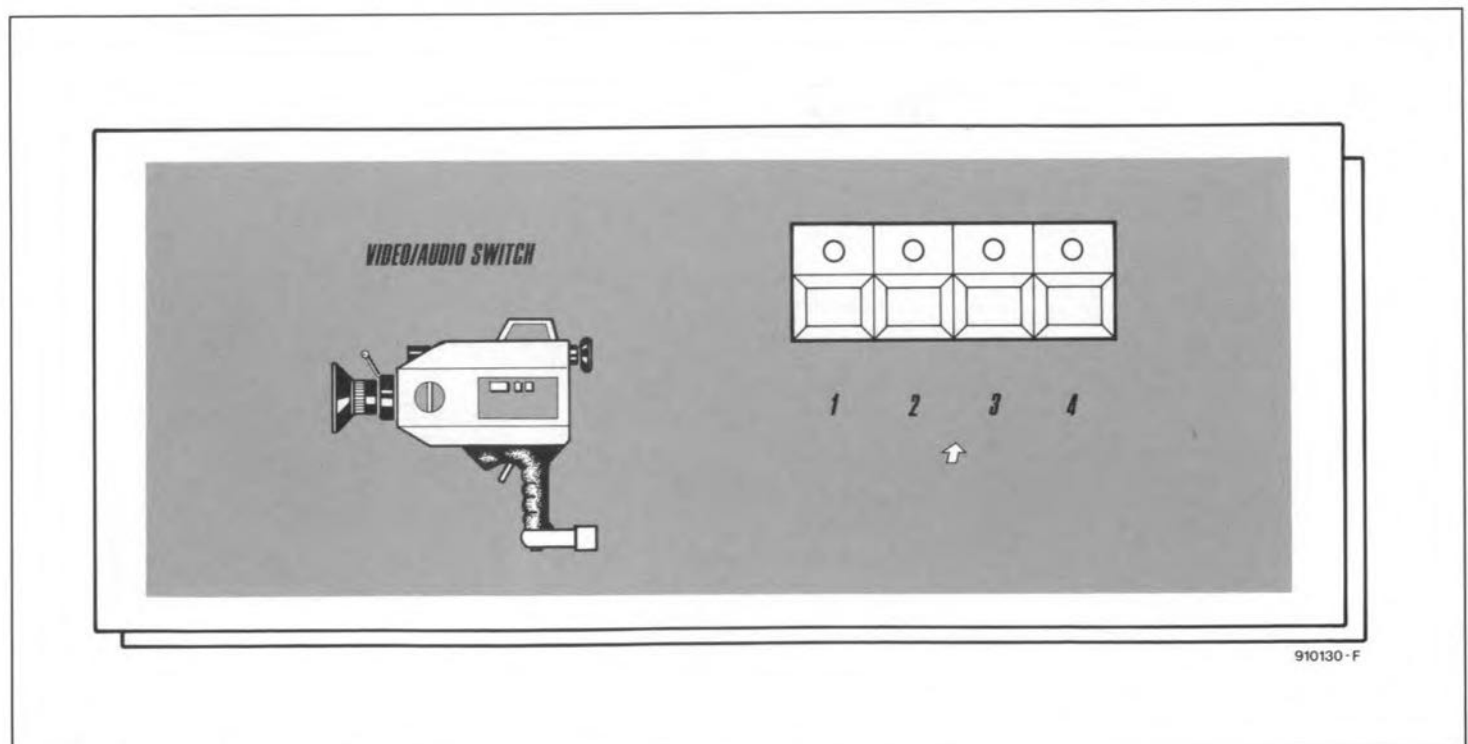
Having constructed the three PCBs in accordance with the component overlay and the components list, they are interconnected

as shown in the wiring diagram, Fig. 4.

The unit is given a neat appearance by using push-buttons with an integral LED (a series of IIT-Schadow 'Digitast' keys is available with this feature). These push-buttons may be mounted on to a small piece of veroboard or stripboard to create a compact controls panel that shows at a glance which video source is connected to the output. As shown in Fig. 5, the prototype is a 2-module version built into a metal enclosure. This unit has an internal mains power supply, and was fitted with a front panel to the layout given in Fig. 6. ■

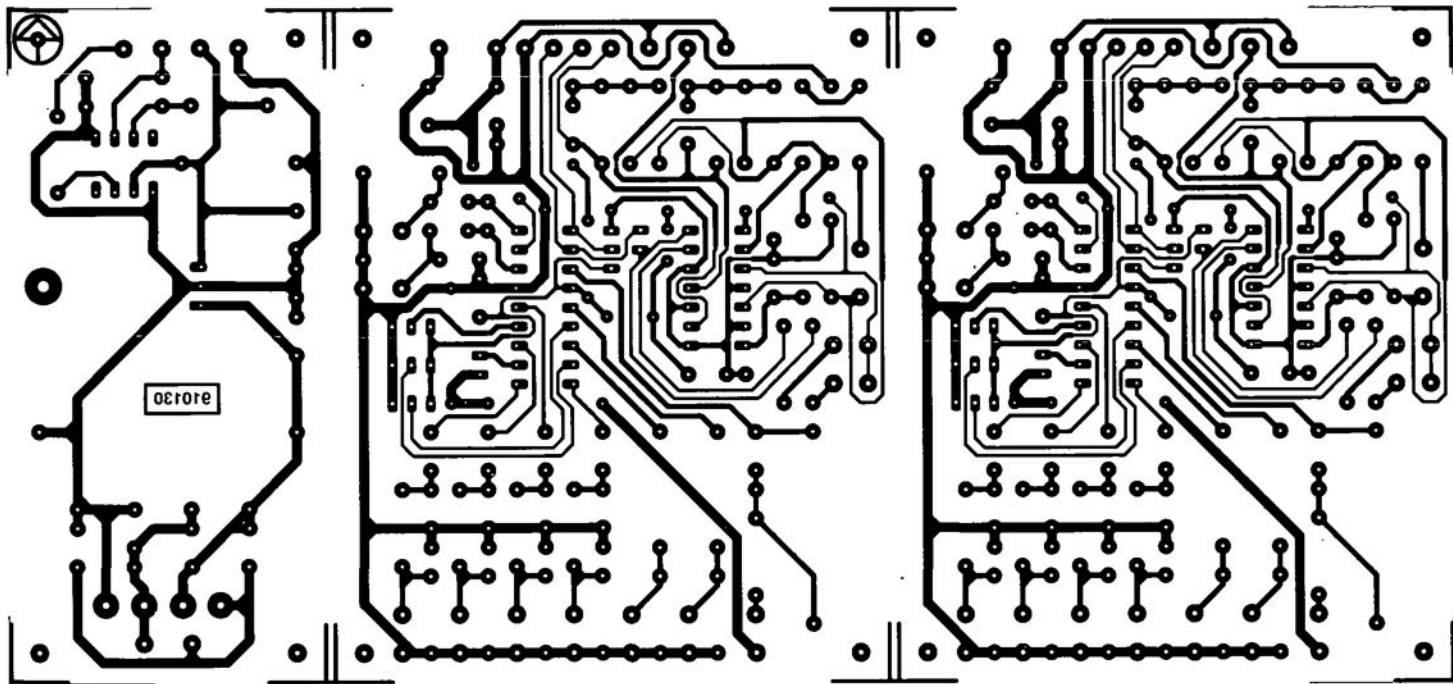
Reference:

1. "8-channel audio/video switch". *Elektor Electronics* March 1991.



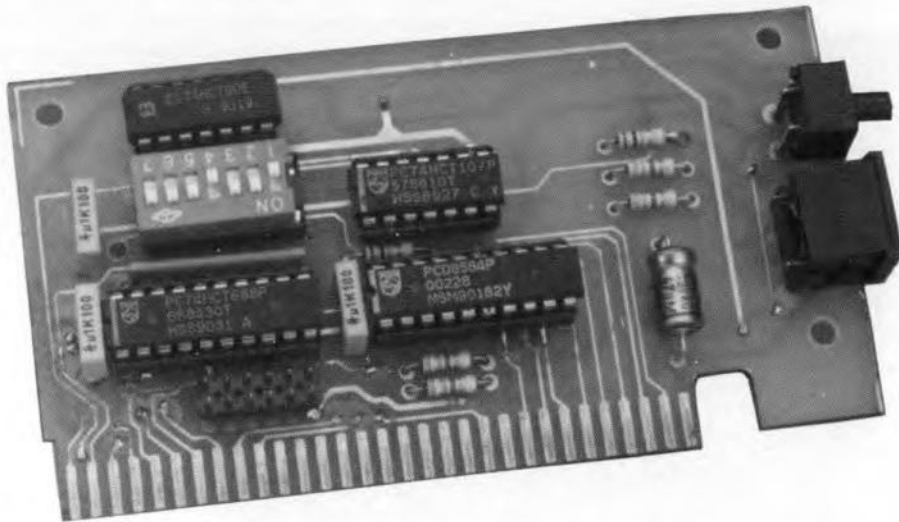
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Fig. 6. Suggested front panel layout.



I²C INTERFACE FOR PCs

FRONT COVER
PROJECT



The I²C interface (Inter-IC Communication) is a Philips invention that has been in use for years to enable ICs to communicate with each other in complex electronic equipment such as radios, video recorders and television sets. The insertion card we present here puts your PC in control of about ten I²C compatible ICs.

by J. Ruffel

IT is a simple but well known fact that the requirement for external pin connections is a limiting factor as regards the complexity of integrated circuits. Also, the more pins on an IC, the more expensive the device becomes to package and mount on a printed circuit board. Further, a large number of connections is inevitably coupled to a higher risk of malfunctions. No wonder IC manufacturers have sought alternative ways to allow complex ICs to communicate with as few as possible interconnections.

The I²C bus designed by Philips is such an alternative: it allows ICs to exchange data via two wires. This type of (serial) communication is particularly suited to relatively slow data transfer, and the protocol certainly does not allow, say, a computer RAM card to be implemented with I²C devices. By contrast, the I²C bus and protocol are perfect for, say, an I/O port or a real-time clock in a video recorder.

The I²C card discussed here allows an MS-DOS compatible PC to communicate with I²C ICs in external application circuits. As such, the card is for all of you who have noticed the large potential of I²C ICs (they are not generally expensive because of mass production), but have so far lacked the means to set up a control link to them. Having built the I²C interface, programming languages such as C, Pascal or assembler may be used to communicate with the I²C ICs as if they were 'external devices'.

I²C, a powerful standard

The I²C bus is a system bus based on three signals: SDA (system data), SCL (system clock) and ground. The SDA and SCL lines

are of the open-drain type, and must be tied to the positive supply line via an external resistor to create a bus structure that allows multiple inputs and outputs to be connected in parallel.

Figure 1 shows the basic electrical configuration. The two communication lines are logic high when they are inactive. The number of ICs connected to the bus is, in principle, unlimited. Note, however, that the lines do have a maximum specification of 400 pF in respect of the load capacitance. The maximum data rate that can be achieved on the I²C bus is about 100 kBit/s.

The definitions used in relation to I²C bus functions are basically as follows:

Master: this is the IC that determines the timing and the direction of a data transfer. This IC is the only one on the I²C bus to supply clock pulses on the SCL line. When multiple master devices are connected to a single I²C bus, this set-up is called a multi-master system.

Slave: this is any IC connected to the I²C bus that is not capable of generating clock pulses. Slave ICs receive commands and clock signals from a master.

Bus free: the bus is free when SDA and SCL are both logic high. A master can access the bus only when this is free.

Start: a master occupies the bus by generating a start condition, which means that SDA is made low while SCL remains high.

Stop: a master can free the bus again by generating a stop condition, which means that SDA is made high while SCL is high.

Data valid: the data on the SDA line must be

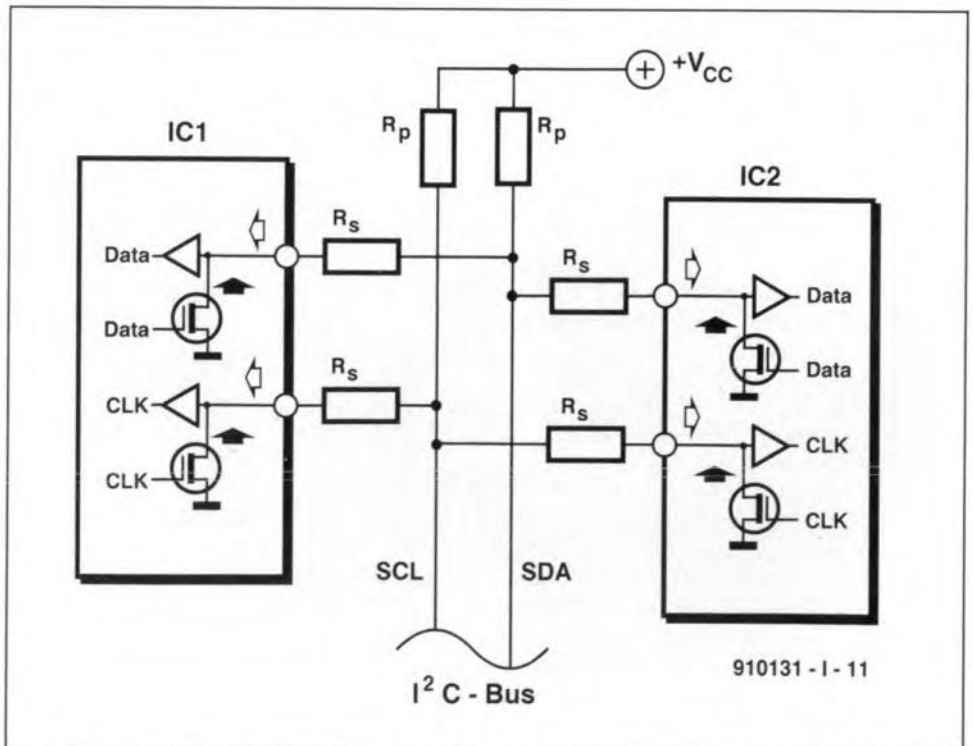


Fig. 1. The I²C bus is basically a 2-wire serial communication system based on open-drain outputs in connected devices, series resistors (R_s) and pull-up resistors (R_p).

stable while SCL is high. This renders the start and stop conditions unique.

Data format: each 'chunk' of information sent via the I²C bus consists of eight data bits (one byte). Each byte is followed by a ninth clock pulse, during which the receiving IC (a master or a slave) is to generate an acknowledge (ACK) pulse. This is done by making the SDA line low during the ninth clock pulse.

IC address: each IC that is suitable for use on the I²C bus has its own, unique, address, which is determined by the manufacturer. In general, this address is 'burned' into the IC, although there are also ICs that allow a part of the address to be set externally. This option allows a number of ICs of the same type to be connected to a single bus. Address 00 is the 'general call address', to which all ICs on the bus respond.

R/W bit: every IC has a 7-bit address. The eighth bit (LSB) that is sent during the addressing operation, is the R/W (read/write) bit. If this bit is '1', a master device reads data from a slave device. When it is '0', a master device writes data to a slave device.

Bus protocol

A protocol has been drawn up to initiate the communication between two ICs on the I²C bus. This protocol is briefly described below.

As soon as the bus is free, a master can occupy it by generating a start condition. The first byte transmitted after the start condition contains the 7-bit IC address and the R/W bit. If the addressed IC is present (i.e., connected to the bus), it responds by returning an ACK pulse. After that, the data exchange may commence.

When the R/W pulse was '0' previously, the master sends data to the slave until it no longer receives ACK pulses, or until all data has been transmitted.

When the R/W pulse was '1' previously, the master generates clock pulses, during which the slave is allowed to send data. After every received byte, the master (which is a receiving device at this stage) generates an ACK pulse. This continues until the master no longer supplies ACK pulses.

The master can free the bus again by generating a stop condition. If it wants to continue communicating, it is possible that the master generates a new start condition instead of a stop condition. This new start condition is called a repeated start, and can be used to address a different IC, or make the R/W bit toggle. Figure 2 shows the start and stop conditions in a timing diagram.

Philips Components and a number of other manufacturers have a wide range of I²C compatible ICs available, including RAMs, EEPROMs, microcontrollers, I/O ports, DTMF encoders, infra-red transmitters and receivers, ADC and DACs, and a real-time clock with calendar (for an overview, see Ref. 1). The range of I²C devices is continuously expanded with new ICs.

The PCD8584 is an I²C bus controller specifically developed to simplify the communi-

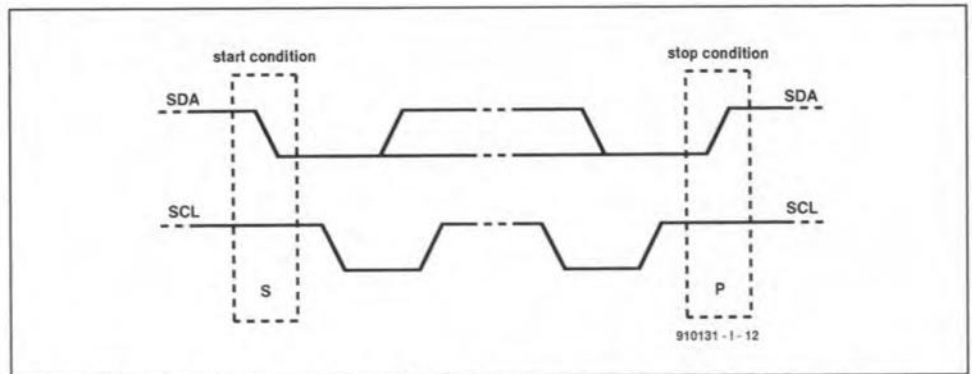


Fig. 2. Signal levels on the SDA and SCL lines mark start and stop conditions.

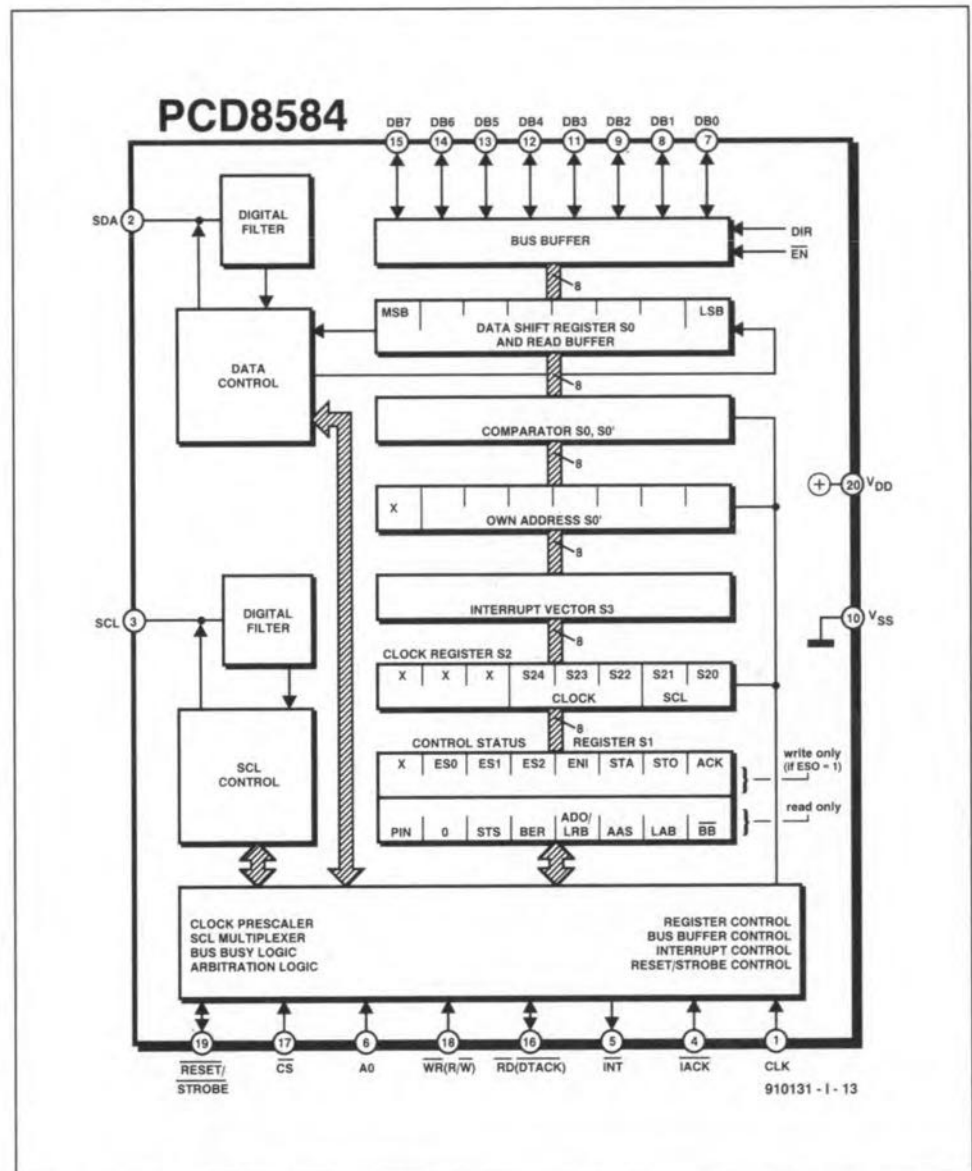


Fig. 3. The PCD8584 I²C controller is a quite complex IC — here, the block diagram is shown.

cation with a parallel port. This IC, which is at the heart of the present interface, arranges all control actions required on the bus, and thus allows ready communication between an I/O port of a PC, and I²C devices.

The PCD8584 I²C controller

The PCD8584 is a powerful, universal, I²C bus controller that forms the link between an 8-bit parallel port of a microcontroller or microprocessor, and the serial I²C bus. The IC supports reading and writing of bytes via

the I²C bus, and is remarkably simple to implement in systems based on a different processor types, such as the 8048/8051 controllers, 80xx processors, and the 68000.

The block diagram of the PCD8584, Fig. 3, shows a number of functional blocks: **Bus buffer.** This is the circuit between the computer bus and the shift register in the PCD8584.

S0', own address. In a multi-master system, this register contains the address to which the controller responds. This address may not be '00' unless the controller is to work in

PCD8584 I²C CONTROLLER

The information below is supplementary to that contained in the PCD8584 data-sheet, and aimed at those of you who intend to work on the device driver.

Initialization

After the circuit has been reset, the controller must be set to the correct mode (80xx or 68000). The default mode is 80xx compatible. On detecting a high-to-low transition at the \overline{WR} input, while \overline{CS} is high, the PCD8584 switches to 68000 mode. Next, a value other than 0 must be put into register S0'. Programming a 0 results in the IC switching to 'passive monitor mode' in which data on the I²C bus is indiscriminately put into the read buffer.

After the mode selection, the internal prescaler is programmed via register S2. This is done with the aid of bits S20 to S24. Table 1 shows the function of the bits in relation to the available clock frequencies.

The I²C interface card uses two addresses in the I/O range of the PC. All registers contained in the controller can be accessed via these two addresses. First, bits ES0, ES1 and ES2 in the 'S1: control' register must be programmed to give the desired settings of the I²C bus controller. The most useful settings are indicated with asterisks in Table 2.

Transmitting data

Before data can be sent via the I²C bus, this must be initialized, and have a state corresponding to a start condition. Bus initialization is achieved by making bit ES0 '1'. The interface can be switched off again by making this bit '0' again. The serial channel may be switched off only after a stop condition has been generated. If this rule is not observed, the controller loses track of the bus operations, and can be synchronized only by a hardware reset.

Before a start condition is generated, the controller finds out if the bus is free by checking if \overline{BB} is '1'. The start condition proper is not generated until the STA bit in the control register is set. Once this is so, an address is sent, complete with a R/ \overline{W} bit. The byte transmitted at this stage must be contained in register S0 when the STA bit is set. After transmitting the device address, the data exchange is initiated.

A repeated start condition is generated in a slightly different manner than a normal start condition. After the bus has been occupied, setting the STA bit has no effect. A start condition is generated, and data is transmitted from S0, only after data is sent to S0. A stop condition is necessary to mark the end of a transmission. This is achieved by setting the STO bit, and writing a value to S0. The actual value written to S0 is irrelevant, since it is ignored.

The transfer of data may commence after the start condition has been set up, and the address, complete with a R/ \overline{W} bit, has been sent. The moment the transmission starts, the PIN (pending interrupt) bit is set. This bit is reset automatically at the end of the transmission. Provided the ENI (enable interrupt) bit had been made '1' by the user, the \overline{INT} output is triggered. The

Table 1. PCD8584 SCL and clock frequency selection

SCL frequency			Internal clock frequency			
S21	S20	f _{SCL} (kHz)	S24	S23	S22	f _{CLK} (MHz)
0	0	90	0	x	x	3
0	1	45	1	0	0	4.43
1	0	11	1	0	1	6
1	1	1.5	1	1	0	8
			1	1	1	12

Table 2. Control/Status register S1

ES0 = 0: serial interface OFF				
A0	ES1	ES2	\overline{IACK}	Function
1	x	x	x	R/W S1: CONTROL *
0	0	0	x	R/W S0': (own address) *
0	0	1	x	R/W S3 (interrupt vector)
0	1	0	x	R/W S2 (clock register) *
ES0 = 1: serial interface ON				
A0	ES1	ES2	\overline{IACK}	Function
1	x	x	1	W S1: CONTROL *
1	x	x	1	R S1: STATUS *
0	x	0	1	R/W S0 (data) *
0	x	1	1	R/W S3 (interrupt vector)
x	0	x	0	R S3 (interrupt vector ACK cycle)
x	1	x	0	long-distance mode

interface card described here does not use the interrupt pulse. Instead, the control software 'polls' the PIN bit. Every polling action requires waiting for the PIN bit to revert to '0' again.

The state of the LRB (last received bit) indicates whether or not a slave has returned an acknowledge condition to indicate correct reception of the data. As soon as PIN is '0' again, the data to be transmitted may be put into register S0. Next, PIN becomes '1' again, the data is transmitted, and PIN becomes '0'. The LRB again indicates that an acknowledge has been generated, which means the data has arrived securely at the slave address. This sequence is repeated as long as the master sends data. The end of the sequence may be marked by a stop condition to indicate that the bus is free, and accessible to other devices. Since the end of the stop condition can not be detected by looking at the PIN bit, the software must wait until \overline{BB} reverts to '1'.

When receiving data, the master must generate an acknowledge condition (i.e., set the ACK bit) after every received byte. The addressing of the master is similar to the transmitting of data. After addressing,

the PIN bit is '0', so that the IC functions as a master/receiver. The controller keeps the SCL line low until the PIN bit is set by a read operation to register S0. This is an indication for the slave/transmitter that the master is ready to receive data. This means that S0 must be read once before the reception of data can commence. The value read from S0 to initiate reception has no significance. The IC subsequently awaits data, generates an acknowledge, and resets the PIN bit. After that, the data may be read from register S0, which resets the PIN, and prepares the IC for reception of the next byte.

Once the desired data has been received, the ACK bit must be reset before S0 can be read. The slave will send one more byte, of which the reception is not acknowledged by the master. After that, the communication may be ended with the aid of a stop condition. The consequence of this protocol is that the slave will transmit at least one byte too many.

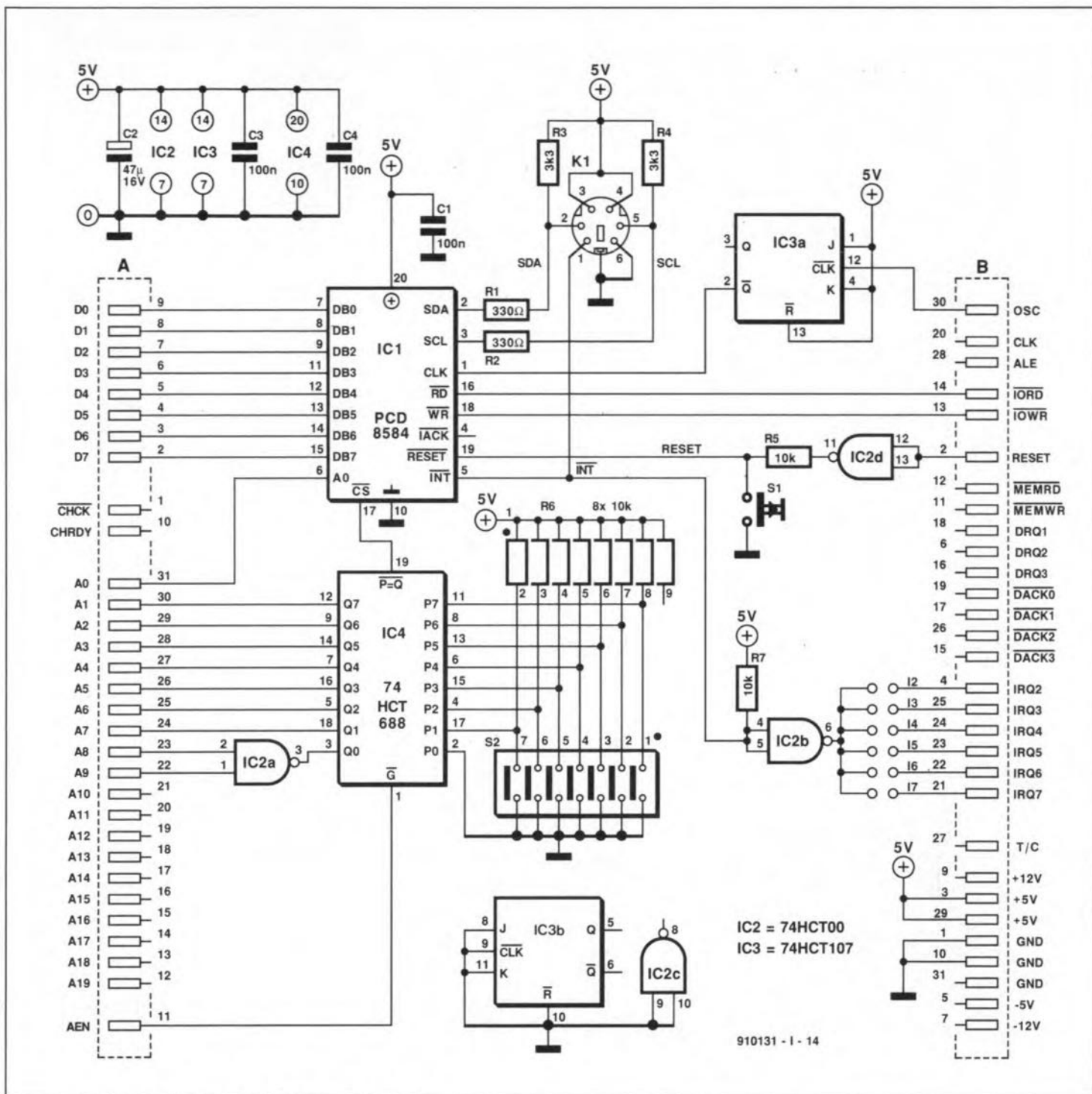


Fig. 4. Circuit diagram of the I²C interface for PCs.

its (mostly passive) 'monitor mode'. On the interface card, this register is of no importance because the PCD8584 is the only master device on the I²C bus, and thus automatically assumes the function of co-ordinating all bus actions.

S1, control/status register. This register is available double. It is addressed when pin A0 is high. All other registers are addressed when pin A0 is low. The selection of the latter depends on bits ES0, ES1 and ES2 in the register 'S1: control'. The selection between control and status register is effected with the aid of the ES0 bit: as long as this is '0', only the control register is accessible for reading from or writing to. Also, the serial interface is switched off. When ES0 is '1', 'S1: control' is written to, 'S1: status' is read from, and the serial interface is actuated. The

remaining bits are not involved until they are significant for the control software.

S2, clock register. The clock pulses on the SCL line are derived from the signal at the CLK input. Bits S20 and S21 in register S2 allow one of four bus clock frequencies (SCL) to be selected: 1.5 kHz, 11 kHz, 45 kHz or 90 kHz. The bits S22, S23 and S24 are used to select the frequency at the clock input of the IC: 3 MHz, 4.43 MHz, 6 MHz, 8 MHz or 12 MHz. The latter is the default after a reset.

S3, interrupt vector. When the controller is used on interrupt basis, it is capable of putting an interrupt vector on the PC bus. This happens when input IACK is made low. Since this function is not used on the PC interface card, S3 has no function in the present application.

Around the controller

In addition to the PCD8584, three other IC and a handful of passive components are required to build the PC interface (see the circuit diagram in Fig. 4).

Circuits IC4 and IC2A form the address decoder, which compares the address on the PC bus with that set on switch block S2. When the two addresses match, the CS line to the PCD8584 is pulled low, so that a read or write command in the I/O range finds its way to the I²C controller. Address signal A0 goes directly to the IC, and needs no further treatment. Since in this case the PC bus is loaded with only one IC input, additional buffering is not required. The IORD and IOWR lines are connected directly to the RD and WR inputs of the IC. The interrupt out-

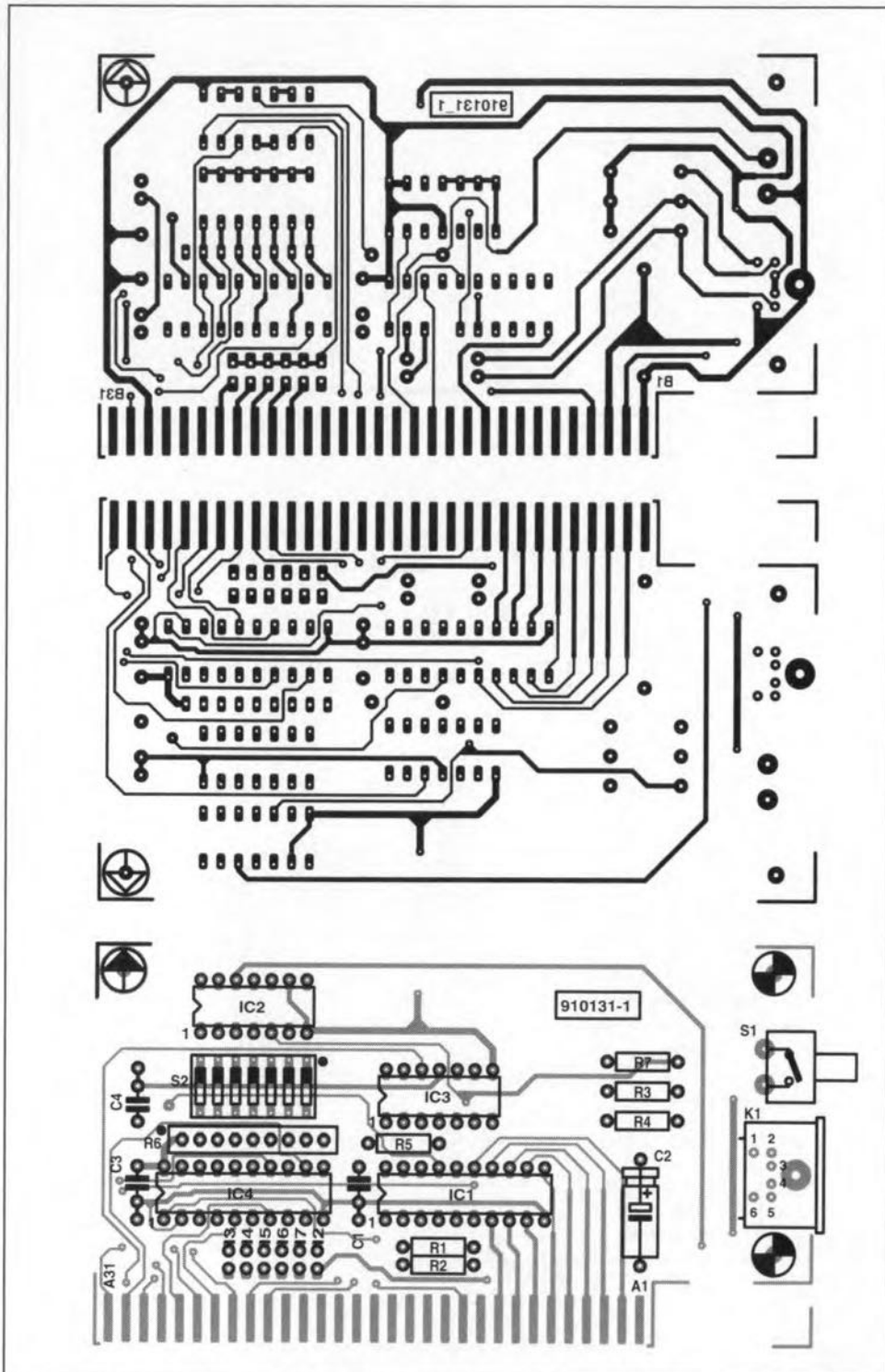


Fig. 5. The printed circuit board for the interface is double-sided and through-plated.

put signal is inverted and buffered by IC2b. Pull-up resistor R7 and connector K1 allow the (open drain) INT outputs of further I²C ICs to be taken up into an interrupt chain with an OR function. A jumper is used to select the interrupt line in the PC. The jumper is not fitted unless interrupts are required (the control software described further on does not use interrupts generated by the interface).

The output of gate IC2d is made short-circuit resistant by resistor R5. This is done to allow the circuit to be reset by the PC as well as manually with the aid of a push-button, S1. This is particularly useful while experimenting, since a soft reset (CTRL-ALT-DEL) on the PC does not actuate the reset line on the bus expansion slot.

Bistable IC3a divides the 14.3-MHz bus clock signal by two. This provides an output clock of 7.16 MHz, which can be handled without problems by the controller when programmed to operate in 8-MHz mode.

Connector K1 is a mini-DIN type on which the SDA, SCL and INT lines are available, and, of course, ground and +5 V. This allows ready connection of the present interface to further (experimental) I²C systems. Resistors R1, R2, R3 and R4 give SCL and SDA the optimum line impedance.

Control software

The control software developed for the present interface will be discussed in detail in relation to our first I²C application circuit, an

COMPONENTS LIST

Resistors:

2	330Ω	R1;R2
2	3kΩ3	R3;R4
2	10kΩ	R5;R7
1	8-way 10kΩ array	R6

Capacitors:

3	100nF	C1;C3;C4
1	47μF 16V	C2

Semiconductors:

1	PCD8584	IC1
1	74HCT00	IC2
1	74HCT107	IC3
1	74HCT688	IC4

Miscellaneous:

1	Push-to-make button with angled terminals	S1
1	7-way DIL switch	S2
1	6-way PCB-mount mini DIN socket	K1
1	Printed circuit board	910131-1

A/D-D/A converter, to be published next month. The software enables the I²C interface to be controlled just as 'easily' as, say, a printer or a graphics card. The driver written for the I²C interface relieves the user of observing bus protocols—all this can be left safely to the software and the bus controller. Interestingly for the programmers among you, the I²C interface driver is supplied with a commented source listing.

Construction

The printed-circuit board (Fig. 5) is a double-sided, through-plated type. The ready-made PCB supplied through the Readers Services has gold-plated PC slot fingers. The construction is straightforward as only relatively few parts are used. As customary with PC extension cards, the board is secured to a support bracket. This is drilled such that the I²C extension connector and the reset switch are accessible on the rear panel of the PC. As it will seldom be required to reset the I²C controller, you may want to omit S1.

The mini-DIN socket carries all the signals required to connect further I²C ICs to the bus.

The DIP switches contained in S2 are set to the desired I/O base address, which lies in the range from 300_H to 3FE_H. Note that the range 300_H to 31F_H is actually reserved for experimental cards. Other ranges may be used, however: for instance, 3D0_H to 3DF_H if you do not have a CGA card in your PC. When the I²C control software is loaded, the interface address is communicated to the driver proper. This arrangement should enable a suitable address block to be found for the card in almost any PC. ■

Reference:

1. "Inter-IC communications", *Elektor Electronics* September 1990.

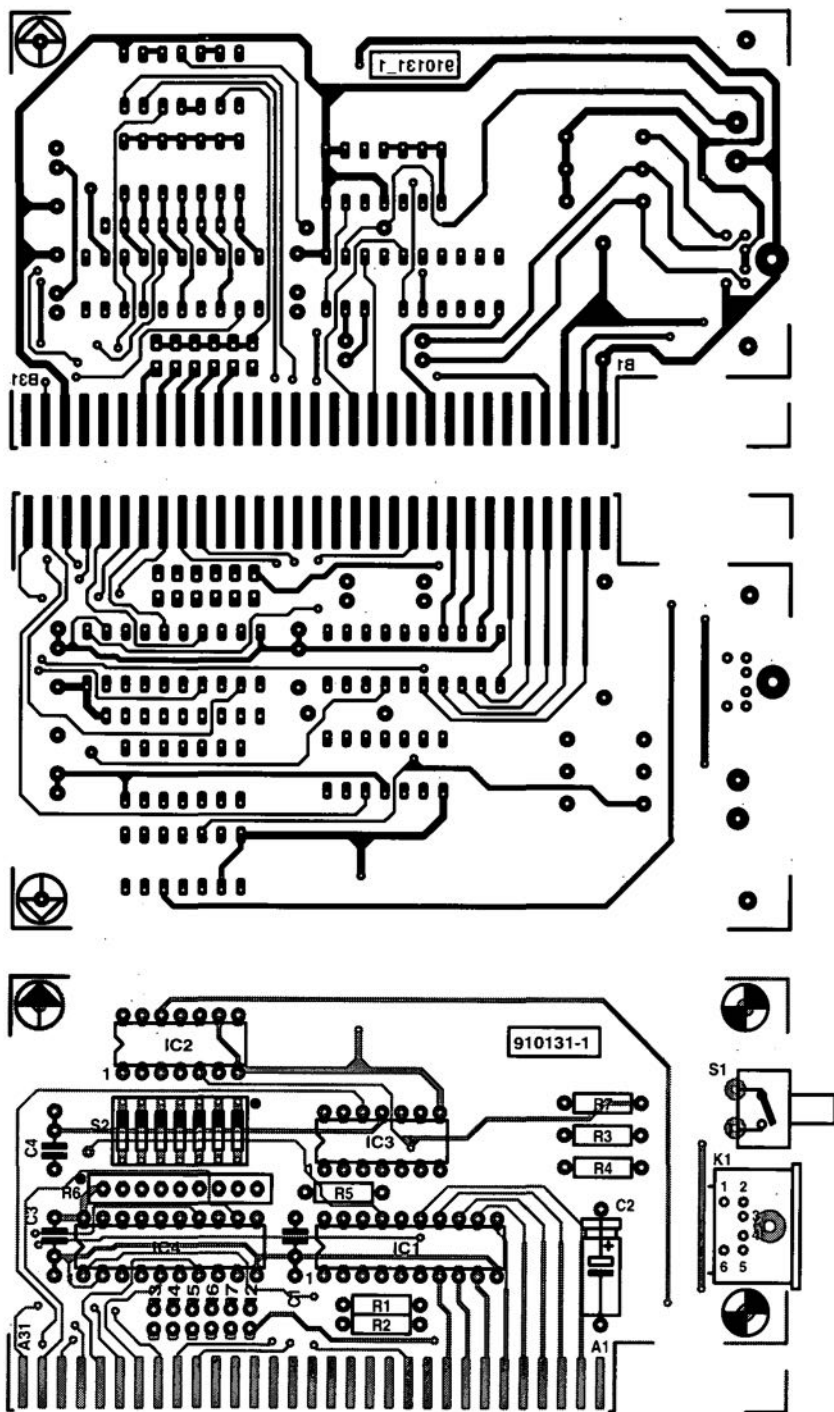


Fig. 5. The printed circuit board for the interface is double-sided and through-plated.

COMPONENTS LIST

Resistors:

2	330Ω	R1;R2
2	3kΩ	R3;R4
2	10kΩ	R5;R7
1	8-way 10kΩ array	R6

Capacitors:

3	100nF	C1;C3;C4
1	47μF 16V	C2

Semiconductors:

1	PCD8584	IC1
1	74HCT00	IC2
1	74HCT107	IC3
1	74HCT688	IC4

Miscellaneous:

1	Push-to-make button with angled terminals	S1
1	7-way DIL switch	S2
1	6-way PCB-mount mini DIN socket	K1
1	Printed circuit board	910131-1

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The mini-DIN socket carries all the signals required to connect further I²C ICs to the

I²C interface for PCs

February 1992

The PCD8584 used in this project is no longer manufactured by Philips Semiconductors, and replaced by the PCF8584. This is a fully compatible IC and only improved as regards the 4-wire long-distance mode, which did not work correctly on the PCD8584.

Real-time clock for 80C32 computer

June 1993

Contrary to what is implied by the description of the parallel connection of the SmartWatch IC pins with the EPROM pins, pin 1 of the SmartWatch

CORRECTIONS AND UPDATES

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

VHF-low converter

June 1993

The parts list should be corrected to

read:

1	2 μ H2	L3
1	0 μ H1	L5

The circuit diagram is correct. The sub-1 μ H chokes used in this project are available from, among others, Cricklewood Electronics.

1.2 GHz multifunction frequency meter

December 1992

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

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.

INDOOR/OUTDOOR TEMPERATURE INDICATOR

by L. Lemon

The circuit presented here is a dual thermometer that indicates the difference between the temperature indoors and out of doors with the aid of two LEDs.

COMBINED thermometers for domestic use appear to be much in vogue these days, and this article looks at the operation and construction of a simplified version of such a unit.

An electronic thermometer is usually a combination of a temperature sensor and a read-out. The sensor and the associated amplifier are easily designed and built based on, say, an NTC (negative temperature coefficient) resistor. Unfortunately, the read-out is a different kettle of fish as it is more complex to design and build. But that is not all: to start with, we are faced with the choice of the type indication: a calibrated moving-coil meter, a LED bar or a pair of 7-segment LED displays. Whichever is used, it has to be implemented double, powered, built, and controlled accurately.

Interestingly, many owners of an indoor/outdoor thermometer seem to find the difference between the two measured values more important than the actual values. In other words: they want to know the effect of opening a window: will this bring in warm or cold air?

What comes to mind is, of course, to omit an absolute readout altogether, and instead create a simple difference indicator. That has been done here, and it allows us to keep the circuit uncluttered and inexpensive.

A difference indicator

The circuit diagram in Fig. 1 shows that the indication is based on two LEDs (light emitting diodes), D1 and D2. When the red LED lights, it is colder outside than inside. The green LED indicates the other condition. Depending on the country you live in, the sea-



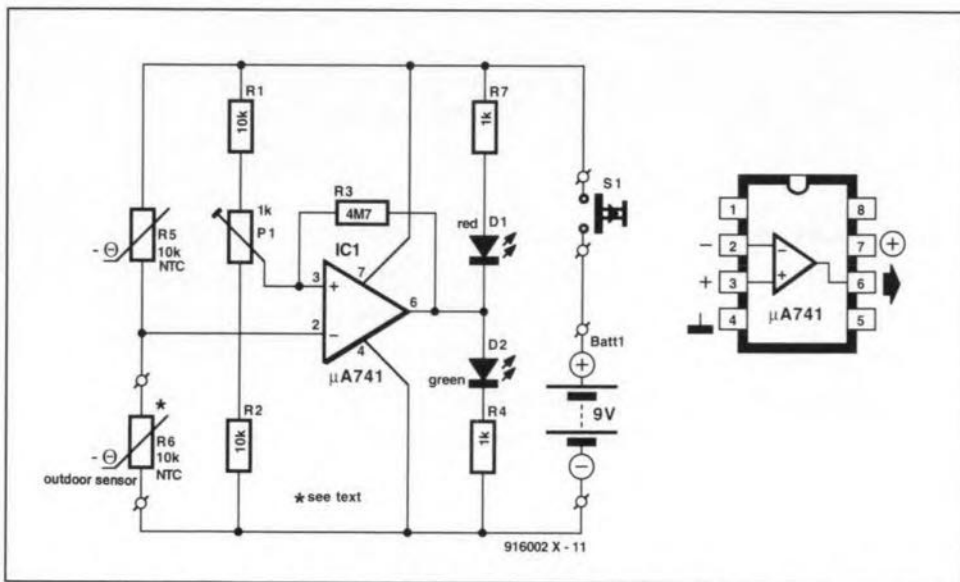


Fig. 1. Circuit diagram of the temperature difference indicator, and pinout of the 741 opamp.

son, and other climatic conditions, you can decide whether or not it makes sense to open a window.

The circuit has four basic elements: a pair of temperature sensors, R5 and R6, a reference voltage source, R1-R2-P1, a comparator, IC1, and a pair of LEDs, D1 and D2, that form the readout.

Here, the familiar and inexpensive Type 741 opamp is used for the comparator function. The 741 works as a comparator because feedback is virtually absent. This means that all voltage differences, however small, between the non-inverting and the inverting input of the device are amplified by a very large factor. Consequently, any voltage difference will cause the opamp output to swing to the positive or negative supply level, depending on whether the input voltage is higher or lower than the reference voltage, respectively. This comparator function is exactly what we require.

Resistor R3 provides some hysteresis, which is necessary to prevent the circuit from oscillating. In practice, a small difference between the voltage at the +input and the -input of the opamp will cause the output to toggle, i.e., swing to the other level. The function of the hysteresis is to prevent this happening at an infinitely small voltage

difference, which will always exist in practice. The upshot is that the opamp output will not toggle until the input voltage difference is large enough to 'overcome' a certain minimum value defined by the 4.7-M Ω feedback resistor.

Once the operation of the comparator is understood, the rest of the circuit is plain sailing. The +input of the opamp is held at about half the supply voltage by a voltage divider, R1-P1-R2. The exact value of this reference potential can be adjusted with the aid of the preset, P1. The variable voltage supplied by the temperature sensors is applied to the -input of the opamp. If this voltage is smaller than that at the +input, the output of IC1 swings to the 'high' level, and the green LED, D2, lights. Conversely, when the voltage exceeds the threshold set with P1, the opamp output is pulled low. Consequently, the green LED goes out and the red LED, D1, lights.

The level of the direct voltage applied to the -input of IC1 is determined by the ratio of the values of R5 and R6, which form a voltage divider. R5 and R6 are identical NTCs, and function as the indoor and outdoor temperature sensor respectively. So long as they measure equal temperatures, their resistance will be roughly equal, and the voltage at

their junction about half the supply voltage. Assuming that the indoor temperature is higher than the outdoor temperature, the resistance of R6 will be smaller than that of R5, since the resistance change in an NTC is inversely proportional to the ambient temperature. Consequently, the voltage at the NTC junction drops below the reference voltage of IC1, so that the green LED lights. Conversely, when R5 measures a higher temperature than R6, the voltage at the NTC junction is higher than the reference, so that the opamp output goes low, and the red

COMPONENTS LIST

Resistors:

2	10k Ω	R1;R2
1	4M Ω	R3
2	1k Ω	R4;R7
2	10k Ω NTC	R5;R6
1	1k Ω preset H	P1

Semiconductors:

1	red LED	D1
1	green LED	D2
1	741	IC1

Miscellaneous:

1	push-to-make button	S1
1	9-V battery	Batt1
1	Printed circuit board	UPBS-1

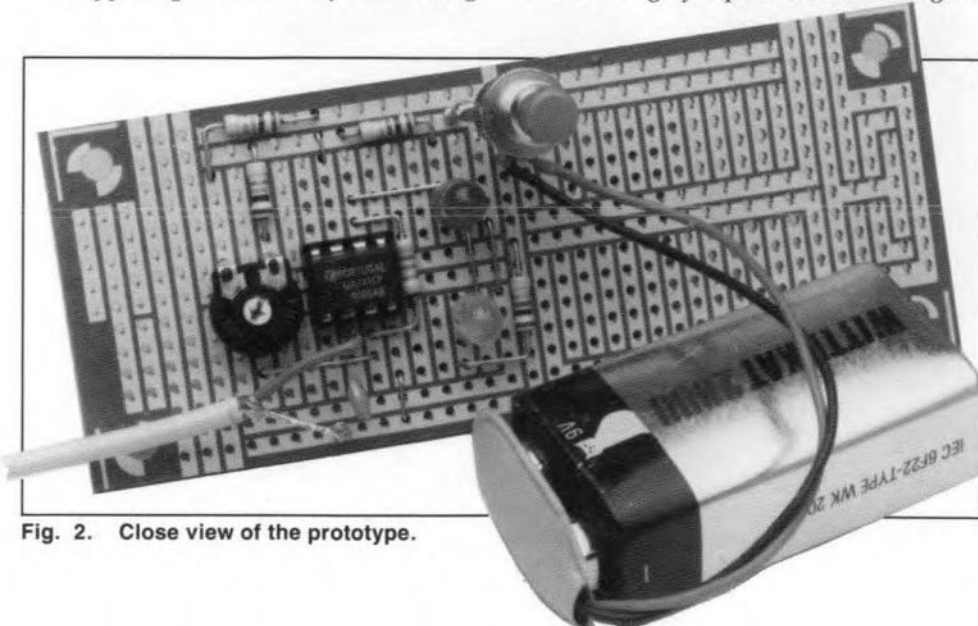


Fig. 2. Close view of the prototype.

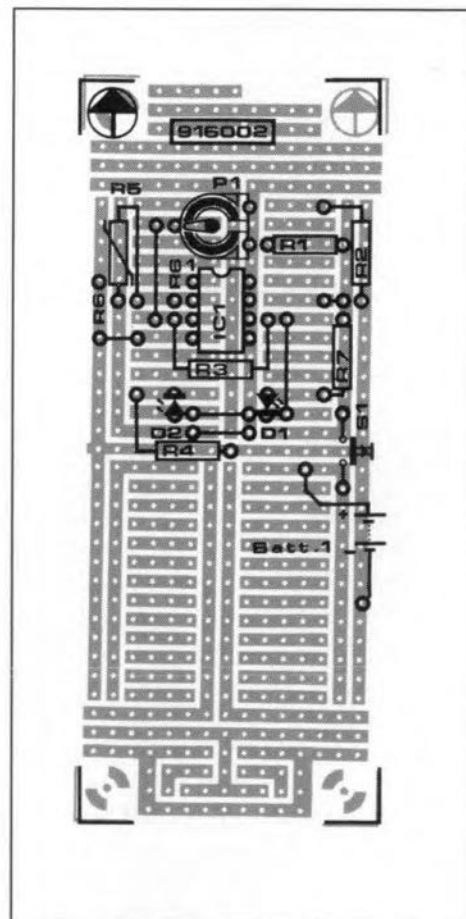


Fig. 3. Suggested component layout on prototyping board size - 1. Note that the outdoor NTC is connected via a screened cable.

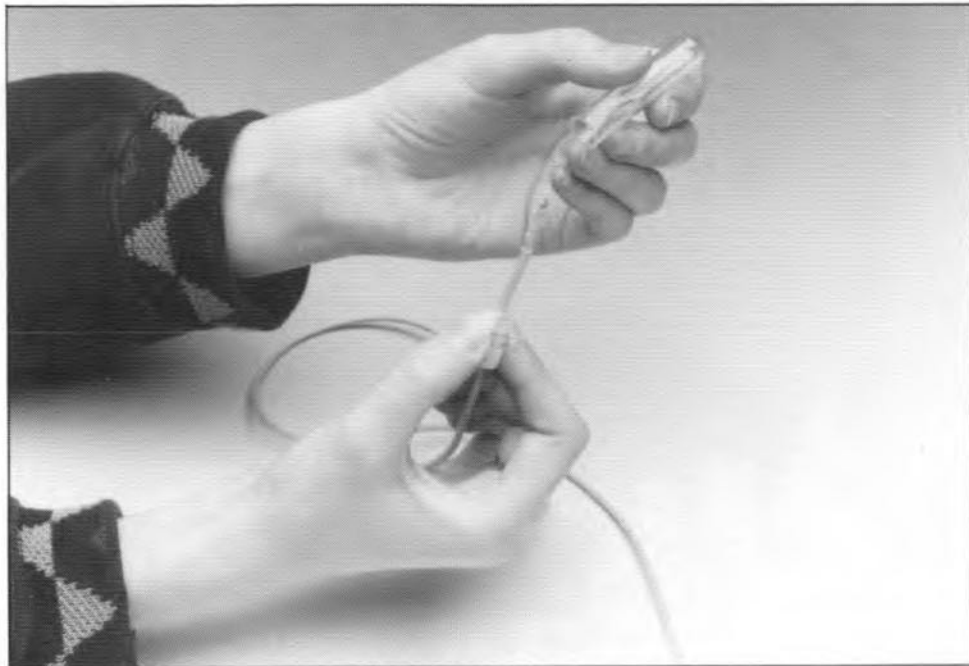


Fig. 4. The outdoor NTC, R6, is fitted in a water-resistant enclosure.

LED, D1, lights.

Resistors R4 and R7 limit the current through the indication LEDs to safe values. The circuit is actuated by pressing push-button S1. The current consumption is so low that a 9-V battery will have sufficient capacity for several tens of hours of use (remem-

ber, the circuit is on only as long as the push-button is pressed to 'take a reading').

Construction

The circuit is built on a universal prototyping board size-1 (UPBS-1), which is available

ready-made through the Readers Services. Figure 2 shows a suggested arrangement of the parts on this board. Construction is not critical, and can be completed within an hour or so. The indoor sensor, R5, is mounted outside the enclosure to enable it to record ambient temperature variations. The outdoor sensor is fitted in a water-resistant enclosure, and connected to the circuit via a length of single-core screened cable (microphone cable is suitable).

The photograph in Fig. 3 shows how the outdoor sensor is fitted into a small transparent plastic tube. The connecting cable is passed through the rubber plug, and sealed with the aid of a small piece of heat-shrink sleeving.

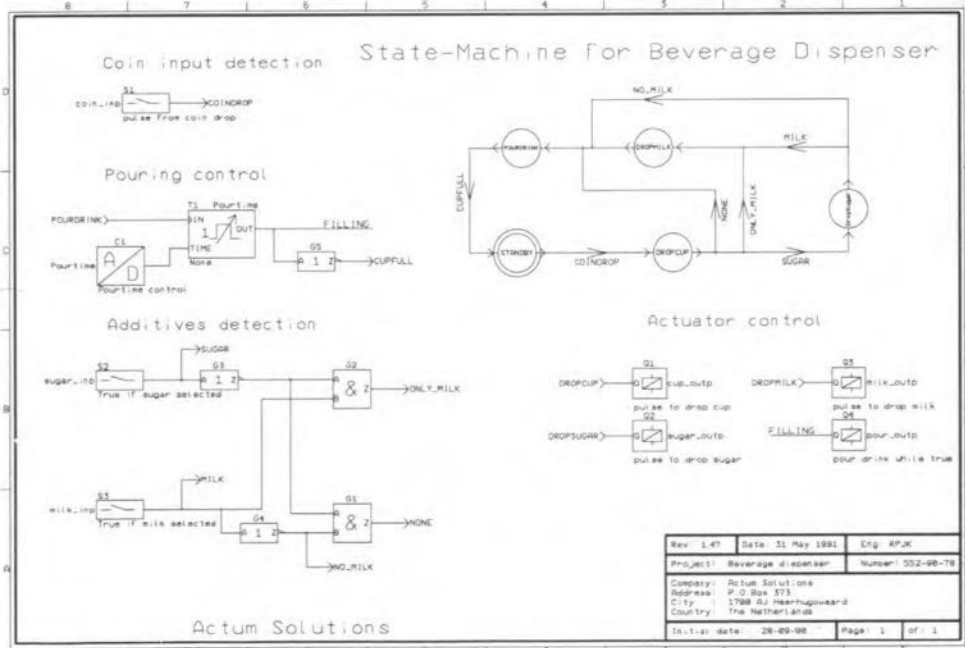
Adjustment

The temperature indicator is simple to adjust. Put the outdoor sensor as close as possible to the indoor sensor, and wait a few minutes to make sure the NTCs are at the same temperature. Hold S1 pressed, and adjust the preset until a point is found where it just switches the LEDs on and off. Next, check that the green LED lights when R6 is heated (use a hair dryer or your soldering iron). Similarly, the red LED should come on when R5 is heated. ■

THE REALIZER — A CASE TOOL FOR REAL-WORLD APPLICATIONS

green
nummer

The Realizer from Actum Solutions is a fully graphics oriented program for IBM ATs and compatibles, that enables embedded systems engineers and industrial automation engineers to convert a schematic diagram into software. A block diagram drawn with the aid of The Realizer consists of standardized industrial control symbols. The Realizer uses this diagram to produce an intermediate piece of software. Microcontrollers, programming languages and PLCs supported by The Realizer include the 68HC05, Siemens PLCs, the 8051, ISO Pascal, ST62 and Hitachi PLC. After generating a program, The Realizer enables you to analyze the behaviour of the block diagram with the aid of a function simulator. After simulation, a postprocessor is started to refine the intermediate software for a specific hardware environment. Interestingly, the program intelligently allocates the available (chip-internal) resources like RAM and I/O. Even multifunction pins of microcontrollers are supported. Apart from standard control-job related symbols, The Realizer supports the use of symbols like timers, counters, decoders and encoders. For higher levels of functionality



an implementation of state-machine (sequencer) design is provided.

For further information on The Realizer, contact

**Actum Solutions • P.O. Box 373 •
1700 AJ Heerhugowaard • Holland.
Telephone: +31 2207 45553. Fax: +31
2207 12457.**

WHY USE LITHIUM CELLS?

by Bill Higgins

MANY electronics designers give little or no thought during their work to the various types of battery now available. Battery manufacturers urge designers to consult them at an early stage of the design, so that appropriate consideration can be given to the battery supply.

One of the newer types of battery, the lithium battery, should be given serious consideration by electronics designers, since these batteries have a number of advantages over other types. These advantages include high energy density, higher cell voltage, lower weight, long shelf life, and wide temperature range of operation. Moreover, lithium batteries are environment-friendly, because they use only small amounts of natural resources to provide energy and, once discharged, the cells are not toxic.

What is lithium?

Lithium, chemical symbol Li, is a silvery white, tough metal with atomic number 3 and a r.a.m. (relative atomic mass) of 6.94. With a relative density (formerly specific gravity) of 0.534, it is the lightest solid known to man to date. It has a melting point of 186 °C, boils at 1360 °C and is highly reactive. The potential electrochemical energy available from it is 3862 Ah kg⁻¹.

The name lithium derives from the Greek lithos, which means stone. Its main use has been in medicine (anti-depressants) and the manufacture of lubricant grease with high resistance to moisture and extremes of temperature.

History of lithium cells

After World War 2, the theoretical fundamentals of lithium cells were established in various laboratories.

- 1949 First patent on lithium cells (Hajek).
- 1957 Second patent: lithium sulphur dioxide (Herbert-Ulam).
- 1960 First laboratory tests (Lockheed-ESB).
- 1974-75 Start of industrial production of cylindrical cells by a number of manufacturers around the world.
- 1978 Start of industrial production of button cells by several manufacturers around the world.

Types of lithium cells

As lithium is highly reactive, it can couple with many different materials. So far, manufacturers have come up with four main types of cell:

- (1) lithium manganese dioxide -- LiMnO₂;
- (2) lithium sulphur dioxide -- LiSO₂;
- (3) lithium thionyl chloride -- LiSOCl₂;

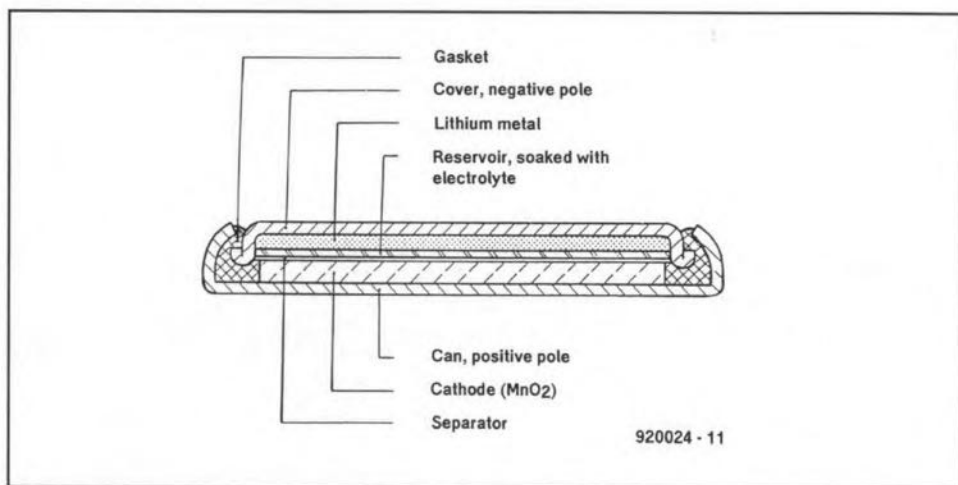


Fig. 1. Cross-section through a flat LiMnO₂ button cell with a large diameter.

Type no.	Capacity (mAh)	Weight (gm)	O/Dia (mm)	Height (mm)
CR927	30	0.48	9.5	2.7
CR1220	35	0.8	12.5	2.0
CR1225	42	0.9	12.5	2.5
CR1620	60	1.2	16.0	2.0
CR2016	75	1.7	20.0	1.6
CR2025	140	2.5	20.0	2.5
CR2032	190	3.0	20.0	3.2
CR2320	120	2.5	23.0	2.0
CR2430	270	3.9	24.5	3.0
CR2450	520	5.7	24.7	5.1
CR2477	1000	8.4	24.7	7.7

Table 1. Typical range of 3 V button LiMnO₂ cells (Renata SA). Lithium manganese dioxide (LiMnO₂) batteries are available in a wide variety of different shapes and sizes. They have a typical shelf life of six years and are suitable for operation over a temperature range of -40 °C to +70 °C. The maximum e.m.f. is 3 V; operation is from 3 V down to 1.4 V. Their capacity ranges from 0.03 Ah to 1.25 Ah. The Ah to mass ratio is typically 0.078. Uses include rescue beacon equipment and life jackets.

Type no.	Capacity (Ah)	Weight (gm)	Dia (mm)	Height (mm)	Volume (cc)
G04	0.45	8	14.2	27.9	4.4
G32	0.9	12	16.3	34.5	7.2
G06	1.0	15	14.2	50.3	8.0
G36	1.7	18	16.3	57.7	12.0
G52	3.3	45	25.6	49.5	25.6
G54	4.3	56	25.6	60.2	31.1
G20	7.75	85	33.3	61.0	53.0
G70	9.5	99	38.6	52.3	61.2
G56	10.5	110	25.6	130.3	67.3
G22	16.5	175	33.3	120.6	104.8
G60	25.0	244	41.7	118.4	161.2
G62	33.0	300	41.7	141.2	192.4
G72	3.9	54	38.6	29.0	33.9

Table 2. Typical range of 3 V cylindrical LiSO₂ cells (Crompton Eternacell). Lithium sulphur dioxide (LiSO₂) batteries are produced in cylindrical and disc packaging. Typical data are enumerated above. These batteries have a shelf life of typically 10 years and operate over a temperature range of -55 °C to +70 °C. The maximum e.m.f. is 2.9 V; operation is from 2.9 V down to 2 V. Their capacity ranges from 0.85 Ah to 8 Ah. The Ah to mass ratio is typically 0.0842. A physical limitation is that external pressure can inhibit their correct functioning.

Type no.	Capacity (Ah)	Weight (gm)	Dia (mm)	Height (mm)
T04	0.85	8.72	14.5	25.2
T06	1.5	16.34	14.5	50.5
T32	1.9	12.87	16.6	33.1
T52	6.0	53.0	26.2	50.0
T20	14.0	105	33.5	61.5

Table 3. Typical range of 3.7 V cylindrical LiSOCl₂ cells (Crompton Eternacell). Lithium thionyl chloride (LiSOCl₂) batteries are available in cylindrical and rectangular housing. Their chemical reaction is $4\text{Li} + 2\text{SOCl}_2 \rightarrow 4\text{LiCl} + \text{SO}_2 + \text{S}$. They have a typical shelf life of 10 years and are suitable for operation over a temperature range of -55 °C to +70 °C. At exhaustion of the battery, it becomes pressurized with SO₂. In certain conditions, the forced reaction between lithium and sulphur can be explosive! This is one of the reasons that extreme caution must always be taken in the use and transport of these batteries. The e.m.f. is 3.7 V and capacity ranges from 0.85 Ah to 14 Ah. The Ah to mass ratio is typically 0.0758. The main use of LiSOCl₂ batteries is in military applications.

Type no.	Capacity (mAh)	Weight (gm)	Dia (mm)	Height (mm)
BR2025	140	2.0	20	2.5
BR2032	190	2.5	20	3.2
BR2320	112	2.5	23	2.0
BR2325	170	2.9	23	2.5
BR2330	260	3.1	23	3.0

Table 4. Typical range of 3 V coin LiCFx cells (Crompton Eternacell) Lithium polycarbonmonofluoride (LiCFx) batteries are produced primarily in disc packaging. They have a shelf life of 10 years and operate over the temperature range -40 °C to +85 °C. They are widely used in industrial and consumer products, especially memory back-up applications. The e.m.f. is 3.3 V and operation ranges from 3.3 V down to 2 V. The capacity ranges from 0.112 Ah to 0.26 Ah. The Ah to mass ratio is typically 0.07.

- (4) lithium polycarbonmonofluoride - LiCFx.

All these are primary cells; since lithium is highly reactive, it cannot be used in rechargeable cells.

Advantages of lithium

Compared with other primary batteries, lithium types offer higher energy density, lower weight, and higher cell voltage. Single cells have an e.m.f. of about 3 V, which, in devices designed to operate from 3 V, can mean a useful saving in space and weight.

Wide operating and storage temperature ranges make lithium batteries prime contenders in applications of extreme cold or heat. Long shelf lives make them useful in standby power applications, for instance, memory storage back-up packs and search and rescue equipment.

Extensive testing procedures by The Underwriters Laboratories at Northbrook, Ill., in the USA, enable manufacturers to receive appropriate recognition for their cells and batteries. The Underwriters Laboratories require that lithium cells and batteries pass a number of abusive environmental, mechanical and electrical conditions.

Disadvantages of using lithium

As lithium reacts with water, it has to be handled with great care. Manufacturers of lithium cells and batteries have constructed special 'dry areas' where the appropriate careful handling can take place. International regulations apply to the transportation of cells with more than 0.5 gm of lithium.

Manufacturers of cells and batteries have had to include adequate venting and internal fuses as safety features in these products. ■

The assistance of the following is gratefully acknowledged.

Renata SA (Switzerland)
Crompton Eternacell Ltd
Duracell Batteries Ltd

APPLICATION NOTE

The content of this note is based on information received from manufacturers in the electrical and electronics industries and does not imply practical experience by *Elektor Electronics* or its consultants.

Dual-output, switch-mode regulator from Maxim (+5 V to ± 15 V or ± 12 V)

THE MAX743 DC-DC converter IC contains all the active circuitry needed to build small, dual-output power supplies. Relying on simple two-terminal inductors rather than transformers, the MAX743 regulates both outputs independently to within $\pm 4\%$ over all conditions of the voltage, temperature and load current.

The MAX743 typically provides 75–82% efficiency over most of the load range. It operates with current-mode feedback at 200 kHz, so it can be used with small, lightweight external components. Also, ripple and noise are easy to filter.

The MAX743 is inherently reliable because of its internal power transistors and monolithic construction. Thermal shutdown prevents overheating, and cycle-by-cycle current sensing protects the power-switch transistors. Other features include overvoltage lock-out and programmable soft-start.

Inductors, capacitors and diodes to complement the MAX743 to construct a complete power supply are available as a kit from Maxim—see Fig. 4. An evaluation kit for use by designers to verify all aspects of performance is also available.

Operating principle

Each current-mode controller consists of a summing amplifier that adds three signals: the current waveform from the power switch FET, a $V_{OUT}-V_{REF}$ error signal, and a ramp signal for AC compensation generated by the oscillator—see Fig. 2. The output of the summing amplifier gates a bistable (flip-flop), which in turn drives the power FET switch.

Both switches are synchronized to the oscillator and turn on simultaneously when the bistables are set. The transistors turn off individually when their switch currents reach a trip threshold determined by the $V_{OUT}-V_{REF}$ error signal. This creates a duty-factor modulated pulse train at 200 kHz, where the on-time is proportional to both the output voltage error signal and the peak inductor current. Low peak currents or high output voltage error signals result in a high duty ratio of up to 90%.

The MAX743 oscillator frequency is generated without external components and is not adjustable. The device is laser-trimmed to

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
Efficiency	$I_{LOAD} = \pm 50\text{mA}$	LX = MAXL001	79		%
		LX = MPP Toroid	82		
Supply Current (Note 2)			20	30	mA
Standby Current	$V_{REF} = +5\text{V}$, includes V_{REF} current		2.2	4	mA
Line Regulation	$V_+ = +4.5\text{V}$ to $+5.5\text{V}$			0.05	%/%
Load Regulation	$I_{LOAD} = 0$ to 100mA			1	%
Reference Voltage	V_{REF}		2.0		V
Oscillator Frequency		170	200	230	kHz
Undervoltage Lock-Out	Measured at V_+	3.8		4.2	V
Thermal Shutdown Threshold			+190		$^{\circ}\text{C}$
LX+ On Resistance (Note 3)			1.2	3.0	Ω
LX- On Resistance (Note 3)			1.0	2.2	Ω
LX+ Leakage Current (Note 4)	$LX_+ = +17\text{V}$, $V_+ = 6\text{V}$			100	μA
LX- Leakage Current (Note 4)	$LX_- = -17\text{V}$, $V_+ = 6\text{V}$			-100	μA
Compensation Pin Impedance	CC+, CC-		10		k Ω
Soft-Start Source Current	SS = 0V	3.0		7.0	μA
Soft-Start Sink Current	SS = 2V, $V_+ = 3.8\text{V}$	0.5	2.0		mA

Note 1: All devices tested to full-load conditions with 50ms pulsed loads using automatic test equipment. In continuous operation, the maximum allowable output current is determined by package thermal characteristics and passive component ratings.
Note 2: Total supply current including inductor current. The worst case for supply current occurs at low input voltage.
Note 3: Guaranteed by design, not 100% tested. Output currents are 100% tested.
Note 4: Tested at wafer level only, not in packaged form.

Table 1. Electrical characteristics.

DESIGN REQUIREMENT				
LOW COST, LOW NOISE	LOWEST COST	WIDE TEMP RANGE	MINIATURE	SURFACE MOUNT
INDUCTOR				
Iron-Powder Toroid	Ferrite Bobbin	Many Types OK	Hi-Flux (MPP) Toroid	Hi-Flux (MPP) Toroid
* 0.6" Diameter * Low EMI * 79% Efficiency	* 0.6" Long * Some Noise * 82% Efficiency		* 0.28" Diameter * Low EMI * 80% Efficiency	* Surface Mount Version
MAXL001 Sold by Maxim G # 51-548	C # 7070-25 G # GA10-103K I # LS4-100	All Listed Inductors	Gowanda Corp. Part # 51-559	Gowanda Corp. Part # 51-560
CAPACITOR				
150 μF Low-ESR Aluminum Electrolytic	220 μF Standard Aluminum Electrolytic	Solid Tantalum (Sintered Anode)	47 μF Tantalum// 4.7 μF Ceramic	33 μF Tantalum// 33 μF Tantalum
* -25 $^{\circ}\text{C}$ to +105 $^{\circ}\text{C}$ * Lowers Ripple * Radial Mount	* 0 $^{\circ}\text{C}$ to +85 $^{\circ}\text{C}$ * 16V or up	* -55 $^{\circ}\text{C}$ to +125 $^{\circ}\text{C}$ * Lowers Ripple	* 0 $^{\circ}\text{C}$ to +70 $^{\circ}\text{C}$ * May Need Additional Compensation	* 0 $^{\circ}\text{C}$ to +70 $^{\circ}\text{C}$
MAXC001 Sold by Maxim Nichicon PL Series	Multiple Suppliers	THF Series Mallory	RPE230 Series Murata Erie	267 Series Matsuo
RECTIFIER				
1N5817	1N5817	1N5802 (UES1102)	MBR030 Motorola	PRLL5817 AmpereX/Phillips
PRINTED CIRCUIT LAYOUT				
See Figure 6	See UM-3 MAX742/743 Application Notes	Modify Layout of Figure 6	Modify Layout of Figure 6	See UM-3 MAX742/743 Application Notes

Table 2. Component design chart.

200 kHz at wafer level, resulting in clean, predictable operation.

Basic circuit

The circuit in Fig. 3 has been tested at all line, load, output current and temperature limits. Guaranteed circuit specifications are listed in Table 1. Many modifications of this basic circuit, such as remote shutdown, re-

duced noise, non-standard passive components, and odd-ball input-output voltages are described in Application note UM-2, MAX742/MAX743 available from Maxim.

Standard 3 W application

With MAXL001 inductors and MAXC001 or equivalent output filter capacitors, output voltage ripple at full load is about 75 mV p-p at

the oscillator frequency (200 kHz). In addition, about 250 mV transient noise occurs at the LX switch transitions. The optional filter shown reduces both these noise components.

The choice of inductor type involves a trade-off, which optimizes either cost and size, or EMI and noise performance. Suggested design approaches for mainstream applications are given in Table 2.

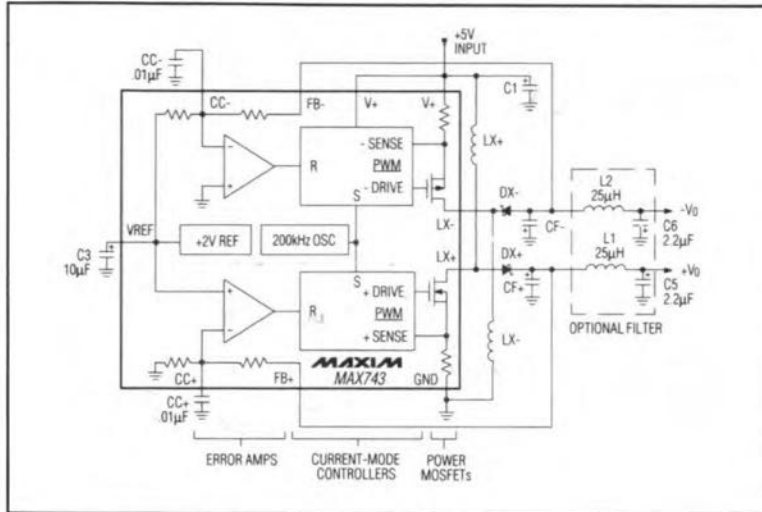


Fig. 1. Functional diagram with optional pi filter.

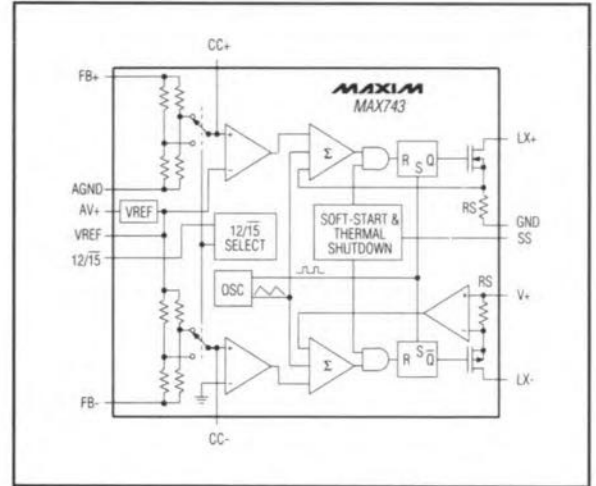


Fig. 2. Block diagram of the MAX 743.



Fig. 4. Kit of components for the circuit in Fig. 1 (available from Maxim).

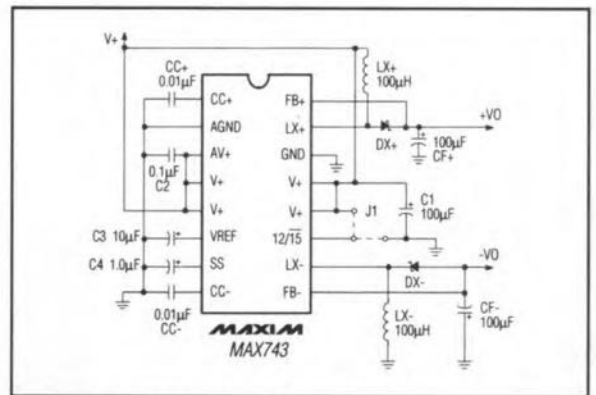


Fig. 3. Basic application circuit.

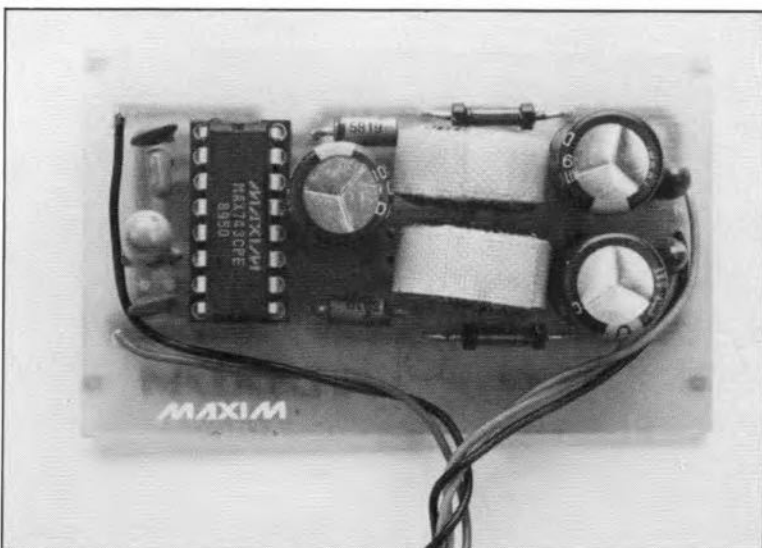


Fig. 5. Circuit in Fig. 2 built on PCB in Fig. 6.

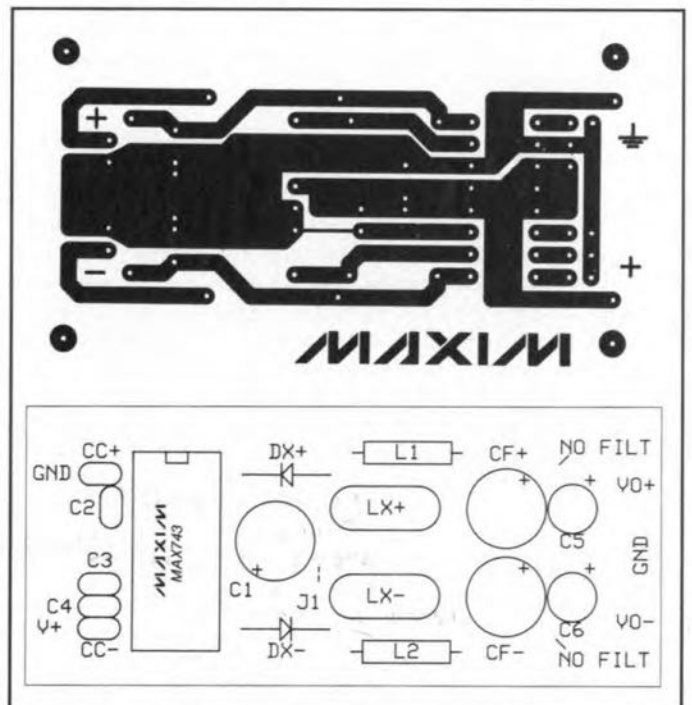
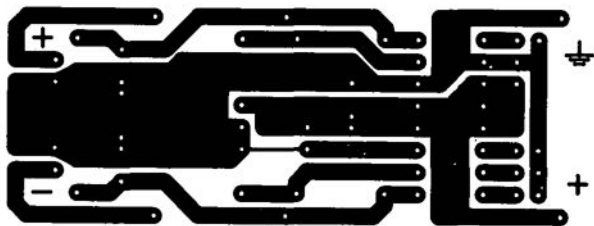


Fig. 6. PCB layout for circuit in Fig. 2 (incl. filter).



MAXIM

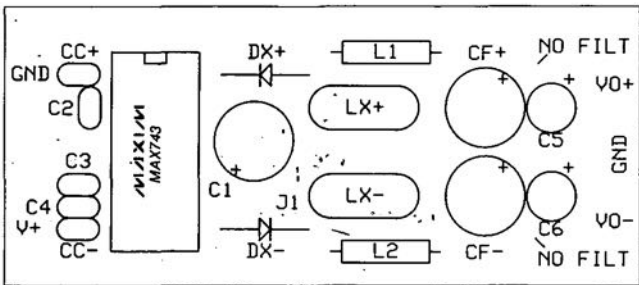


Fig. 6. PCB layout for circuit in Fig. 2 (incl. filter).

SCIENCE & TECHNOLOGY

GREATER CAPACITY IN SUBMARINE OPTICAL CABLES

by Bill Presdee, BSc, CEng, MIEE

EXPERIMENTS undertaken by British Telecom (BT) research engineers portend a dramatic increase in the communications carrying capacity of optical telephone links and a means of checking the viability of an optical fibre link in service without disrupting communications.

There are two separate studies. The first is concerned with increasing to 2.4 Gbit s⁻¹ the capacity of a single fibre in the existing submarine fibre optic cable link, currently operating over 132 km (80 miles) between the United Kingdom and Guernsey in the Channel Island. The second is a method of detecting faults on the link without interrupting its data transmission.

When optical fibres were introduced in the 1970s, they promised a means of transmitting data at a high rate of Gbit s⁻¹ over many kilometres of cable with minimal attenuation. These dreams of an ultimate method of transmission were, to some extent, dashed by the discovery that, although there was a small but significant loss of signal over tens of kilometres, the main enemy of effective high-speed transmission was pulse broadening.

This can be attributed mathematically to group velocity dispersion, which is the change in pulse propagation velocity with change in light frequency. The broadening is proportional to the length of fibre traversed, so eventually a maximum cable length is reached where the pulse width is sufficient for pulses to overlap, and effective communications ceases.

Learning from water

Attempts to reduce this effect by tuning the laser light source (nominally 1.30 μm) to the wavelength at which the group velocity dispersion reaches a minimum (1.55 μm for standard fibre) were not particularly effective. However, other phenomena were evident in fibre technology that could be exploited to provide a solution. These included the slight non-linearity of the refractive index, n , and the *soliton** effect. This was first recorded about a century ago by a cleric watching the passage of barges along a canal.

He noticed that at times a single wave was generated that maintained its shape and ve-

locity for several kilometres along the canal, and one can visualize him galloping along the towpath making his observations. The supporters of the general system theory, who subscribe to the portability of the elements of systems between disciplines, will be pleased to know that this effect occurs also in fibre optics.



Tests in progress on the submarine optical link between the United Kingdom and the Channel Islands.

The effect of the non-linearity in the refractive index can be expressed as the difference between its nominal value and that at an arbitrary low intensity that is proportional to the light intensity. The introduction of light power measured in watts into a single mode fibre of area, say, 10^{-6} cm^2 , represents a translation to an intensity of MW mm^{-2} . The effect of this non-linearity can be allowed to build up over hundreds or even thousands of metres of fibre length.

Dispersion effect

The non-linearity leads to an effect known as self-phase modulation in which, owing to phase and frequency shift (or chirp) as the pulse traverses the fibre, the frequencies in the leading half of the pulse are lowered, while those in the trailing half are raised. The effect of dispersion is to advance the trailing half of the pulse while

retarding the leading half and, provided the chirp is large enough, dispersion—formerly the pulse broadener—leads to pulse narrowing.

The considerations necessary in formulating a soliton pulse for communications in optical fibres are vested in complex mathematics. Basically, however, the pulse-broadening effects of dispersion have to be balanced exactly by the pulse-narrowing effects of non-linearity.

Of course, apart from arriving at the right formulation for development of a high-capacity optical communications system, it is necessary to have an advanced technology with which to implement it, and the experimental work to prove its efficacy.

The future traffic requirement of unrepeaters submarine systems, most of which currently operate at below 140 Mbit s⁻¹, is likely to be as high as 2.4 Gbit s⁻¹. Accordingly, BT engineers have mounted a series of field trials to demonstrate that the technology developed by them can meet this requirement and that the British mainland can be linked to its off-shore islands and its continental neighbours with very high capacity, repeaterless optical cables.

Successful demonstration

The UK-CI 7 cable running between Stoke Fleming in southwestern England and St Peter Port in Guernsey, which is 132 km (80 miles) long and has been in operation at 140 Mbit s⁻¹ since early 1989, was selected for this purpose. An initial three-day trial with a frequency shift keying (FSK) system demonstrated successfully that the system capacity could be increased eight-fold to 1.2 Mbit s⁻¹ with the use of commercially available components giving a 2 dB operating margin. The trials were continued to investigate the feasibility of a further increase in cable capacity to 2.488 Gbit s⁻¹ and a greater operating margin.

In direct intensity modulated systems, wavelength chirping of the distributed feedback (DFB) laser diode produces an unacceptable dispersion penalty. The FSK system was used to give a narrow width spectral source in preference to a phase shift keying (PSK) system, which would be more complex, or an external modulation system, which would introduce high coupling losses and need a high electrical drive signal. An FSK direct detection system can be implemented economically, reduce the effect of fibre dispersion and permit a high

* a stable particle-like solitary wave state that is a solution of certain equations for propagation in plasmas, fluid mechanics, lasers, solid-state physics and elementary-particle physics. *Editor*

launch power with only a small electrical signal.

Fibre amplifier

FSK modulation, with the mark and space separated by a few GHz, was applied to a commercial DFB laser with an integral optical isolator. The FSK signal was demodulated at the transmitter with the aid of an optical self-homodyne technique via a fibre-based Mach-Zehnder interferometer. Care was taken to minimize thermal drift and to keep the laser centre frequency and the interferometer aligned.

To provide sufficient optical power at the receiver, an erbium-doped fibre amplifier was pumped with light from a high-power multi-quantum well semiconductor laser to boost the transmitted signal level. Two complete transmitters and avalanche pho-

todiodode (APD) receivers were used to allow simultaneous two-way transmission of a pseudo-random bit sequence (PRBS) test across the submerged cable.

As an additional demonstration, a surface-acoustic wave electronic data regenerator was installed at the Guernsey terminal to loop the received data back to the mainland. At 2.488 Gbit s⁻¹ (equivalent to 32 broadcast television channels or 35 000 telephone calls), the demonstration showed the feasibility of upgrading the capacity of existing submarine links.

Fault location

Optical time domain reflectometry (OTDR) is a well-established diagnostic method of evaluating various features of an optical cable and locating discontinuities and junctions such as breaks and splices.

The UK-CI 7 cable was again used as a test-bed to explore the possibility of conducting such OTDR tests without removal of traffic from the line and consequent loss of revenue. The OTDR test unit, operating at 1514 nm, and a 1546 nm DFB laser, modulated with 565 Mbit s⁻¹ PRBS data, were coupled to the system fibre via a nominally 50:50 passive fibre coupler.

At the OTDR end and the APD receiver end, 3-nm full-width half-maximum filters mounted in low back-reflection fibre optic beam expanders were used to prevent reflections back into the DFB laser.

The field trials showed that the data channel does not affect the OTDR measurements. Conversely, bit error rate measurements conducted with and without the OTDR operating showed that it did not degrade the performance of the 565 Mbit s⁻¹ data channel. ■

THE LIVING COMPUTER

920026

by C.C. Whitehead

MANY people have the false idea that computers have only come into being in the world of today. However, they have always been here, even before the dawn of the history of mankind: *living computers*.

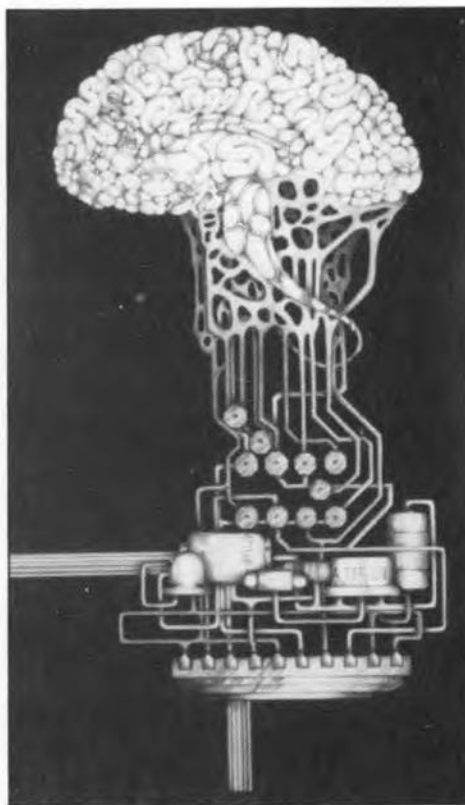
Living computers exist in many different forms, as do non-living (electronic and electromechanical) ones designed by them. To understand this fully, it is necessary to know what is contained in a computer so that it can perform a variety of tasks.

The first essential is the gathering of information: the inputs of the computer. At this stage, these are said to be in *analogue* form: sound and vision in the case of a human observer.

At the next stage, the inputs are usually converted into *digital* form and stored in *memories*, of which there are usually more than one. In the case of the living memories in the human brain, there are two: long-term and short-term.

Memories are necessary because time is involved: the time when information is received and the time when it is processed are not necessarily the same. The time scale in a computer and that in the outside world are not usually the same. The non-living computer has a clock to control its speed of operation and this gives it a great advantage over the living computer: in it, the digital units of information form a time-space pattern that is unique for every discrete item of information.

The clock produces pulses of energy at precisely regular intervals. Each computer has a specific clock-rate: in the human brain this is about 16 pulses per second. An electronic computer can have a clock-rate a million or more times faster and can, therefore (in theory) perform a task in seconds that would take the human brain as many weeks. However, the human brain has a little understood at-



tribute not possessed by the non-living computer: emotion*, which, among other functions seems capable of greatly increasing the clock-rate when an important situation arises so that during the emergency time seems to pass more slowly.

What is perceived in the outside world plus the information already in the memories, all

* In our opinion, a distinction should be made between *brain* and *mind*, the 'brain' being the hardware and the 'mind' the software; emotion is not a manifestation of the brain, but of the mind. *Editor*

in digital form, is passed to the 'thinking' unit in the computer, the *logic unit*, where it is processed into a form that is assumed to be acceptable to the individual or organization interested in the output(s) of the computer.

After it has been processed, the information, still in digital form, is passed to the output unit where it is reconverted into the form required at the outputs, which is usually analogue.

Another feature that is common to all computers is that they have to be *programmed*, that is, made to be able to deal with the information in a manner that is acceptable to the individual or organization using the computer. Programming in the kinds of computer with which we are dealing in this paper is always carried out by a human being. In the case of non-living computers, this is done by (replaceable) software; in living computers, as stated in an earlier article (Ref. 1), it is done by the process called *education*.

That body of great educators, the Jesuits, knew all about this and claimed that the essential elements of programming were imparted during the first seven years of the individual's (conscious) life. Even today, the government in the UK lays great stress on the importance of the first seven years in school.

All of this raises interesting philosophical questions. For instance, what is the nature of time, which in the human consciousness is associated with the velocity of propagation of electromagnetic radiation? Who or what programmes the computer in the case of living creatures other than human beings? Most religions have an answer to this: God. That is a perfectly reasonable concept with which, however, not everybody is in agreement. ■

Reference 1.

'A world of computers' by C.C. Whitehead, *Elektor Electronics*, December 1991.

DESIGN IDEAS

The content of this column is based solely on information supplied by the author and does not imply practical experience by *Elektor Electronics*

PWM temperature controller

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

PRECISION temperature control is required from egg incubators to large industrial furnaces. If the temperature controller malfunctions, the product gets damaged. In this article a precise, proportional temperature controller with built-in safety limit is discussed. The application of the controller is left to the reader's imagination.

Temperature control systems

Basically, there are three different ways of controlling temperature.

- **On/off control.** This is the simplest way to control temperature. When the temperature of the medium to be heated is below the set point, the heating element is switched ON and turned OFF when the temperature goes above the set point. The medium temperature fluctuates from the set point by 5–10 °C. Figure 1a shows the control signal and the temperature rise. Typical applications are for cooking oven, laboratory hardening processes and sealing equipment.
- **Proportional control.** With this, the heater is switched ON and OFF in a series of pulses, PWM signals, supplying less heat to the system as it approaches the set temperature. The actual temperature varies by 1–4 °C from the set point as shown in Fig. 1b. This tolerance is called *offset* or *droop*. Typical applications include ovens, annealing processes and seal processing packing machinery.
- **Proportional integral control.** This is similar to proportional control, but introduces an integral control action that reduces the offset level as shown in Fig. 1c. The actual temperature differs from the set point by 0.5–3 °C. This type of (expensive) control is used for sensitive laboratory temperature-humidity ovens and for industrial injection machinery.

Temperature to voltage

One practical way of converting temperature to voltage is the use of β thermistors. A thermistor is made of a metal alloy whose resistance varies inversely with temperature, that is, it has a negative temperature coefficient (NTC). Thermistors are available

in different ranges (100 Ω to 500 k Ω).

When a thermistor is configured in a measuring bridge, such as a Wheatstone bridge, the ambient temperature can be sensed and

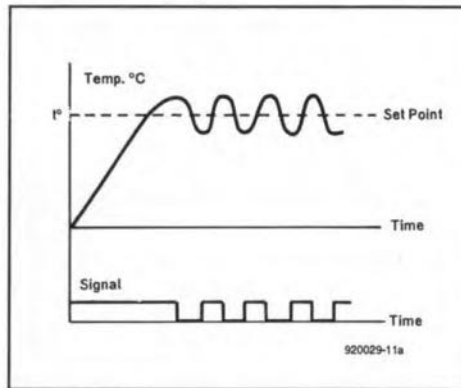


Fig. 1a. On/off control.

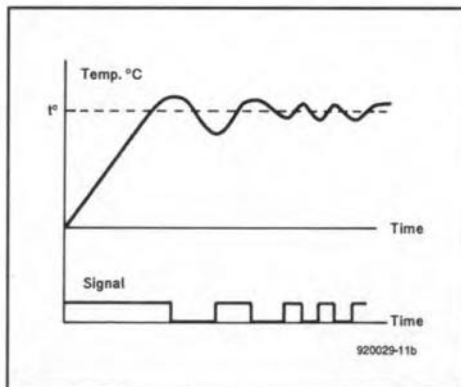


Fig. 1b. Proportional control.

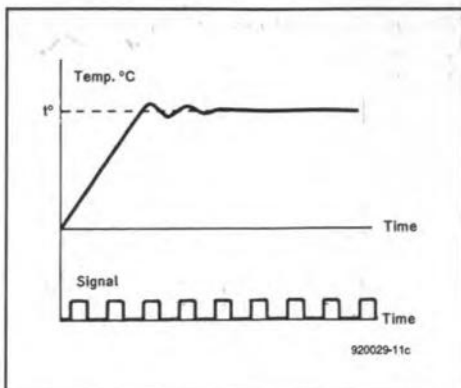


Fig. 1c. Proportional integral control.

measured as a differential voltage across the bridge terminals. In practice, a resistor of suitable value is connected across the thermistor to improve linearity of the temperature-voltage gradient.

A more accurate accurate linear temperature sensor is the LM335Z or LM35, whose output is directly proportional to the absolute temperature and provides an output signal of 10 mV K⁻¹.

To limit self-heating in temperature sensors, the current flowing into the sensor should be restricted to a value of 1–5 mA.

Proportional temperature controller

Figure 2 shows a simplified diagram of the basic ingredient of a low-cost proportional temperature controller. The sensed temperature, measured as a variable voltage, is compared with a reference voltage in the comparator. The output of the comparator is used to control a train of pulses generated by the pulse-width modulated (PWM) circuitry. This in turn controls the state of the switch that connects the heater to the power supply.

The complete circuit is shown in Fig. 3 and Fig. 4. It is powered by a 12 V d.c. regulated supply—see Fig. 5.

The temperature sensor chosen in the design is a disk thermistor that has a value of 180 Ω at 25 °C. It is connected as one arm in the balanced bridge formed by R₅, R₆, and R₇. The sensitivity of the bridge, or temperature set point, is adjusted with P₁. The

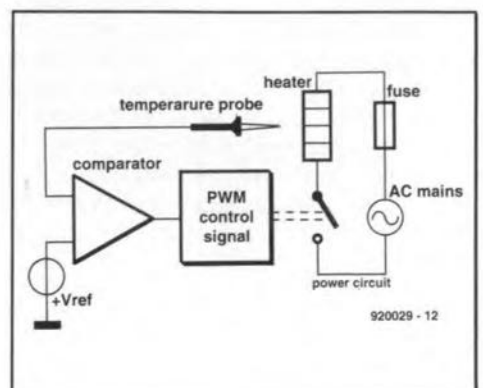


Fig. 2. PWM temperature controller.

values of the three resistors are calculated to limit the sensor current to 5 mA at 50 °C (maximum temperature to be reached).

The voltage difference across the bridge is applied to IC_{1a}. The output, pin 1, of this comparator goes high when the temperature sensor falls below the set point. The gain of IC_{1a}, determined by R₈, is large enough (up to 54 dB) to sense the small changes in the input voltage.

An analogue-to-digital pulse-width modulated train of pulses is generated by IC_{1c}-IC_{1b}. The frequency of the sawtooth signal at pin 8 of IC_{1c} is determined by R₉-C₂. The relation between the sawtooth signal, tempera-

ture signal (pin 1) and the generated PWM signal is shown in Fig. 6.

Over-temperature protection

Malfunction of the control circuit or the triac can cause a significant rise in temperature, which, in case of unattended temperature-controlled equipment, could be serious. The circuit is wired so that during normal operation, that is, with the temperature below the safety limit set by P₂, the output (pin 14) of IC_{1d} is high enough to switch on T₁, so that relay Re₁ is energized as long as the temperature does not exceed the safety limit.

When the temperature rises appreciably, the potential across the thermistor, and thus the output of IC_{1d}, drops. This causes T₁ to switch off, so that the relay is deactuated and the mains supply is disconnected from the heater and IC₂. This condition is indicated by the sounding of the buzzer.

When the temperature drops again, the voltage drop across the thermistor increases and T₁ is switched on again.

Note that the relay contact must be rated at the maximum current through the heater element.

screened cable and inserted two thirds of the way into the tube. After satisfactory tests (see below) the probe can be filled with silicon sealant and finished as required.

Circuit testing

Compare the waveforms at pins 1, 7 and 8 of IC₁ with those in Fig. 6. Remember that the circuit carries mains voltage!

Connect a 100 W bulb in place of the heater and adjust P₁ to check that the bulb can be *gradually switched ON* and *OFF*.

Using a hair drier, direct hot air at the probe and check that the bulb is switched off *instantly*.

To test the safety limit, short the sensor terminals: Re₁ will be energized and the buzzer will sound. Remove the short, blow hot air at the sensor and adjust P₃ until the alarm sounds. Measure the temperature and repeat until the upper temperature limit is set.

Temperature calibration

There are several ways of calibrating the circuit. If the medium used is water and the probe is made waterproof, a simple thermometer can be used to read the temperature of the water. Potentiometer P₁ is then varied until the triac changes state; the temperature is marked and so on for a range of temperatures.

Further reading:

Motorola: *Optoelectronics Device Data*, Fourth Edition, 1989.

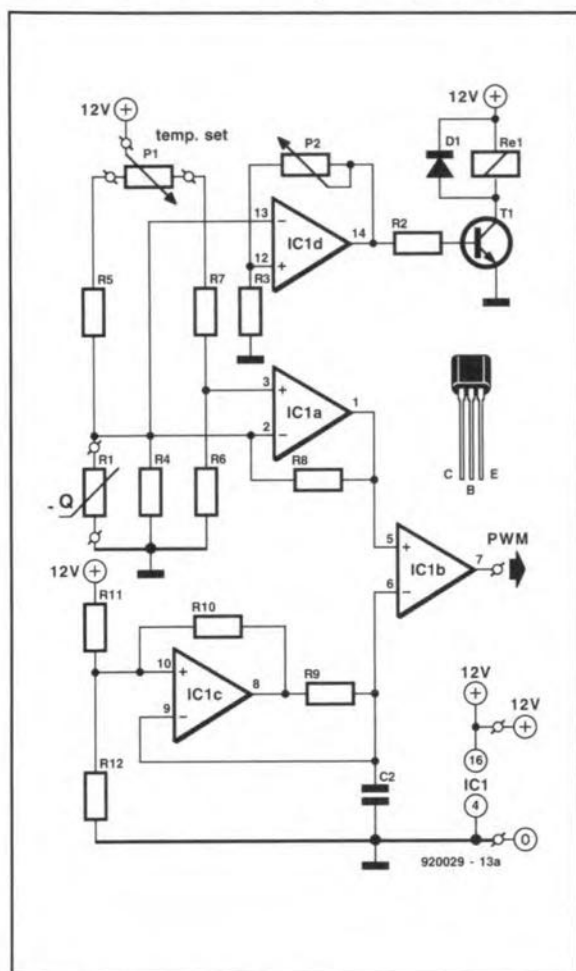


Fig. 3. Circuit of the temperature controller.

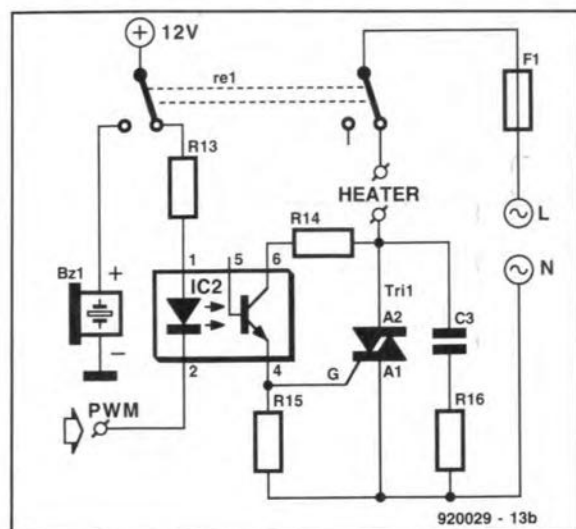


Fig. 4. Optoisolator triac driver.

Temperature probe

The probe is made of a 100 mm long, 20 mm dia., brass tube, of which one end is crimped.

The thermistor is connected by

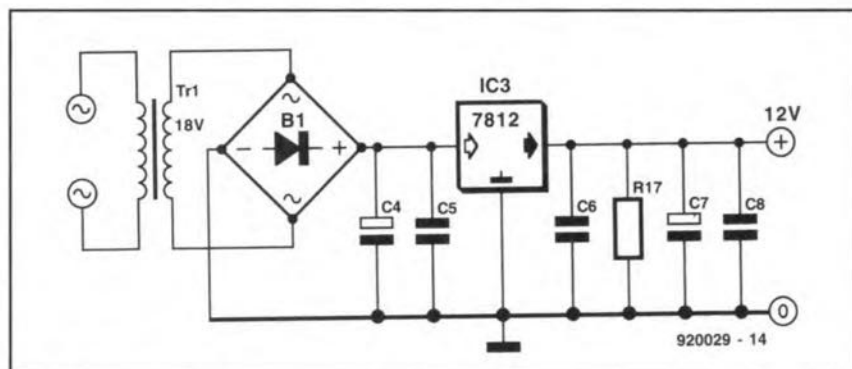


Fig. 5. Regulated 12 V power supply.

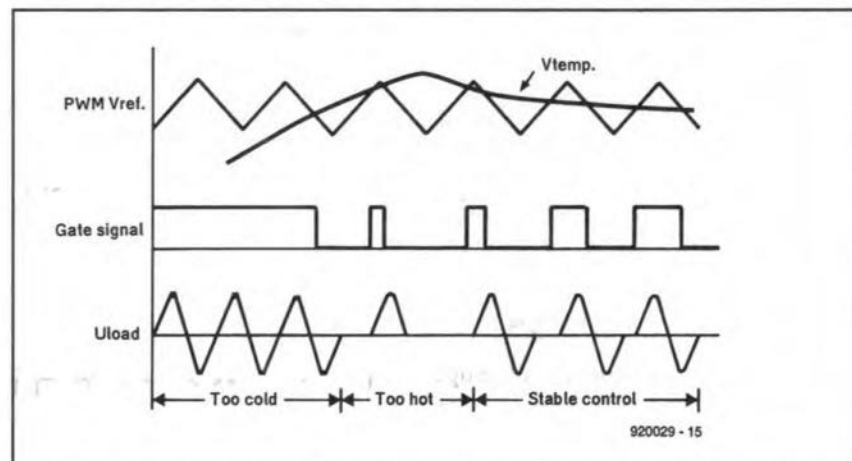


Fig. 6. Proportional temperature control.

DIRECT DIGITAL SYNTHESIS — DDS

Recent developments in integrated circuit technology have made it possible to build all-digital frequency synthesizers. Following a short remedial course on traditional synthesizer circuits, this article looks at the operation of DDS, its advantages and disadvantages.

by Dipl. Ing. G. Kleine

PLL and synthesizer circuits

SYNTHESIZER circuits are used to generate AF and RF signals with a stable frequency. They are usually based on a phase-locked loop (PLL) circuit as illustrated in Fig. 1. The output frequency is supplied by a VCO (voltage-controlled oscillator). A phase detector compares the frequency (or the phase) of the VCO output signal with a reference frequency, f_{ref} . The output signal of the phase detector is an error signal that is filtered before it is applied to the VCO as a control voltage. When the PLL is locked, the VCO is phase-locked to the frequency reference.

Since it is usually required for the VCO to cover a certain frequency range or channel raster, the basic PLL is extended with a programmable frequency divider. This results in a simple frequency synthesizer circuit (Fig. 2). The VCO allowing, every frequency between f_{ref} and $N_{max}f_{ref}$ can be generated by appropriate setting of the programmable divider, N . The factor N_{max} is the maximum divide ratio that can be set on the divider. Provided the loop filter is correctly dimensioned, the stability of the output frequency equals that of the reference frequency.

Where relatively high output frequencies are required (say, >100 MHz) a fast prescaler ($\div N$) is used ahead of the programmable divider ($\div N'$). Since the use of a prescaler increases the step size of the synthesizer from f_{ref} to $N \cdot f_{ref}$, the reference frequency may have to be lowered to achieve the same channel raster.

The drawbacks normally associated with a low reference frequency are negated by a modulo-2 prescaler, whose scaling factor can be switched between N and $N+1$. Figure 3 shows the block diagram of a such a synthesizer. The counters 'A' and 'M' are clocked by the output signal of the prescaler, and count up to a preset value. Counter 'A' determines the rate at which the prescaler divides by $N+1$. On reaching its preset (end) value, it switches the pres-

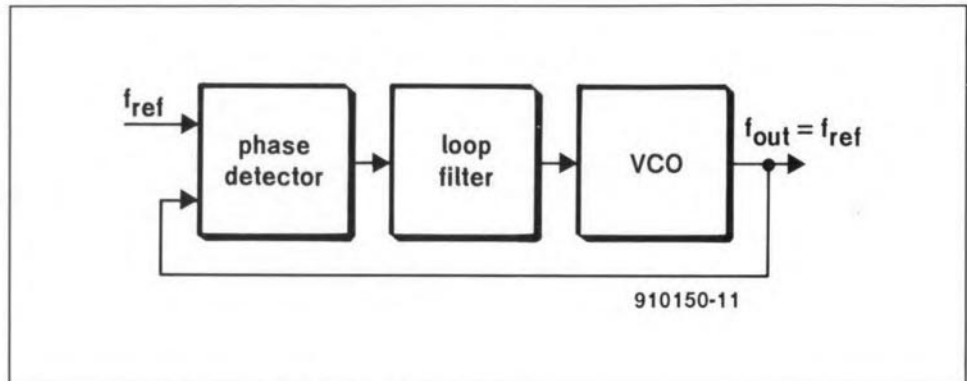


Fig. 1. Block diagram of a classic phase-locked loop.

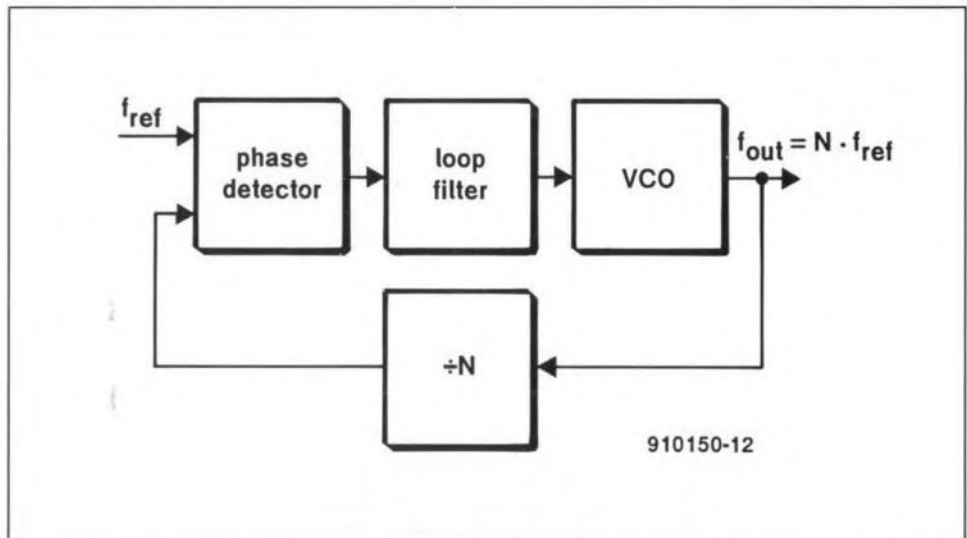


Fig. 2. Block diagram of a simple synthesizer.

caler to scaling factor N . The prescaler then divides by N until counter 'M' also reaches its preset (end) value. When this happens, counter 'M' resets itself as well as counter 'A', and the prescaler is switched back to division by $N+1$.

It will be clear that the modulo-2 principle works only when M is greater than A . The two counters give rise to a time-averaged scaling factor, P , which is calculated from

$$P = (N + 1) \cdot A + N \cdot (M - A)$$

$$P = N \cdot M + A$$

The modulo-2 principle thus allows chan-

nel rasters with frequency steps of f_{ref} to be achieved at relatively high frequencies. As an example, consider a synthesizer with a $\div 10/\div 11$ prescaler (i.e., $N=10$), $A=0$ to 9, $M=10, 11, 12$, etc. This results in $P=10 \cdot M + A$, so that all scaling factors starting with 100 can be used without 'gaps'.

A disadvantage of the synthesizer circuits discussed so far is that the output frequency is invariably a multiple of f_{ref} . This can be overcome by the so-called fractional- n system (Fig. 4). As in the modulo-2 system, scaling factors are switched, which results in an average scaling factor that is not an integer (e.g., 145.23). This is achieved with the aid of an accumulator,

whose contents are incremented by F on every clock pulse. When the value L is reached, the accumulator switches the prescaler to division by $N+1$ during one clock pulse. The number of clock pulses over and above L remain in the accumulator as the new start value. It can be shown that this system gives rise to an average scaling factor

$$P = N + F/L.$$

This means that every required frequency resolution of the synthesizer can be achieved by appropriate choice of N and L . As an example, consider a system in which $N=10$, $F=0$ to 9, and $L=10$. Hence, $P=10+F/10$, so that scaling factors such as 10.1, 10.2, etc. to 10.9 can be set via the increment, F .

An extensive discussion of the operation and design of the PLL and synthesizer circuits discussed above may be found in Ref. 1. This background literature also provides a type classification of PLL circuits and phase detector circuits. In addition, an overview is given of available PLL and synthesizer ICs.

Direct digital synthesis — DDS

A new, all-digital way of generating AF and RF signals is offered by DDS. The principle is shown in Fig. 5. A phase increment register with high resolution (e.g., $L=32$ bits) is provided with a value F that corresponds to the rate at which the phase of the clock signal f_{clk} is changed. Next, F is added to the existing phase value stored in a latch, and the result ends up in the latch again. When an overflow occurs in the adder, the L -bit-wide result is stored in the latch. This means that there is no carry on position $L+1$. The phase value held in the latch forms the address for a sine function ROM, whose data are applied to a D-A converter. As with

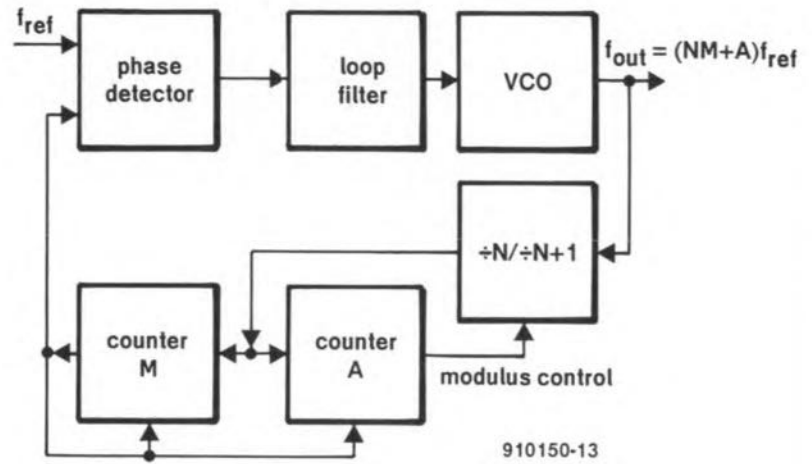


Fig. 3. Block diagram of a synthesizer with a modulo-2 prescaler.

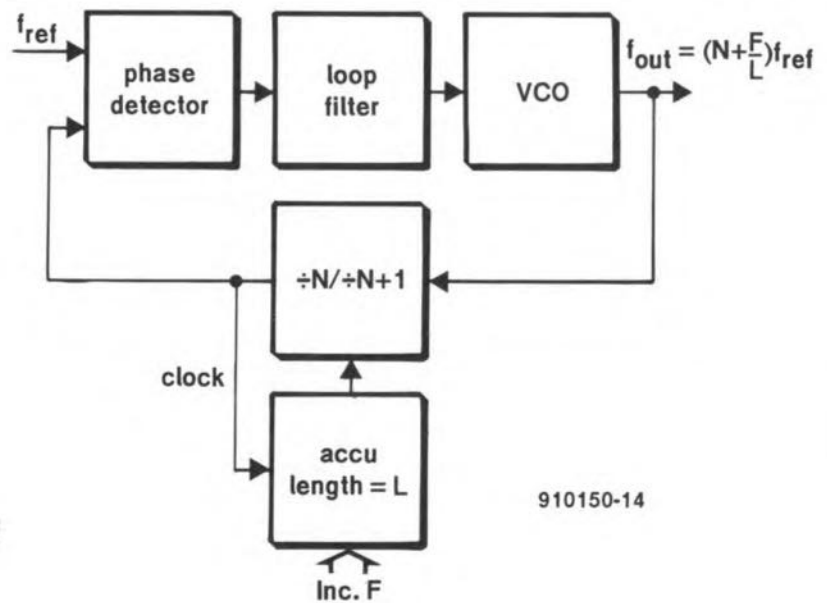


Fig. 4. Block diagram of a fractional- n synthesizer.

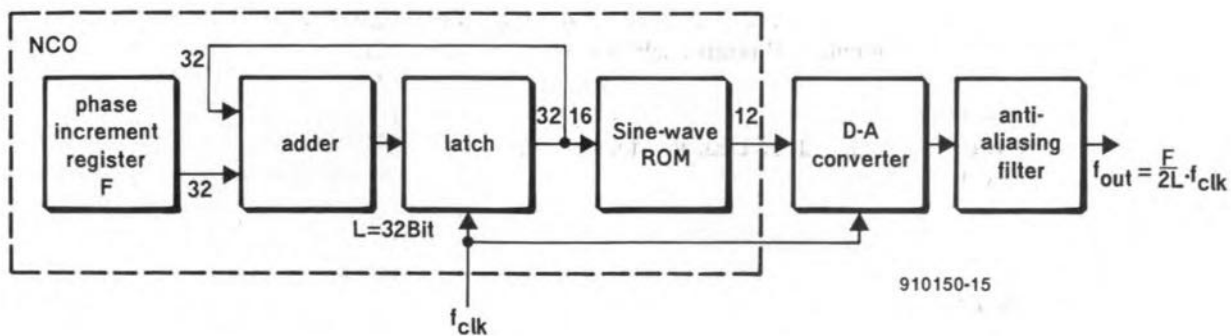


Fig. 5. Block diagram of a synthesizer based on the DDS principle.

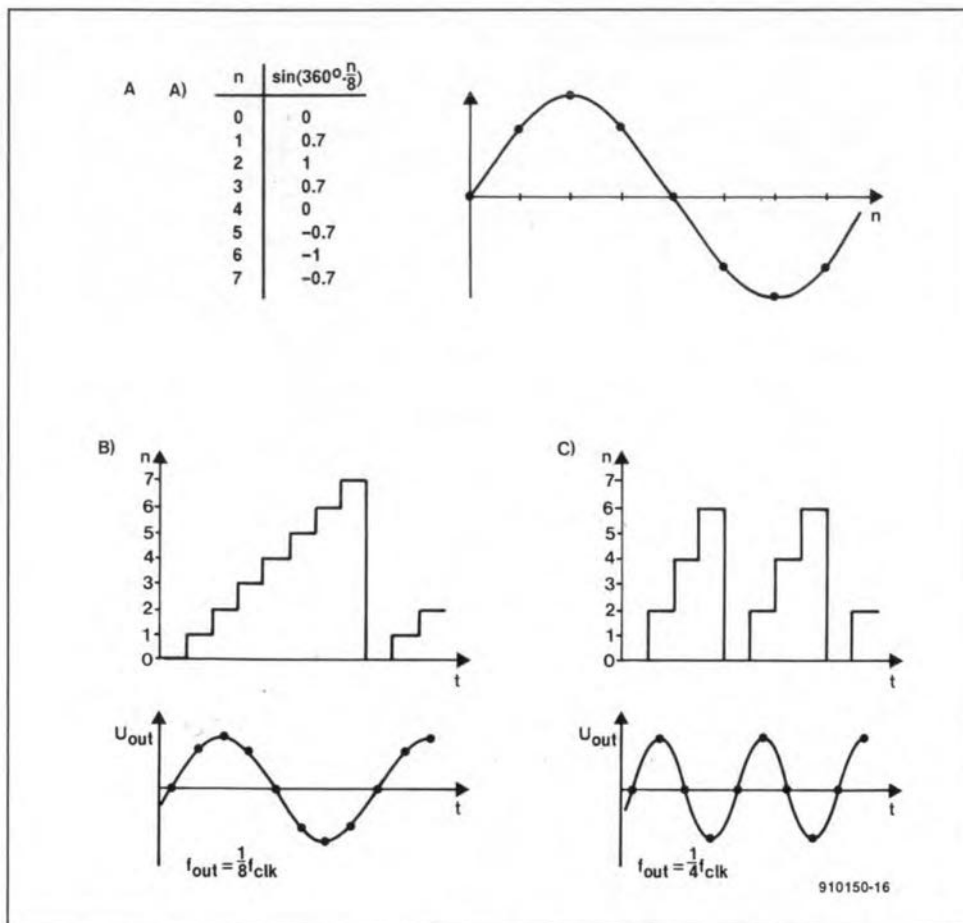


Fig. 6. (a) Contents of the sine function ROM at a resolution of 3 bits. (b) Waveform generated with a phase increment of $1 \times 360/8$. (c) Ditto for $2 \times 360/8$.

any other sampling system, this D-A conversion must be followed by an anti-aliasing filter that serves to suppress spectral components above $f_{clk}/2$.

The power of DDS arises mainly from the high resolution of the phase increment register, the adder and the latch. In Fig. 5, for example, the resolution is represented by L , which has a width of 32 bits. The full resolution need not be used throughout the system, however. The sine function ROM is addressed by, for instance, the most-significant 16 bits of L , while the resolution of the DAC is still lower at 12 bits.

In theory, the maximum output frequency of a DDS-based synthesizer is $\frac{1}{2}f_{clk}$. In practice, however, values of $0.2f_{clk}$ to $0.4f_{clk}$ are achieved because of the anti-aliasing filter (see Fig. 7a). The phase increment, F , and the phase resolution, L , determine the output frequency, f_o , of the DD synthesizer:

$$f_o = (F/2^L) \cdot f_{clk}$$

where a frequency resolution of

$$f_{clk}/2^L$$

is achieved. As an example, consider a DDS in which $f_{clk}=100$ MHz and $L=32$ bits. This offers an impressive resolution of 0.0233 Hz at f_o .

The operation of the sine function ROM is illustrated in Fig. 6. Figure 6a shows the output values of the ROM at a resolution of 3 bits. One period of the sine function consists of 8 discrete steps ($n=0$ to 7), of which the corresponding values are stored in digital form at address n . Figure 6b shows the value of n in the latch output signal for a phase increment, F , of 1. Shown below is the signal after the D-A converter and the anti-aliasing filter. The amplitude values of this signal correspond to the entries in the sine function ROM table (Fig. 6a). The output frequency, f_o , equals $\frac{1}{8}f_{clk}$.

The effect of increasing the phase increment, F , to 2 is shown in Fig. 6c. It is seen that the range of phase values (or ROM addresses), n , is cycled through twice as fast, which results in f_o doubling to $\frac{1}{4}f_{clk}$.

The output frequency spectrum of a DD synthesizer will inevitably contain certain spurious components (Fig. 7). A number of these are alias components caused by the sampling operation (Fig. 7a). These components occur at

$$f_{+alias}(i) = i \cdot f_{clk} + f_o$$

and at

$$f_{-alias}(i) = i \cdot f_{clk} - f_o$$

where i is 1, 2, 3, The component writ-

ten as $f_{-alias}(1) = f_{clk} - f_o$ is the critical component for the anti-aliasing filter. Figure 7a shows this for a DDS in which $f_o=15$ MHz and $f_{clk}=40$ MHz.

The other spurious components in the output spectrum are caused by the limited resolution and non-linearity (quantization errors etc.) of the D-A converter. These components are multiples of the output frequency as well as intermodulation products caused by 'imaging' with the y -axis and the component $f_{clk}/2$. Fig. 7b shows these additional spurious products at $f_o=15$ MHz and $f_{clk}=40$ MHz. A number of components formed by imaging with $f=0$ and $f=f_{clk}/2$ are indicated.

The level of the alias components depends directly on the resolution of the D-A converter, whose non-linearity in addition determines the level of the other spurious components. As in any other real-time sampling system, the alias components are $(\sin x)/x$ weighted (Fig. 7a), so that their level drops with increasing frequency.

The digital circuit between the phase increment register and the sine function ROM is usually referred to as an NCO (numerically controlled oscillator) (Refs. 2; 3). A number of currently available DDS ICs have an on-board DAC—others require an external DAC. Where the address lines of the ROM are accessible, a different conversion table may be used to implement other waveforms, e.g., a triangular wave. There exist also DDS ICs that contain a sine and a cosine function in ROM. These ICs enable two output signals with a phase difference of exactly 90° to be generated. Such signals are often required as carriers representing I- (in phase) and Q (quadrature) components in complex modulator circuits (Refs. 2; 4; 5).

The output signal may be modulated by extending the block diagram in Fig. 5. Inserting an adder stage between the latch and the sine function ROM enables either phase modulation (PM, PSK, BPSK, QPSK) or frequency modulation (FM, FSK) to be realized by adding phase values. Similarly, amplitude modulation (double-sideband AM with carrier) is achieved by inserting a multiplier between the sine function ROM and the DAC. Since all three modulation systems are implemented with digital means, they are in principle purely linear. More information on analogue and digital modulation systems may be found in Ref. 4.

DDS: outlook

Although DDS components are still relatively expensive, they have some advantages that give them a potential for wide practical use. In addition to the high resolu-

tion of the output frequency, the short lock (settling) time is particularly valued. By contrast, traditional synthesizers have a loop filter whose response to the VCO control voltage slows down the speed at which the synthesizer can change its output frequency.

As already mentioned, DDS offers the possibility of pure linear modulation of the output signal. It should be noted however, that the all-digital modulation causes certain spurious products in the output signal. The level of these products can be reduced by increasing the resolution of the phase increment register and the D-A converter.

A further advantage of a DD synthesizer is its wide output frequency range of several decades. Again by contrast, this is almost impossible to achieve with traditional VCOs. Finally, the stability of the output frequency depends on the reference clock, f_{clk} , only. This means that ageing effects in the clock source are ruled out thanks to the all-digital principle of operation. ■

References:

1. Theorie und Anwendung des Phase-locked Loops (Theory and application of phase-locked loops). Published by AT Verlag.
2. The DDS Handbook (second edition). Published by Stanford Telecom *.
3. Design notes published by Rohde & Schwarz, Stanford Telecom/Alfatron GmbH, Plessey, Tricom Mikrowellen GmbH, Qualcomm Inc.

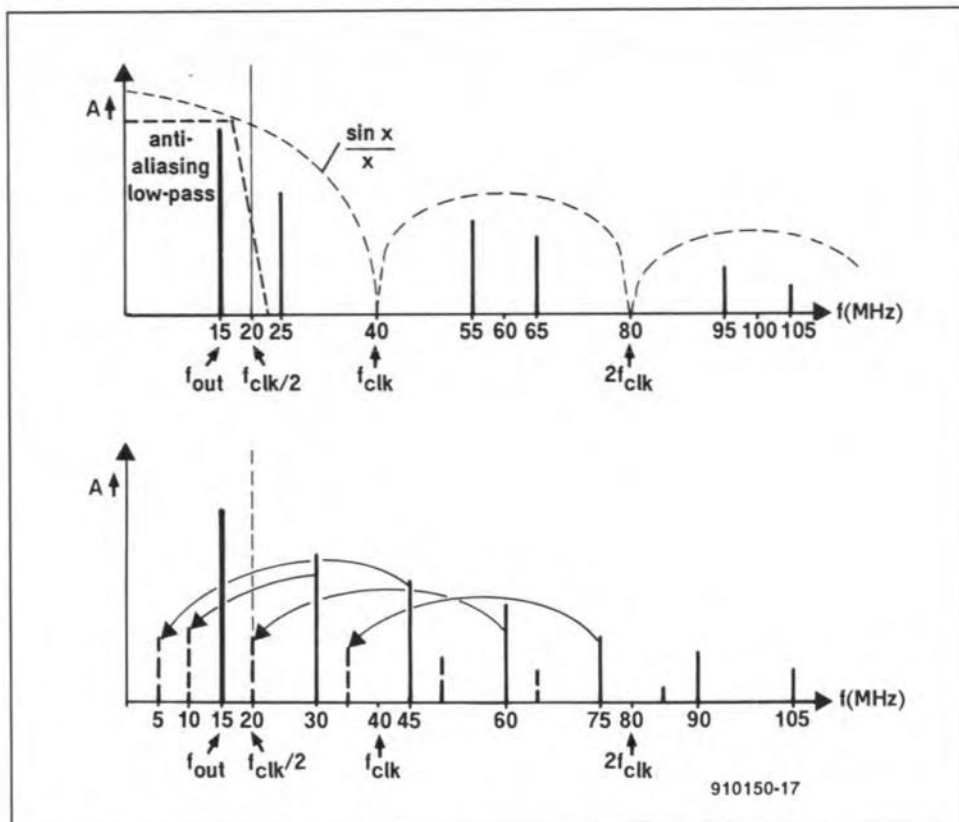


Fig. 7. Alias and spurious components in a DDS spectrum. The solid vertical lines represent multiples of f_{out} . The dashed vertical lines represent image products relative to $f=0$ or $f_{clk}/2$.

4a. Analoge Modulationsverfahren (Analogue modulation systems). By R. Mäusl. Published by Hüthig Verlag.

4b. Digitale Modulationsverfahren (Digital modulation systems). By R. Mäusl. Published by Hüthig Verlag.

5. Audio spectrum shift techniques. *Elektronics* October 1991.

* This book may be ordered from Alfatron GmbH, Stahlgruberring 12, 8000 München 82, Germany. Telephone: +49 89 420491-0. Fax: +49 89 420491-59.

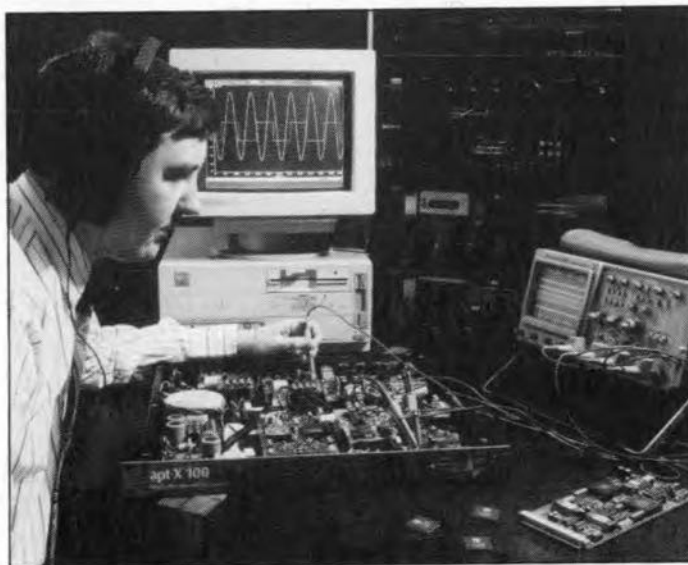
AN INNOVATIVE MUSIC COMPRESSION SYSTEM

gen nummer

Dr Ian Mack, a researcher in the Department of Electrical and Electronic Engineering, Queen's University, Belfast, Northern Ireland, is seen evaluating the noise performance of the APT-X 100 digital audio data compression system, which was developed by Audio Processing Technology. It is thought to be the first of its kind in the world which quadruples the supply of digital music signals stored on a compact disc.

The brains of the system are the high-speed digital signal processor (DSP) chips, shown in the foreground, to code and compress from 16 to four, the binary digits (bits) which represent the audio signals with no loss of quality.

The breakthrough not only has far-



reaching implications for the music industry, but also in the field of satellite communications. Since it would mean the reduction to a quarter of the current num-

ber of sound signals transmitted by satellite, it would reduce to a quarter the bill for expensive satellite time.

The company was founded jointly in 1988 by Stephen Smyth as a result of research undertaken by him while he was a student at Queen's University, and QUBIS, a holding company of the university, that provided the initial funding. In 1989, Solid State of Oxford became the majority shareholder.

For further information, contact

Paul Smith, Publicity Coordinator, Audio Processing Technology Ltd., 21 Stranmillis Road, Belfast, Northern Ireland BT9

5AF. Telephone: (0232) 662714. Fax: (0232) 382208.

8051/8032 ASSEMBLER COURSE

PART 1: INTRODUCTION

This is the first part of an assembly-level programming course for microcontrollers in the popular MCS51 series from Intel. MCS51 processors have been at the heart of many circuits described over the past few years in this magazine. As far as the courseware is concerned, although almost any MCS51-based computer or microprocessor system can be used to run the assembler and the example programs offered, the course is tailored to the 80C32/8052-BASIC computer described about a year ago in this magazine.

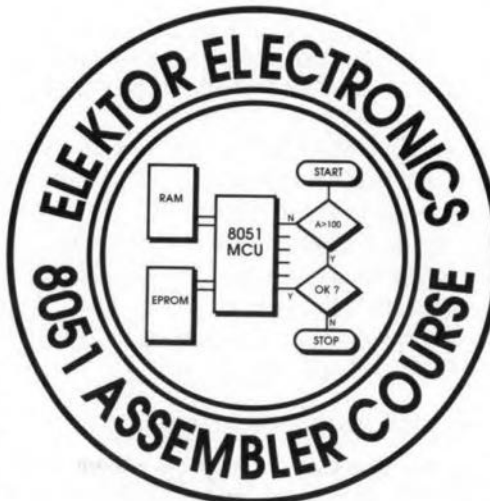
by Dr. Ing. M. Ohsmann

THE aim of this course is to provide an introduction into programming and applying the widely used microcontrollers in the MCS51 family from Intel. In addition, the connection of a number of hardware extensions is discussed in relation to assembly-language control programs. As to the 'minimum requirements' on your part, the course is at a fairly basic level, both as far as the software and the hardware is concerned. Knowledge of BASIC is assumed, and you should have some experience in microprocessor circuit construction to be able to build the hardware extensions proposed during the course. The description of these hardware extensions will be chiefly in line with other projects in the magazine, which means that ready-made printed circuit boards will be available.

On completing this course, you are, hopefully, able to implement microprocessor control in a wide range of circuits, be it an alarm clock with a special 'wake up' tune and a birthday memory, an RS232 (V24) serial interface analyzer with LC display, or a power meter with a serial interface and a true-rms processor. All of these applications require a small extension to the 80C32 single-board computer (Ref. 1) and some made-to-size software.

Software

The subject of courseware was already mentioned. To enable you to reach your 'goal' quickly, the subject matter discussed



in the article instalments is supported by a number of software utilities written to run on IBM PCs and compatibles (versions for the Atari ST are being developed). Remember, the PC (or the Atari) has the double function of (1) a terminal that communicates with the 80C32/8052 computer, and (2) a device used to develop MCS51 assembly code for downloading to the 80C32/8052 computer.

The diskette offered in relation to this article contains all the programming tools required to follow the course:

- a 8051 assembler, EASM51, complete with source code written in Pascal;
- a terminal program, V24COM, complete with source code written in Pascal;
- a 8051 monitor program, EMON51;
- a number of 8051 programming examples written in assembler;
- short documentation (.DOC) files for

the examples.

In particular the collection of programming examples on the disk offers the 'beginner' in assembly language programming the possibility to obtain hands-on experience with a number of programming techniques.

Hardware

As already mentioned, the course is best followed if you have a 80C32/8052 computer (Ref. 1) up and running. Since this computer can be built in different versions, it is necessary to first agree on a certain standard configuration required for the course. To make sure that the clock rates, timing intervals and the communication speed with the PC are defined, a 12-MHz quartz crystal must be used. If fitted, inductor L1 has to be removed to prevent the crystal operating at an overtone frequency. Next, wire link (or jumper) 'A' is installed to enable the microcontroller to address EPROM IC7. The following microcontroller types may be used on the board: 8051, 80C51, 8052, 80C52, 8031, 80C31, 8032 or 80C32.

For a first test of the completed computer, its serial interface is connected to the PC (or Atari), the power is switched on, and ... wait! Nothing works without a system EPROM!

Monitor in EPROM

No computer system can do anything that makes sense without a program that is executed at power-on. This program is often called the bootstrap. After completing the bootstrap routines, the processor usually enters the monitor program.

The monitor program used during this course may be obtained in two ways: by purchasing the ready-programmed EPROM available through the Readers Services, or by programming your own EPROM using the EMON51.HEX file on the course disk. In the latter case, you will also need an EPROM programmer.

The monitor EPROM contains a number of useful programming tools ('utilities') and subroutines to assist the beginning programmer. These parts of the monitor program will be examined and used at a number of moments during the

course. They include:

Utilities:

- program downloading from the PC;
- display RAM/EPROM contents;
- register modification.

Subroutines:

- input/output;
- arithmetic.

To test the monitor EPROM, insert it into the EPROM socket on the computer board. Next, connect a loudspeaker, a resistor and a switch to the 64-way DIN a-c connector as shown in Fig. 1. Check the pinning of the connector: particularly when an angled type is used, the pin numbers may differ from those shown in Fig. 1.

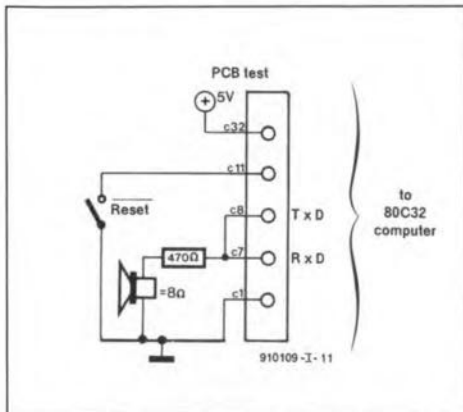


Fig. 1. A switch and a loudspeaker are used to run an initial test on the 80C32 board.

Switch on, and press the reset button. The loudspeaker should produce a short tune. When not, you are in for a round of debugging. Try to isolate the problem. The 'tune' test should still work when the following components are removed from the board: the RAM, IC₆; the external address driver, IC₉; and the external bus driver, IC₁₀. If necessary, remove these ICs one by one to locate the source of the trouble. If you have an oscilloscope, check that pin 30 of the microcontroller supplies clean ALE pulses. Clean should be taken to mean that the waveform is stable, rectangular, and formed by 166-ns long pulses, and 833-ns long pauses (a duty factor of about 0.2). The level of the ALE pulses should be about 4 V. Any disruptions or jitter observed in the pulse train nearly always point to problems with the quartz crystal.

The PC connection

The PC communicates with the microcontroller via the V24 serial interface on the computer board. Although the V24 interface used here works with TTL levels, the connection to an RS232 port on a PC should not cause problems. The system comes to life when you run the terminal

emulation program on the PC, and the monitor program on the 80C32 computer. This arrangement allows the PC to send data to the 80C32 computer, and vice versa, and will be particularly valued when debugging your own programs is in order. The combination of the 80C32 board and the PC thus forms a kind of development system.

To check if it all works, connect the PC and the computer as shown in Fig. 2. Run V24COM, configured for COM1:, on the PC. The serial port is selected in the associated .CFG file.

Switch on the 80C32 board, and press the reset button. The computer should report with the message 'EMON 51 Version ...' on the PC screen. If this works, you are ready to start programming.

Hardware extensions

Simple as they may be, the first tests as discussed above already indicate the flexibility of the 80C32 board in regard of different hardware configurations. All the hardware used during this course is accommodated on a single extension board, which will be discussed in next month's issue, along with part 2 of the course. This board features:

- LEDs for status indication;
- input keys;
- an LCD readout;
- a digital-to-analogue converter;
- an analogue-to-digital converter;
- a MIDI interface.

It will be clear that these extensions can be fitted as required. For instance, if you do not expect to ever develop a MIDI application, the relevant part of the extension circuit may be omitted. Alternatively, if you require one specific extension only, say, the LCD readout, this may also be built

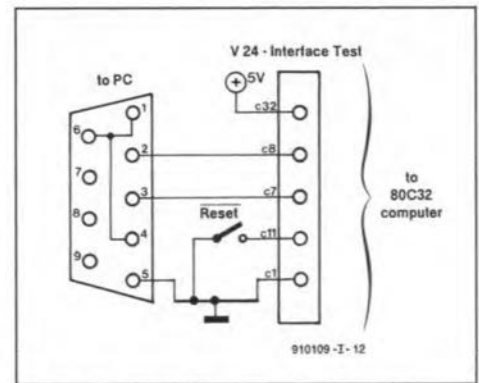


Fig. 2. Serial interface connection.

separately on a piece of veroboard or strip-board.

Assembly language programming

The listing in Fig. 3 is an example of an assembly language program written with the aid of a word processor. In fact, what you are looking at is XAMPLE01.A51 contained on the course diskette. In this so-called source code program, each line contains, in principle, one instruction only. The way in which the instructions are written is called mnemonics, which is a kind of shorthand for the instruction in words. In addition to the lines that result in the actual instructions for the microcontroller, there are lines containing assembler commands, which are necessary to manage the assembler program itself. Further, the listing is completed with comment lines that serve as an aid in understanding the operation of the program. The use of comment is entirely optional, but considered good practice for the obvious purpose of documenting one's programs. More information on the commands that may be used is found in the file EASM51.DOC on the course diskette.

```

; FILE XAMPLE01.A51
;
START   ORG   4100H                ; start object code at 4100H
        MOV   DPTR,#text          ; DPTR points to text
        LCALL STXT                 ; MONITOR Routine print text
        RET                        ; return to MONITOR

text    DB   'The FIRST Program ...',13,10,0
;-----
; MONITOR INTERFACE
;
ccSTXT  EQU 2                      ; MONITOR command to send text
COMMAND EQU 030H                  ; MONITOR command memory location
MON     EQU 0200H                 ; MONITOR entry address
;
STXT    MOV  COMMAND,#ccSTXT      ; MONITOR set command
        LJMP MON                  ; MONITOR call it!
;-----
;
END

```

Fig. 3. This is what a typical source code file looks like. Written with the aid of an ASCII compatible word processor, it is used by the assembler to produce a block of object code. The program shown here, XAMPLE01.A51, may be found on your course disk.

The function of the assembler, EASM51.EXE, is to use the source code file to 'assemble' a sequence of machine code instructions, data and addresses, that forms an executable program (object code). This is called the assembly phase. In other words, the assembler builds a new file, in which the mnemonics are replaced by opcodes, and all comment is removed. The output file has the extension .HEX (for example, XAMPLE01.HEX), and is nearly impossible to interpret for us humans, as it is a long series of two-digit numbers. To the controller, however, these numbers are meaningful as they represent instructions, operands, data and addresses, which together form a program that can be executed when it is contained in the program memory. The object code is generated on the PC, and sent to the 80C32 board by means of a download operation carried out with the aid of the V24 terminal program.

In addition to the object code, EASM51 also generates a list file with the extension .LST (see XAMPLE01.LST on the course disk). The list file shows the object code produced by the assembler alongside the associated line of source code.

Figure 4 illustrates shows the file types and operations encountered as the programming session evolves. Although the list file can not be used by the microcontroller, it is valuable for the programmer as it allows him or her to check the function of the assembler right up to the last bit of the object code.

Those of you keen on getting started straight away with the assembler are advised to read the information contained in XAMPL01.DOC.

First steps in programming

Following the initial hardware and software tests, we have arrived at a point where programming theory can be tackled. A complete description of all the hardware aspects and the full instruction set of the MCS51 family of microcontrollers is beyond the scope of this course, and readers are advised to consult the databooks for this (Ref. 2).

This course follows a different approach, as it discusses hardware and software aspects in a step-by-step manner, completing the picture gradually. The dis-

cussions of the various subjects are supported by examples found on the course disk. First, we tackle the various registers and memories that can be addressed internally and externally by the 8051.

Processor registers

The 8051 normally processes data with a size of one byte. The processor does this with the aid of a number of internal registers, each of which is capable of holding one byte. The accumulator, A, is the central register capable of logic or arithmetic operations on two operands. For example, if you want to add the value '10' to a byte, the byte is loaded into the accumulator, and the constant '10' is added. The result of this addition, the sum of 10 and the byte, is stored in the accumulator.

In addition to the accumulator, there are other 8-bit registers, which can be used, for instance, for the control of the interfaces and the internal timers. In most microprocessors and microcontrollers, these registers are addressed with special instructions. By contrast, the designers of the 8052 use a quite different, more elegant, approach. To enable all registers to be accessed with as few as possible instructions, each of them is assigned a special address with a value greater than $07F_H$. The accumulator, for instance, has the internal address $0E0_H$. The registers are accessed as so-called Special Function Registers (SFRs). This allows the devices in the MCS51 family to have different numbers of SFRs, and yet use the same instruction set. When an SFR address is given in combination with internal addressing (which concerns the 128-byte internal RAM), the content of the relevant SFR is taken as the operand.

The ports and the serial interface of the 8051 are addressed just like SFRs. An overview of all SFRs will be given in next month's instalment.

In addition to the SFRs there are eight general purpose registers identified as R0 to R7. These registers form a register bank and are used for intermediate storage of data. There are no fewer than four of these banks, which will be particularly valued when dealing with interrupt processing. During this course we will use register bank '0' only. The register banks are contained in the internal data memory. After resetting the controller, register bank '0' is automatically selected.

Program, data and internal memory

The microcontrollers in the MCS51 family have different memories for programs and data.

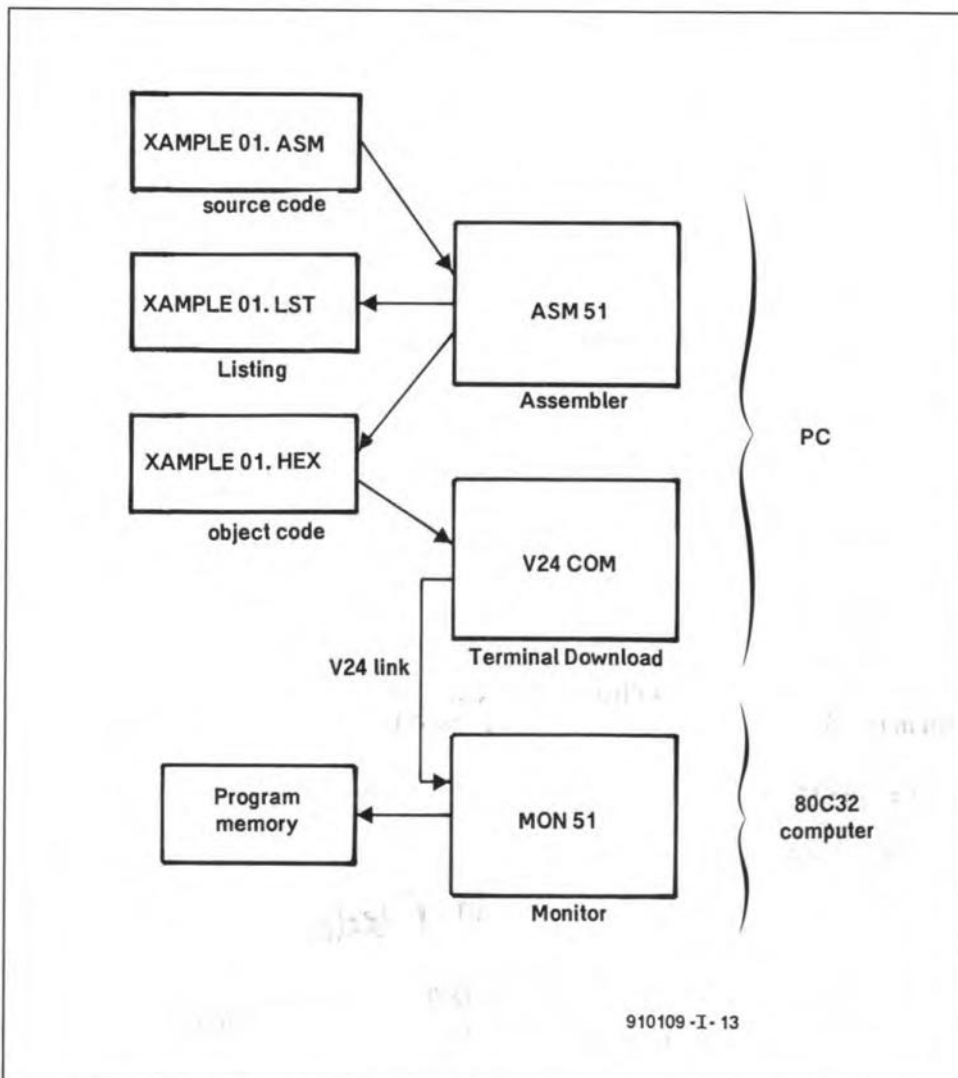


Fig. 4. From source code (written on the PC) to object code (loaded into the program memory of the 80C32 board).

Program memory

The program memory can have a size of up to 64 KByte, and holds the instructions for the controller. It can also be used to store look-up tables and constants. In this course we will use the controller with external program memory only, and that is why the external access (\overline{EA}) terminal, pin 31, is made permanently low. This disables the program storage function of the internal memory (if available). Access to the external program memory is signalled by \overline{PSEN} (pin 29) going low.

The program memory need not always be ROM or EPROM. Part of the address space may be formed by RAM or external I/O circuits. The program memory addresses 00000_H to 04000_H correspond to the lower addresses 00000_H to 04000_H of the EPROM, IC7, while program addresses 04000_H to 08000_H correspond to the area 04000_H to 08000_H in the system RAM, IC6. This means that instructions in the latter range are fetched from RAM, which allows new programs to be loaded into RAM, and executed from there, in spite of the presence of a programmed EPROM.

Data memory

MCS51 controllers are capable of writing to and reading from a data memory with a maximum size of 64 KByte. The data memory is used chiefly for the storage of large amounts of data. The RAM on the 80C32 computer board, IC6, may be used for data as well as program storage. It is located in the address area 00000_H to 08000_H , which means that if you store a program beginning at 04000_H in the data memory, it can be run equally as a program from 04000_H onwards, since access to program memory at these addresses selects the RAM also.

Most microcontroller applications can be run with very little data memory, and consequently the internal RAM is often used instead of an external data memory. The internal RAM has a size of 128 bytes in the 8031 and the 8051, and 256 bytes in the 8032 and the 8052. Since in this course a maximum of 128 bytes of internal RAM will be used, all these controllers can be used without problems.

The internal memory is also used to contain the controller registers. The stack, used for subroutine management, is also located in the internal RAM, so that the actual number of bytes free for use by the programmer is usually smaller than 128.

The internal RAM range between 20_H and $2F_H$ is the so-called bit-addressable range, which is used by certain bit manipulation instructions to address, change or call up any single bit. Finally, the monitor EPROM occupies a few bytes in the inter-

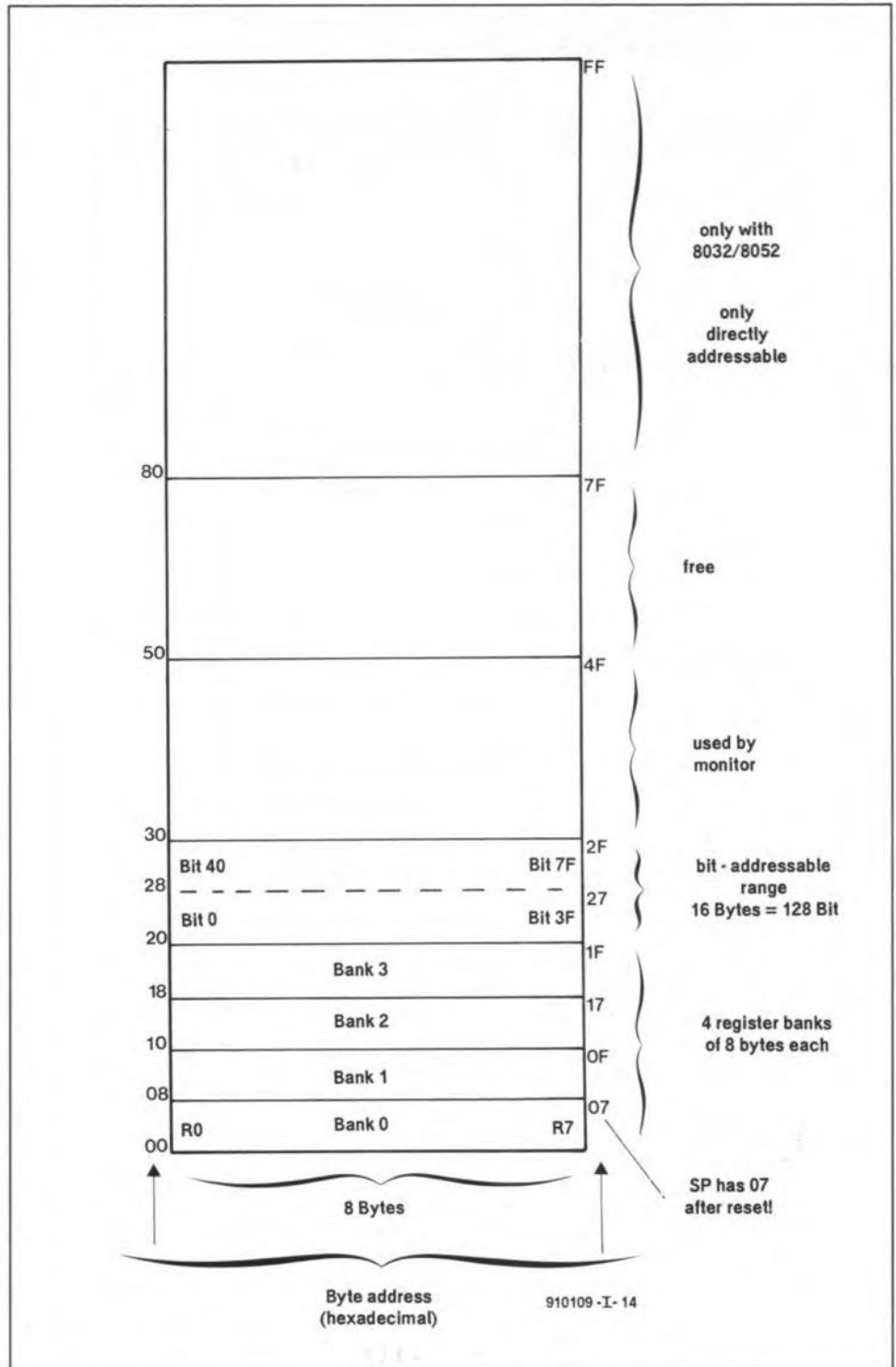


Fig. 4. Map of the internal RAM memory (128 or 256 bytes, depending on the processor type).

nal RAM, of which the address division is shown in Fig. 5.

What's next?

The next instalment in this series will show a short overview of the MCS51 instruction set and the SFRs. The first programming examples shed light on the processing of external signals and the data transfer via the V24 (serial) interface. Also, a few important monitor routines will be explained that can be put to good use in your own programs. □

References:

1. "80C32/8052-AH BASIC single-board computer". *Elektor Electronics* May 1991.
2. 8-bit embedded controller handbook, 1990. Intel Corp. order number 270645.

The course diskette (5¼-inch 360 KByte MSDOS format) mentioned in this article is available under order number **ESS 1661**, and the EPROM containing the monitor program EMON51 under order number **ESS 6091**. Prices and ordering information concerning these items may be found on the Readers Services page elsewhere in this issue.

MINI SQUARE-WAVE GENERATOR

by K. Schoenhoff

A handy square-wave generator is described which provides fifty-seven stable frequencies for the testing of digital circuits. It is battery operated and its output signal is TTL compatible.

AMONG the products of Seiko-Epson is a series of programmable frequency-generator ICs with square-wave output. These CMOS circuits, housed in a 16-pin DIL case, contain a stable crystal oscillator and two programmable dividers.

The difference between the 11 members of the series lies primarily in the fundamental oscillator frequency and in the operation of the second divider stage. Oscillator frequencies of 1 MHz, 796 kHz, 600 kHz, 153.6 kHz, 100 kHz, 96 kHz, 60 kHz, and 32.768 kHz, are available. The first divider is set by the three bits at inputs CTL1–CTL3 and provides

divisors of 1, 2, 3, 4, 5, 6, 10 and 12. The second divider is programmed by the three bits at inputs CTL4–CTL6 and provides either eight decadic divisors in the range 1–10⁷ or eight binary divisors in the range 1–128. The series also contains a type (86500) that has no oscillator and is thus nothing but a programmable divider.

All ICs in the series require the same supply voltage (5 V), and have identical pinouts. Most of them draw a current of only about 0.5 mA, but the 1 MHz, 768 kHz and 600 kHz types draw up to 2.0 mA. The output of all types is TTL compatible, and is in most cases



a true square-wave (duty factor=50%). Only when divider 1 is set to 3 or 5, and divider 2 to 1, for instance, to obtain an output of 333.3 kHz or 200 kHz, does the output signal become asymmetrical.

All ICs have a direct oscillator output (FOUT, pin 11), which can be used independently. This output is disabled by making R(reset) pin 14 logic 1.

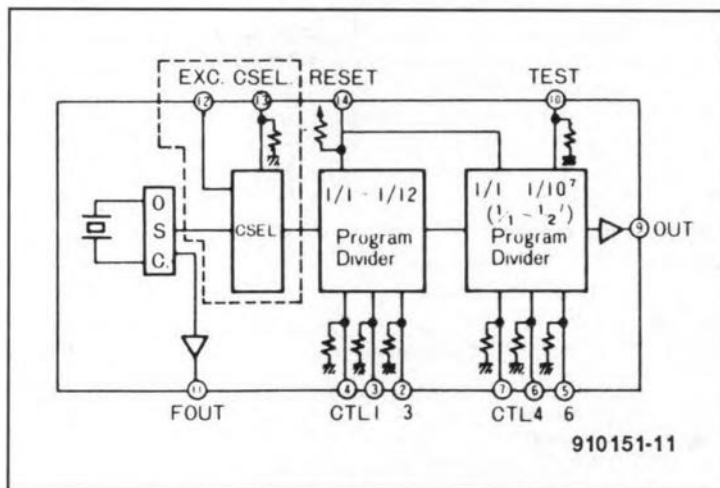


Fig. 1. Block schematic of the frequency-generator IC.

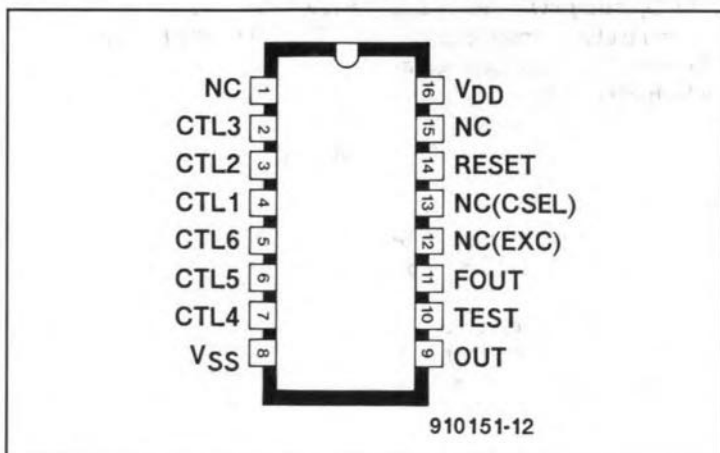


Fig. 2. Pinout of the frequency generator IC.

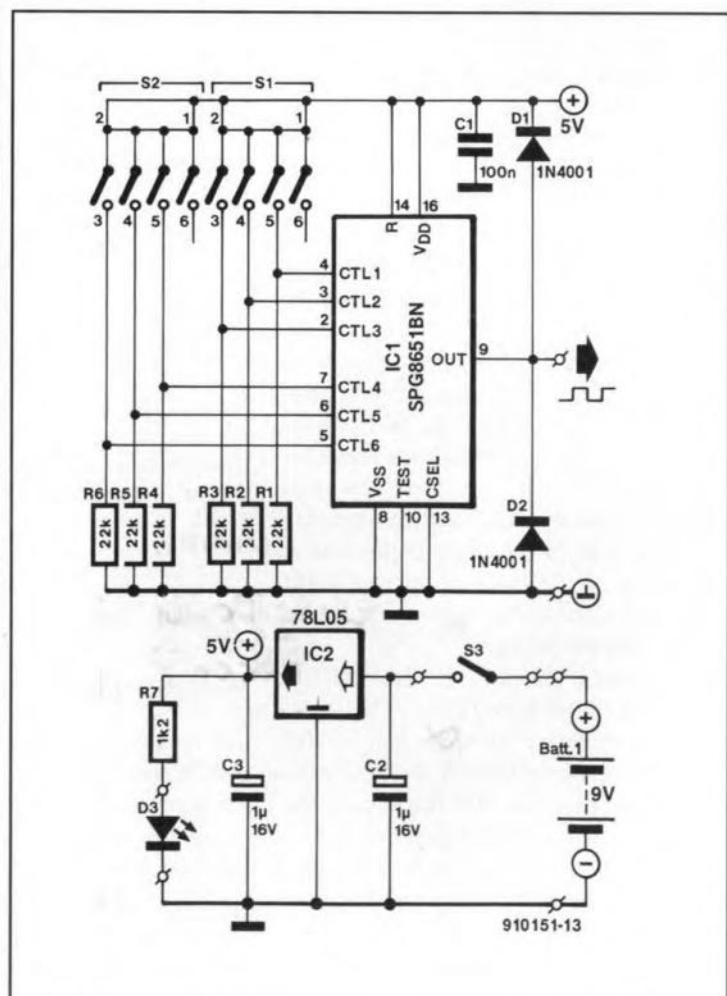


Fig. 3. Circuit diagram of the mini square-wave generator.

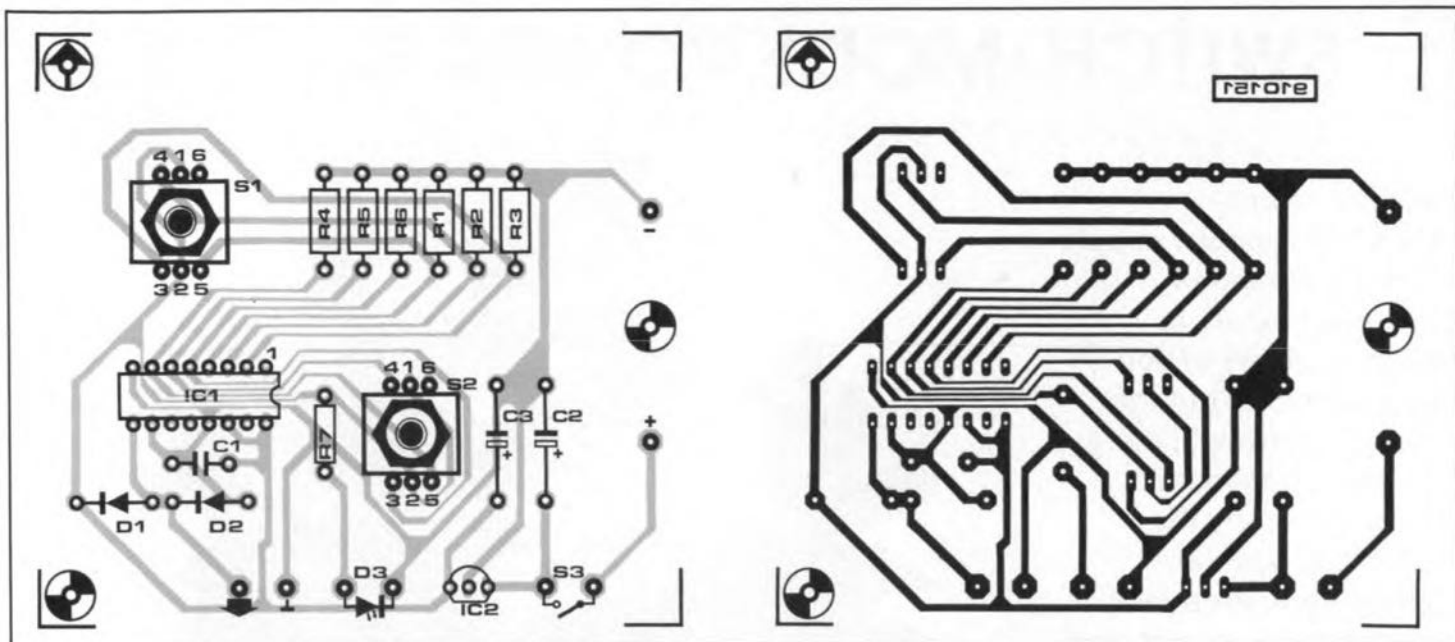


Fig. 4. Printed-circuit board for the mini square-wave generator.

Resistors:

R1-R6 = 22 kΩ

R7 = 1.2 kΩ

Capacitors:

C1 = 100 nF

C2, C3 = 1 μF, 16 V

PARTS LIST

Semiconductors:

D1, D2 = 1N4001

D3 = LED, 3 mm. high efficiency

IC1 = SPG8640BN (see text)

IC2 = 78L05

Miscellaneous:

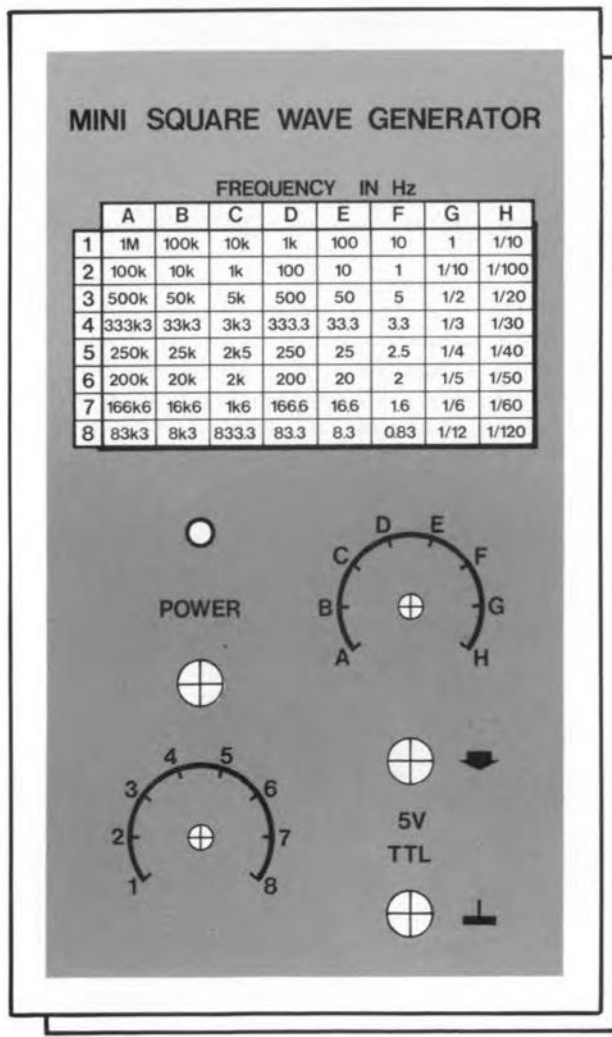
S1, S2 = BCD switch for PCB mounting

S3 = single-pole, single-throw switch

Bt1 = 9 V battery with connecting clip

Man-made fibre housing

PCB 910151



910151-F

Fig. 5. Front panel foil for the generator.

Pin 12 (NC - EXC) of the SPG8650 is an input for an external clock oscillator; the logic level at pin 13 (NC - CSEL) determines which clock is used (1=extern; 0=intern).

In the present circuit, it is possible to use either the Type SPG8640BN ($f_{max}=1$ MHz) or the SPG8651BN ($f_{max}=100$ kHz). In the prototype an 8640BN was used; the circuit of this is shown in Fig. 1 and its pinout in Fig. 2. This type is ideal for use in a universal test instrument, since it provides 57 stable frequencies in the range 1-8.33 MHz.

The circuit diagram of the generator in Fig. 3 is, of course, dominated by IC1. The only other items are six pull-down resistors, R1-R6, the unavoidable binary-coded decimal (BCD) coding switches, S1 and S2, and a simple battery power supply.

The low-drop voltage regulator, which ensures low power dissipation, provides a stable 5 V output. The two capacitors, C2 and C3, ensure smooth operation of the 78L05.

On-off indicator D3 must be a high-efficiency type that, in spite of the relatively high bias resistor, lights brightly. A normal LED would draw as much, or more, current as the rest of the circuit together.

Since it is required to keep the unit as small as possible, and independent of the mains supply, the choice of battery poses a slight problem. The use of a 9 V (PP3=6F22)

battery and 78L05 regulator, as used in the prototype, offers the best solution. It is possible to use four 1.5 V HP7=AA=RG batteries, but these take up more room than a 9 V type. Also, their discharge characteristic is against them: the regulator does not work with an input below 5.5 V. An up-to-date, but rather more expensive, solution would be to use two 3 V lithium batteries.

Construction

The unit is best built on the printed-circuit board shown in Fig. 4. Populating it should not present any problems.

The coding switches are upright types that can be obtained with screwdriver or knob control

The foil for the front panel—see Fig. 5—carries a copy of the frequency table from the Seiko-Epson data sheet. When different ICs from the series are used, the table from the relevant data sheet should be cut out to replace that in Fig. 5. ■

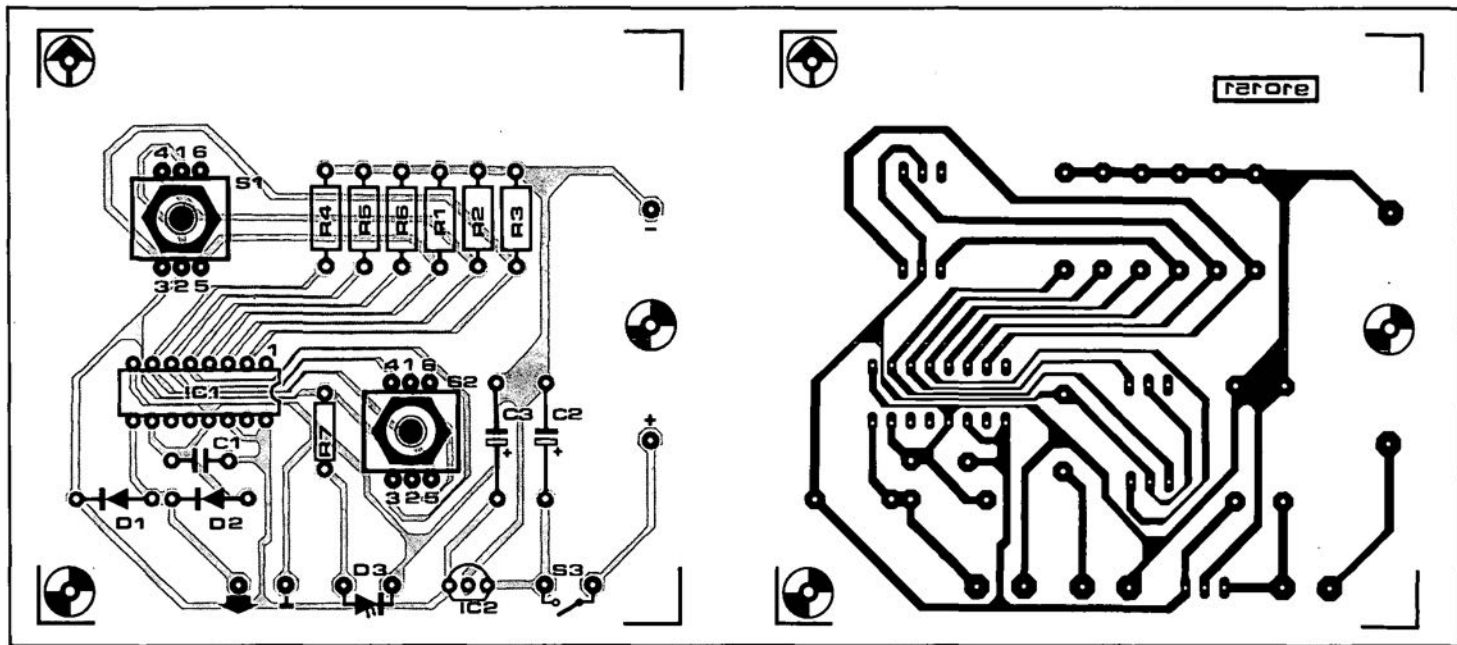


Fig. 4. Printed-circuit board for the mini square-wave generator.

CORRECTIONS & TIPS

Low-frequency counter

(January 1992, p. 44)

The parts list on p. 44 erroneously states that capacitors C_{11} and C_{12} are tantalum types. Since the polarity of the voltage across these capacitors may be inverted, the capacitors should be bipolar aluminium types.

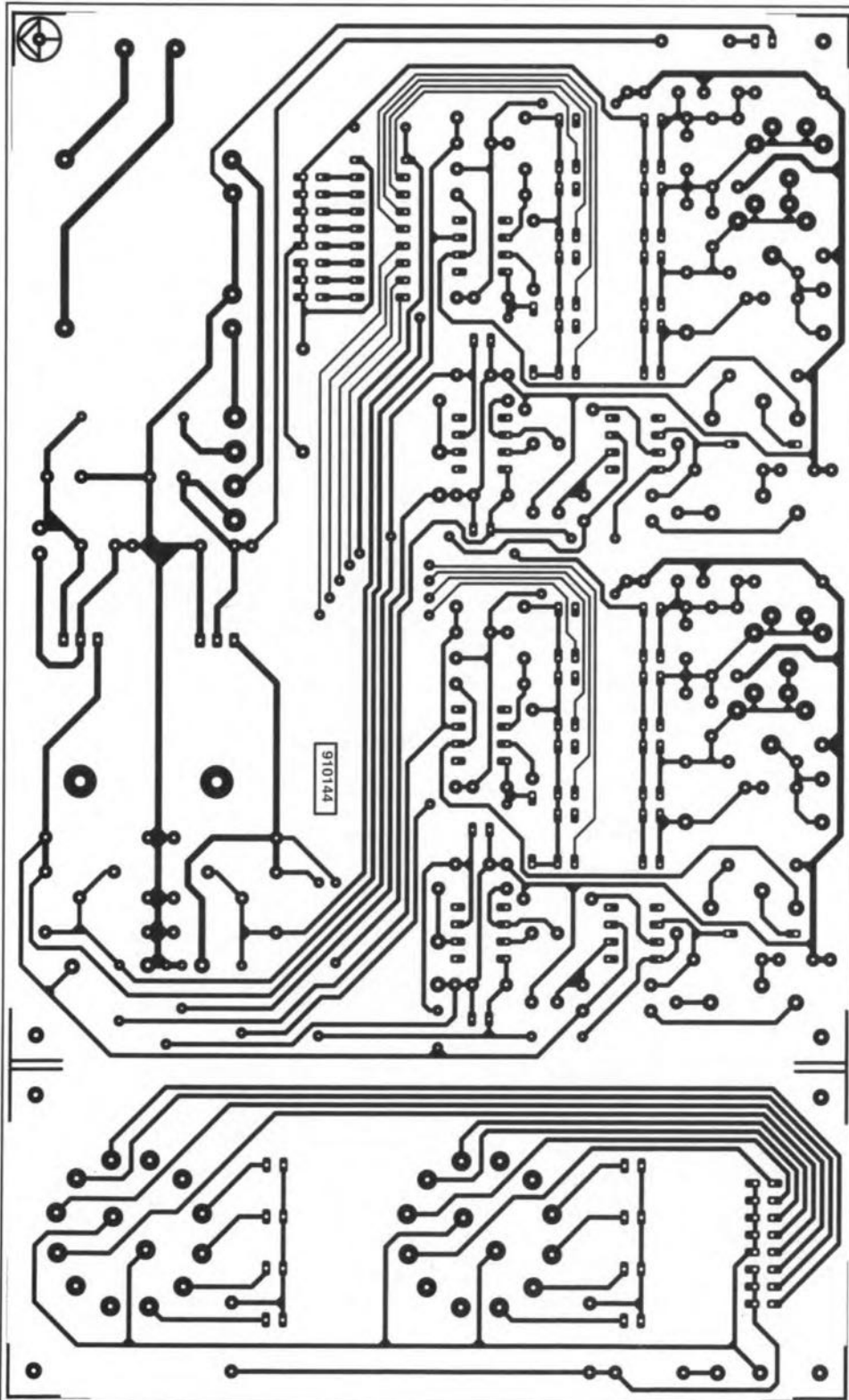
Under 'Construction' on p.45, it is stated in the penultimate paragraph that 'the connection between the input socket and C_{10} must be single screened cable'. In fact, the connection is so short that screened cable is not

necessary.

Measurement amplifier

(February 1992)

Owing to a misunderstanding, the track side of the printed circuit board (p.23) was not included with the article in our February 1992 issue. Our apologies for this oversight. The missing drawing is shown below.



Automatic cycle lights

(July/August 1991, p. 49)

Sir—In the construction of 'Automatic cycle lights', I have encountered three problems.

1. Triggering of IC_{1b} at input $-T$ (pin 11) and R (pin 13). A trailing edge at $-T$ triggers the IC if R is high. It is, however, possible that R is still low or is just changing state. A (not very elegant) solution to this is to connect the line from \bar{Q} (pin 7) to R to junction R_2-D_1 (+ battery) via a $1\text{ k}\Omega$ resistor.
2. A short pulse caused by the switching on of the battery triggered input $+T$ of IC_{1b} (pin 4), which switched the battery off again. This was cured by connecting a 470 nF capacitor between $+T$ and earth.
3. Triggering at $+T$ of IC_{1b} was so sensitive that even a tiny movement of the bicycle causes the battery to be switched off. In other words, if you don't hold the bicycle absolutely still, its lights will flash on and off. The sensitivity can be made variable by replacing resistor R_4 by a $100\text{ k}\Omega$ preset

Helge Bergmann, Hannover

Mini square-wave generator

(February 1992, p. 60)

Sir—On page 61 of 'Mini square-wave generator', you refer to the 7805 regulator as a "low-drop regulator, which ensures low power dissipation". I would disagree with that description: in my books that regulator is definitely not a low-drop type.

P. Thompson, Bristol

You are right: we apologize for that error.

Editor

Build a compact-disc player

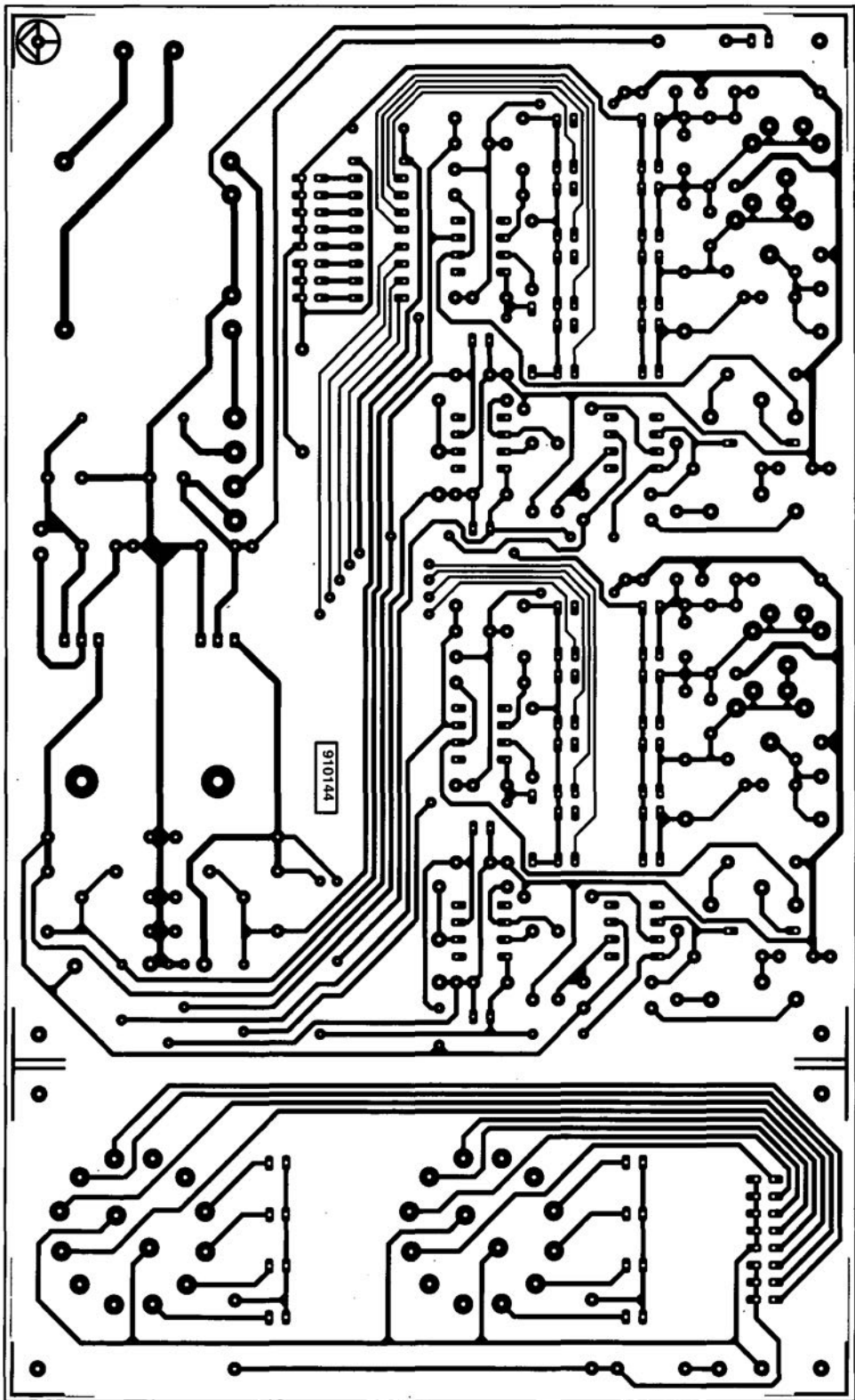
(January 1992, p. 36)

Sir—I think that your contributor, T. Giffard, in his article 'Build a compact-disc player' must be a lot more wealthy than I am if he considers £249 "an affordable price" (available from only one advertiser in your January issue!). Especially as Philips' own personal compact disc player complete and ready made costs only £149 (from Argos).

J. Easton, Watchet, Somerset

The price mentioned in the advertisement is HFL (Dutch guilders) 249 (equivalent to about £80.00).

Editor

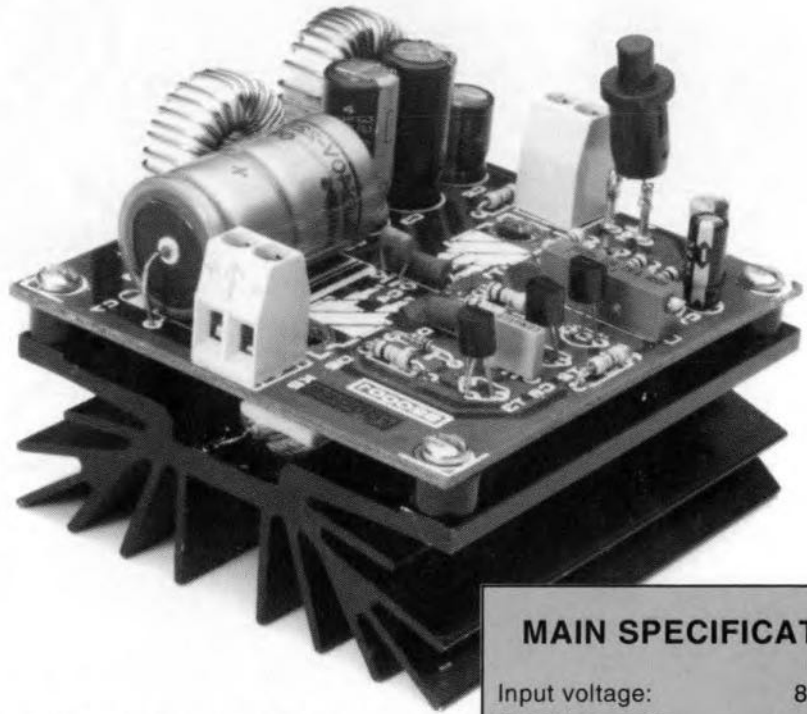


910144



SWITCH-MODE POWER SUPPLY

A reliable, sturdy switch-mode power supply is not the easiest of electronic circuits to design, as many of you may have found out the hard way. However, there is no reason to remain forever put off by the thoughts of special transformers, the mains potential in the circuit and RF interference. The circuit we present here is aimed at providing a hands-on introduction to working on an SMPSU. What we have in mind is a versatile circuit that is safe and based on inductive components that are available ready-made.



MAIN SPECIFICATIONS

Input voltage:	8–35 V d.c.
Output current:	max. 4.5 A
Output voltage:	4.8–20 V
Output ripple:	<100mV _{pp}
Efficiency:	66–88%
Dimensions:	75×70×55 mm

by J. Ruffell

MODERN electronic equipment generally needs one or more d.c. power supplies. Depending on the type of equipment, either a linear or a switch-mode stabilizer is used. Compared to the linear supply, the SMPSU (switch-mode power supply) has some distinct advantages, including smaller size and higher efficiency for the same output power. In spite of these advantages, the SMPSU is by no means an ideal direct voltage source, because the output voltage is usually not very 'clean'. In practice, however, an SMPSU is often the cheapest, and in a few cases, the only, alternative. These days, SMPSUs are applied in, among others, computers, TV sets and car radio boosters. The power supply described here efficiently converts a high, unregulated, voltage into a low, regulated, voltage.

The control IC

The heart of the SMPSU is an integrated circuit originally developed to function as a reset controller in microprocessor systems. Although generating a reset signal is far simpler than regulating a direct voltage, this IC contains everything to perform the latter function quite well.

For convenience, the block diagram of the control IC (IC1) is shown in the circuit diagram of the power supply (Fig. 1). The plastic TO-226 enclosure contains a reference source, a comparator, an attenuator and a

switching transistor. Three points in the internal circuit of the IC are bonded out to pins for connection to external parts. The internal voltage divider is dimensioned such that the toggle point of the comparator lies at 4.61 V. When the input voltage (applied to pin 2) exceeds this threshold, the open-collector output of the IC (pin 1) switches to high impedance. Because of the comparator's hysteresis, the switching transistor does not start to conduct until the input voltage is lower than 4.59 V. The potential at pin 3 of the MC34064 must be at least 1 V below that at the analogue input. This difference is necessary because the IC (peculiar but true) is powered via its input. The input/supply current is about 400 μ A at a voltage ($U_{2,3}$) of 5 V.

The control loop

In power supply terminology, the present SMPSU is classified as a non-continuous automatic regulation loop, which is basically the same as an on/off regulator. The main characteristic of this class of regulator is that the regulating element (T1) has only two states: 'on' (conduct current) and 'off' (block current).

MOSFET T1 switches under the control of IC1. However, because of the small output current (only 10 mA) and the presence of an on-chip diode between the input and the output, the open-collector output of the con-

trol IC can not drive the gate of the p-channel MOSFET directly. Hence, a two-transistor driver, T2-T3, is required between the regulator and the power FET. In the configuration shown in Fig. 1, the two transistors have the double function of driver and inverter.

IC1 measures the stabilized voltage via an adjustable voltage divider, P1-R2-R3. The setting of the preset potentiometer determines the divide ratio and thereby the level of the output voltage. Capacitor C9 provides local decoupling for the resistors, which is essential for the stability of the switching frequency.

Capacitor C1 speeds up the response of the regulator to output voltage variations. A value of 22 μ F is sufficient to reduce the level of the triangular-shaped ripple across C5 to 50 mV_{pp} or so.

Two passive filters

The network consisting of D2, C5 and L1 is typical for step-down (buck) switch-mode power supplies. The oscilloscope screen dump in Fig. 2 shows the input and output voltage of this network, as well as the current that flows through the inductor (see Fig. 3). The direct voltage applied to connector K1 is first 'chopped to pieces' with the aid of T1. Next, the D-C-L network 'calculates' the average value of the pulsating drain-voltage. This results in a direct voltage across C5 that is proportional to the ratio $T_{on}/(T_{on} + T_{off})$.

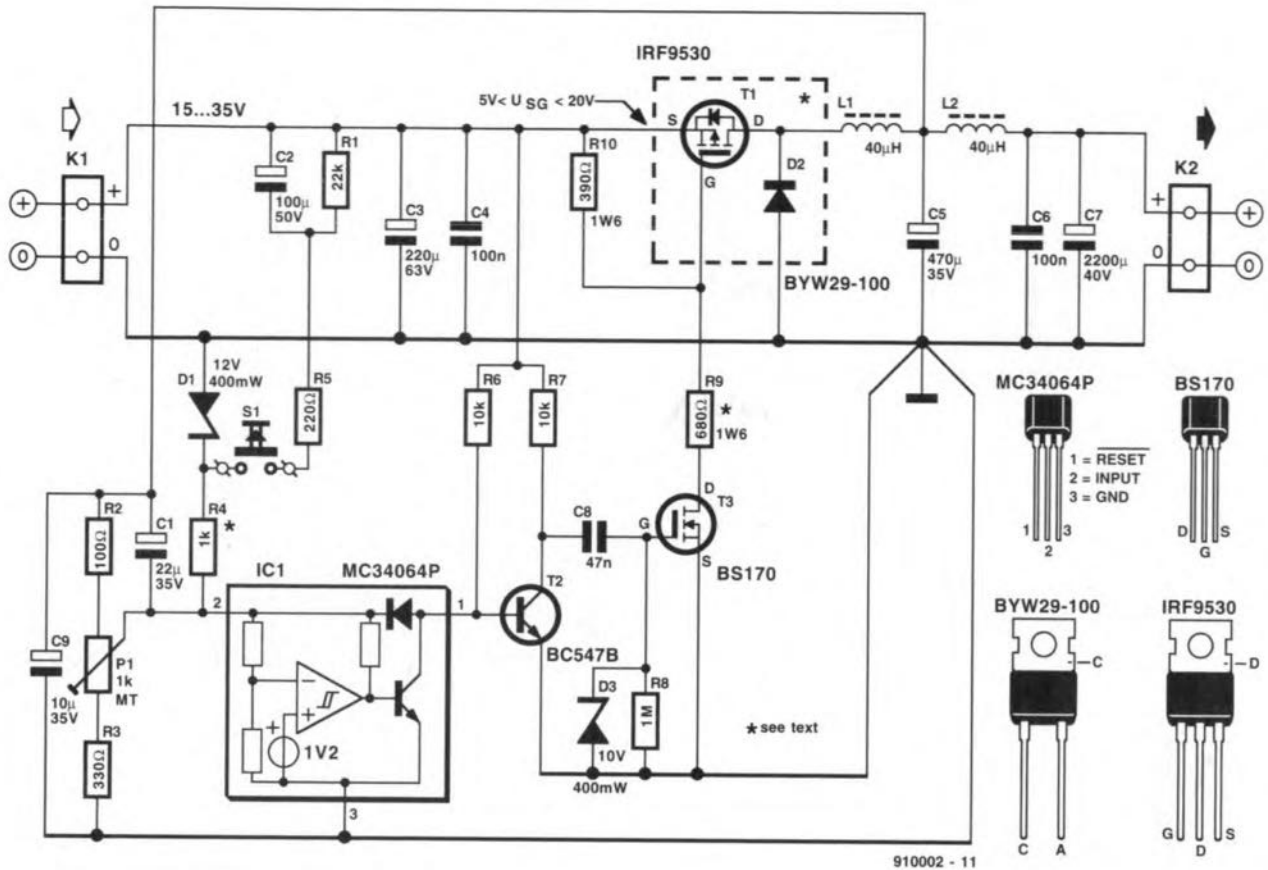


Fig. 1. Circuit diagram of the switch-mode power supply.

The remaining ripple across C_5 is inherent to the non-continuous regulation. Fortunately, the frequency of this ripple is relatively high, which makes it easy to suppress with the aid of a second-order passive filter. Here, this network consists of L_2 and C_7 . Even at the lowest frequency of about 15 kHz, the ripple is still attenuated by a factor of more than 300.

The conductance (which is the reciprocal of the internal resistance) of the inductor is about 40 siemens. This causes a small voltage loss that can not be compensated by the regulation system.

The components that form the $D-C-L$ filter must meet a number of special requirements. Firstly, the inductor must not reach

saturation, because that would mean a sudden reduction of the self-inductance, which in turn causes the current to rise to an unacceptably high value. The linear characteristic of i_{L1} (see Fig. 2b) indicates that the core used here shows no sign of saturation even at maximum load. The core material also determines the efficiency. If the core of L_1 is made

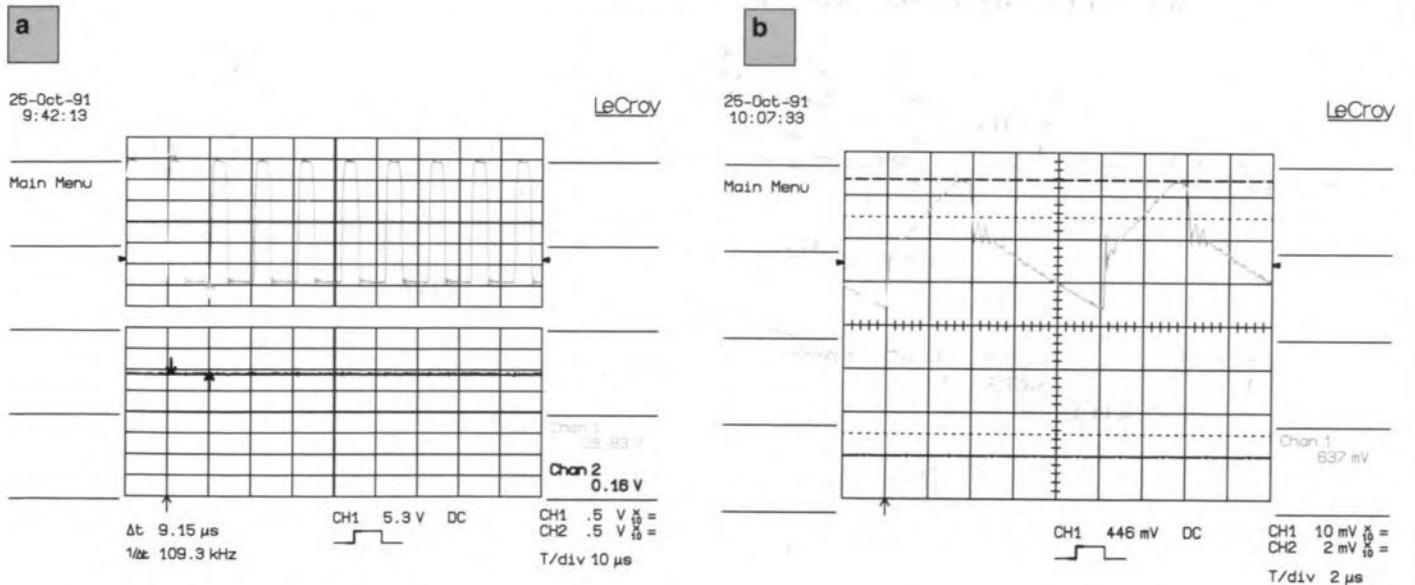


Fig. 2. Input and output voltage of the DCL -filter (oscilloscope a), and the current through the inductor (oscilloscope b).

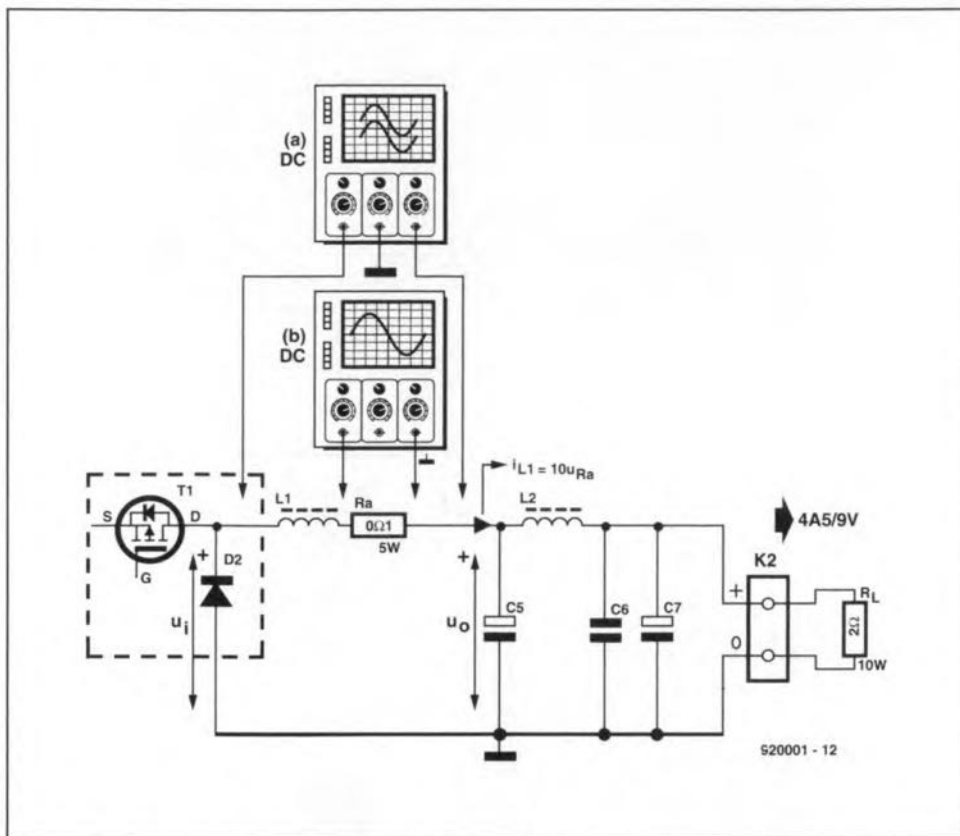


Fig. 3. The inductor current is measured indirectly, i.e., via the voltage across resistor R_a . This test set-up produces the oscillograms shown in Fig. 2.

of pure iron, and thus forms a conductor, the changing magnetic field will cause eddy losses. By contrast, ferromagnetic materials (or ferrites) are insulators in which eddy currents occur to a much smaller extent, resulting in greater efficiency. These materials do have a problem, though, in that a high saturation field strength can only be achieved when the ring core has an air gap. The high magnetic resistance of air increases the number of turns required to achieve a certain inductance, which means that the size and volume of the inductor increase also. Furthermore, the field caused by leakage

emission around the air gap calls for extra attention in relation to certain radiation level restrictions. Arguably, winding such an inductor yourself is really out of the question, and that is why a ready-made type is used in the present supply. The Type SF-T10-30 inductor used here has a ferrite core, and its low cost certainly makes up for a slightly lower efficiency.

Diode D2 closes the current loop through L1 via C5 (and the load), and back to the inductor. The diode functions as a flyback device, and conducts only when the MOSFET is cut off. The forward current is of the order

of a few amperes, so that the dissipation depends strongly on the forward voltage drop. The continuous switching causes additional losses, because the diode will not switch off the moment the MOSFET starts to conduct again. This means that a reverse recovery current flows briefly (t_{rr}) when the cathode-anode voltage is relatively high. Evidently, this causes power loss, and the highest efficiency is therefore achieved only when the diode has a low forward voltage specification, and a short recovery time (t_{rr}). The BYW29-100 used here meets this combined requirement amply with the respective parameters of 0.7 V (at 5 A) and 25 ns.

Starting

The supply voltage for IC1 is derived directly from the voltage across C5. Initially, C5 will have no charge, so that the IC is not immediately powered when a direct voltage is connected to K1. The open-collector output then forms a high impedance, so that the SMPSU will not start on its own. To bring life to the oscillator, switch S1 must be briefly pressed. This closes a circuit, and produces a voltage at pin 2 of IC1 that enables the switching transistor to start conducting. Consequently, the power FET, T1, receives a negative gate-source voltage via T2 (an n-p-n transistor) and T3 (a VMOS FET), so that the resistance of the source-drain channel drops to about 0.3Ω , and the voltage across C5 rises. Within a very short time, the regulation loop stabilizes the output voltage at the set level. The input voltage of IC1 then toggles continuously between the two switching levels, 4.59 V and 4.61 V. The toggle rate is determined chiefly by the input voltage level.

Capabilities

There are a number of alternatives to the circuit as shown in Fig. 1. For instance, the supply may be made suitable for lower input

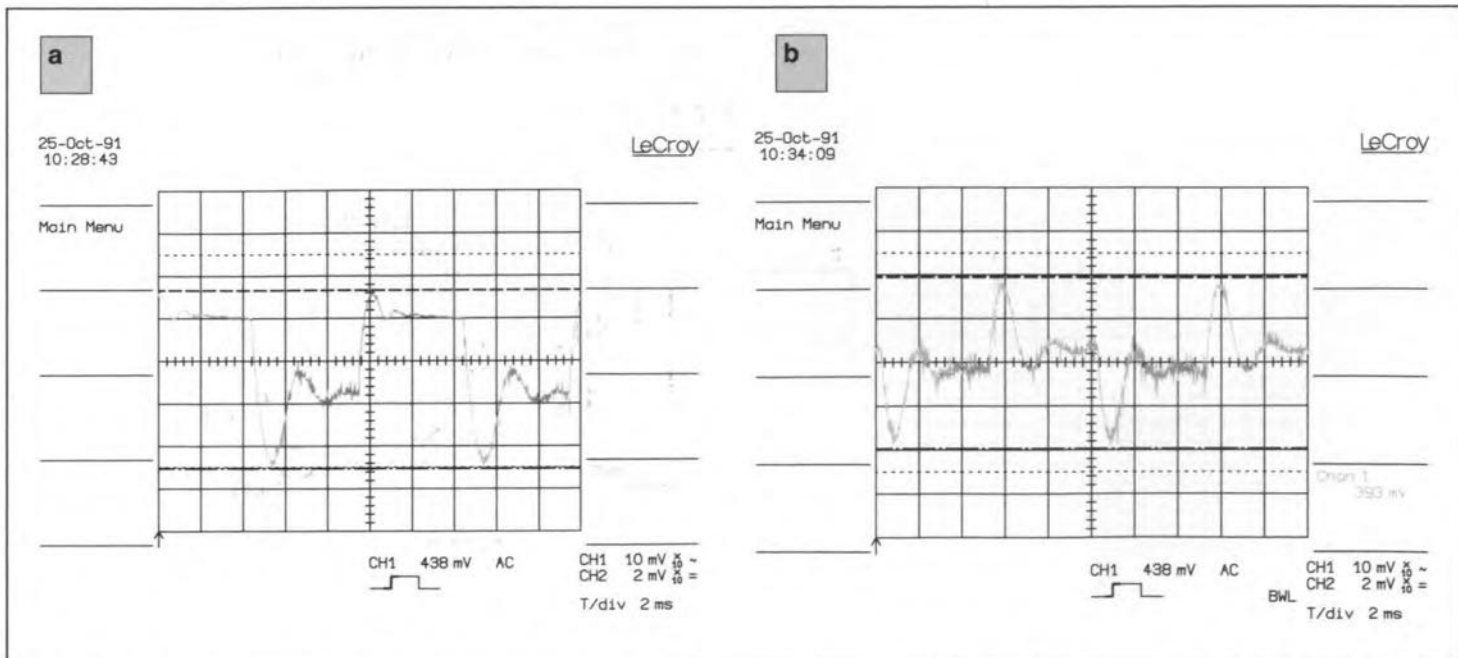


Fig. 4. The response of the SMPSU to fast load variations was measured twice: once with $R_{L1}=8\Omega$ (oscillogram a), and once with $R_{L1}=\infty$ (oscillogram b). See also Fig. 5.

voltages by changing some of the component values. More about this later—first, we have a look at the capabilities of the 'standard' circuit as drawn in Fig. 1.

The output voltage is adjustable between 4.8 V and 20 V with the aid of a multiturn preset potentiometer, P1. Although the power supply is suitable for input voltages from 15 V to 35 V, the input voltage must always be 2 V higher than the desired output voltage. The SMPSU module is capable of supplying a continuous output current of 4.5 A, and a peak output current of 8 A.

Table 1 shows an overview of the efficiency measured at a number of output voltages. A separate column gives the corresponding values that may be expected from a typical linear power supply. Clearly, the SMPSU is more efficient, particularly when there is a relatively large difference between the input and the output voltage. Hence, a smaller, less expensive, heat-sink may be used. The temperature of the heat-sink indicated in the parts list rises about 30 °C above the ambient temperature when the SMPSU is fairly heavily loaded (in: 30 V; out: 5 V/4.5 A).

The level of the switching noise across the output capacitor, C7, is independent of the set voltage. On our prototype, we measured a noise level of only 70 mV_{pp} at an input voltage of 30 V and an output voltage of between 5 V and 20 V at 4.5 A. Provided the load is adequately decoupled, this noise level will rarely cause problems.

The oscillograms in Fig. 4 show the output voltage variation as a result of a sudden load change (this is the so-called dynamic behaviour of the supply). The circuit used for this test is shown in Fig. 5.

Input voltage section

The basic version of the SMPSU may be connected directly to a basic rectifier circuit. The circuit diagram of the input voltage section consisting of a transformer, a bridge rectifier and a reservoir capacitor, is shown in Fig. 6. This circuit must be built and wired separately, observing safety precautions in regard of the mains voltage present on the primary winding of the transformer.

The two 12-V secondary windings of the toroid transformer are connected in series, so that a direct voltage of about 30 V is obtained across C1. The input, K1, of the SMPSU is connected to the terminals of the reservoir capacitor, C1, via a fast 5-A glass fuse. The fuse protects the MOSFET and the transformer from being overloaded. When the output of the supply is accidentally short-circuited, fuse F2 usually remains intact because the oscillator then stops working virtually instantly.

Lower input voltages

It is possible to use the SMPSU module with input voltages lower than 15 V, for instance, with a 12-V vehicle battery. Only two resistors then need to be changed. The first is R9, which must be made smaller to ensure that

Switch-mode power supply					Linear power supply			
U_o (V)	I_{in} (mA)	P_{in} (W)	P_o (W)	η (%)	I_{in} (mA)	P_{in} (W)	P_o (W)	η (%)
20	1 110	27.8	24.4	88	1 300	32.5	24.4	75
10	648	16.2	12.2	75	1 300	32.5	12.2	38
5	372	9.3	6.1	66	1 300	32.5	6.1	19
$U_{in} = 25V$					$I_o = 1.22A$			

Table 1. Performance of the switch-mode power supply compared to that of a typical linear supply.

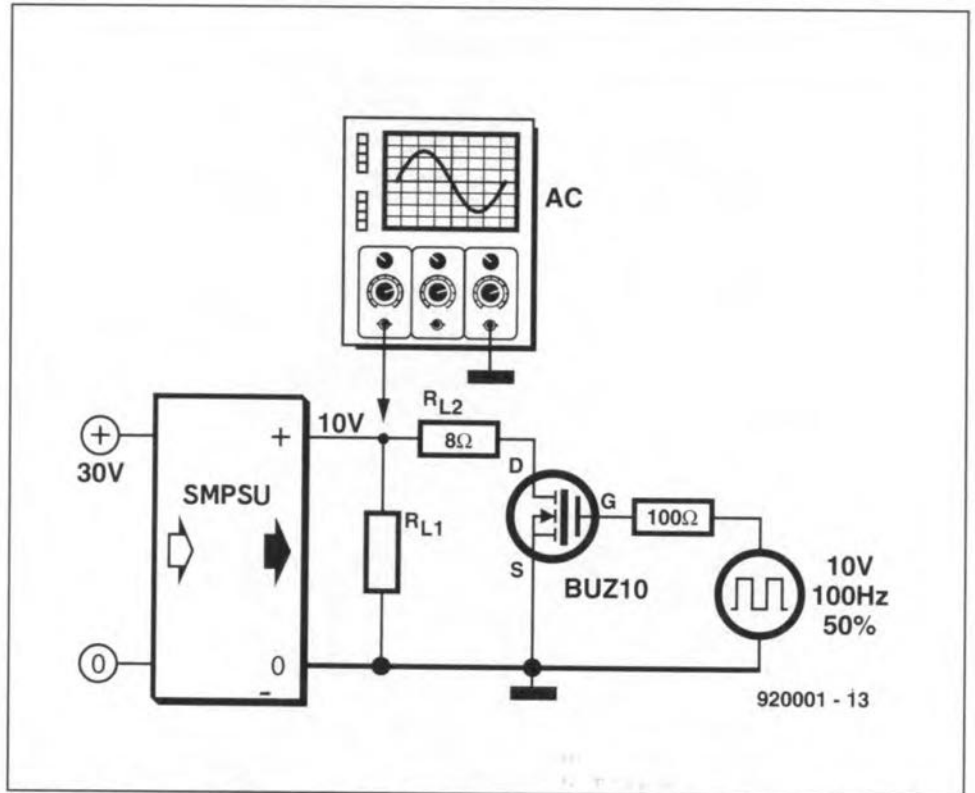


Fig. 5. This circuit allows us to check the response of the power supply to fast load variations. The results of the tests are shown in Fig. 4.

T1 is provided with a sufficiently high gate-source voltage. This means that, in principle, R9 could be replaced by a wire link, since the maximum value of U_{g-s} (20 V) is never reached at an input voltage of, say, 12 V.

However, experiments with the supply have shown that a value of 0 Ω has a negative effect on the quality of the output voltage. As a result of the almost unlimited discharging current that flows into the gate capacitance,

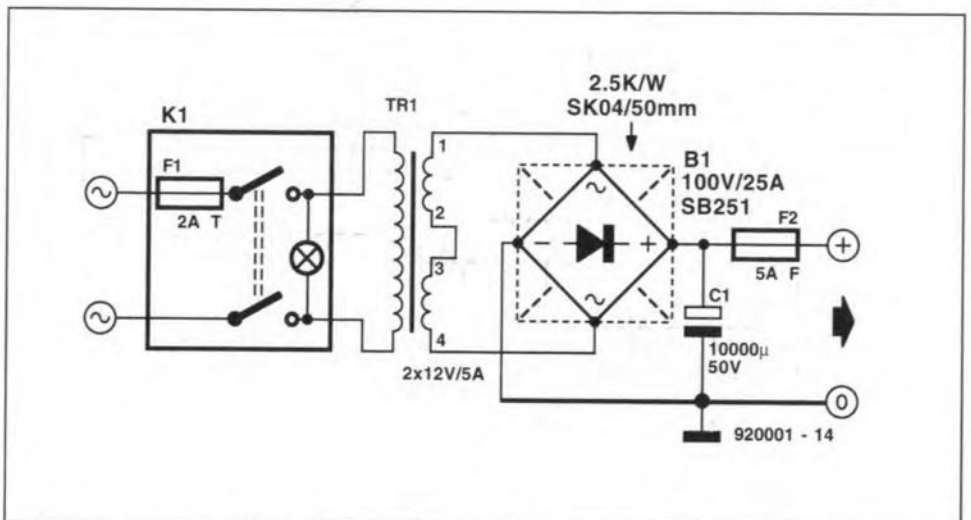


Fig. 6. This rectifier circuit supplies a direct voltage of about 30V, which is applied to the input of the SMPSU. The bridge rectifier must be fitted on a heat-sink.

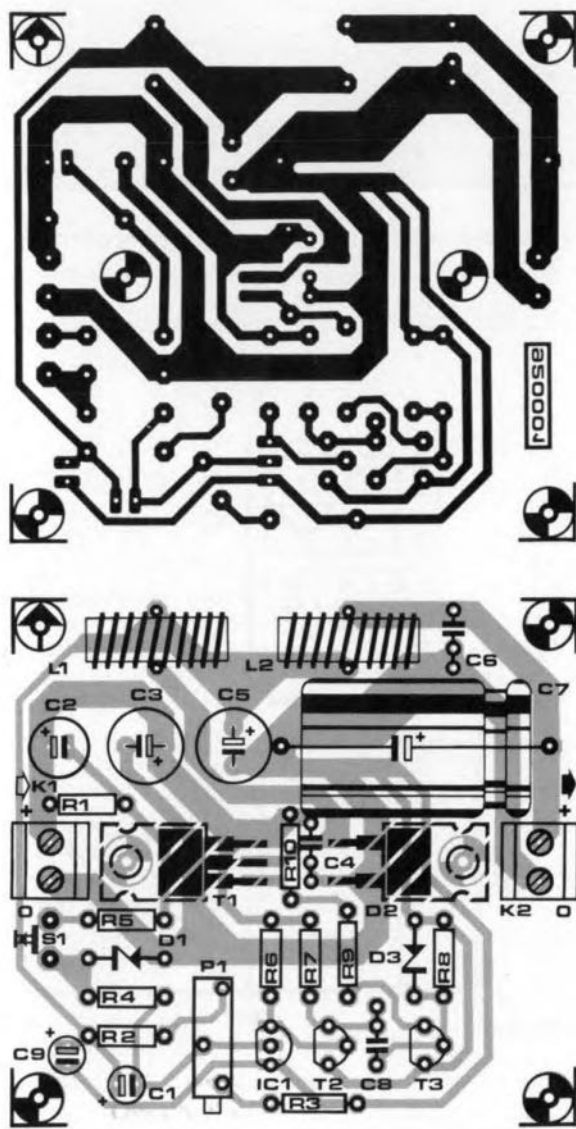


Fig. 7. Track layout (mirror image) and component mounting plan of the PCB designed for the SMPSU.

the level of the switching noise rises by a factor of three to four. Table 2 shows an overview of the recommended values of R9 for a number of input voltages. Also shown in Table 2 are the associated measured noise (ripple voltage) levels.

The second component, R4, need not be changed unless the supply refuses to start when S1 is pressed. If this happens frequently, R4 is best changed to 470 Ω at input voltages smaller than 10 V.

Construction hints

The printed circuit board designed for the switch-mode power supply makes the construction very easy. This is mainly because all dimensions are geared to the size of the heat-sink. When the board is completely populated, it is simply bolted on to the heat-sink to give a compact, robust unit. The PCB for building the SMPSU is available ready-made through our Readers Services.

U_{in} (V)	R9 (Ω)	U_{ripple} (mV _{pp})	U_{ripple} (mV _{pp})
8-10	0	45	55
10-12	120	35	50
12-15	150	45	60
15-35	680	55	70

Table 2. Suggested values of resistor R9 depending on input voltage.

Having acquired all the components, it is best to start with the preparation of the heat-sink (a Type SK64 from Fischer). First, place the bare board on top of the flat side of the heat-sink. Next, mark the locations of the six drill holes. Drill the holes at 2.5 mm (0.1 in) and tap a 3-mm ($\frac{1}{8}$ in) inside thread.

Continue with the component placement and soldering work on the board. Transistor T1 and diode D2 are held over till the last.

COMPONENTS LIST

Resistors:

1	22k Ω	R1
1	100 Ω	R2
1	330 Ω	R3
1	1k Ω *	R4
1	220 Ω	R5
2	10k Ω	R6;R7
1	1M Ω	R8
1	680 Ω 1.6W* (PR37)	R9
1	390 Ω 1.6W (PR37)	R10
1	1k Ω multiturn preset	P1

Capacitors:

1	22 μ F 35V radial	C1
1	100 μ F 50V radial	C2
1	220 μ F 63V radial	C3
2	100nF	C4;C6
1	470 μ F 35V radial	C5
1	2200 μ F 40V	C7
1	47nF	C8
1	10 μ F 35V radial	C9

Inductors:

2	SF-T10-30** (40 μ H/5A)	L1;L2
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Semiconductors:

1	12V 0.4W zener diode	D1
1	BYW29-100**	D2
1	10V 0.4W zener diode	D3
1	IRF9530**	T1
1	BC547B	T2
1	BS170	T3
1	MC34064P**	IC1

Miscellaneous:

2	2-way PCB terminal block (pitch 5mm)	K1;K2
1	push-to-make button	S1
1	SK64/75** Heat-sink (2.3K/W)	
1	Printed circuit board	920001

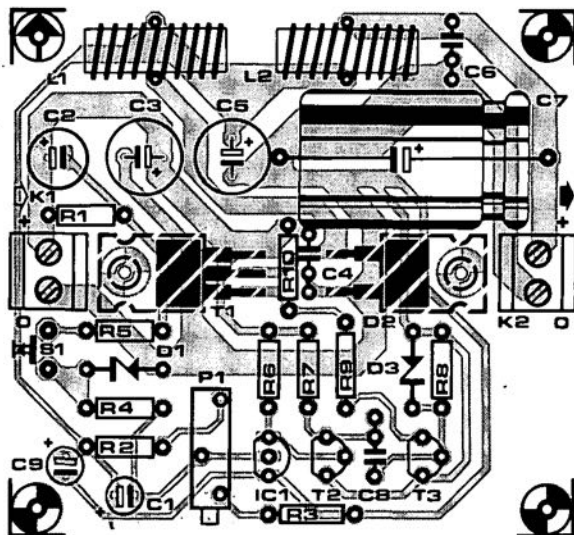
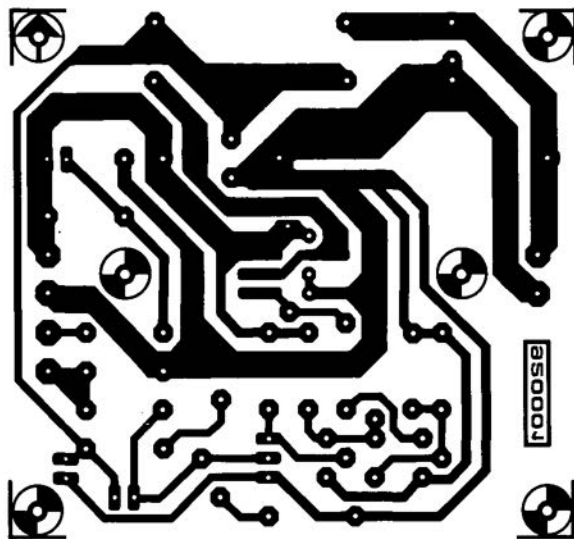
* See text.

** Available from C-I Electronics, P.O. Box 22089, 6360 AB Nuth, Holland. Fax: +31 45 241877.

Mount inductor L1 and resistors R9 and R10 at a small distance from the board to allow for their dissipated heat.

The power semiconductors, T1 and D2, are fitted at the track side of the board. Apply a small amount of heat-conducting paste to the backs of the metal tabs of T1 and D2 before bolting the devices to the heat-sink. Insulating washers are not strictly required, although when they are omitted the heat-sink is at a potential with respect to ground.

The operation of the circuit may be checked by performing a few basic measurements. Mind you: certain problems may not come to light until the supply is connected to a fairly heavy load. A good example of this is intermittent operation ('stuttering') of the unit as a result of insufficient decoupling of R2, R3 and P1 (check C9 in this case). In all cases, be sure to test the supply thoroughly at an output current of 3 to 4 A. ■



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