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ELEKTOR ELECTRONICS

THE INTERNATIONAL ELECTRONICS MAGAZINE

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AUDIO & HI-FI • COMPUTERS & MICROPROCESSORS • DESIGN IDEAS • RADIO, TELEVISION & COMMUNICATIONS • SCIENCE & TECHNOLOGY • TEST & MEASUREMENT

80C535 EXTENSION BOARD

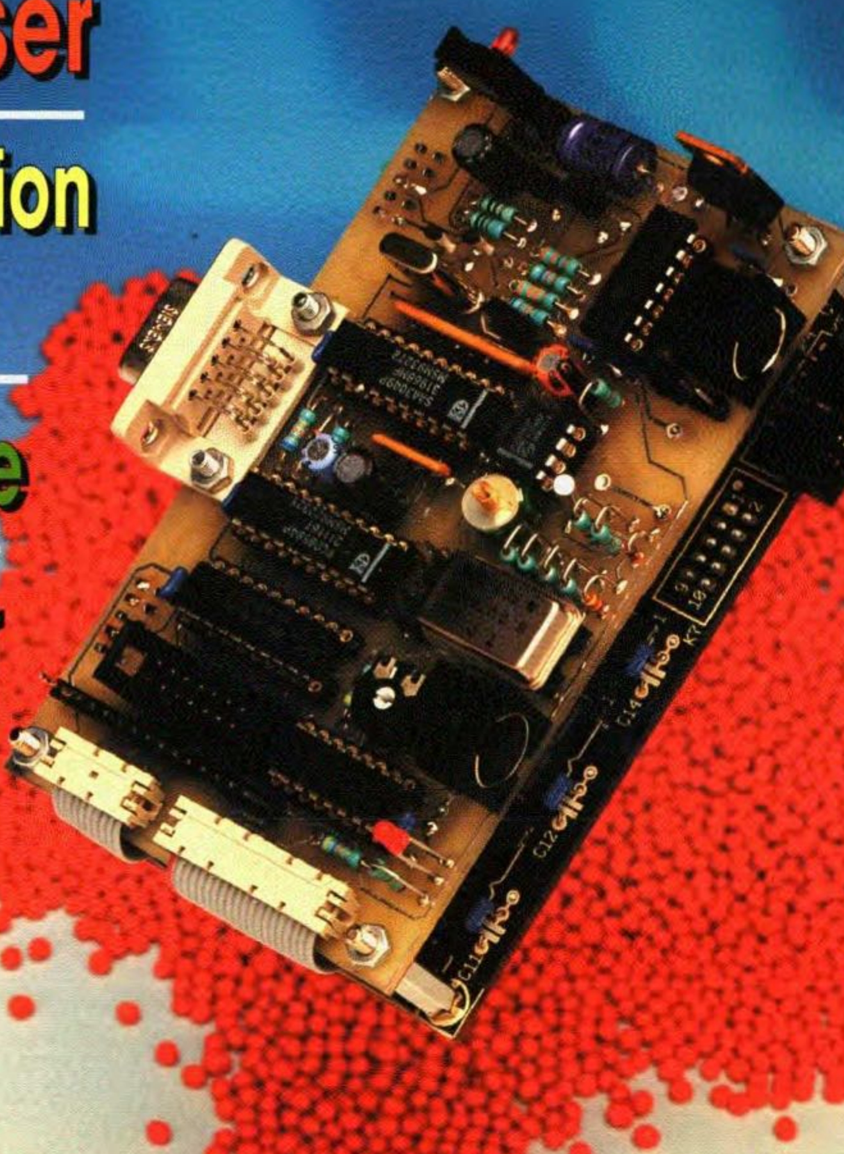
Intelligent EPROM eraser

Fuel consumption monitor

RS485 interface

I²C bus booster

IL300 linear optoisolator



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ELEKTOR ELECTRONICS

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Number 223
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  plus
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- Car battery monitor
- IR controlled switch
- Solid-state temperature sensor
- MIDI volume pedal (1)
- Electric charge meter
- Resettable fuse for caravans

  and more for your continued interest

Front cover
The popular 80C535 single-board computer we published a few months ago has a powerful microcontroller and can be used for a multitude of control applications. The extension card described on page 8 gives the 80C535 a number of new interfaces mainly intended for communications with the real world.

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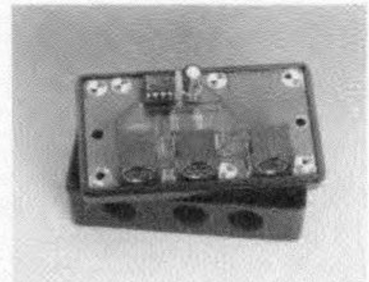
BOOK CREDIT VOUCHER. On **Page 73** we publish the last voucher that will entitle the holder to up to 30% discount on the purchase price of any Elektor Electronics book.



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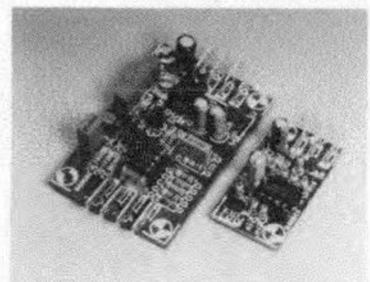

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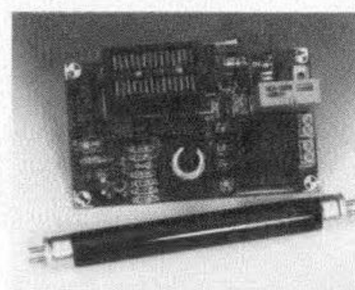

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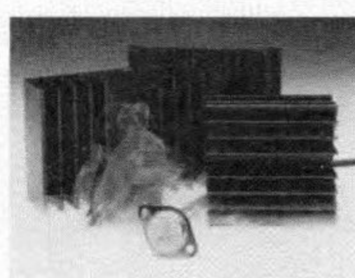

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ELECTRONICS SCENE

NEW TELEPHONE SERVICE FROM ENERGIS

In the UK's liberalised telecommnications environment, opportunities exist for those able to provide innovative services cost-effectively, a situation that has evolved since Mercury Communications was set up. As Mercury has grown to become a successful company over the past ten years, it has provided a spur to BT (British Telecom), especially with regard to business services, to the ultimate benefit of users throughout the country.

Recently, a new company, Energis[1], has received a full licence to operate telecommunications services throughout the UK. The company is owned by the National Grid Company PLC (NGC), in turn owned by the 12 regional electricity distribution companies in England and Wales. The power network of its parent company already consists of 7000 km of overhead powerlines and 600 km of underground ducts that can be used by Energis as the backbone upon which to build the network.

Whereas Mercury had to build its network from the ground up, Energis is able to take advantage of an existing high quality infrastructure. Thus it can be seen that the company is ideally placed to build up its telecommunications network.

The National Grid's 400-kV power network is managed from a national control centre which is as advanced as any in the world as, after all, continuity of electricity supply is of vital national importance.

The NGC network is linked to the twelve Regional Electricity Companies (RECs) and to Scottish Power. Through their ownership of wayleaves (the right-of-way rented to an electricity company to enable it to span private land) and networks, there is access to every business and home in the country.

Retail chains. The RECs also have customer databases, customer service systems and are already geared to bill customers quarterly. In addition, many have chains of retail stores, telemarketing centres and an established sales force. Most of these assets would have to be built from scratch for any other new operator.

Energis claims that it will employ the UK's first national 100% synchronous digital hierarchy (SDH) network to deliver its services. The SDH-based fibre optic cables for this are being installed on to the National Grid Company's electricity pylons at the rate of 80 km per week. This is being done, in some places, by replacing the existing earth wire with a new composite earth wire incorporating 24 optical fibres. In other areas, a 24-fibre optic cable is being wrapped around the existing earth wire. This is a well-proven technique employing a machine that pulls itself along the earth wire around which it simultaneously winds the optic cable.

Engineers carry out the dangerous task of working on live lines, whilst installing fibre-optic cable on an earth wire of the electricity transmission network in Britain

The SDH network will be operating at 2.4 Gbit s^{-1}. This is four times the speed of the majority of today's fibre optic networks and, combined with the fact that 24-fibre cables are being installed, it provides a major opportunity to reduce costs.

Network integrity. As the SDH network provides the option of transmitting the same information via two diverse routes simultaneously, a high level of network integrity can be achieved. This is further enhanced by the fact that all network and individual customer lines will be software defined, keeping human/mechanical intervention to a minimum, with the result that the network will be inherently more reliable. In fact, the pilot network which has been on trial for the past two years has consistently achieved levels of 99.98% reliability.

Energis has won its first customer. It has signed a ten year contract to provide the BBC[2] (British Broadcasting Corporation) with a Managed Broadcast Network for the distribution of TV and radio services throughout the UK. The BBC runs a nation-wide broadcast network to provide TV services for BBC1 and BBC2 as well as radio services for Radios 1, 2, 3, 4 and 5. The facility provided by Energis will link the BBC's studios and programme production centres to their TV and radio broadcast centres, so providing the final link to the UK audiences.

It is investing over £ 100 million in the first phase of its network, which connects twenty of the largest towns and cities including London, Bristol, Birmingham, Cardiff, Liverpool, Manchester and Newcastle and so encompasses 70% of the UK population. Then, by the coming autumn, it aims to have doubled the size of its backbone network so as to reach most of the major towns and cities. An investment of 'a multiple of £ 200 over the next decade is foreseeable' according to the company.

Potential market. As well as targeting high users in the consumer market (via cable and other operators), another of its potential markets is the large and multi-national customers. With over 2000 named customers generating £ 2 billion from switched telephony and £ 1.5 billion from private circuits, this will be a key target.

Energis proposes to offer a range of services including 2 Mbit s^{-1} and above leased circuit and managed bandwidth, in addition to standard voice and data services.

The strategy adopted by Energis is to be, primarily, a marketing and services company that will run its network to deliver communications and related information serices to meet the requirements of UK customers. Consequently, it will focus on these core activities and sub-contract all others. To this end, it has already awarded contracts worth £ 40 million to Northern Telecom[3] and Digital[4] and has appointed them as prime contractors for the network and the Information System. Thus, it can be seen that as well as innovative technology it is necessary to adopt an innovative strategy towards liberalisation.

1. Energis Communications Ltd, Carmelite House, 50 Victoria Embankment, London EC4W 0DE. Telephone 071 936 5555. Fax 071 936 5500.
2. British Broadcasting Corporation, Broadcasting House, London W1A 1AA. Telephone 071 580 4468. Fax 071 840 6607.
3. Northern Telecom Europe, Stafferton Way, Maidenhead, England SL6 1AY. Telephone (0628) 812 000. Fax (0628) 812 496.
4. Digital Equipment Co. Ltd, Digital Park, Imperial Way, Reading, England RG2 0TE. Telephone (0734) 868 711. Fax (0734) 575 027.

SCOTTISH PROFESSOR HEADS US SIGNAL PROCESSING SOCIETY

A Scottish university professor has become the first non-US engineer to be elected President of the Signal Processing Society of the Institute of Electrical and Electronics Engineers (IEEE) of America.

Professor T. Durrani from the electronics and electrical engineering department at Strathclyde University in Glasgow not only breaks the US domination of high office in the Signal Processing Society but also becomes the first engineer from Scotland to achieve eminence in the 100-year history of the New York-based IEEE.

The IEEE is the largest professional organisation of engineers in the world with a world-wide membership of over 320 000, and its Signal Processing Society is the fourth largest out of the 35 societies of the IEEE – after computing, power engineering and communications. It is also one of the fastest growing societies of the IEEE.

Signal processing is concerned with the acquisition, analysis and extraction of information from data, and as such it plays a key role in the development of information technology. Strathclyde has one of the largest research groups on signal processing in Europe.

University of Strathclyde, Glasgow, Scotland G1 1XQ. Telephone 041 552 4400. Fax 041 552 0775.

SERC INTO EUROPE

Britain's Science and Engineering Research Council (SERC) has joined the European Union of Physics Research Organisations (EUPRO). This union, formed in 1992, originated from a SERC initiative to establish closer links between the national research councils in the European Community (now European Union – EU) member countries responsible for mainstream or small-scale physics, and to pursue coordinated actions of mutual benefit in areas of research and postgraduate training. It extends the membership beyond the EC research councils to other European Free Trade Area (EFTA) countries and major organisations that engage in physics research.

Science and Engineering Research Council, Polaris House, North Star Avenue, Swindon, England SN2 1ET. Telephone (0793) 411 256; fax (0793) 411 468.

ADAPTING SPACE TECHNOLOGY FOR TERRESTRIAL MOBILES

With the projected increase in mobile traffic and restrictions on the radio spectrum allocated to such services, methods for increasing capacity without occupying more of the available spectrum are becoming increasingly important. New personal communications network (PCN) services, designed to operate in difficult urban environments, also require base stations whose coverage can be tailored to the local environment.

Engineers say this can be achieved through the use of multi-beam adaptive antenna technology, which involves the use of phased array antennas combined with advanced signal processing methods. This technology enables each mobile equipment user to be allocated a dedicated 'beam' that can be directed towards him and track his movements. This means that more users can occupy a given frequency band and interference from adjacent mobiles operating on the same frequency is reduced by the increased isolation between the beams.

This technology has already been used successfully for military and space communications, and now the plan is to apply it to the commercial sector. The two years of research needed to achieve this is now to be carried out by a consortium of European organisations led by Britain's ERA Technology from Leatherhead. Others involved are Alcatel SEL and Hagenuk from Germany, Detycom and the University of Catalonia from Spain, Aalborg University in Denmark and Bristol University in the UK.

ERA Technology Ltd, Cleeve Road, Leatherhead, England KT22 7SA. Telephone (0732) 374 151; fax (0732) 374 496.

RECHARGING THE BATTERY INDUSTRY

Scientists in Scotland are working on a new electrode material that could open the door to rechargeable high-energy batteries with longer life and much less weight.

Rechargeable batteries are now required for a vast range of electronic products, and presentday conventional nickel cadmium (NiCd) batteries not only constitute a large part of the weight and volume of the electronic product, but also contain poisonous cadmium heavy metal that can leak out once the battery is thrown away.

Dr. Peter Bruce and Prof. Colin Vincent from the electrochecmical and materials science centre at St. Andrews University have spent several years developing a lithium battery containing a new manganese oxide compound that is less harmful to the environment.

Another disadvantage of the current rechargeables is that they suffer from self-discharge even if they are not used, and the energy drains away very quickly. The new lithium batteries being developed at St. Andrews can store up to three times the energy compared with NiCd cells and no self-discharge occurs. This will allow devices such as mobile phones to last much longer between charges.

Prototypes of the St. Andrews battery are already on test and the technology involved has been patented. Meanwhile, the scientists are looking even farther into the future by investigating the feasibility of all-solid-state batteries where the liquid electrolyte is replaced by a polymer. This could lead to many new applications in consumer products, medicine and ultimately in electric vehicles.

Dr. P. Bruce, University of St. Andrews, College Gate, St. Andrews KY16 9AJ, Scotland. Telephone (0334) 6250; fax (0334) 62570.

DIRECTLY HEATED CRYSTAL OSCILLATOR

Available from Sematron is a directly heated crystal oscillator (DHXO) that has been designed to provide the same performance as an oven-contained oscillator with the size, weight and power consumption parameters of TCXOs.

The DHXO is available in a frequency range of 7–20 MHz. Frequency over temperature is 2×10^{-7} over –20 °C to +70 °C. Ageing is 1×10^{-7} per year. DC power is less than 1 W at +12 V d.c.

Sematron (UK) Ltd, Lattice House, Baughurst, Basingstoke, England RG26 5LL. Telephone (0734) 819 970. Fax (0734) 819 786.

NEW SMD SELECTION GUIDE

Designers of high density power supplies, motor controller and other types of power circuit will find the SMD selection guide from International Rectifier, covering its complete range of surface mount devices (SMDs), to be particularly relevant to their needs.

The guide contains shortform data on the full range of surface mount products, which includes HEXFETs (power MOSFETs), Schottky diodes, IGBTs and MOS-gate driver ICs.

International Rectifier, Holland Road, Hurst Green, Oxted, England RH8 9BB. Telephone (0883) 713 215. Fax (0883) 714 234.

STARTER RADIO & VIDEO PROJECT

Maplin's Starter Radio & Video project is a beginner's step-by-step construction guide on a VHS video tape. By following the instructions on video, the constructor will be introduced to the basic techniques of project building.

By the end of the video, a simple AM radio receiver circuit will have been constructed, ready to use. Certain troubleshooting advice is given at the end of the video.

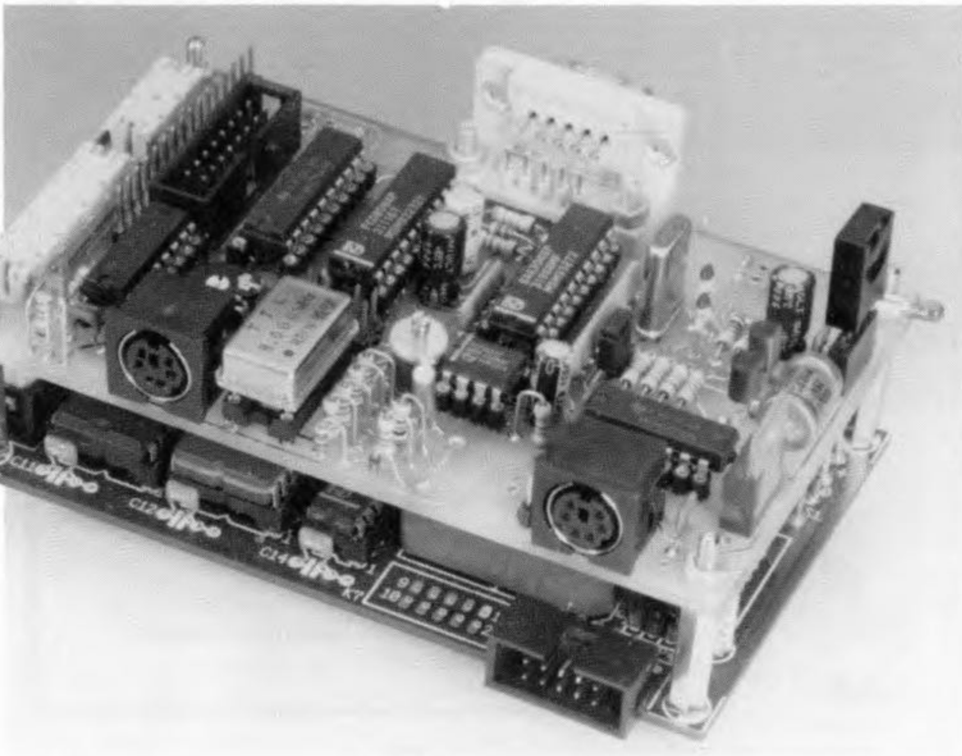
Additional kits of parts are also available without the video tape, so that several students can work from one video tape. However, these extra kits can not be used unless at least one video tape is available.

Prices are: Start Radio & Video £16·95; Start Radio – no Video £9·95.

Maplin Electronics, PO Box 3, Rayleigh, England SS6 8LR. Telephone (0702) 554 161. Fax (0702) 553 935.

80C535 SBC EXTENSION CARD

The popular 80C535 single-board computer described a few months ago in this magazine has a powerful microcontroller, and can be used for a wide variety of control applications. The extension card described in the present article gives the 80C535 computer a number of new interfaces mainly geared to communication with the real world.



NEW INTERFACES

- LC display
- IR decoder (RC5 standard)
- I^2C interface
- real-time clock
- RAM memory with battery backup
- DCF77 clock connection
- Standard RS232 connection
- All extensions supported by assembler code examples

Design by W. Hackländer and S. Furchtbar

DESPITE the many interfaces available on the 80C535 microcontroller board (Ref. 1), there remain a number of connection options to be desired. The present extension adds the following interfaces to the 80C535 card: (1) an RC5 decoder which allows a standard (Philips) infra-red remote control to be used in application programs; (2) a real-time clock with battery back-up and an (optional) DCF77 interface for perfectionists; (3) an LC display interface which supports displays with one or two lines, and 16 up to 40 characters, with or without a backlight.

The new interfaces are supported by a number of elementary subroutines written in assembler code. This software collection is supplied on a 3.5-inch diskette which may be obtained through our Readers Services.

The hardware

The circuit diagram of the extension card, **Fig. 1**, may appear a little unstructured because it is basically a collection of small sub-circuits. At the heart of the circuit is an address decoder type 74HCT138, IC_5. Its inputs are connected to address lines A12 through A15, and to a combination of the $\overline{RD}$ (read) and $\overline{WR}$ (write) signal made by an OR function. Only three output signals of IC5 are used. The signal at output $\overline{Y7}$ is used to select an 8-bit bus buffer, IC_3, at address 03000_H. The $\overline{Y6}$ signal selects the LC display (at address 02000_H-02003_H), and the $\overline{Y5}$ signal selects I^2C controller IC_4 (at addresses 01000_H and 01001_H). In practice, the address ranges created by IC_5 are much larger (01000_H locations) than would appear from this overview. Fortunately, the 'wide' addressing is not a problem, and has the advantage of requiring far less components than a decoder for the exact addresses required.

LCD interface

The LCD is taken up in the controller's I/O address range at the four addresses 02000_H through 02003_H. Two types of pin header, a single-row type (K_{12}) and a double-row type (K_9), are available to cater for the most commonly used connectors on LCD modules. Preset P_1 is available for the contrast setting, while resistor R_{11} determines the intensity of the LED-based backlight, if that is feature of the LCD module you are using.

Infra-red receiver

The RC5 compatible infra-red remote control receiver is built around IC_2, a SAA3049 from Philips Semiconductors. A complete infra-red receiver device type SFH505, SFH506 or IS1U60 is connected to pin header K_1. These IR receiver devices convert infra-red light received from a remote control unit into electrical signals which can be processed by IC_2. Capacitors C_1 and C_{14} decouple the supply voltage of the receiver IC. A LED, D_1, lights to indicate that the SAA3049 detects a valid (RC5) command from your infra-red remote control. The LED driver is formed by transistors T_1, T_2 and a handful of resistors. The LED control signal is also fed to dataline D7 of the 80C535 card. The line is held low for about 15 ms after each received code. This enables the microcontroller to detect the reception of valid data. Output TO (toggle) changes state every time a key has been pressed on the remote control. The microcontroller can interrogate the state of this output via dataline D6 of IC_3.

The RC5 decoder, IC_2, requires a clock of 4 MHz, which is supplied by the oscillator built around quartz crystal X_1. The RC5 decoder IC is powered via R_6. This resistor, together with capacitors C_5 and C_6, ensures that the supply voltage is sufficiently decoupled. The decoder, like all RC5 components, requires a unique address. This is set with the aid of K_{11}, which consists of five solder spots located across pins 7, 8, 15, 16 and 17 of the IC. For

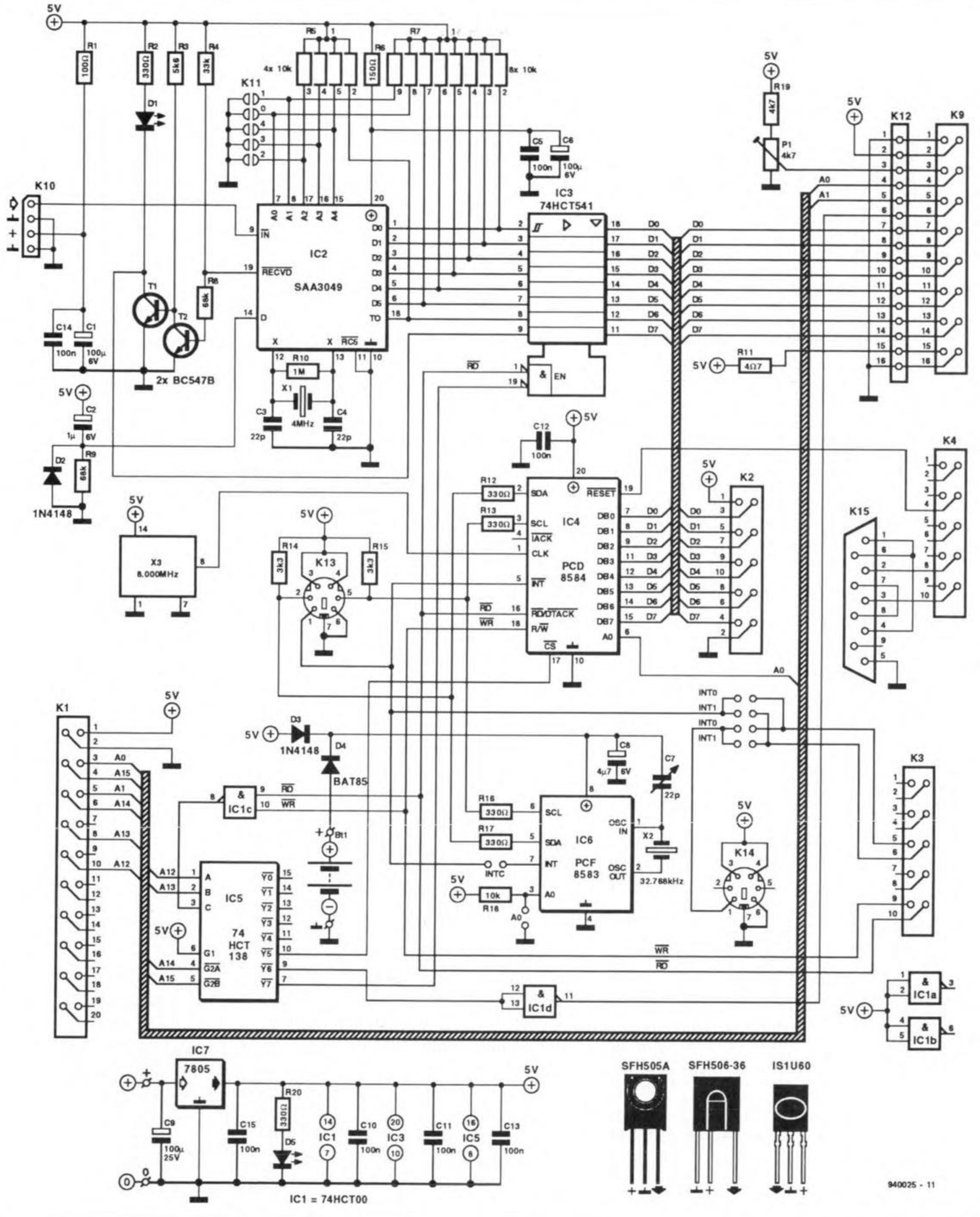


Fig. 1. Circuit diagram of the 80C535 extension card. Note the different connections of the three infra-red receivers that may be used.

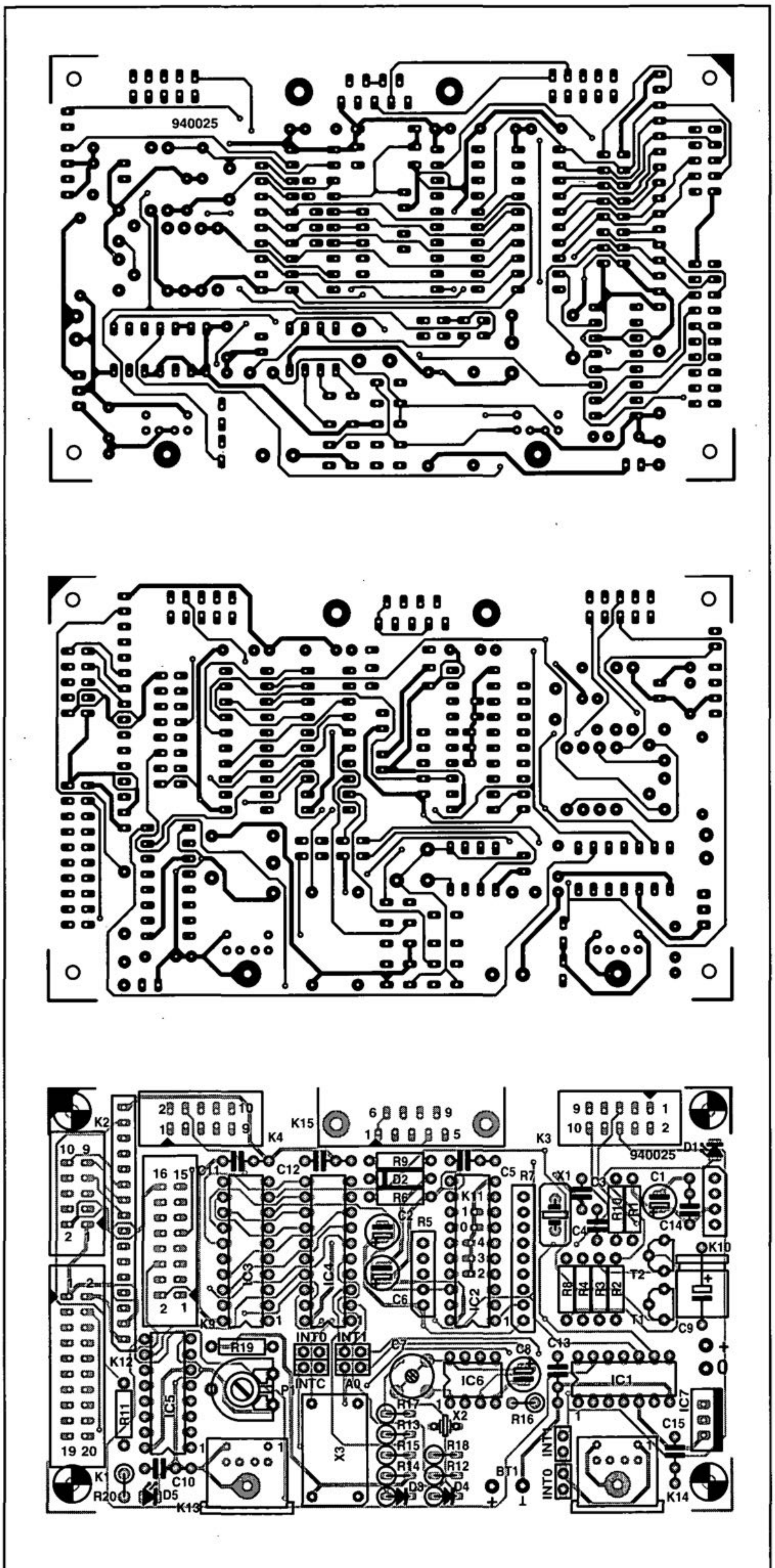


Fig. 3. Artwork of the double-sided through-plated board designed for the extension circuit. This extension board is designed to fit on top of the 80C535 computer board.

COMPONENTS LIST

Resistors:
R1 = 100Ω
R2,R12,R13,R16,R17,R20 = 330Ω
R3 = 5kΩ6
R4 = 33kΩ
R5 = 4-way 10kΩ array
R6 = 150Ω
R7 = 8-way 10kΩ array
R8,R9 = 68kΩ
R10 = 1MΩ
R11 = 4Ω7
R14,R15 = 3kΩ3
R18 = 10kΩ
R19 = 4kΩ7
P1 = 4kΩ7 preset H

Capacitors:
C1,C6 = 100µF 6V radial
C2 = 1µF 6V radial
C3,C4 = 22pF
C5,C10-C15 = 100nF
C7 = 22pF trimmer
C8 = 4µF7 6V radial
C9 = 100µF 25V

Semiconductors:
D1,D5 = LED
D2,D3 = 1N4148
D4 = BAT85
T1,T2 = BC547B
IC1 = 74HCT00
IC2 = SAA3049 (Philips Semiconductors)
IC3 = 74HCT541
IC4 = PCD8584 (Philips Semiconductors)
IC5 = 74HCT138
IC6 = PCF8583 (Philips Semiconductors)
IC7 = 7805

Miscellaneous:
K1 = 20-way boxheader.
K2,K3 = 10-way boxheader.
K9 = 16-way boxheader.
K10 = 4-way SIL socket.
K12 = 16-way SIL header.
K13,K14 = 6-way mini-DIN socket.
K15 = 9-way sub-D socket, PCB mount, angled.
X1 = 4-MHz crystal.
X2 = 32.768-kHz crystal.
X3 = 8-MHz TTL oscillator.
Bt1 = 3V6 battery.
Infra-red receiver SFH505A, SFH506-32 (Siemens) or IS1U60 (Sharp).
Printed circuit board 940025 and diskette 1941. Set order code: 940025 (see page 70).
The diskette is also available separately as order code 1941 (see page 70).

note: K4-K8 are on the 80C535 SBC, not on the extension board.

example, if the decoder is assigned the address '0', it will only respond to commands received from a TV remote control. Address '0' is selected by closing all wire links with a drop of solder tin.

The result of the decoding operation is a 5-bit code which appears in inverted form at RC5 address 0 on outputs D0-D5 of the SAA3049. Since the outputs are of the open-collector type, they are fitted with pull-up resistors. Next, IC_3 buffers the data signals for the microcontroller to read at address 03000_H.

I²C bus interface

The central part in this sub-circuit on the 80C535 extension card is a PCD8584 parallel-to-I²C converter (Ref. 2). The real-time clock type PCF8583, IC_6, is controlled via the I²C bus created by the PCD8584. To enable I²C modules described in previous issues of *Elektor Electronics* to be connected without problems, the bus is terminated into a 6-way mini-DIN socket on the extension card. The clock signal for the I²C bus interface is supplied by a crystal oscillator module, X_3. The reset signal is furnished by the 80C535 card, and reaches the PCD8584 via connector K_4. Apart from the $\overline{RD}$ and $\overline{WR}$ signals, the interface requires an address selection signal, and address line A0. The address selection signal may be found at output $\overline{Y5}$ of IC_5, and is active in the range between 01000_H and $01FFF_H$. Although the interrupt option is not used in any of our previous I²C projects, it is supported on the present card. If you wish to use them, interrupt lines INT0 and INT1 can be selected via the corresponding jumper on connector K_3. The SDA and SCL lines of the I²C controller are also taken to the on-board real-time clock, IC_6. This clock has its own quartz crystal, X_2, and may be located at one of two addresses.

Depending on the presence of a jumper on connection A0, the real-time clock is at I²C address $A0_H$ (jumper fitted) or $A2_H$ (jumper not fitted). Diode D_4 enables a battery to take over the supply of the clock if the mains supply is switched off. Diode D_5 then ensures that only IC_6 is supplied by the battery.

Power supply

A conventional regulated power supply is built around IC_7. LED D_5 functions as a 'power on' indicator.

Construction

The track layouts and component mounting plan of the double-sided through-plated printed circuit board designed for the extension card are shown in **Fig. 3**. The board is designed to fit on top of the 80C535 computer card. All the necessary connections can be made using a few pieces of flat-cable. The extension card and the main computer card form a compact unit.

Construction is easy if you use the ready-made printed circuit board supplied through our Readers Services. Thanks to the solder mask on the ready-made board, the risk of short-circuits between adjacent solder spots is minimal. None the less, use solder tin sparingly. If a certain extension sub-circuit is not required, the relevant parts may simply be omitted.

Optionally, the connections of K_3 and K_4 may be made using a couple of discrete wires. This is possible because only a limited number of connections is used on each connector. On the down side, the discrete wires make disassembling the project a little more complicated, for instance, when a fault has developed. A final tip as regards soldering: secure the 9-way sub-D connector wit the aid of two M3 bolts before soldering the pins. In that way, you prevent strain on the solder connections if a plug is inserted.

The circuit is ready for use once all components have been fitted, the interconnections made, and the board is mounted on top of the 80C535 card.

Preset P_1 can not be adjusted until the circuit is powered, and an LCD is connected. The optimum contrast setting is readily found.

Trimmer capacitor C_7 enables the clock to be adjusted should this run too fast or too slow.

Software on disk

Diskette **1941** supplied through our Readers Services contains two programs (and their source files) which should prove helpful to test and adjust the extension card. These programs are loaded and run (from address 4100_H) using the EMON51 or EMON52 monitor discussed in Ref. 3. An I²C display driver program is avaialble separately as order code 946197-1.

The RC5 extension may be tested with the aid of the program IRTEST.A51. First, the display shows the text 'RC5 demo" on the top line. The lower line displays: 'waiting .....'. If you press a number key on your RC5 compatible remote control, the corresponding value appears on the LCD. The '+' function of the 'volume' key on the remote control selects a kind of 'shift' function. From then on, the second function of each key is used. pressing the '–' key takes you back to the standard functions. If other keys than the numbers are pressed on the remote control, the LCD shows '255'.

The real-time clock and the I²C interface may be tested by running the program CLK535.A51 on the disk. The time is displayed on the LCD.

Apart from enabling you to test the respective extension functions, the programs also give you enough information to write your own software for the extended 80C535 computer. If you have developed an interesting application, let us know!

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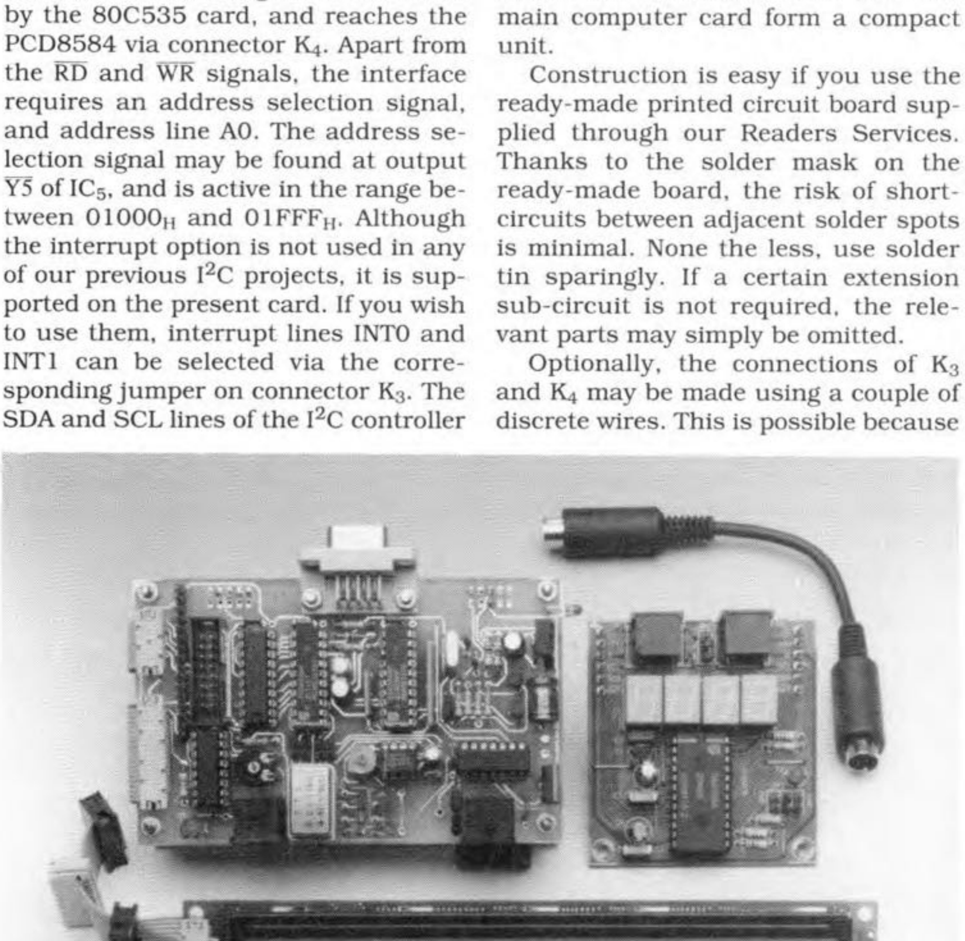


Fig. 2. 80C535 extension card with two ancillaries that can be connected straight away: the *Elektor Electronics* I2C controlled LED module (right) and a 2x40 character LCD (below). Software to drive the LED display unit as a seconds counter is available on disk as order code 946197-1 (see page 70 for price and ordering details).The program runs under EMON51/52.

References:

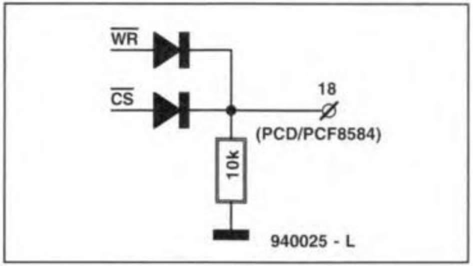
1. 80C535 single-board computer, *Elektor Electronics* February 1994.
2. I²C interface for PCs, *Elektor Electronics* February 1992.
3. 80C535 Hardware/assembler course, *Elektor Electronics* March through June 1994.

CORRECTIONS AND UPDATES

80C535 Extension card

June 1994, p. 8-11

The PCD8584 may be switched to '6800' mode if a WR signal arrives without a CS signal. The problem may be solved by combining WR and CS in a diode-AND gate as shown below. Pin 18 if IC_4 is taken out of the IC socket and connected to ground via a 10-kΩ resistor. The WR signal is found on socket pin 18, and CS on pin 10 of IC_5. Also note that the PCD8584 is currently supplied as the PC**F**8584.



Dual-purpose LED display

December 1994, p. 90

Resistor R_{33} should be connected to ground, not to +12 V as shown in the circuit diagram.

Experimentation board for PICs

July/August 1994, p. 74.

In the circuit diagram, the signals on pins 7 and 8 of both connectors K_1 should be swapped. MCLR is on pin 8, and RTCC on pin 7. The relevant PCB is all right.

Mains signalling system (2)

May 1994, p. 10-14.

The instructions for command 'T' should read: 'T' must be followed by the address in ASCII, and terminated with a semicolon (';').

The baudrate for the communication software should be set to 300, format: 8 bits, 1 startbit, 1 stop bit, no parity.

Electronic fuse

March 1994, p. 56.

To prevent transistor T_2 from burning out when the reset switch is pressed during an overload condition, switch S_1 should be connected between the collector of T_1 and the base of T_2.

FUEL CONSUMPTION METER

FOR FUEL-INJECTION ENGINES

Design by F. Ebert Eberl

To meet market demands as well as environmental requirements, automobile engineers are continually striving towards minimal fuel consumption and cleaner exhaust systems. That quest has resulted in most modern cars being fitted with multi-valve engines, fuel injection, electronic engine management and catalytic converters. Surprisingly, however, few production cars are fitted with a fuel consumption meter: an omission which this article attempts to put right.

In contrast to the situation in carburettor engines, it is fairly easy in the case of fuel injection engines to obtain information as to fuel consumption.Since the pressure in the supply line in a correctly working system is controlled by a fuel pressure regulator, the consumption is directly proportional to the number and length of the injection pulses. Using these pulses to charge a capacitor results in a potential which rises when more fuel is used. Applying this potential to a voltmeter results in a sort of consumption monitor. However, this would only indicate the amount of fuel being consumed, whereas a fuel consumption meter gives an indication of how many litres per 100 km are being used. The correlation between litres per 100 km and m.p.g. (a more usual concept in many English-speaking countries) is shown in the table. This means that information is needed as to the distance covered in a time unit. Fortunately, most modern cars are equipped with an electronic tachometer, the number of pulses produced by which can be used to divide the number of injection pulses. The principle of this is shown in **Fig. 1**. The injection pulses are used in conjunction with a current source to charge a capacitor. The potential developed across the capacitor is applied to the voltmeter via a buffer stage. This gives a indication of the fuel consumption. The pulses from the tachometer are given a specific length by a pulse shaper and are then used to operate an electronic switch. Every time the switch is closed, the capacitor is partially discharged. The higher the pulse rate, the more frequently the capacitor is discharged and the lower the voltmeter reading. In other words, the voltmeter reading has become a function of both the injected amount of fuel and the distance travelled. If a tachometer is not fitted, pulses may be derived from the speedometer (see circuit description).

ℓ/100 km	M.P.G.
3	91.8140
4	68.8579
5	55.0805
6	45.9035
7	39.3480
8	34.4290
9	30.6031
10	27.5423
11	25.0385
12	22.9518
13	21.1861
14	19.6733
15	18.3617
16	17.2140
17	16.2013
18	15.3012
19	14.4958
20	13.7713
21	13.1154
22	12.5192
23	11.9751
24	11.4761
25	11.0168

Note that if the car is at a standstill with the engine idling, the indicated fuel consumption will be high: this is because if there is no distance travelled, the quotient of consumption and zero (distance) is infinitely high.

Circuit description

The capacitor that is charged and partially discharged is C_1. Charging takes place via current source T_2. The required constant base potential for this transistor is obtained with the aid of D_1. The level of the charging current can be set within wide limits with P_1. Since the quiescent drive voltage, U_i, for the injection valves is high, the current source is controlled by 'inverting switch' T_1. When U_i is high, T_1 is on and short-circuits D_1, so that T_2 is off. When U_i is low (that is, during injection of fuel), T_1 is off and T_2 functions as required.

The signal from the tachometer, U_t, consists of rectangular pulses whose width and amplitude vary appreciably. For accuracy's sake, the pulses are given a fixed shape: their rate is, of course, not affected. The shaping is effected by differentiating network R_7-C_3, peak limiter D_2 and monostable IC_{1b}. The pulses at the output of the monostable have a fixed length of 1 ms. They control 'discharge switch' T_3. Thus, C_1 is discharged via R_5 for a fixed period at every pulse contained in U_t. The total discharge period per time unit is thus directly proportional to the rate of the U_t pulses.

The potential across C_1 is applied to voltmeter M_1 via buffer IC_{1a}. To prevent the meter needle vibrating with small voltage variations, these are smoothed by shunt capacitor C_2.

If a tachometer is not fitted in the vehicle, the U_t pulses can be derived from the speedometer. The speedometer cable is coupled to the instrument on the dashboard by a rotating magnet. The magnet rotation can be used to generate pulses with the aid of a magnetoresistive sensor. The outputs of the sensor used, IC_4, are linked to a differential amplifier, IC_{3b}, which amplifies ×100 each unbalance caused by magnetic induction in the measuring bridge.

The output of IC_{3b} is applied to the positive input of comparator IC_{3a}, which uses the differential signal averaged by R_{18}-C_8 as its reference. Thus, the output of the comparator is a rectangular pulse train, whose frequency is directly proportional to the speed of the vehicle. Diode

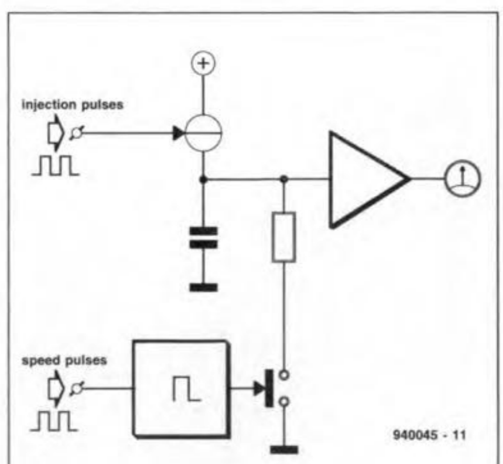
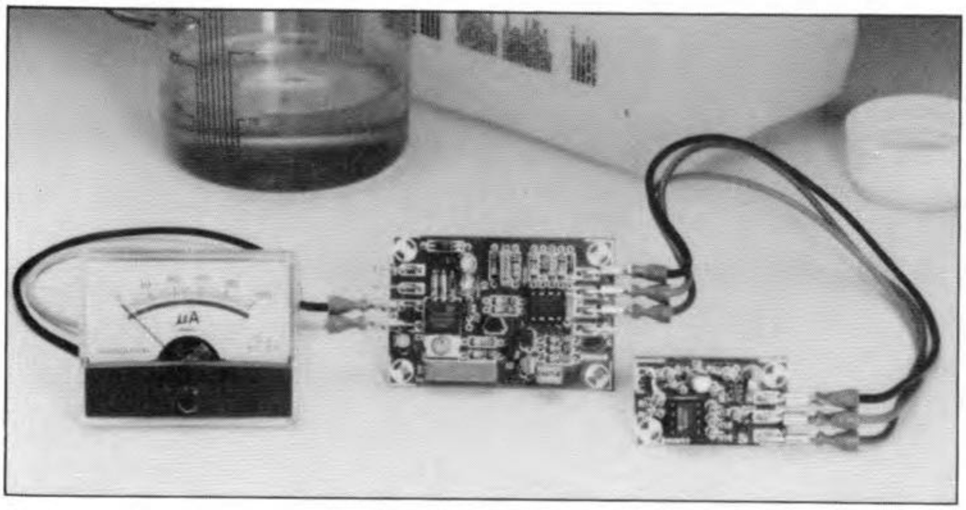


Fig. 1. Charging a capacitor in rhythm with the fuel injection pulses and partially discharging it in rhythm with the tachometer pulses gives an average potential across the capacitor that is representative of the fuel consumption over a certain distance.



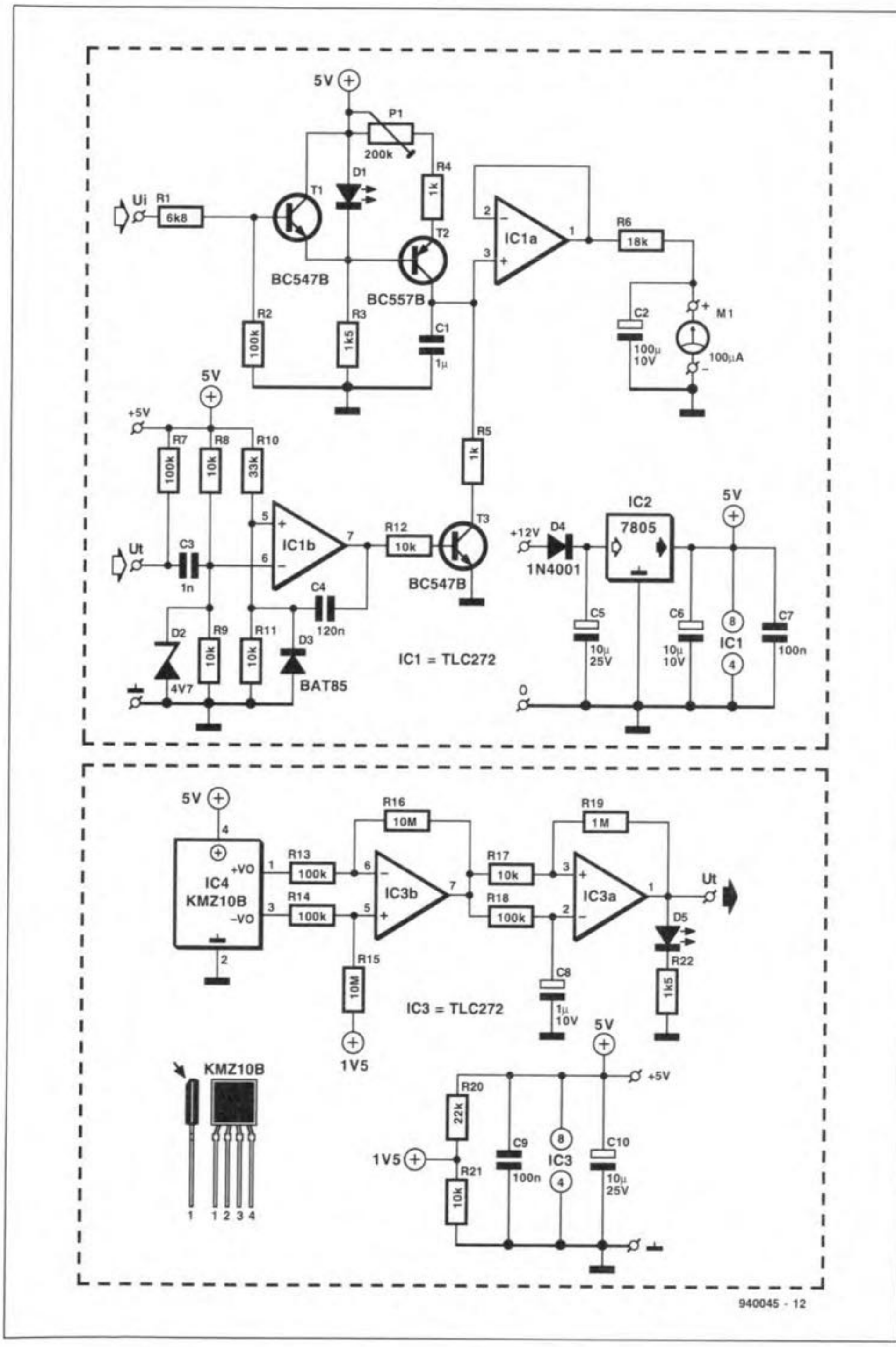


Fig. 2. Circuit diagram of the fuel consumption meter. The lower section is for use with a speedometer.

D_5 indicates whether output pulses are being produced.

The sensitivity of the sensor was found to be more than adequate. In the prototype, reliable pulses were produced even when the distance between IC_4 and the casing of the speedometer was a few centimetres (up to an inch).

Construction

The printed-circuit board for the meter is shown in **Fig. 3**. It is easily cut into two: the larger portion for the meter section and the smaller for the the magnetoresistive section if required. Populating the board(s) is straightforward. Connections between the board(s) and the tachometer are by right-angle car-type terminals as shown in **Fig. 4**.

The voltmeter and the 12 V battery supply are connected to the meter board.

If the magnetoresistive sensor is used, the relevant board is best fitted in a small synthetic enclosure mounted directly on to the speedometer. The best position for it must be determined by trial and error as most speedometers are slightly different from one another. During a short test drive it can be determined from D_5 whether the sensor is in a strong enough magnetic field. If the distance between this board and the meter board is more than about 30 cm (1 foot), the interconnection between them should be made in screened cable.

The positions from where to take the U_i and U_t signals are best found out from a car dealer or from the car's technical handbook. However, the position of the injection valves is easily found. One of the terminals on the valve carries 12 V: the U_i pulse train is taken from the other terminal.

The take-off point for the tachometer is normally situated near the drive shaft or the differential: there is only one terminal. Whether this is the correct terminal can be checked by driving the car for a short distance after having taken the socket off the terminal. If the tachometer does not work, it is the terminal from which the U_t pulse train is to be taken. Reinsert the socket on to the terminal once the tap has been made.

Calibration

The fuel consumption meter is intended primarily to facilitate driving with fuel economy in mind. Therefore, an indication as to fuel economy was deemed more important than an exact reading of the fuel used. It is for that reason that the read-out is not a complex digital one, but a simple moving-coil instrument. Moreover, his type of meter is not so distractive for the driver: a safety plus point. Nevertheless, some calibration is necessary to set a guideline.

It may be asked whether only one calibrating device, P_1, is sufficient in all cases. The answer is: almost certainly in 99% of them. The range of the preset is fairly large and, moreover, the difference in aspects with which this article is concerned between a variety of cars is not very great. For instance, the injection time in most cars is in the range 1–15 ms. Furthermore, the relationship between tachometer rate and distance covered is determined by the final drive ratio, which in most cases is not far from 4:1. At the end of this article is a formula for the exact calculation of the average potential across C_1.

Three things are required for the calibration: a chauffeur, the car's instruction book and a wind-still day.

In the instruction book, find and note the manufacturer's consumption figures: these normally include that at a constant 56 m.p.h. and 70 m.p.h.

When the engine has reached its normal running temperature, keep the speed at constant 56 m.p.h. and adjust P_1 for a meter deflection of about $^1/_4$ f.s.d. Mark this point with the figure given in the instruction book, say, 9.00 (ℓ/100 km) or 30 (m.p.g.). Then, accelerate to 70 m.p.h. and check that the meter reading increases and mark that point with the figure given in the book, say, 11.00 (ℓ/100 km) or 25 (m.p.g.).

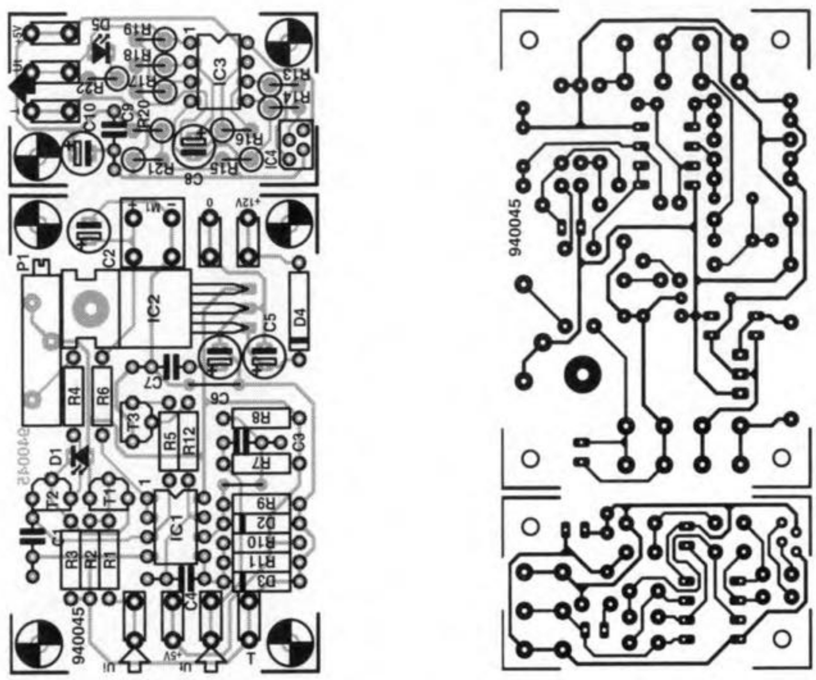


Fig. 3. The printed-circuit board for the fuel consumption meter is easily cut into two. The smaller section is for use with a speedometer if a tachometer is not fitted.

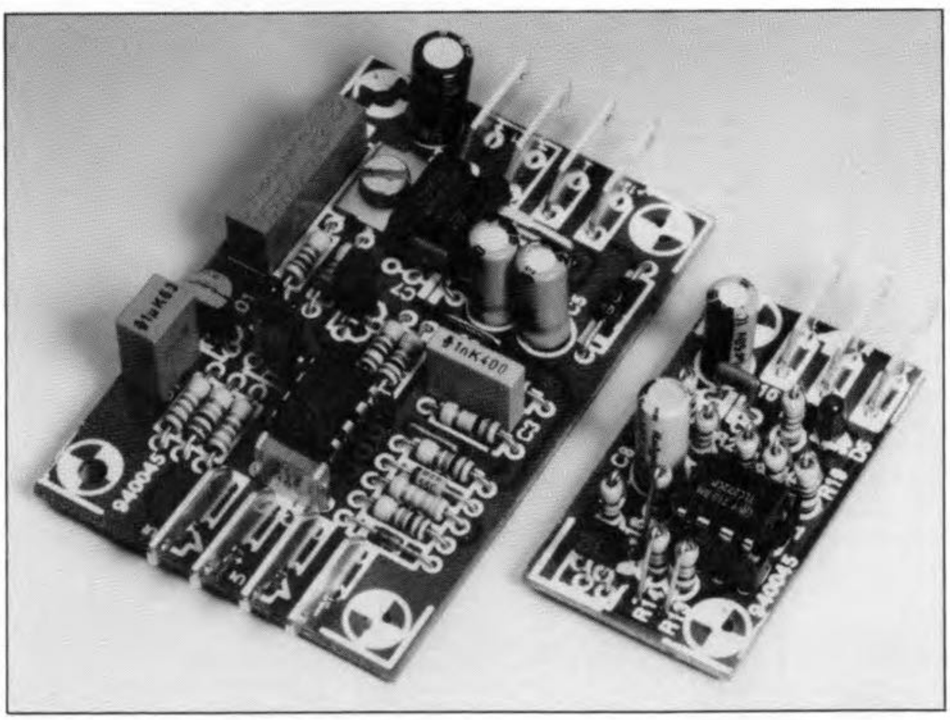


Fig. 4. The completed boards clearly show the car-type right-angle terminals.

Average capacitor voltage

The average value of the voltage, U_{C1}, across capacitor C_1 is given by:

$$U_{C1} = \frac{f_i t_i I_i (1+e^{-t_t/R_oC})}{2 f_t C (1-e^{-t_t/R_oC})}$$

where

f_i = injection rate (n/60 Hz, where n = engine speed in r.p.m.);
t_i = injection duration (1–15 ms);
f_t = tachometer rate in Hz;
t_t = monotime of tachometer (ms);
R_o = discharge resistor ($R_5 = 10^3\ \Omega$);
C = integrator capacitance ($C_1 = 10^{-6}$ F);
I_i = integrator current (can be set between 6 µA and 10 mA).

Parts list

Resistors:
R_1 = 6.8 kΩ
R_2, R_7, R_{13}, R_{14}, R_{18} = 100 kΩ
R_3, R_{22} = 1.5 kΩ
R_4, R_5 = 1 kΩ
R_6 = 18 kΩ
R_8, R_9, R_{11}, R_{12}, R_{17}, R_{21} = 10 kΩ
R_{10} = 33 kΩ
R_{15}, R_{16} = 10 MΩ
R_{19} = 1 MΩ
R_{20} = 22 kΩ
P_1 = 200 kΩ multiturn preset (horizontal)

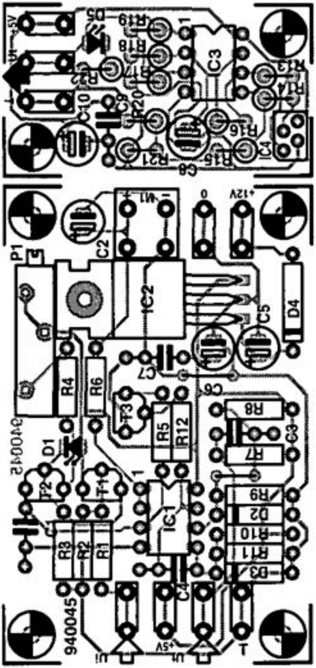
Capacitors:
C_1 = 1 µF, polyester
C_2 = 100 µF, 10 V, radial
C_3 = 1 nF
C_4 = 120 nF
C_5, C_{10} = 10 µF, 25 V, radial
C_6 = 10 µF, 10 V, radial
C_7, C_9 = 100 nF
C_8 = 1 µF, 10 V, radial

Semiconductors:
D_1, D_5 = LED, red, low current
D_2 = zener, 4.7 V, 500 mW
D_3 = BAT85
D_4 = 1N4001
T_1, T_3 = BC547B
T_2 = BC557B

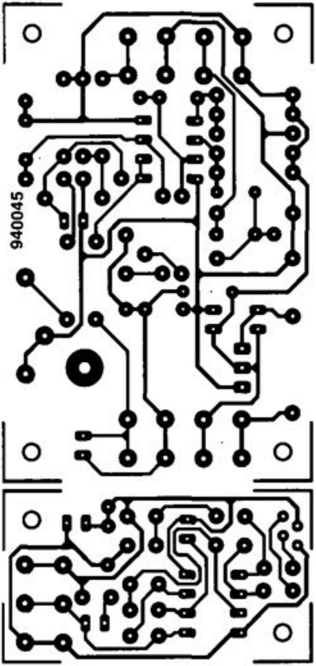
Integrated circuits:
IC_1, IC_3 = TLC272CP
IC_2 = 7805
IC_4 = KMZ10B (Philips)

Miscellaneous:
11 off car-type terminals, right-angled with matching sockets
M_1 = moving-coil meter, 100 µA, 1.9 kΩ
PCB Ref. 940045

[940045]

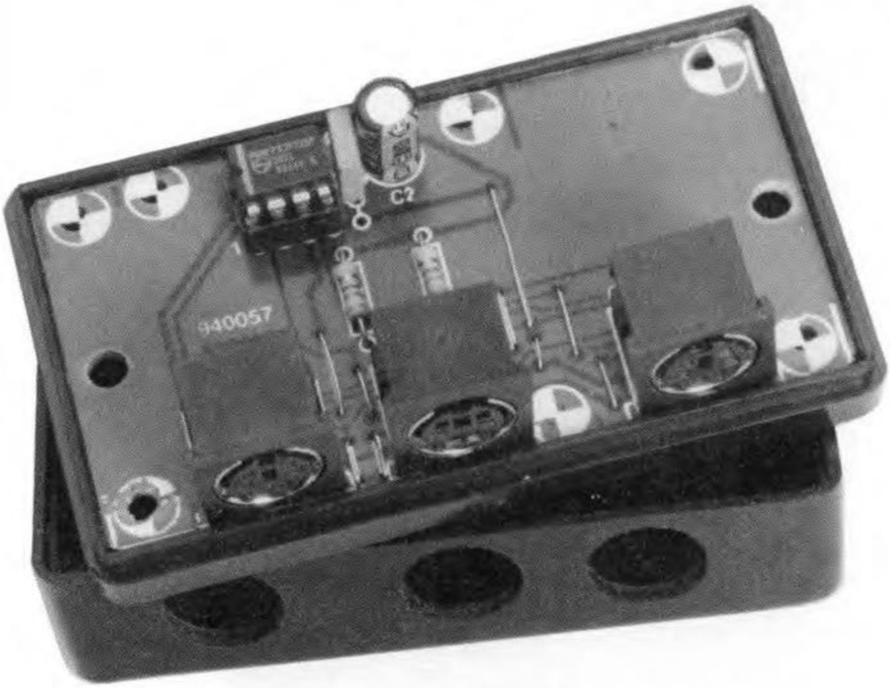


940045
P1
C2
IC2
C5
D4
+12V
0
M1
C7
C6
R4
R6
T3
R5
R12
R8
R7
C3
D1
T2
T1
IC1
R9
D2
R10
R11
D3
C4
C1
R3
R2
R1
+5V
T



940045

I^2C BUS BOOSTER



The I^2C interface was originally developed by Philips Components to enable complex integrated circuits to communicate using a simple bidirectional bus. In practice, I^2C compatible ICs may be located up to a few meters apart. In some cases, for instance, with remote data logging, longer connections will be desired. The bus booster described here increases the maximum distance between I^2C devices by a factor 10, bringing small networking applications within easy reach.

Design by K. Walraven

THE abbreviation I^2C stands for Inter Integrated Circuit (IIC), which already hints at a bus system which allows ICs to talk to each other. Usually, this means that connections are established 'across the board' or between boards, that is, over distances of a couple of tens of centimetres. Effectively, the I^2C bus makes it possible for many complex ICs on the same board, or different modules (for instance, in a TV set) to share a common bus for data and command traffic. To ensure the integrity of the information conveyed via the bus, the total capacitance of all connected devices is subject to certain restrictions.

The design of the I^2C bus is based on a maximum load capacitance of 400 pF. Consequently, in practice distances of up to a few meters may be covered. By using a special I^2C bus driver, the 82B715, the maximum bus capacitance may be increased by a factor 10. This is achieved by stepping down the impedance of the SDA (data) and SCL (clock) lines with the aid of an electronic transformer. As illustrated in **Fig. 1**, a single 82B715 contains two current transformers which effectively multiply the transformer 'primary' (i.e., device input) current by a factor 10. The structure of a buffered I^2C link is shown in **Fig. 1a**, while **Fig. 1b** illustrates how two I^2C outputs are interconnected via a common pull-up resistor. The current transformer (an all solid-state device!) is inserted between the pull-up resistor and the output. If the open-collector output of port A is actuated, a current I_b flows from the pull-up resistor to the actuated output via the current transformer. The output transistor of buffer A_b boosts the current through the buffered I^2C line by a factor 10. The 'excess' current ($9 \times I_b$) is shunted back into the supply by the transformer output transistor. The logic low level which is then on the I^2C bus line is copied 1-to-1 to the input of port B via buffer B_b. Since the rule that only one output may be low at a time is still valid, it is possible to connect several ports to the buffered bus. Also, buffers may be connected in parallel because they have the same open-collector structure at their outputs as any other I^2C compatible IC. It should be noted, though, that each I^2C line may have only one 330-Ω pull-up resistor. Consequently, all resistors (i.e., pull-up resistors as well as current limiting resistors) should be removed from the entire path between the existing I^2C connection, and that of the I^2C buffer.

On the existing I^2C modules described in *Elektor Electronics*, the series resistors on the bus lines served to afford ESD (electrostatic discharge) protection. They should now be removed to prevent problems caused by the voltage they introduce. If ESD resistors were used throughout the circuit, a logic low level would correspond to a voltage between 0.5 V and 2 V, which exceeds the specifications of the buffer device. You need not worry about removing the ESD protection because the buffer IC itself contains protective circuitry.

Speak of capacitance

The fact that the maximum capacitive load on an I^2C bus line is about 400 pF says very little in itself — arguably, the practical consequences of this restriction are far more important. When ordinary PCB material is used, each port represents a load of about 10 pF. The wiring capacitance is about equal. Consequently, a maximum of 20 ICs may be connected to a single bus line. Longish cables, however, reduce this number considerably, because they introduce a relatively large stray capacitance.

Based on a maximum clock frequency of 100 kHz, the *RC* time may not exceed 1 µs. At this timing, the pulse edges have a length of about 4 µs. Assuming that a pull-up resistor of 330 Ω is used, the maximum load capacitance equals about 300 pF (1 µs/3.3 kΩ). Consequently, a maximum load capacitance of 3,000 pF may be present at the buffered side if a 330-Ω pull-up is used. It is hard to say how this can be translated into cable length, since that depends largely on the type of cable used. In any case, a cable of a couple of meters will not present problems. If you want to 'squeeze' the driver for maximum drive capacity, the value of the pull-up resistor may be lowered to 153 Ω, at which the maximum permissible current of 30 mA flows.

If a lower clock frequency is used, the maximum permissible capacitance that can be driven increases proportionally.

The oscilloscope screendump in **Fig. 2** shows the SDA signal (upper trace) as well as the SCL signal (lower trace). The start condition is shown to the left. The falling (negative-going) edge is clearly steeper than the rising (positive-going) edge. This is caused by the fact that the rising edge depends on the relatively high value of the pull-up resistor. An acknowledge pulse is generated at the ninth signal edge. At the time the screendump was made, a 330-Ω resistor was still present in series with the output of a module. As a result, the '0' level is at about 0.5 V here.

Practical circuit

The circuit diagram of the I²C bus booster is given in **Fig. 3**. All and sundry will agree that this is a very simple circuit indeed. The heart of the booster is formed by the 82B715 in position IC_1. Both outputs of the IC are fitted with a 330-Ω pull up resistor. The I²C connection to be buffered is linked to connector K_1. The booster chip is supplied with input signals, supply voltage and ground via this connector. The interrupt signals are not boosted, and passed straight on to the outputs of the circuit, connectors K_2 and K_3. It is therefore better not to use interrupts.

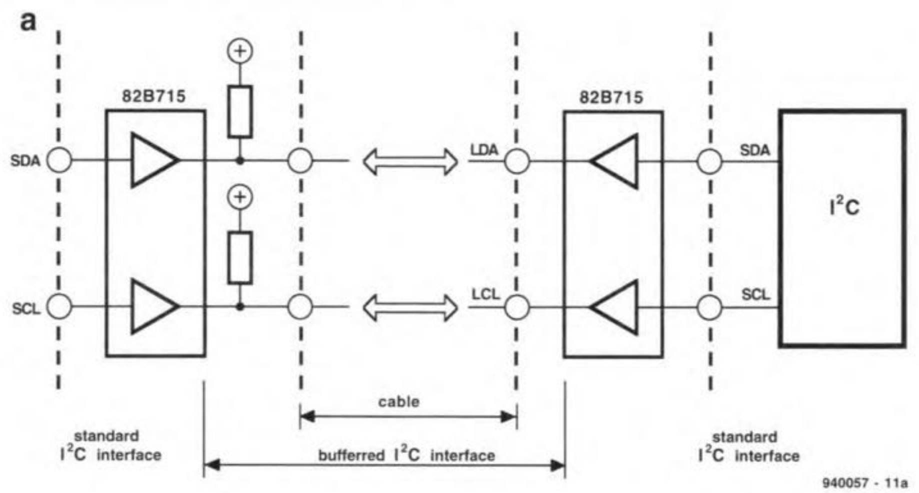
The presence of two connectors makes chaining the bus much easier. If chaining is not required, you may want to omit K_3. Capacitors C_1 and C_2 are added to afford sufficient decoupling of the supply line.

Construction

The artwork of the printed circuit board designed for the bus booster is shown in **Fig. 4**. Since it seems logical to use at least two buffers, the PCB enables you to build two buffers at a time. The two PCBs are best separated before they are populated.

Because of its small size, the circuit is easily built into a compact case. The construction itself is entirely straightforward, and requires hardly any comment. Just make sure you work carefully, and use a solder iron with a fine tip.

Having completed the bus booster it is time to remove all pull-up resistors



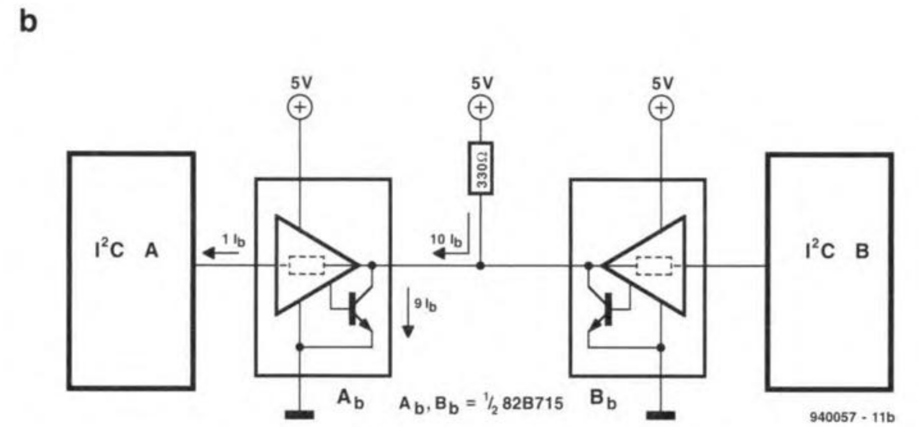


Fig. 1. Structure of a buffered I²C bus. The current transformers in the 82B715 are used to buffer the bus lines.

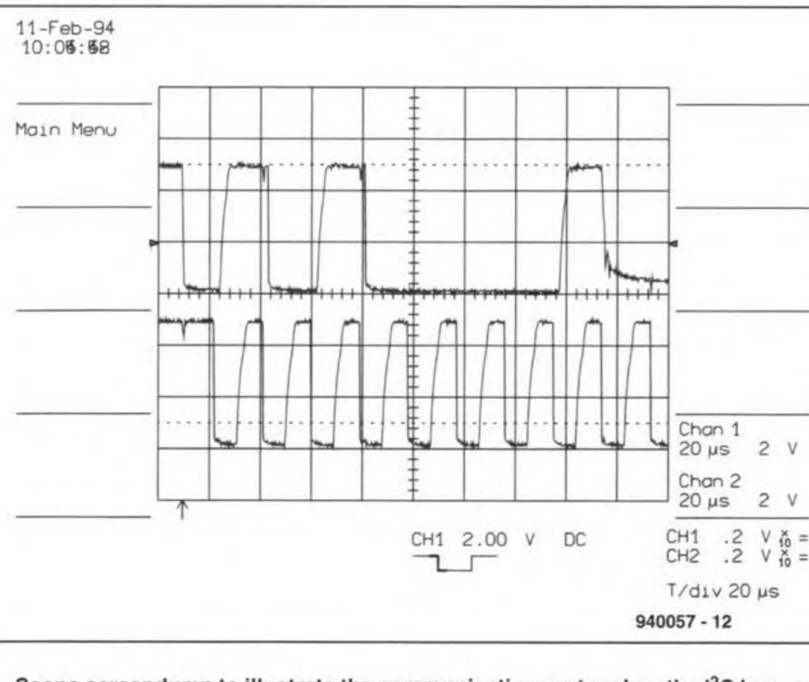


Fig. 2. Scope screendump to illustrate the communication protocol on the I²C bus, and the effect of the 330-Ω resistor in series with an I²C line.

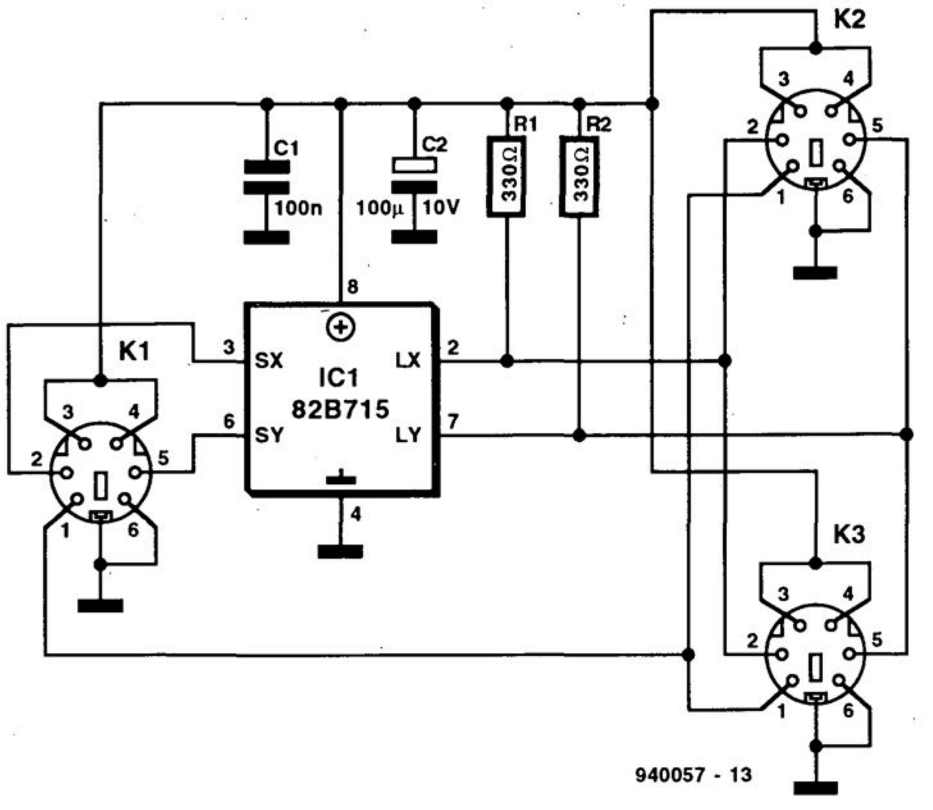


Fig. 3. Circuit diagram of the I²C bus booster: one integrated circuit and a handful of passive parts do the trick.

from the existing I²C bus, as well as any 330-Ω ESD protection resistors. Disconnect the cable, and connect a bus booster at both ends. From then on, the buffering on the bus could not be better, and much larger distances can be covered reliably. By the way, do remember that only one set of buffer resistors is necessary in each I²C bus system. This means that resistors R_1 and R_2 need to be fitted on one of the two booster boards only.

(940057)

I²C projects described in previous issues of this magazine:

Inter-IC communications, September 1990.
Video digitizer, July/August 1991.
I²C interface for PCs, February 1992.
I²C LED display, June 1992.
Speech/sound memory, December 1992.
I²C alphanumerical display, July/August 1993.
I²C bus fuse, July/August 1993
I²C opto/relay card, february 1993.
I²C power switch, december 1993.
I²C tester, January 1994.

I²C driver software for PCs, on disk: order code 1821 (see page 70).

COMPONENTS LIST

Resistors:
R1,R2 = 330Ω

Capacitors:
C1 = 100nF
C2 = 100µF 10V

Semiconductors:
IC1 = P82B715PN (Philips Semiconductors)

Miscellaneous:
K1,K2,K3 = 6-way mini-DIN plug.
1 plastic case, inside dimensions 75x40x15mm (e.g., Conrad 52.26.51).
Printed circuit board 940057 (see page 70).

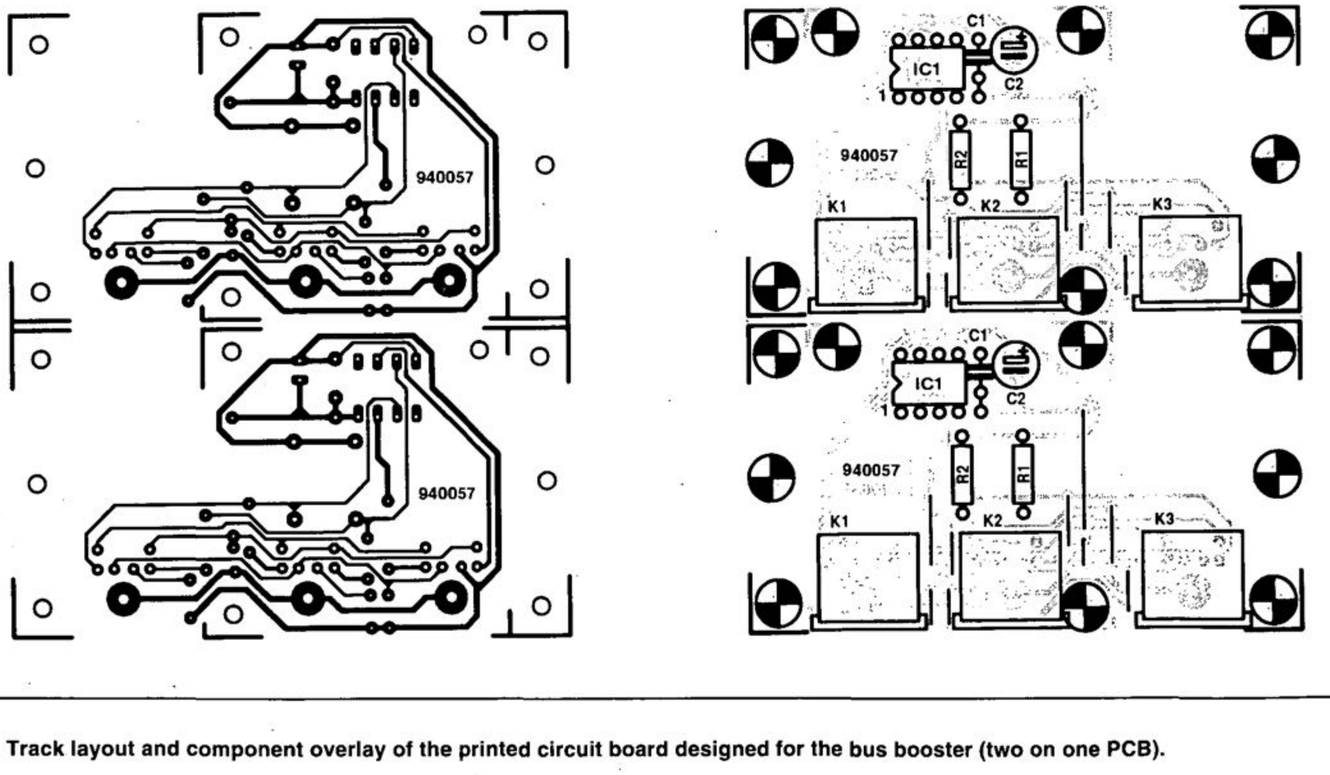


Fig. 4. Track layout and component overlay of the printed circuit board designed for the bus booster (two on one PCB).

RC ACTIVE FILTERS

By H. Kroeze

Could the authors of 'A Practical Method of Designing RC Active Filters' which appeared in the March 1955 issue of *IRE Transactions on Circuit Theory*[1] have expected that the unity-gain low-pass and high-pass sections would still be highly popular 40 years later? Perhaps. What is certain is that this popularity is not misplaced: only two resistors, two capacitors an an operational amplifier are needed to construct a unity-gain, second-order filter with a high input impedance and low output impedance. Several of such second-order sections can be cascaded to obtain a higher even-order filter[2]. Formulas for calculating the component values to obtain given cut-off frequencies and filter types abound[3]. Unfortunately, almost without exception, these are intended for calculating the capacitor values and assume equivalent resistors.

Figure 1 shows an example of a fourth-order Butterworth low-pass filter (f_c = 1000 Hz) that has been computed with the aid of such formulas. It is clear that they give rise to awkward values of capacitors: 172 nF, 147 nF, 416 nF and 60.9 nF. Since capacitors not falling in the E12 series of values are virtually unobtainable, they have to be constructed from series/parallel combinations or have to be rounded to the nearest E12 value. Rounding off often results in a serious deviation of the wanted characteristic, while combinations of capacitors cost more components.

There is, however, another way. The characteristic of the filter in **Fig. 2** is identical to that of the section in **Fig. 1**. However, the filter in **Fig. 2** uses capacitors that have standard E12 values. The price for this is that the resistors are no longer equivalent and have no E12 values. This does not matter, however, because 1% metal film resistors in the E96 series, which are freely available, can be used. Rounding off to the nearest E96 value results in negligible deviations from the wanted characteristic.

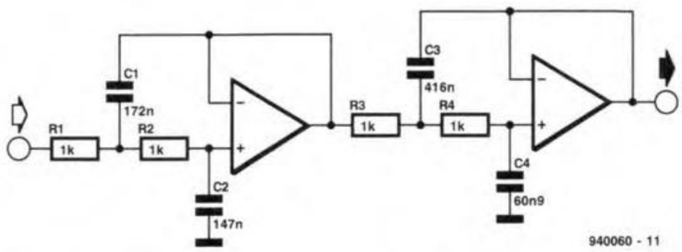


Fig. 1. Fourth-order Butterworth low-pass filter (f_c = 1 kHz) designed in the classical manner, that is, with equivalent resistors.

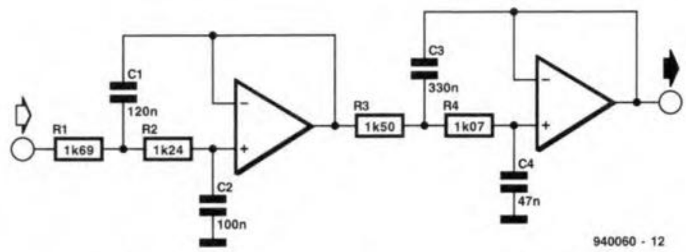


Fig. 2. The same filter as in Fig. 1 but now with capacitors with E12 values.

100	102	105	107	110	113	115	118
121	124	127	130	133	137	140	143
147	150	154	158	162	165	169	174
178	182	187	191	196	200	205	210
215	221	226	232	237	243	249	255
261	267	274	280	287	294	301	309
316	324	332	340	348	357	365	374
383	392	402	412	422	432	442	453
464	475	487	499	511	523	536	549
562	576	590	604	619	634	649	665
681	698	715	732	750	768	787	806
825	845	866	887	909	931	953	976

Standard values in the E96 series.

Computation

Another result of using E12 value capacitors is that computing the component values has become rather more difficult. For this reason, Table 1 simplifies the design of Butterworth or Bessel sections of the second, fourth, sixth and eight order. The table is normalized for 1 Hz; the resistor and capacitor values are in ohms and fards respectively. The required number of sections is given for each order; each section can be selected from three sets of component values. The capacitor values in successive sets differ roughly by a factor 2, so that in almost all cases an optimum choice can be made.

The computation is commenced by selecting a component set for each section of the wanted order. Divide the values of the capacitors in that set by a suitable power of 10, say, 1 000 000, to obtain values in µF or nF. Multiply the resistance value by the same power of 10 and divide the result by the cut-off frequency (in Hz). **Figure 3** gives an example of an eighth-order Bessel filter with a cut-off frequency of 15 Hz; the selected sets are indicated in **Table 1**.

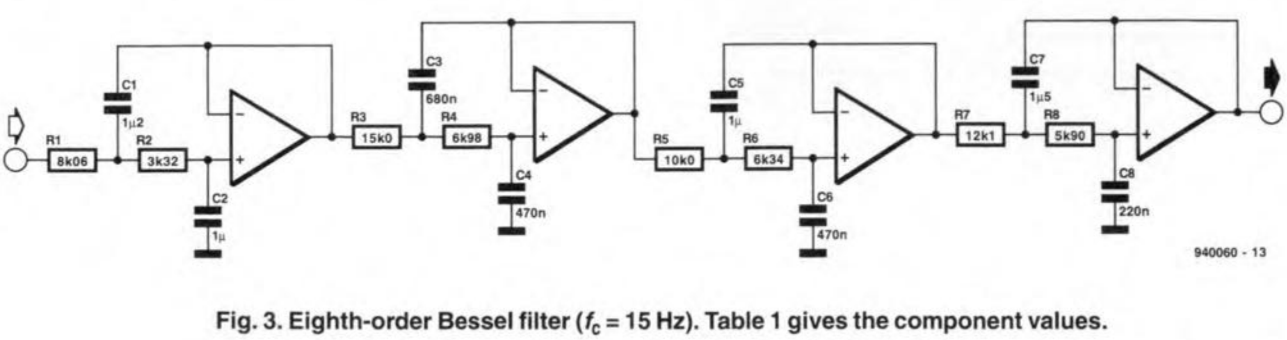


Fig. 3. Eighth-order Bessel filter (f_c = 15 Hz). Table 1 gives the component values.

Other filters

Figure 4 shows the basic design of a third-order section[3]. Several books publish the design formulas, but these give the same problems with the capacitor values as mentioned earlier.

Selecting E12 values for the capacitors comes down to a solving a non-linear set of three equations with peripheral conditions. This is virtually impossible by conventional means[6], but is fairly straightforward with modern mathematics software such as MathCad. This program gives a fairly easy to use design table for Butterworth and Bessel sections—see **Table 2**. From this table, a set of the wanted type is selected, after which the values of the capacitors and resistors are determined as before. **Figure 4** shows a 50 Hz Butterworth filter designed with the aid of this table.

It is also possible to design a fourth-order section with only one opamp[6]—see **Fig. 5**. The solution of ther associated set of equations is typified by the large ratio of C_4 to the other capacitors. This limits the application of this

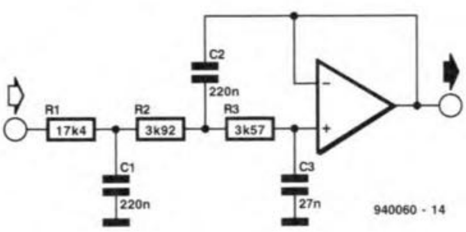


Fig. 4. Third-order Butterworth filter (f_c = 50 Hz)

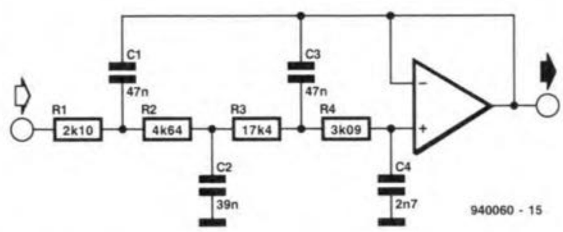


Fig. 5. Fourth-order Bessel filter (f_c = 1 kHz)

LP order 2 amp 1	Butterworth section 1 R1 C1	R2 C2	Bessel section 1 R1 C1	R2 C2
set 1	6.4078 0.047	3.8230 0.022	6.5671 0.033	3.2835 0.022
set 2	2.981 0.1	1.8079 0.047	2.9511 0.068	1.6598 0.047
set 3	1.4647 0.22	0.7861 0.1	1.4448 0.15	0.7224 0.1

LP order 4 amp 2	Butterworth section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4	Bessel section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4
set 1	8.1075 0.027	5.2598 0.022	3.8944 0.18	1.6425 0.022	6.4648 0.027	3.2232 0.022	3.9217 0.068	1.6795 0.022
set 2	3.5330 0.056	2.7240 0.047	1.5106 0.33	1.0811 0.047	2.9313 0.056	1.6051 0.047	1.8774 0.15	0.7445 0.047
set 3	1.6967 0.12	1.2441 0.1	0.8582 0.82	0.3600 0.1	1.3893 0.12	0.7428 0.1	0.7373 0.27	0.4950 0.1

LP order 6 amp 3	Butterworth section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4	section 3 R5 C5	R6 C6	Bessel section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4	section 3 R5 C5	R6 C6
set 1	9.4750 0.027	4.5007 0.022	6.4078 0.047	3.8230 0.022	2.0019 0.33	1.7428 0.022	6.1376 0.027	2.7008 0.022	3.7208 0.033	3.2864 0.022	2.3757 0.1	1.3359 0.022
set 2	4.3076 0.056	2.2342 0.047	2.9810 0.1	1.8079 0.047	1.2094 0.82	0.5434 0.047	2.8020 0.056	1.3351 0.047	2.2613 0.082	1.0186 0.047	1.1503 0.22	0.5870 0.047
set 3	2.0398 0.12	1.0348 0.1	1.4647 0.22	0.7861 0.1	0.4404 1.5	0.3834 0.1	1.3253 0.12	0.6191 0.1	0.8186 0.15	0.7230 0.1	0.5429 0.47	0.2736 0.1

LP order 8 amp 4	Butterworth section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4	section 3 R5 C5	R6 C6	section 4 R7 C7	R8 C8
set 1	9.8702 0.027	4.3205 0.022	7.1515 0.033	4.8787 0.022	5.4726 0.082	2.5657 0.022	1.9579 0.68	0.8648 0.022
set 2	4.5072 0.056	2.1353 0.047	2.8591 0.068	2.7721 0.047	2.6197 0.18	1.1429 0.047	0.9384 1.5	0.3829 0.047
set 3	2.1317 0.12	0.9902 0.1	1.5733 0.15	1.0733 0.1	1.0036 0.33	0.7648 0.1	0.3614 2.7	0.2596 0.1

LP order 8 amp 4	Bessel section 1 R1 C1	R2 C2	section 2 R3 C3	R4 C4	section 3 R5 C5	R6 C6	section 4 R7 C7	R8 C8
set 1	5.6550 0.027	2.3841 0.022	4.9607 0.033	2.0954 0.022	3.2100 0.047	2.0005 0.022	1.8128 0.15	0.8839 0.022
set 2	2.5867 0.056	1.1763 0.047	2.2564 0.068	1.0465 0.047	1.4922 0.1	0.9467 0.047	0.8708 0.33	0.3915 0.047
set 3	1.2229 0.12	0.5457 0.1	1.0914 0.15	0.4610 0.1	0.7365 0.22	0.4098 0.1	0.3978 0.68	0.1955 0.1

Table 1. Component values for designs with 1–4 operational amplifiers. Component numbers as in Fig. 3.

design to a narrower frequency range. Also, because of component tolerances and parasitic effects, such as the finite output impedance of the opamp, the deviation from the wanted characteristic is rather larger than in the two opamp design. Nevertheless, this design may prove useful in situations where accuracy is not of paramount importance.

The conversion to E12 values of the capacitors in the fourth-order section, as shown in Table 3, is optimised for minimum capacitor ratios. It appears that Butterworth filters can be designed with a capacitor ratio of 50, whereas in Bessel filters the ratio should not be greater than about 15. **Figure 5** gives a worked example of a Bessel filter with a cut-off frequency of 1 kHz.

An interesting application of the one opamp design is that it enables replacing a second-order Butterworth filter in an existing equipment by a fourth-order Bessel filter. This requires no more than changing the values of the existing resistors and capacitors and adding two resistors and two capacitors. The graph in **Fig. 6** shows that the frequency characteristic of a fourth-order Bessel section is better than that of a second-order Butter-worth filter. Moreover, the Bessel filter is phase-linear and does not produce ringing of pulse-shaped signals.

LP order 3 amp 1	Butterworth			Bessel		
	R1 C1	R2 C2	R3 C3	R1 C1	R2 C2	R3 C3
set 1	1.9499 0.1	4.4880 0.1	3.8390 0.012	1.5400 0.1	1.8916 0.1	2.2694 0.022
set 2	0.8786 0.22	1.9594 0.22	1.7919 0.027	0.6367 0.22	1.0456 0.22	0.8061 0.056
set 3	0.4159 0.47	0.9667 0.47	0.8105 0.056	0.2972 0.47	0.4929 0.47	0.3746 0.12

Table 2. Component values for third-order filter designed as in Fig. 4.

LP order 4 amp 1	Butterworth				Bessel			
	R1 C1	R2 C2	R3 C3	R4 C4	R1 C1	R2 C2	R3 C3	R4 C4
set 1	1.4520 0.1	4.9960 0.056	12.860 0.1	5.5790 0.0022	0.9708 0.1	2.3380 0.082	3.5570 0.1	2.7100 0.0068
set 2	0.6496 0.22	2.3760 0.12	5.3660 0.22	2.8380 0.0047	0.4406 0.22	1.0660 0.18	1.6270 0.22	1.2230 0.015
set 3	0.3122 0.47	1.0390 0.27	2.3460 0.47	1.4140 0.01	0.2111 0.47	0.4647 0.39	1.7100 0.47	0.3126 0.027

Table 3. Component values for fourth-order filter designed as in Fig. 5.

Addendum

The transfer function of a second-order filter is

$$T(s)=\frac{\omega_c^2}{s^2+\alpha\omega_c\cdot s+\omega_c^2}$$

Where ω_c is $2\pi f_c$ (natural frequency), a and b are arbitrary factors, α = magnification, and $s = j\omega$.

Let $R_1 = aR$; $R_2 = R$; $C_1 = bC$; and $C_2 = C$, then:

$$\omega_c=\frac{1}{RC\sqrt{ab}}$$

and

$$\alpha=\frac{1+a}{\sqrt{ab}}$$

$$1+a = \alpha\sqrt{(ab)}$$

$$a^2+2a+1 = \alpha^2 ab$$

$$a^2+a[2-\alpha^2 b]+1 = 0$$

$$a = \{(\alpha^2 b-2)+\sqrt{D}\}/2$$

The denominator polynomial, $D_{(s)}$, is

$$D_{(s)} = [2-\alpha^2 b)^2-4 \qquad [1]$$

$$D_{(s)} = \alpha^4 b^2-4\alpha^2 b > 0$$

This can be solved only if $D_{(s)}$ is positive.

$$\alpha^2 b[\alpha^2 b-4] > 0$$

with α and b positive, so that

$$\alpha^2 b > 4$$

$$b > 4/\alpha^2$$

$\therefore$ α determines a minimum for b.

Select a slightly larger b from the E-12 series, calculate a from [1] and then C_1, C_2, R_1 and R_2.

[940060]

References

1. 'A Practical Method of Designing RC Active Filters' by R.P. Sallen & E.L. Key; *IRE Transactions on Circuit Theory*, March 1955, pp. 74–85.
2. *Operational Amplifiers* by G.E. Tobey, J.G. Greame, L.P. Huelsman; pp. 295–321.
3. *Filter Handbook* by Stefan Niewiadomski, pp.74–75; Heinemann Newnes 1989.
4. *Function Circuits* by Y.J. Wong & W.E. Ott, pp. 228–236.
5. *Electronic Filter Design Handbook* by Arthur B. Williams & Fred J. Taylor; McGraw-Hill 1988.

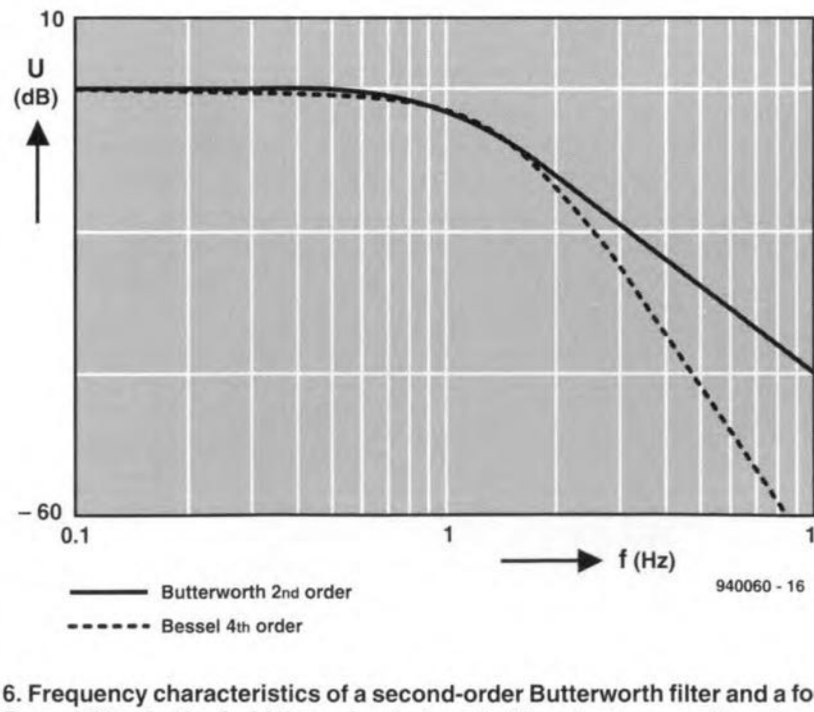
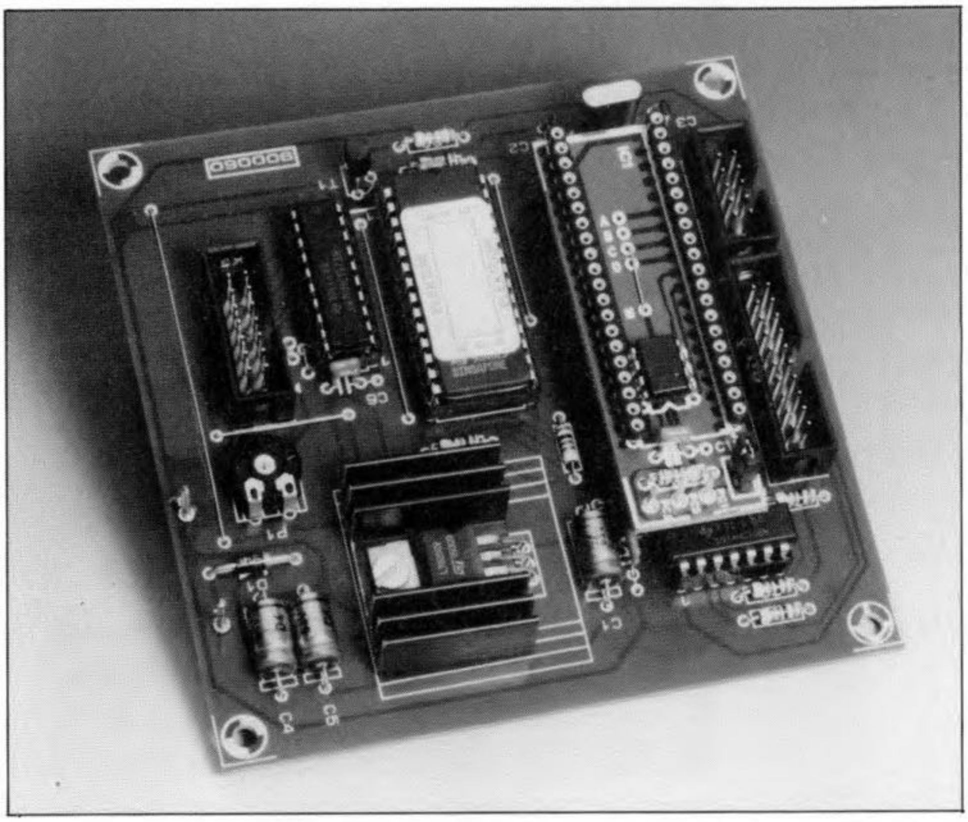


Fig. 6. Frequency characteristics of a second-order Butterworth filter and a fourth-order Bessel filter, both of which can be designed with only one operational amplifier.

RS485 INTERFACE FOR MCS-51 MICROCONTROLLER SYSTEMS



RS485 MAIN SPECIFICATIONS

Driver:

Capacity:	max. 32 receivers
Output impedance:	120 Ω
Leakage current:	< 100μA
Diff. voltage:	1.5 to 5 V
Output current:	150 mA into ground
	250 mA at –7 to +12 V

Receiver:

Input impedance:	12 kΩ
Max. common-mode voltage:	–7 to +12 V
Diff. sensitivity:	±200 mV
Bit rate:	max. 32 Mb/s
ESD protection:	to 2.5 kV

Practically all of today's computer systems are equipped with an RS232 interface. That trusty and widely known serial interface enables a full-duplex link to be set up between two devices. Unfortunately, the applications of the RS232 interface remain limited to just that function. By contrast, the RS485 standard creates many more, and more advanced, possibilities. This article describes the operation of the RS485 bus, and how existing MCS51 microcontroller cards can be upgraded for small networking applications.

Design by W. Hackländer

PHYSICALLY, the RS485 interface is an improved version of the familiar RS232 interface. Introduced in 1983, the RS485 interface is mainly suited to implementing control systems and data distribution. An important improvement over the RS232 interface is the use of a symmetrical ('twisted-pair') link. However, this means that the two-wire link is only suitable for half-duplex connections. All transmitters and receivers are connected to the same two wires. Only one transmitter can be active at a time; all other connections are either in high-impedance mode, or listen to the information conveyed via the link.

In industrial applications, the RS485 system is often used as a local bus called P-net (a derivate of the Field-bus). In the PC environment, RS485 is used in SCSI and IPI interfaces. In the field of domestic electronics, it could be used to create a kind of home bus, with some interesting specifications: a maximum data rate in excess of 10 Mb/s (up to 30 Mb/s under favourable conditions, and using a symmetrical cable), cable lengths of up to 1,200 m, and reduced susceptibility to noise thanks to the use of differential inputs on the receivers. Also, the RS485 bus generates very little noise by virtue of the low signal levels in combination with the differential inputs. Not surprisingly, an increasing number of professional users turn to RS485 when it comes to implementing serial connections.

RS485: differential to combat noise

The major difference between a differential connection and a 'normal' single-ended one is that two signal wires are used as opposed to one. The signals on the two wires are always in anti-phase. Noise induced in the cable is always in phase, i.e., it has the same phase on the two signal wires, and is thus automatically cancelled at the receiver's differential inputs. Furthermore, the two anti-phase signals automatically cancel each other's electrical and/or magnetic fields. Consequently, the emission of noise is reduced to a minimum.

As always in transmission technology, the integrity of the data conveyed via the link is best protected by ensuring that the cable is terminated with the characteristic impedance. In practice, this means that terminating resistors are fitted at both ends of the cable. The value of these resistors is equal to the characteristic impedance of the network. As a rule of thumb, this impedance is the same as the output impedance of the transmitter, which is 120 Ω in the present case.

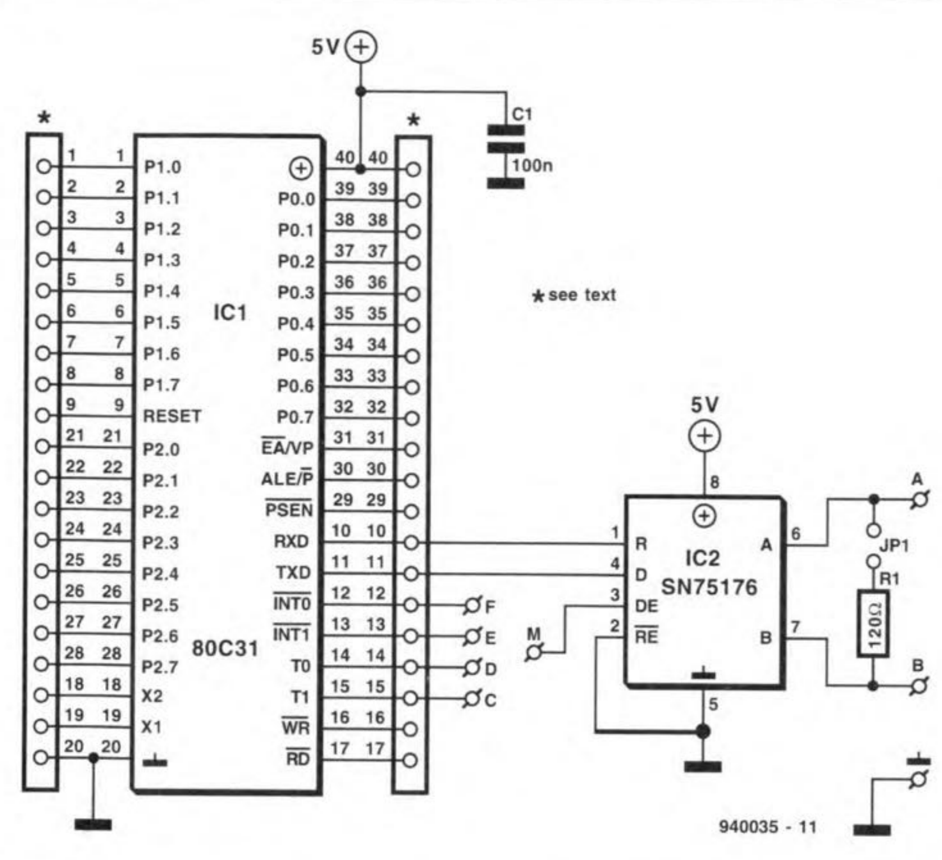


Fig. 1. Circuit diagram of the RS485 interface.

That also restricts the length of the link. Even good-quality copper wire of, say, 0.25 mm^2 cross-sectional area represents a certain resistance. A practical value is about 80 Ω per kilometre. Consequently, a cable with a length of between 1,200 and 1,500 m has an impedance which is equal to the characteristic impedance, which automatically gives us the maximum cable distance that can be used.

Another aspect which sets restrictions to the link concerns the stray capacitance and inductance inherent to any electrical cable. Since data is conveyed at relatively high speeds, steep edges appear in the signal, along with their associated higher harmonics. As the length of the cable increases, the harmonics level drops, so that the signal edges are less steep. The result is easily guessed: data is corrupted, and the serial link is no longer reliable if a certain cable length is exceeded.

RS485 hardware

The simplicity of an RS485 interface is illustrated by the drawing in **Fig. 1**. The SN75176 contains a complete transmitter and a receiver in a DIL-8 package. The standard version of this IC is marked by the suffix 'A', and has a typical current consumption of 35 mA. The 'B' version is much faster, but has a higher current consumption: about 42 mA. Finally, a low-power version is available, the SN74LBC176, with a current consumption of only 1.5 mA. Remarkably, this lower current consumption has virtually no negative effect on the performance of the IC.

The circuit diagram indicates that the RS485 interface replaces an existing RS232 interface. Furthermore, a microcontroller output is needed to actuate the driver. This is achieved by pulling the DE input logic high. A jumper allows you to determine which port line of, say, a 80C31 or 80C32, is used for that purpose. There is a snag, however: the input/output lines on all MCS51 family microcontrollers are

FUNCTION TABLE (RECEIVER)

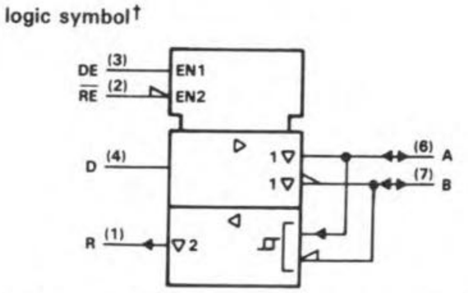
DIFFERENTIAL INPUTS A – B	ENABLE $\overline{RE}$	OUTPUT R
$V_{ID} \geq 0.2$ V	L	H
$-0.2\text{ V} < V_{ID} < 0.2$ V	L	?
$V_{ID} \leq -0.2$ V	L	L
X	H	Z

H = high level, L = low level, ? = indeterminate,
X = irrelevant, Z = high impedance (off)

FUNCTION TABLE (DRIVER)

INPUT D	ENABLE DE	OUTPUTS A	OUTPUTS B
H	H	H	L
L	H	L	H
X	L	Z	Z

logic symbol†



†This symbol is in accordance with ANSI/IEEE Std 91-1984 and IEC Publication 617-12.

logic diagram (positive logic)

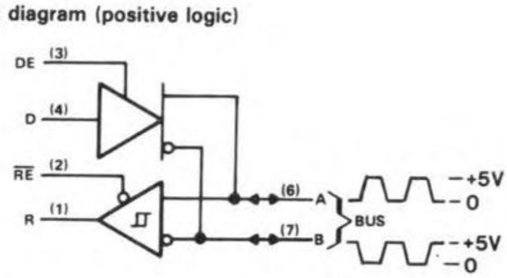


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Fig. 2. Internal schematic and function table of the SN75176B RS485 bus transceiver.

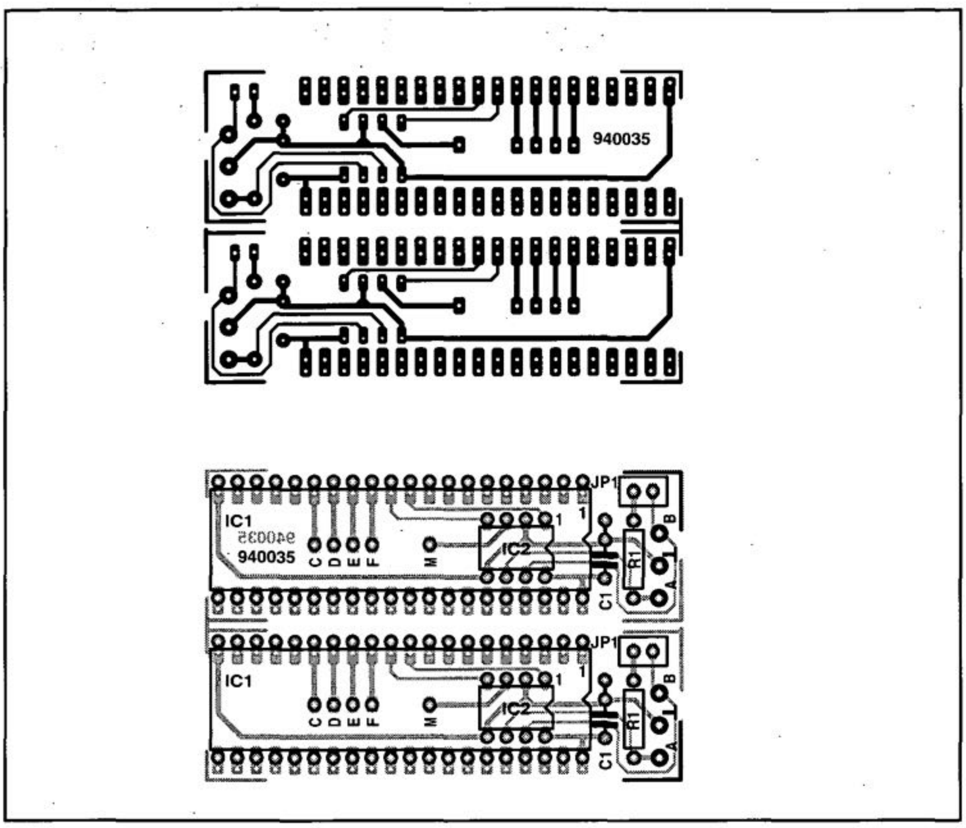

Fig. 3. Track layout (direct reading) and component overlay of the double-function (receiver/transmitter) PCB designed for the RS485 interface. The PCB is designed to upgrade existing MCS51 microcontroller systems with an RS485 interface.

COMPONENTS LIST

Resistors:
R1 = 120Ω

Capacitors:
C1 = 100nF

Semiconductors:
IC1 = existing MCS51 processor
IC2 = SN75176BP or equivalent

Miscellaneous:
JP1 = jumper
3 PCB terminals
Printed circuit board 940035 (see page 70).

logic high after a reset. Consequently, the interface would be enabled immediately. This can be prevented only by programming, i.e., writing a '0' to the relevant port line as part of the software initialization.

Resistor R_1 is used in conjunction with jumper JP_1 to connect a terminating resistor on the line (if necessary; see below).

The RS485 protocol

A single RS485 driver may be loaded with up to 32 receiver inputs. Since all receivers and transmitters in an RS485 network ('mini LAN') are connected to one and the same twisted-pair, only one transmitter may be active at a time. If one transmitter is active, all others must be in high-impedance mode. This is ensured by the protocol drawn up for the RS485 interface. You may even write your own protocol. If you have a system which has an RS485 interface available, only appropriate software is required to set up a local network.

An example of a protocol that may be taken as a basis is found in the European standard for mains signalling. No message conveyed via the network may take longer than 1 second. After such a message, a pause must be observed with a minimum length of 125 ms. During the first 80 ms of this pause, no device is allowed to take control of the bus. After this part of the pause, i.e., between 85 ms and 115 ms, any transmitter may attempt to take control of the bus. To prevent bus conflicts, all stations have slightly different delays, which enables a certain priority system to be set up. Transmitters with a high priority, for example, wait 85 ms, while transmitters with a low priority wait longer, say, 110 ms. No transmitter has successfully taken control of the bus until it is capable of reading back a perfect copy its own transmitted data. Next, the communication with the desired receiver(s) may be initiated. A receiver must be capable of detecting whether or not data on the bus must be accepted or ignored. Consequently, the header must contain at least one address for the relevant receiver to be selected.

Construction

The printed circuit board, **Fig. 3**, has been designed for compatibility with a microcontroller system based on an MCS51 device. An example of such a system is the 80C32 microcontroller-board described in **Ref. 1**.

Since an RS485 link consists of at least one transmitter and one receiver, the two functions are combined on one printed circuit board which is supplied through the Readers Services. The two sub-boards are separated before they are populated. Position IC_1 receives an IC socket with long pins, which are inserted into the controller socket on the MCS51 computer board. A wire-wrap socket may be used for this purpose, but do make sure to stack an additional IC socket onto the one on the MCS51 board, because the wire-wrap pins are so thick that they can only be inserted once into the spring-loaded contacts of a normal IC socket.

Fit the wire link to select the microcontroller port line you wish to use to enable the RS485 interface (connect M to C, D, E or F). Remove the microcontroller from its socket on he MCS51 board, and insert a standard IC socket instead. In this socket insert the pins of the RS485 extension board. Then insert the microcontroller into the socket on the extension board.

Any existing hardware of the RS232 interface on the original MCS51 board should be disabled or even removed. If not, you will be faced with a bus conflict at the controller's RxD input.

Towards your own network

When wiring up the network, be sure to connect all A's to A's, and B's to B's. The terminating resistor must be used at two points only, namely the start and the end of the link. The jumper is fitted only at these points — on all other boards, it is omitted. The circuit is then ready for use, and perhaps unwittingly you have made the first step towards setting up a local network.

(940035)

Reference:
1. Radio data system decoder, *Elektor Electronics* February 1991. Board order code: 900060, see *Product Overview* elsewhere in this issue.

HEATSINKS: HOW AND WHEN TO USE THEM

By our technical staff

Anyone who has ever designed or built a power supply or power amplifier has experienced the problem of heat dissipation. The heat produced in such equipment can be quite substantial. Heatsinks are normally used to improve the thermal capacity of power semiconductors and therefore to enable them to dissipate the heat generated when they are in operation. With some background information, it is not too difficult to calculate the size of heatsink required for a particular application.

Thermal resistance

The thermal resistance, R_{th} (or ϕ), of a semiconductor device can be expressed by

$$R_{th} = \Delta T / \Delta P, \qquad \text{[Eq. 1]}$$

where R_{th} is expressed in **thermal ohms**, T in **kelvin** and P in **watts**. A kelvin (K) is equivalent to a °C, so that formulas may be encountered that use °C instead of kelvin. The idea of thermal resistance may become clearer with reference to **Fig. 1**. This shows a closed space, such as a room, which is warmed by a heater to a temperature T_1. Outside the room, the temperature is T_2. If the thermal resistance of the walls of the room is known, it can be calculated how much heat is lost from the room to the outside. A well-insulating wall has a high thermal resistance, so that relatively little heat is lost to the outside ($\Delta P = \Delta T / R_{th}$). From Eq. 1 it appears that there is no heat loss when $\Delta T = 0$. In **Fig. 1**, this means, of course, that the temperature in the room is the same as that outside.

Most buildings nowadays have cavity walls with the space between the two layers of bricks often filled with insulating material—see **Fig. 2**. The thermal resistance of such a wall actually consists of several thermal resistances in series—see **Fig. 3**.

Series resistances

Just as in the case of a cavity wall, the thermal resistance of a semiconductor mounted on a heatsink consists of a number of thermal resistances in series. In **Fig. 4**, a power transistor is mounted on a heatsink with the aid of an isolating washer – note that the distance between these are shown exaggerated. The first thermal resistance, $R_{th\,j\text{–}mb}$, is between the semiconductor junction (j) and the mounting base (mb). The second, $R_{th\,mb\text{–}h}$ is that between the mounting base (mb) and the heatsink (h). The value of the latter depends on the material of the isolating washer and whether there is heat conducting paste present. Then there is the thermal resistance, $R_{th\,h\text{–}a}$, between heatsink (h) and the ambience (a). A schematic representation of the three resistances is given in **Fig. 5**. This figure also shows a source of heat, T_j-T_a. The flow of heat is indicated by P.

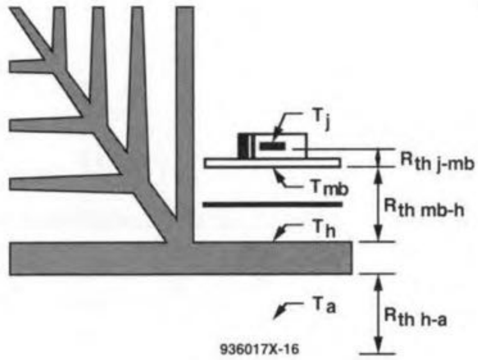

Fig. 4.

Some guidelines

Determining the parameters of a heatsink is normally started at the hottest point of the chain of resistance: the semiconductor. Data in transistor databooks normally include several thermal resistances. **Table 1** gives the thermal resistance of a number of popular transistor cases. $R_{th\,j\text{–}a}$ is the thermal resistance between the semiconductor and the ambience; this value applies only if no heatsink is used. If a heatsink is used, thermal resistance $R_{th\,j\text{–}mb}$ applies.

Another important property needed for the calculation is the maximum semiconductor temperature, T_j. This is normally given in the relevant databook and is usually not higher than 200 °C.

Thermal resistance $R_{th\,mb\text{–}h}$ is determined by the isolating material, if any, between the semiconductor mounting base and the heatsink. **Table 2** shows a number of different mounting methods and the associated thermal resistances. If, for instance, the semiconductor is mounted directly on the heatsink, the thermal resistance between the two is fairly low, but can be lowered further by the use of heat conducting paste between them.

When a semiconductor is mounted on a heatsink, electrical isolation is often required and this can be obtained by the use of isolating washers. These may be made from a variety of materials: beryllia is the most expensive, but has the highest thermal conductivity and dielectric strength, followed by hardened anodised aluminium washers which have good thermal conductivity and dielectric strength. Mica washers used to be popular but they can crack and peel with time. High-temperature plastics, such as Kaplon and Mylar, have lower dielectric strength than mica, but they are cheaper.

The value of the third thermal resistance, $R_{th\,h\text{–}a}$, is given in the heatsink data. This resistance, given in K W^{-1} (occasionally in °C W^{-1}), refers to a black-bodied heatsink that is mounted vertically (fins upwards). **Table 3** shows what happens to the thermal resistance of a heatsink in adverse conditions. If the heatsink

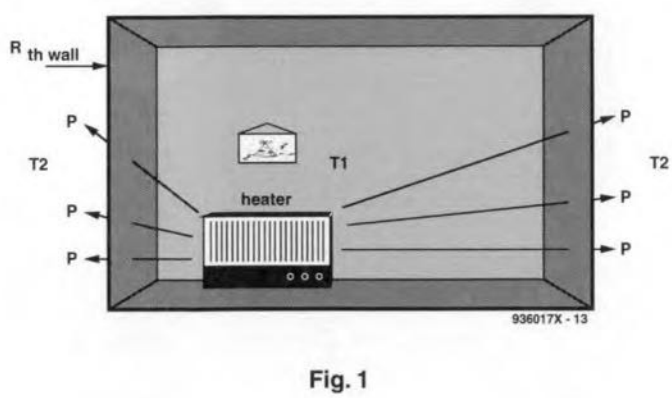

Fig. 1

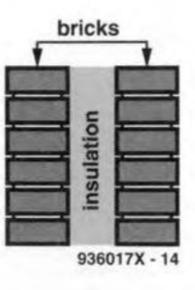

Fig. 2

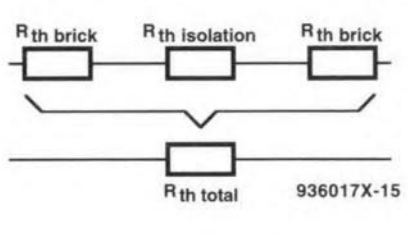

Fig. 3

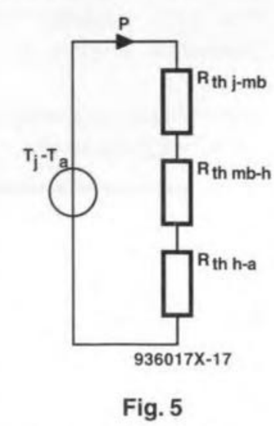

Fig. 5

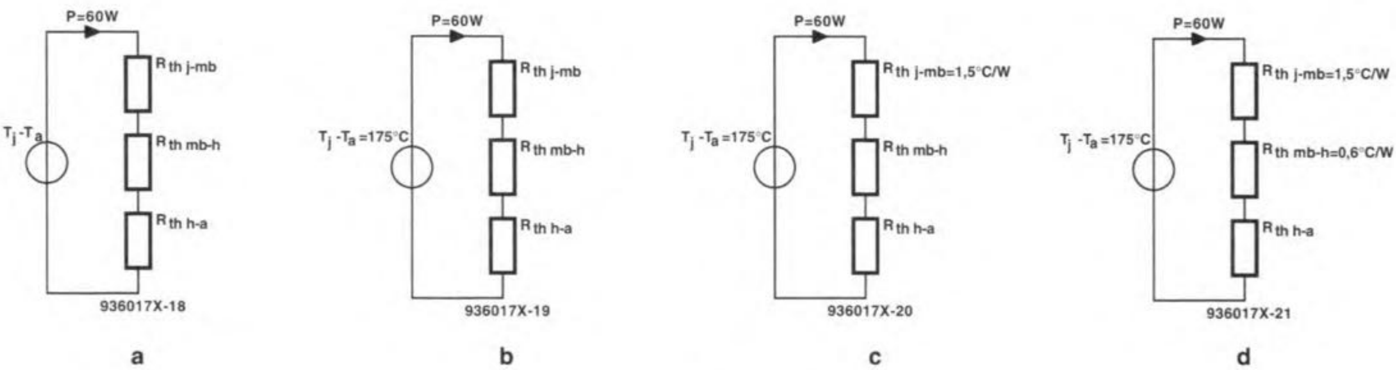

Fig. 6

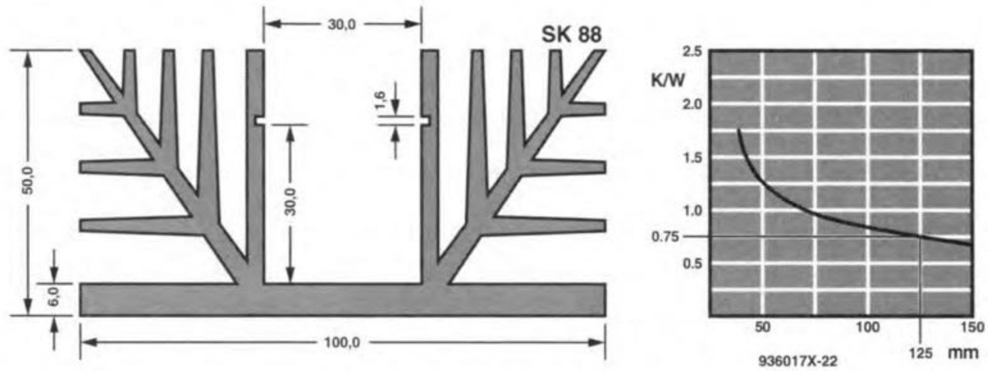

Fig. 7.

is not black, but natural aluminium, its thermal resistance increases by 10–15%. If the heatsink is mounted horizontally instead of vertically, its thermal resistance increases by 15–20%.

Practical calculation

To calculate a heatsink for a power transistor with a TO-3 case, assume that the maximum voltage at the emitter-collector junction is 20 V and that the emitter current at this voltage is 3 A. The dissipation is thus 3×20 = 60 W. This is the first data to be put into the heat flow diagram in **Fig. 6a**. The transistor databook states that the maximum temperature, T_j, of the semiconductor must not exceed 200 °C. This is a worst-case figure: if the transistor were really that hot in practice, touching it would be very serious indeed.

Case	$R_{th\ j\text{-}a}$	$R_{th\ j\text{-}mb}$
TO-18	500	200
TO-92	250	150
TO-39	200	12.5
TO-126	100	5
TO-220	70	2
TO-3	40	1.5

Table 1

Next, assume an ambient temperature, T_a, of 25 °C. The temperature difference, T_j–T_a, is then 175 °C. This value is also entered into the heat flow diagram—**Fig. 6b**.

The total thermal resistance, $R_{th\,total}$, is calculated from Eq. 1:

$$R_{th\ total} = \Delta T / \Delta P = 175/60 = 2.92\ \mathrm{K\ W^{-1}}.$$

The sum of all thermal resistances must, therefore, not be higher than this value.

For a TO-3 case, **Table 1** gives a value for $R_{th\ j\text{-}mb}$ of 1.5 K W^{-1}, and this value is entered into the heat flow diagram in **Fig. 6c**.

The best isolation between the transistor and the heatsink is, according to **Table 2**, an aluminium oxide washer in conjunction with heat conducting paste. This gives a thermal resistance, $R_{th\ mb\text{-}h}$, of 0.6 K W^{-1} (worst case), which is also entered into the heat flow diagram—see **Fig. 6d**.

From these data, the thermal

Isolating material	$R_{th\ mb\text{-}h}$
None	0.05–0.2
Heat conducting paste	0.005–0.1
Aluminium oxide and paste	0.2–0.6
Mica (0.05 mm) and paste	0.4–0.9
Silicon rubber, no paste	0.84–0.88

Table 2

Natural aluminium	$R_{th\ h\text{-}a}$ + 10–15%
Horizontal fitting	$R_{th\ h\text{-}a}$ + 15–20%

Table 3

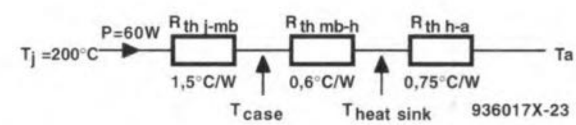

Fig. 8

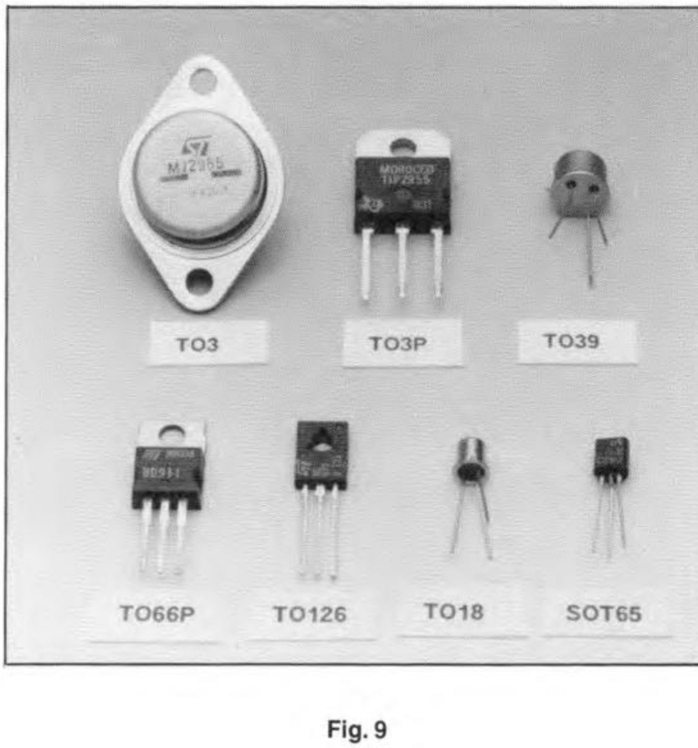

Fig. 9

resistance, $R_{th\,h\text{-}a}$, of the heatsink is calculated:

$$R_{th\,h\text{-}a} = 2.92 - (1.5 + 0.6) = 0.82\ \mathrm{K\,W^{-1}}.$$

Figure 7 shows a suitable heatsink with manufacturers' specifications. The graph shows that the heatsink, cut to a length of 125 mm, has a thermal resistance of about 0.75 K W^{-1}, which is slightly lower than the required 0.82 K W^{-1}. This figure applies, as stated before, to a black-bodied heatsink mounted with its fins upwards.

Safety and isolation

In the calculation, the maximum junction temperature, T_j, was taken as 200 °C. In practice, it is advisable to build in a safety margin and take a value for T_j of 100–150 °C.

From the data established in the calculation, the temperature of the case and the heatsink at full load can be calculated with reference to the redrawn heat flow diagram in **Fig. 8**. The temperature of the case, T_{mb}, is equal to the junction temperature, T_j, less the temperature difference between junction and case, $T_{j\text{-}mb}$. The temperature developed across $R_{th\,j\text{-}mb}$ is the value of the resistance times the heat flow:

$$T_{j\text{-}mb} = R_{th\,j\text{-}mb} \times P = 1.5 \times 60 = 90\ °\mathrm{C}.$$

Thus, the temperature of the case, T_{mb}, is:

$$T_{mb} = T_j - T_{j\text{-}mb} = 200 - 90 = 110\ °\mathrm{C}.$$

This shows at once that it is dangerous to touch the transistor case, so that during construction great care must be taken to ensure that this component can

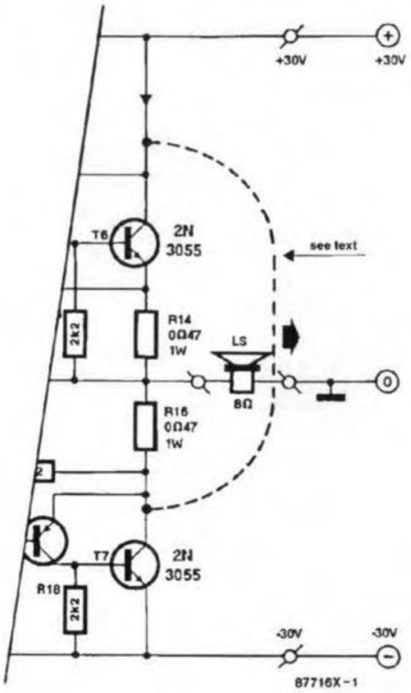

Fig. 10

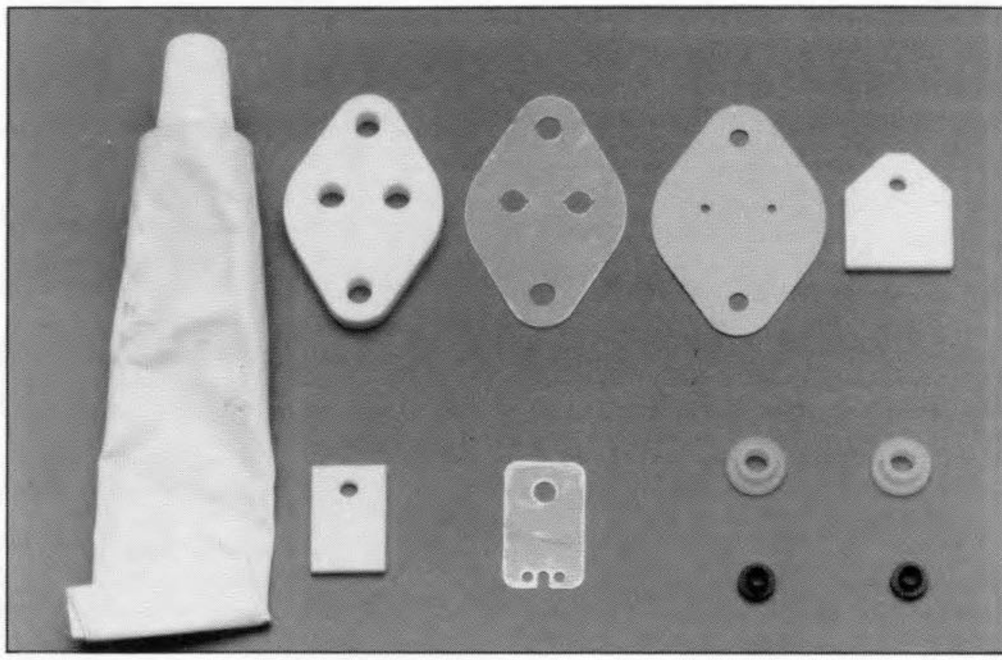

Fig. 11

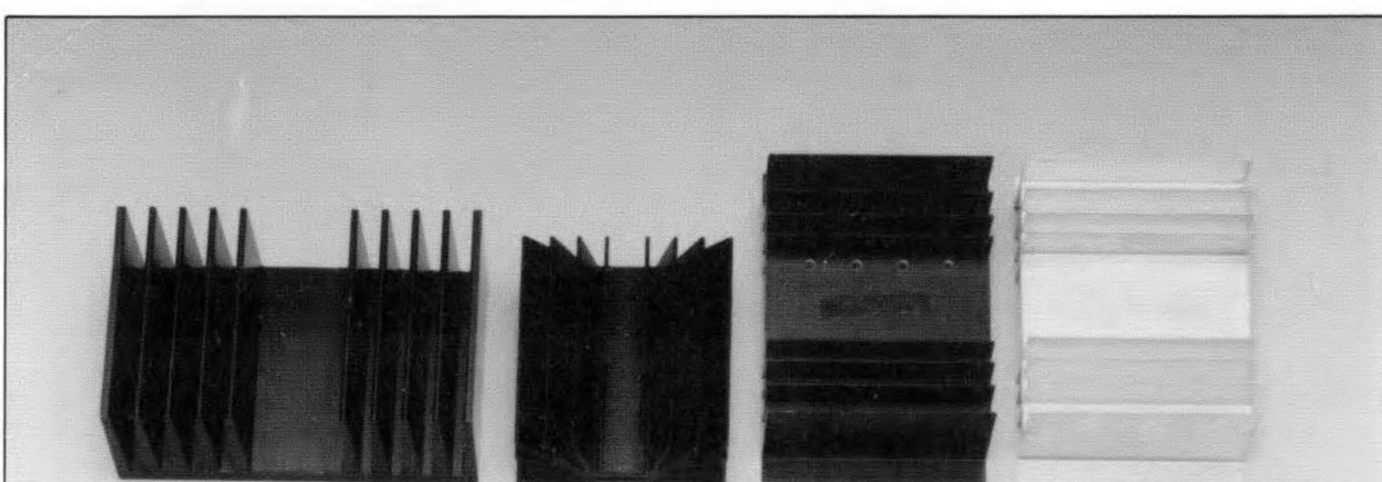

Fig. 12

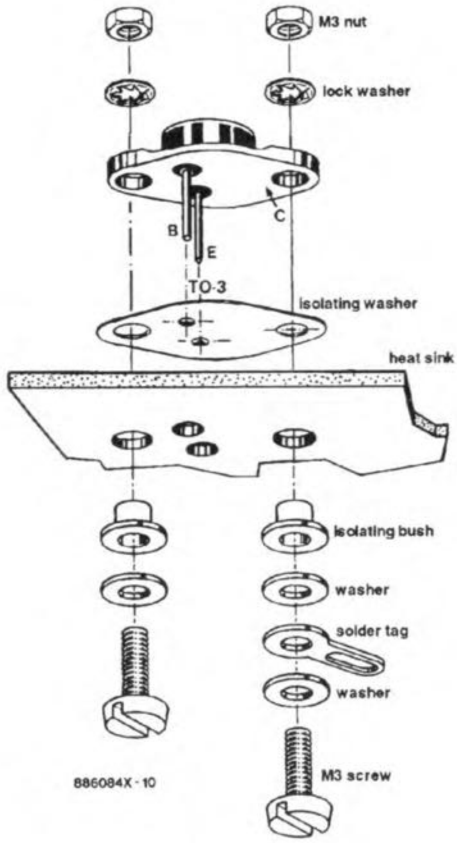

Fig. 13

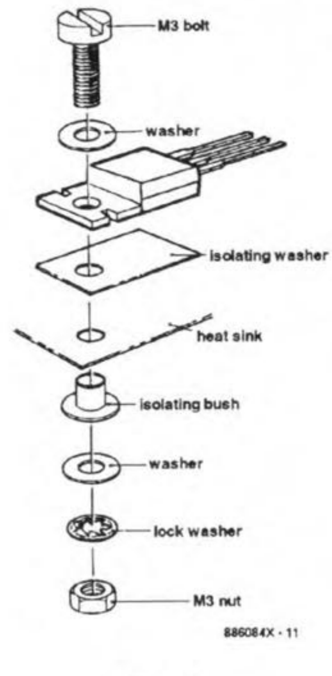

Fig. 14

not be touched accidentally.

The temperature, $T_{mb\text{-}h}$, developed across the isolating washer and heat conducting paste is:

$$T_{mb\text{-}h} = R_{th\ mb\text{-}h} \times P$$
$$= 0.6 \times 60 = 36\ ^\circ C.$$

Thus, the temperature of the heatsink is:

$$T_h = T_{mb} - T_{mb\text{-}h}$$
$$= 110 - 36 = 74\ ^\circ C.$$

The temperature, $T_{h\text{-}a}$, developed between the heat sink and the ambience is:

$$T_{h\text{-}a} = R_{th\ h\text{-}a} \times 60$$
$$= 0.75 \times 60 = 45\ ^\circ C.$$

Thus, the ambient temperature, T_a, may rise to

$$T_a = T_h - T_{h\text{-}a} = 74 - 45 = 29\ ^\circ C.$$

When it rises above this value, the earlier calculated values must be reappraised. Even in temperature climates, the ambient temperature during the summer months may rise to 35 °C.

In practice, and for safety's sake, the semiconductor temperature, T_j, should be taken between 100 °C and 150 °C. The temperature, T_h, of the heatsink should be not higher than the maximum ambient temperature plus 25 °C, that is, in temperate climates, not higher than 60 °C.

Practicalities

Figure 14 shows the most frequently encountered semiconductor cases with their coding. The thermal resistance of most of these cases is given in **Table 1**.

To see why it is essential in many cases to isolate the semiconductor from the heatsink, consider **Fig. 10**. Output transistors T_6 and T_7 are housed in a TO-3 case, which needs to be mounted on a heatsink. The collectors of the transistors are connected to the (metal) case, as is also often the case with voltage regulators. If the transistors were mounted on the heatsink without isolating washers, they would be interlinked so that the positive supply line would be short-circuited. However, even when there is only one semiconductor on a heatsink, it is advisable to isolate the two. This may be important if, for instance, the fixing terminal carries a voltage of >42 V, which can be lethal for human beings.

Figure 11 shows a number of isolating parts; the tube at the left is heat conducting paste which appreciably improves the heat transfer between semiconductor case, isolating washer and heat-sink.

The light-coloured washers are made of ceramic aluminium oxide, which has a low thermal resistance; only beryllia (beryllium oxide) is better, but much more expensive.

The transparent ones are of mica, which has a fairly good thermal conductivity, but whose mechanical properties are not so good, as stated earlier.

The grey washers are of silicon rubber. This material does not have such a good thermal resistance, but it offers several advantages. First, the washers do not need heat conducting paste, since this makes very little, if any, difference. Second, the elasticity of the rubber ensures good thermal contact even if the surface of the heatsink is rough. Mechanically, these washers are very durable.

The isolating bushes at the bottom righthand side are for use around the screws with which the semiconductors are fixed into place.

Figure 12 shows a number of heat-sinks. The thermal resistance of some of them (SK71, SK64, SK01, SK59) depends on their length (they can be cut to the required size). The others can not be altered, so their thermal resistance is fixed. Note the difference in thermal resistance between the black-bodied Type SK71/SK75 and the same size aluminium SK01/AL75 beside it.

Fixing heatsinks

The fixing of a heatsink often starts with drilling the required fixing holes. This does not apply, of course, to predrilled types. If drilling is required, it is best to use the isolating washer as template. In the larger heatsinks, the holes for the fixing screws may be tapped to avoid (protruding) nuts and washers at the back of the heatsink.

Figures 13 and **14** show fixing details of two of the most commonly encountered semiconductors.

TO-3-cased transistors are normally fixed between the fins of the heatsink. Since the metal case of these devices forms the third terminal, a solder tag must be fitted under one of the fixing screws for connection to the remainder of the circuit. Safety is increased, and the risk of short-circuits reduced, by fitting a small sheet of aluminium or perspex into the slots provided—see **Fig. 15**. There are also special covers available for TO-3 cases.

Finally

Once familiarity with the foregoing has been gained, calculating the parameters of heatsinks will have become as easy as working with Ohm's law.

[936017]

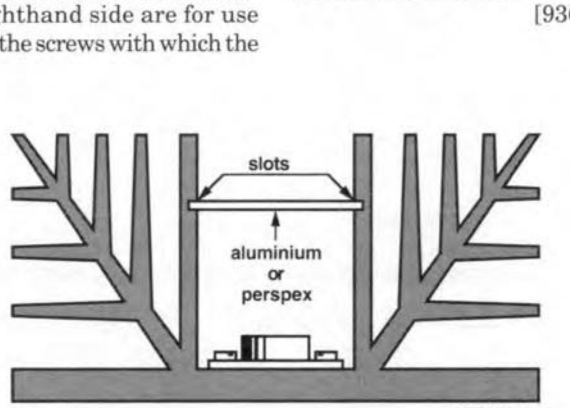

Fig. 15

920152 - II A

80C535 ASSEMBLER/HARDWARE COURSE

PART 4 (FINAL): TIMER-2 PROGRAMMING AND COMPARE/CAPTURE UNIT (CCU)

Software by Dr. M. Ohsmann

Timer-2 interrupts

Timer-2 can also generate interrupts. An interrupt is generated if it has been enabled (bit 5 in register IEN0, see Fig. 7), and if subsequently the TF2 or EXF2 bit has been set (compare Fig. 9). It is, therefore, possible to have an interrupt generated on an overflow, or a timer reload caused by an external event. How this can be achieved is demonstrated by the example below.

Let us assume timer-2 is to generate an interrupt every 500 µs. The interrupt routine should toggle port line P1.6 on each interrupt. Consequently, the main program must initialize timer-2 accordingly, and may then enter into an endless loop. The interrupt routine, triggered by timer-2, then does the actual work.

The assembler code program written for the above function is shown in **Fig. 12**. Lines 18 through 21 of the program ensure that interrupt routine INTT2 is executed when a timer-2 interrupt occurs (compare the EMON51/52 documentation). Lines 23, 24 and 25 cause the reload register to be loaded with a value of 500. Next, line 27 puts timer-2 into the right mode: use the 1-MHz internal clock, and automatic reloading. Next, the timer is actually started.

By setting the corresponding bits in interrupt enable register IEN0 (line 28), interrupts generated by timer-2 are allowed. The endless loop is entered in line 29.

The interrupt routine is short — a typical three-liner. Line 32 toggles port line P1.6. Line 33 resets the timer-2 overflow flag, while the RETI instruction in line 34 ends the interrupt processing.

The program, which may be found on your course disk as file 535XMP05.A51, is sent to the 80C535 SBC using the V24 utility, and then started as usual. If you wish to convince yourself of the operation, hook up the little circuit shown in **Fig. 13**, which will make the 1-kHz tone audible.

Next, a couple of questions, to spur you into modifying the program, and so deepen your programming skills. What is the simplest way to change the program so that it generates a 500-Hz tone? What are the lowest and highest frequencies that can be generated? A further interesting point: is the mark/space ratio of the signal at P1.6 exactly 1:1, and is this independent of the reload value and the main program?

The above questions/assignments already hint at other, often more complex, aspects of timer programming. In the following example, timer-2 is used in conjunction with the compare-capture unit, CCU, to have the 80C535 perform an accurate time measurement.

```
****** EASM52 ASSEMBLER LISTING (535XMP05) ******
LINE LOC  OBJ         T        SOURCE
   1 0000                  ; 535XMP05.A51
   2 0000                  ;                                   ; Special Function Registers:
   3 0000                  DPL   EQU 082H
   4 0000                  DPH   EQU 083H
   5 0000                  CRCL  EQU 0CAH
   6 0000                  CRCH  EQU 0CBH
   7 0000                  T2CON EQU 0C8H
   8 0000                  IEN0  EQU 0A8H
   9 0000                  IRCON EQU 0C0H
  10 0000                  P1    EQU 090H
  11 0000                  ;                                   ; MONITOR INTERFACE
  12 0000                  COMMAND EQU  030H                   ; MONITOR: command location
  13 0000                  MON     EQU  0200H                  ; MONITOR: jump address
  14 0000                  ccLINK  EQU  040H                   ; Interrupt chaining routine
  15 0000                  ;
  16 0000                          ORG   4100H                 ; Program starts at 4100H
  17 4100                  MAIN    EQU   $
  18 4100 74 06    [1]             MOV   A,#6                  ; Index for TF2 interrupt
  19 4102 90 41 1C [2]             MOV   DPTR,#INTT2
  20 4105 75 30 40 [2]             MOV   COMMAND,#ccLINK
  21 4108 12 02 00 [2]             LCALL MON                   ; Chain INT TF2
  22 410B                  ;
  23 410B 90 FE 0C [2]             MOV DPTR,#-500              ; 500 microseconds (16 bit value)
  24 410E 85 82 CA [2]             MOV  CRCL,DPL               ; Set reload value
  25 4111 85 83 CB [2]             MOV  CRCH,DPH
  26 4114
  27 4114 75 C8 11 [2]             MOV  T2CON,#00010001B       ; T2 mode: reload at 1 MHz clock
  28 4117 75 A8 A0 [2]             MOV  IEN0,#10100000B        ; Enable T2 interrupt
  29 411A 80 FE    [2] NOPE        SJMP NOPE                   ; Main program does nothing else
  30 411C                  ;
  31 411C                  ;                                   ; TIMER 2 interrupt routine
  32 411C B2 96    [1] INTT2       CPL P1.6                    ; Generate output signal
  33 411E C2 C6    [1]             CLR IRCON.6                 ; Reset T2 interrupt flag
  34 4120 32       [2]             RETI                        ; Done
  35 4121                  ;
  36 4121                          END
********* SYMBOL TABLE (14 symbols) **********
      DPL :0082       DPH :0083     CRCL :00CA     CRCH :00CB
    T2CON :00C8      IEN0 :00A8    IRCON :00C0       P1 :0090
  COMMAND :0030       MON :0200   ccLINK :0040     MAIN :4100
     NOPE :411A     INTT2 :411C
```

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Fig. 12. Timer 2 as a 16-bit reload timer. Example 535XMP05.A51 on your course disk.

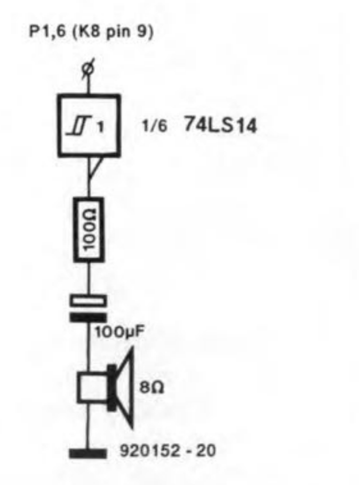

Fig. 13. External circuit to make the tone generated by the program listed in Fig. 12 audible.

Receiving time marks

The schematic structure of the CCU is

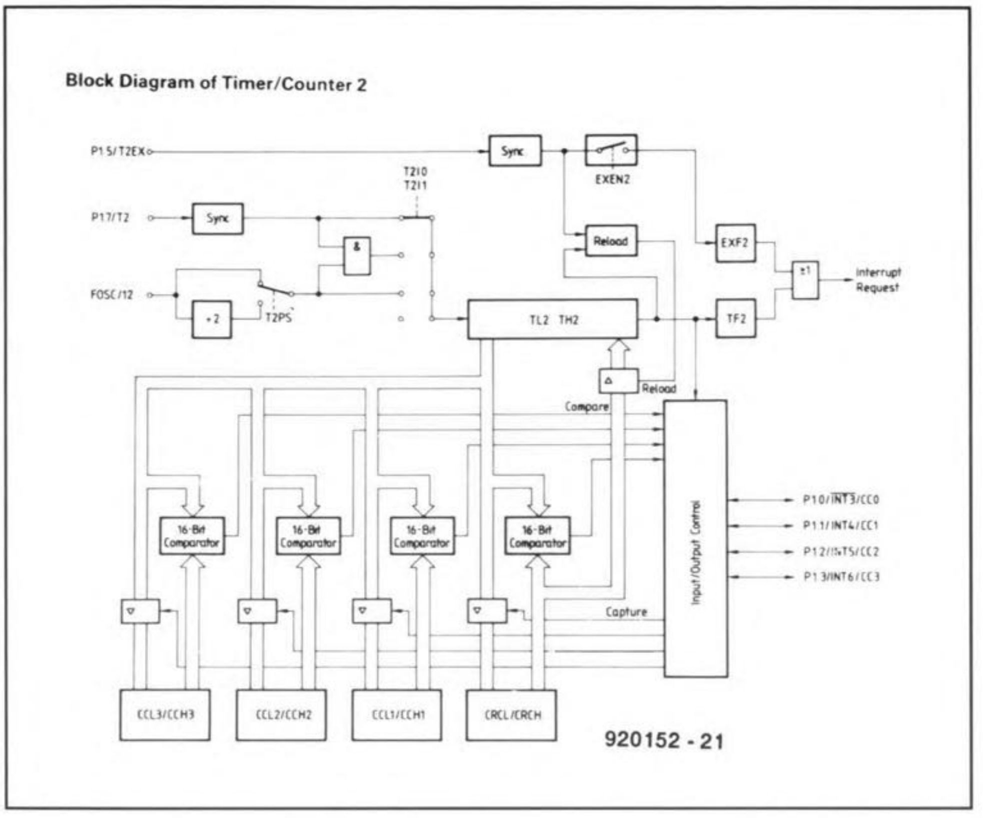

Fig. 14. Overview of CCU structure.

80C535 PROGRAMMING

You need the following to be able to follow the 80C535 hardware/assembler course:

Hardware:
- An 80C535 single-board computer as described in *Elektor Electronics* February 1994.
- An IBM PC or compatible with a serial port on COM1: or COM2:.

Software:
- Courseware disk (5.25 inch, 360K MSDOS format).
- EMON52 system EPROM, fitted on 80C535 SBC.

The courseware disk and EPROM are available as a set under order code **6221**. The disk is also available separately under order code 1811. For prices and ordering information, see page 70.

JOIN THE COURSE!

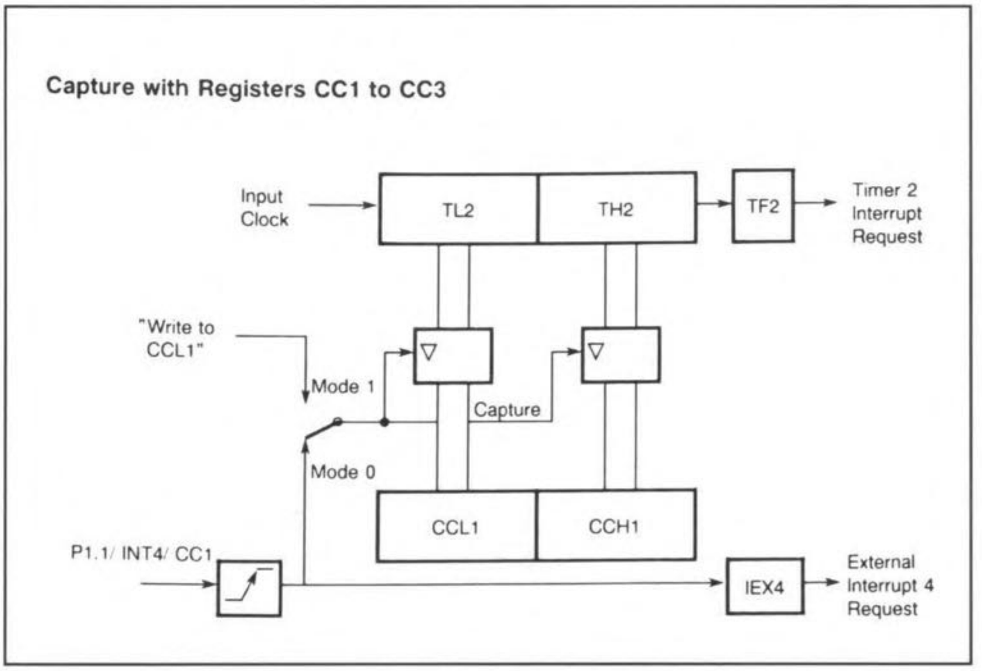

Fig. 15. Capture operation using register CC1 (or CC2, CC3).

shown in **Fig. 14**. Timer-2 is complemented by three auxiliary 16-bit registers: CCL1/CCH1, CCL2/CCH2 and CCL3/CCH3. In 'capture' mode, these registers serve to latch the state of timer-2 on the occurrence of a signal. Timer-2 however is not stopped. This allows the instant the latching signal occurs to be timed with some precision. If you do not use the reload option of timer-2, the 16-bit register CRCL/CRCH may be used as a fourth capture register. If you wish to use capture register CC1, the signal path shown in **Fig. 15** should be observed. Two capture modes are available. In **mode 0**, a rising (positive) edge at P1.1 (INT4; CC1) is used as the capture event. When a positive-going signal edge is detected, the current state ('count') of timer-2 is copied into the capture register. At the same time, flag IEX4 (bit 3 in the IRCON register) is set. Provided interrupt 4 has been enabled, an interrupt is actually generated. In **mode 1**, the capture event is controlled by a write action into register CCL1. This enables the current state of timer-2 to be 'captured' without having to stop the timer. Similar signal path options apply to capture registers CC2 and CC3; only other bits on port P1 and other interrupts are used (compare **Fig. 14**). If you wish to use the CRC register as a capture register, Fig. 15 should be exchanged with **Fig. 16**. This gives you the additional option of using the positive or the negative signal edge at port P1.1 as the capture event. This selection is made via bit 6 in the T2CON register. Since the same registers are used for the 'compare' mode of the CCU (described further on), the register functions and modes must be laid down beforehand. This is achieved by programming the CCEN register at address 0C1H, whose bit functions are shown in **Table 4**.

Capture in practice

As an example, **Fig. 17** shows a circuit which is used to delay a TTL signal. Your assignment is to write a program to determine the delay of the positive signal edge. To begin with, a long lasting low level is programmed on port line P1.7. This signal should last until C1 is fully charged. Next, a positive edge is generated at P1.7, and timer-2

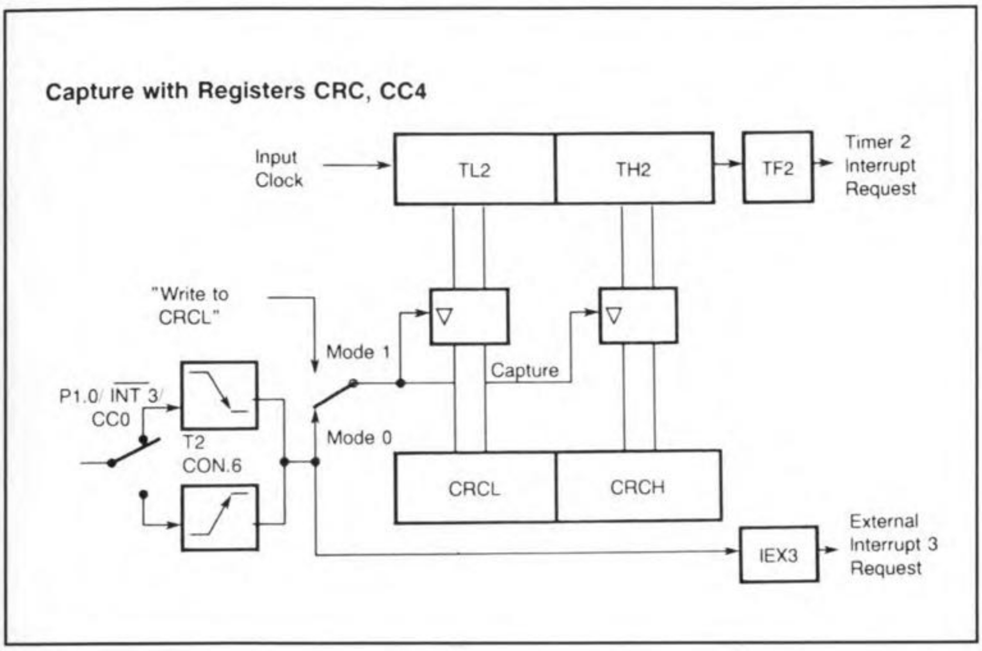

Fig. 16. Capture using reload register CRC.

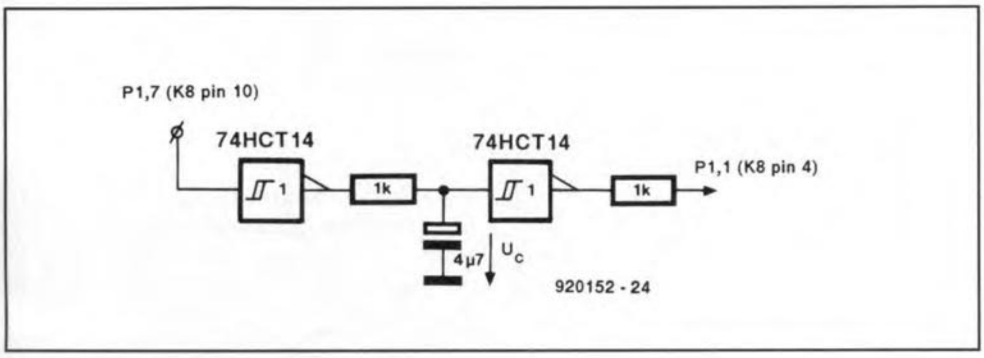

Fig. 17. This circuit delays the P1.7 signal. The delay is measured by the microcontroller.

BITS 7	6	5	4	3	2	1	0	
								CRC register
						0	0	Compare/capture disabled
						0	1	Capture on signal edge (selectable pos. or neg.) at P1.0
						1	0	Compare enabled
						1	1	Capture on CRCL Write
								CC1 register
				0	0			Compare/capture disabled
				0	1			Capture on rising edge at P1.1
				1	0			Compare enabled
				1	1			Capture on CCL1 Write
								CC2 register
		0	0					Compare/capture disabled
		0	1					Capture on rising edge at P1.2
		1	0					Compare enabled
		1	1					Capture on CCL2 Write
								CC3 register
0	0							Compare/capture disabled
0	1							Capture on rising edge at P1.3
1	0							Compare enabled
1	1							Capture on CCL3 Write

Table 4. Bits in the compare-capture-enable register, CCEN at address 0C1H.

is started at a clock of 1 MHz (without a reload value). The counter state is captured by the delayed edge at port line P1.1. The occurrence of the capture event is flagged by the set IEX4-bit. Next, the value contained in capture register CC1 must be output in decimal notation. The value equals the signal delay in microseconds, with a negligible error.

Figure 19 lists the program written to perform the complex function summarized above. To begin with, the input signal to the circuit in **Fig. 17** is made low. In line 18, timer-2 is stopped, and a delay of about 30 ms is inserted. Next, timer-2 is reset to 0 in lines 22 and 23. Line 24 causes register CC1 to function as a capture register, whereupon flag IEX4 (the capture flag of CC1) is reset. Also, the timer-2 overflow flag, TF2, is cleared (line 26). The measurement proper starts in line 27 by starting timer-2 (how is it programmed at that time?), and by having P1.7 generate a rising edge at the input of the circuit in Fig. 17. In lines 29 and 30, the processor waits for the capture event to occur (IRCON.3 becomes 1), or a timer-2 overflow (IRCON.6 becomes 1). When timer-2 produces an overflow before the capture event occurs, the overflow flag is reset, and a new measurement is attempted. This happens when, for instance, the test circuit (Fig. 17) is not yet connected to the 80C535 computer, or if C_1 is so large that the delay of the signal edge lasts longer than 65,535 µs.

Normally, however, the capture event will happen first, whereupon the processor will start to execute the instructions after the label 'capture'. First, the IEX4 (IRCON.3) flag is reset, and the capture value is stored into internal RAM, CCL1/CCH1, as a 16-bit value. Next, the value is sent to the hex-to-decimal conversion routine in EMON51/52 (lines 37, 38 and 39), from where it is sent to the terminal via V24. As a guide, the test circuit produced a value of 14,000 µs in several experiments carried out by the author. Next, a new measurement is started. Note that the input signal had already been reset to low in line 36.

Questions and assignments: how can the program of Fig. 19 be modified to measure the delay of a negative signal edge? Why would you want to use the CRCL/CRCH pair as a capture register? If you are capable of mastering this assignment (the result of which should be verified using an oscilloscope), it is likely that you have understood the basic functions and programming of the CCU in 'capture' mode.

One more assignment: how do you go about measuring times in excess of

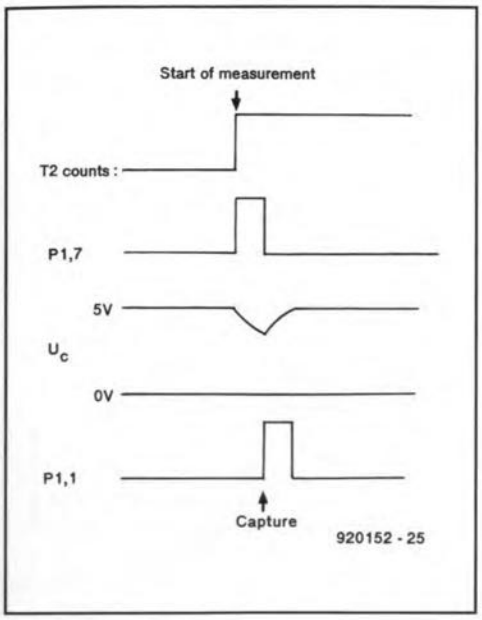

Fig. 18. Signal timing for the delay measurement performed by the program in Fig. 19.

65,535 µs? Tip: set up a software counter which is increased by one every time a timer-2 overflow occurs. This counter is stopped when the capture event occurs.

CCU in 'compare' mode

The CCU may be used in 'compare' mode to generate pulsewidth-modulated signals. Via the CCEN register, each of the register units CC1, CC2, CC3 and, in some cases, CRC also, may be individually switched to 'compare' mode. In compare mode, the contents of the relevant compare/capture register (for instance, CC1) is compared to the contents (state) of timer-2. A compare signal is generated if the two 16-bit values are equal. Depending on the mode selection, this signal can cause two different processes:

Mode 0

In mode 0, the compare signal is used to toggle one of the output bits. This may be used, for instance, to generate a pulsewidth-modulated signal. For this mode of the CCU to be selected, bit 2 (T2CM) in the T2CON register must be set to 0.

In the following discussion, it is assumed that CC1 is used. Consequently, the signal generated in compare mode appears at port line P1.1. The structure of the output driver is then as shown in **Fig. 20**. The compare signal sets the output bistable, which is reset when timer-2 produces an overflow.

Now, let us tackle pulsewidth modulation in practice. To begin with, timer 2 is programmed to operate in reload mode. The reload value (contained in the 16-bit CRC register) then determines the period of the signal. If the reload value is, for instance, 0FF00H, the period has a length of 256 µs if the internal 1-MHz clock is used. Assuming that the CC1 register contains the value 0FF01H, the output bistable is set immediately after the reload, and is not reset until the next reload operation. In this way, a signal is produced which is at logic 1 for 255 of

```
****** EASM52 ASSEMBLER LISTING (535XMP06) ******
LINE LOC  OBJ          T          SOURCE
   1 0000                  ;                                    ; 535XMP06.A51
   2 0000                  T2CON   EQU   0C8H                   ; Special Function Registers
   3 0000                  TL2     EQU   0CCH
   4 0000                  TH2     EQU   0CDH
   5 0000                  CCEN    EQU   0C1H
   6 0000                  CCL1    EQU   0C2H
   7 0000                  CCH1    EQU   0C3H
   8 0000                  IRCON   EQU   0C0H
   9 0000                  P1      EQU   090H
  10 0000                  ;
  11 0000                          ORG 050H                  ; Start address in internal RAM above EMON51
  12 0050                  nLOW    DS 2                        ; 16 bit variable
  13 0052                  ;
  14 0052                          ORG   4100H                 ; Program starts at 4100H
  15 4100                  MAIN    EQU   $
  16 4100 D2 91     [1]            SETB  P1.1                  ; To make sure, enable for use as input
  17 4102 C2 97     [1] LOOP       CLR   P1.7                  ; Signal = Low
  18 4104 75 C8 00  [2]            MOV   T2CON,#0              ; Stop TIMER/COUNTER 2
  19 4107 90 00 1E  [2]            MOV   DPTR,#30              ; Wait a while (30 ms)
  20 410A 75 30 21  [2]            MOV   COMMAND,#ccLTIME
  21 410D 12 02 00  [2]            LCALL MON
  22 4110 75 CC 00  [2]            MOV   TL2,#0                ; Reset TIMER/COUNTER to 0
  23 4113 75 CD 00  [2]            MOV   TH2,#0
  24 4116 75 C1 04  [2]            MOV   CCEN,#00000100B       ; Capture mode for CC1
  25 4119 C2 C3     [1]            CLR   IRCON.3               ; Clear capture flag IEX4
  26 411B C2 C6     [1]            CLR   IRCON.6               ; Clear T2 overflow flag
  27 411D 75 C8 01  [2]            MOV   T2CON,#00000001B      ; Start TIMER/COUNTER 2 (no Relaod)
  28 4120 D2 97     [1]            SETB  P1.7                  ; Signal = High
  29 4122 20 C3 08  [2] WAIT       JB    IRCON.3,CAPTURE       ; Wait for capture-flag
  30 4125 30 C6 FA  [2]            JNB   IRCON.6,WAIT          ; If neccessary wait for T2 overflow
  31 4128 C2 C6     [1]            CLR   IRCON.6               ; Clear overflow flag
  32 412A 02 41 02  [2]            LJMP  LOOP                  ; No capture within 65536 microsec
  33 412D C2 C3     [1] CAPTURE CLR     IRCON.3              ; Reset capture flag
  34 412F 85 C2 50  [2]            MOV    nLOW,CCL1            ; Capture register CC1 as 16 bit value
  35 4132 85 C3 51  [2]            MOV    nLOW+1,CCH1          ; Store in internal RAM
  36 4135 C2 97     [1]            CLR    P1.7                 ; Signal = LOW
  37 4137 78 50     [1]            MOV    R0,#nLOW             ; Output capture value decimally via V24
  38 4139 75 30 05  [2]            MOV    COMMAND,#ccdR016
  39 413C 12 02 00  [2]            LCALL  MON
  40 413F 90 41 4B  [2]            MOV    DPTR,#TXT2           ; Send text
  41 4142 75 30 02  [2]            MOV    COMMAND,#ccSTXT
  42 4145 12 02 00  [2]            LCALL  MON
  43 4148 02 41 02  [2]            LJMP   LOOP                 ; And do again
  44 414B 0D 0A 00      TXT2       DB     13,10,0
  45 414E                  ;                                   ; MONITOR calls etc.
  46 414E                  ccSTXT  EQU  002H                   ; Send text
  47 414E                  ccdR016 EQU  005H                   ; Send 16 bit value @R0 decimal
  48 414E                  ccLTIME EQU  021H
  49 414E                  COMMAND EQU  030H                   ; MONITOR: command location
  50 414E                  MON     EQU  0200H                  ; MONITOR: jump address
  51 414E                          END
********* SYMBOL TABLE (19 symbols) **********
    T2CON :00C8       TL2 :00CC       TH2 :00CD      CCEN :00C1
     CCL1 :00C2      CCH1 :00C3     IRCON :00C0        P1 :0090
     nLOW :0050      MAIN :4100      LOOP :4102      WAIT :4122
  CAPTURE :412D      TXT2 :414B    ccSTXT :0002   ccdR016 :0005
  ccLTIME :0021   COMMAND :0030       MON :0200
```

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Fig. 19. Listing of the propagation delay meter, a practical example of how the CCU's capture mode is used. This is example program 535XMP06.A51 on your course disk.

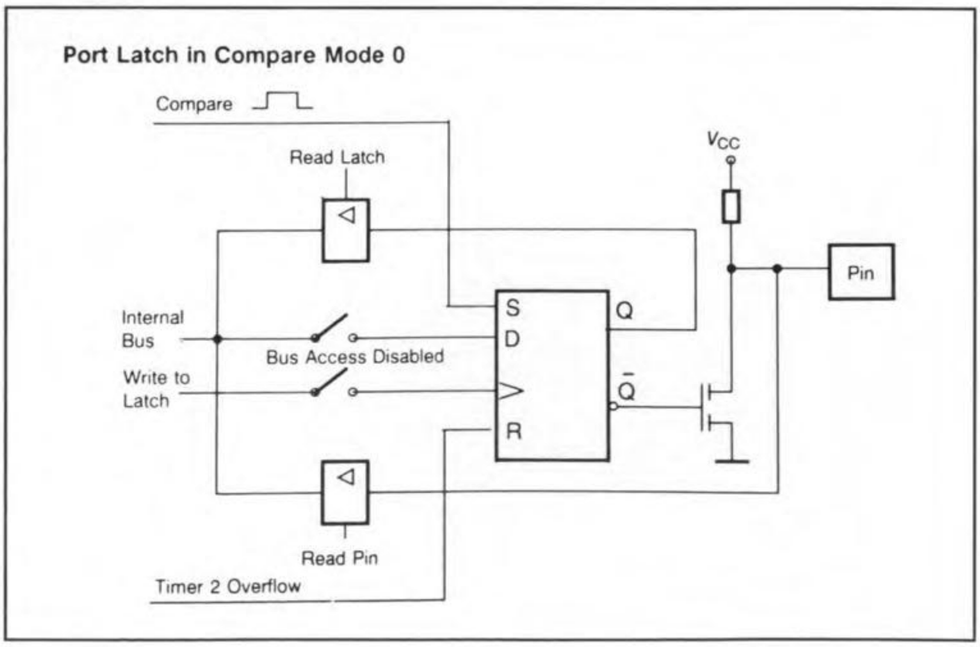

Fig. 20. Output port bit circuitry in CCU compare mode 0 (PWM mode).

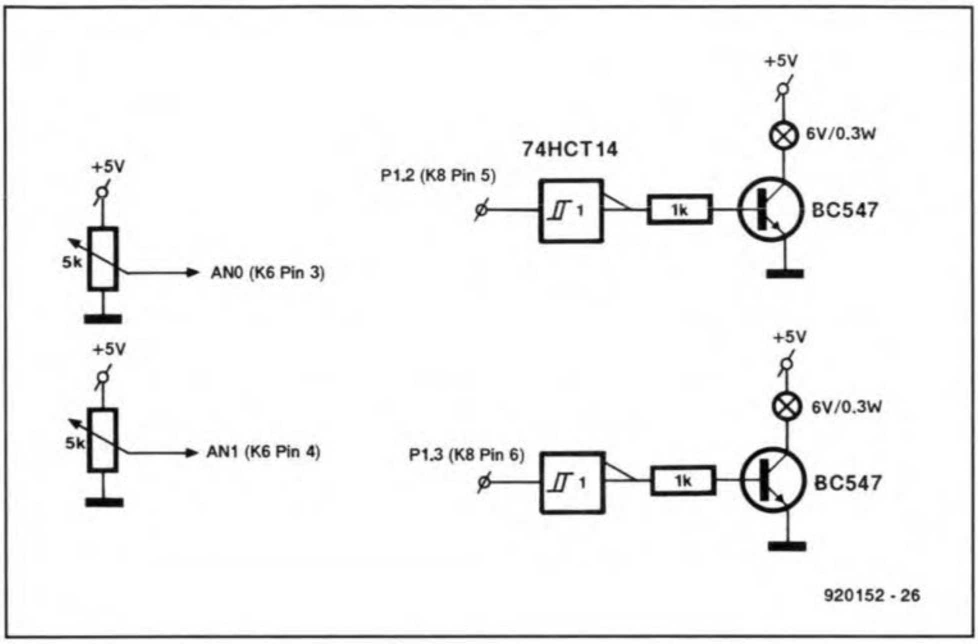

Fig. 21. This circuit and the listing in Fig. 22 implements a lamp brightness control using pulsewidth modulation.

256 µs. If the CC1 register contains, for instance, the value 0FF80H, the resulting signal has a mark/space ratio of 1:1, i.e., 128:128 µs. Thus, the pulsewidth is a function of the value written into the CC1 register.

Moving to more practical aspects of our example, two potentiometers, P_1 and P_2, are to be used to control the intensity of two lamps. As shown in **Fig. 21**, the wiper voltage of the pots has a range of 0 to 5 V. These voltages are digitized with the aid of the ADC contained in the 80C535 microcontroller (channels 0 and 1). The converted values (between 0 and 255) are to be used to generate two PWM signals at pins P1.2 and P1.3. For this purpose the CC2 and CC3 compare registers are used, with the respective PWM outputs on pins P1.2 and P1.3. The complete program is listed in **Fig. 22**.

The period of the PWM signals is fixed at 256 µs in lines 23 and 24. The high-order bytes of the CC2 and CC3 compare registers are set to 0FFH (255 decimal). The low-order byte is reserved for the result of the A-to-D conversion. In line 30, CC2 and CC3 are programmed to operate in compare mode. Next, timer-2 is programmed in line 31. At the same time, the entire CCU is switched to compare mode 0. From then on, PWM signals are automatically generated.

The loop starting at label LOP only serves to make sure that the compare registers are always loaded with the values after the A-to-D conversion. First, the voltage at channel 0 is converted (lines 32 and 33, and subroutine ADC1). The result is written into the CCL2 register. Next are ADC channel 1 and CCL3, and then the program starts again by looping back to LOP.

As a suggestion for an assignment, try your hand at something really complex: program a mains-synchronized three-phase control dimmer for 12-V halogen lamps using CC1, CC2 and CC3.

Mode 1: generating exact signals

So far we have restricted ourselves to a discussion of the CCU's compare mode 0. The other mode, mode 1, is selected by setting bit 2 (T2CM) in the T2CON register. This selects the output port configuration shown in **Fig. 23**. It is assumed that CC1 is to be used. In this mode, writing to port line P1.1 does not copy the bit value directly to the output latch, but into a shadow register, where it is latched. The bit is not copied to the output register until the next compare signal is supplied by CC1. This mode serves to enable jitter-free signals to be generated under software control. The software must always write the new bit value into the shadow register in time, that is, before the next compare signal appears.

Since there is only one bit, T2CM, to make the selection between compare mode 0 and mode 1, all compare registers have to be used in the same mode. Individual CC registers may, however, be used in capture mode, irrespective of the compare mode selected. This is

```
****** EASM52 ASSEMBLER LISTING (535XMP07) ******
LINE LOC  OBJ          T         SOURCE
   1 0000                ;  535XMP07
   2 0000                ;                                   ; Special Function Registers
   3 0000                ACC    EQU 0E0H
   4 0000                DPL    EQU 082H
   5 0000                DPH    EQU 083H
   6 0000                ADCON  EQU 0D8H
   7 0000                ADDAT  EQU 0D9H
   8 0000                DAPR   EQU 0DAH
   9 0000                ;
  10 0000                CCEN   EQU 0C1H
  11 0000                CCL2   EQU 0C4H
  12 0000                CCH2   EQU 0C5H
  13 0000                CCL3   EQU 0C6H
  14 0000                CCH3   EQU 0C7H
  15 0000                CRCL   EQU 0CAH
  16 0000                CRCH   EQU 0CBH
  17 0000                ;
  18 0000                T2CON  EQU 0C8H
  19 0000                TL2    EQU 0CCH
  20 0000                TH2    EQU 0CDH
  21 0000                ;
  22 0000                        ORG   4100H               ; Program starts at 4100H
  23 4100 75 CA 00 [2] MAIN      MOV   CRCL,#0             ; Reload value is 0FF00H
  24 4103 75 CB FF [2]           MOV   CRCH,#255
  25 4106 75 C4 00 [2]           MOV   CCL2,#0             ; Compare value=0FF00H on CC2 and CC3
  26 4109 75 C5 FF [2]           MOV   CCH2,#255
  27 410C 75 C6 00 [2]           MOV   CCL3,#0
  28 410F 75 C7 FF [2]           MOV   CCH3,#255
  29 4112
  30 4112 75 C1 A0 [2]           MOV   CCEN,#10100000B     ; CC2 and CC3 to COMPARE
  31 4115 75 C8 11 [2]           MOV   T2CON,#00010001B    ; Reload at 1 MHz clock
  32 4118 74 00    [1] LOP       MOV    A,#0               ; Channel 0
  33 411A 12 41 28 [2]           LCALL ADC1                ; convert A-to-D
  34 411D F5 C4    [1]           MOV    CCL2,A             ; and use as LSB of COMPARE value
  35 411F 74 01    [1]           MOV    A,#1               ; Channel 1
  36 4121 12 41 28 [2]           LCALL ADC1
  37 4124 F5 C6    [1]           MOV    CCL3,A             ; provides LSB of CC3
  38 4126 80 F0    [2]           SJMP  LOP
  39 4128                ;
  40 4128 44 10    [1] ADC1      ORL    A,#00010000B       ; Set start ADC bit
  41 412A F5 D8    [1]           MOV    ADCON,A            ; Set A-D control
  42 412C 75 DA 00 [2]           MOV    DAPR,#0            ; Start A-D conversion 0-5 Volt
  43 412F 20 DC FD [2] BSY       JB     ADCON.4,BSY        ; Wait until ready
  44 4132 E5 D9    [1]           MOV    A,ADDAT            ; Fetch result
  45 4134 22       [2]           RET
  46 4135                        END
********* SYMBOL TABLE (20 symbols) **********
     ACC :00E0       DPL :0082       DPH :0083     ADCON :00D8
   ADDAT :00D9      DAPR :00DA      CCEN :00C1      CCL2 :00C4
    CCH2 :00C5      CCL3 :00C6      CCH3 :00C7      CRCL :00CA
    CRCH :00CB     T2CON :00C8       TL2 :00CC       TH2 :00CD
    MAIN :4100       LOP :4118      ADC1 :4128       BSY :412F
```

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Fig. 22. Two-channel pulsewidth modulation is demonstrated by this example program, 535XMP07.A51, on your course disk.

achieved by setting the relevant bits in the CCEN register.

The description of the CCU in the 80C535 datasheets covers quite a few pages. Only the most elementary aspects are covered in this last instalment of the course. The exact timing of the individual steps, for instance, is not covered here since it is a too extensive subject.

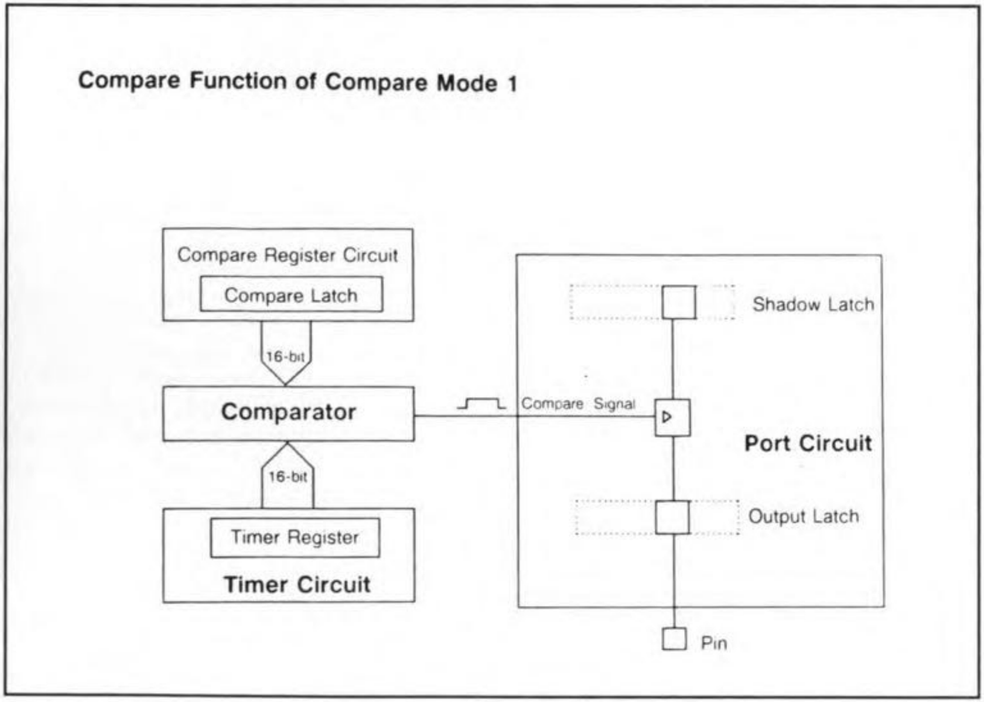


Fig. 23. Output port bit circuitry in CCU compare mode 1.

Outlook

The discussion of the main functions and programming options of the CCU brings us to the end of the present short course. The questions, suggestions and assignments may be used as exercises in practical programming of the 80C535. In all cases, the programs found on your course disk may be used as a starting point for your own experiments. (920152-4)

For further reading:
Siemens, SAB80C517/80C537 Single-Chip Microcontroller Data Book.
Siemens, 80C515/80C535 datasheets, in Microcomputer Components, Microcontrollers, Data Catalog 1990.

Note:
in response to readers' enquiries we would suggest the following suppliers of the SAB80C535 microcontroller:
» Electrovalue, Unit 3, Central Trading Estate, Staines TW18 4UX, England. Tel. (0784) 442253, fax (0784) 460320.
» C-I Electronics, P.O. Box 22089, 6360-AB Nuth, Holland. Fax: (+31) 45 241877 (IC supplied incl. of free datasheet).
C-I Electronics also supply kits of the 80C535 computer. *tech. ed.*

INTELLIGENT EPROM ERASER

Time is money, and the shorter one has to wait for an EPROM to be erased, the better. The circuit described here will save you time because it keeps a constant eye on the contents of the EPROM, leaving the UV lamp on no longer than necessary for the device to be completely erased.

Design by A. Rietjens

EPROMs are ideal devices for software developers because they can be programmed fairly quickly, and retain their contents for ten years or longer. On the down side, erasing an EPROM, because it contains an error made during the software development phase, can be time consuming. On average, an EPROM will have to be exposed to ultra-violet light for a period of between five and fifteen minutes before you can be sure that it is blank again. In practice, this time has to be established by trial and error, and large differences may exist between devices. In most cases, the actual time required to erase the EPROM is never known. In other words, the UV exposure period is always too long, and precious time is wasted.

The present circuit tackles EPROM erasing in a novel way. The EPROM to be erased is inserted into the socket of the eraser. Next, the erasure process is started by the circuit lighting an UV lamp or small tube. The contents of the EPROM are continuously read, until all cells read FF_H. Just to make sure you understand the principle: an EPROM is fully erased when all its memory cells (which are basically bistables) are at logic '1'. All memory locations then read back 'FF_H'. A cell is programmed by making it logic '0'.

Once all cells read 'FF_H', the lamp is left on for about one minute. This is done to make sure that all cells are properly cleared, including those who are at an undefined state between erased and still programmed. In this state, a cell can 'relapse' into the programmed state shortly after the lamp has been switched off. The extra erasure time guarantees that can not happen.

The circuit

The circuit shown in **Fig. 1** is reasonably simple. The central part is formed by position IC5, the EPROM to be erased. The EPROM is powered, and its $\overline{CS}$ and $\overline{OE}$ inputs are permanently actuated. This causes the EPROM outputs to be enabled during the entire erasure period. For a complete check to be made on the EPROM contents, the entire address range has to be scanned. This is done with the aid of a 16-bit address counter, IC_1-IC_2. The first, IC_1, is a combined counter/oscillator. Because the counter's output range is not continuous (spanning Q3 through Q13, but with Q10 missing), a second counter is cascaded. Outputs Q3-Q9 of IC_1 drive address lines A0-A6. Output Q9 is also used to clock the second counter, IC_2, which drives the remaining address lines, A7-A15, as well as the lamp interface, via output Q11.

The circuit diagram clearly indicates that there is no direct connection between address lines A14-A15 and the EPROM to be erased. This enables EPROMs with different memory capacities to be erased. The entire 16-bit address range is used by the 27x512 only. Depending on the selection made with switch S_1, 27x256 and 27x128 devices may be handled also.

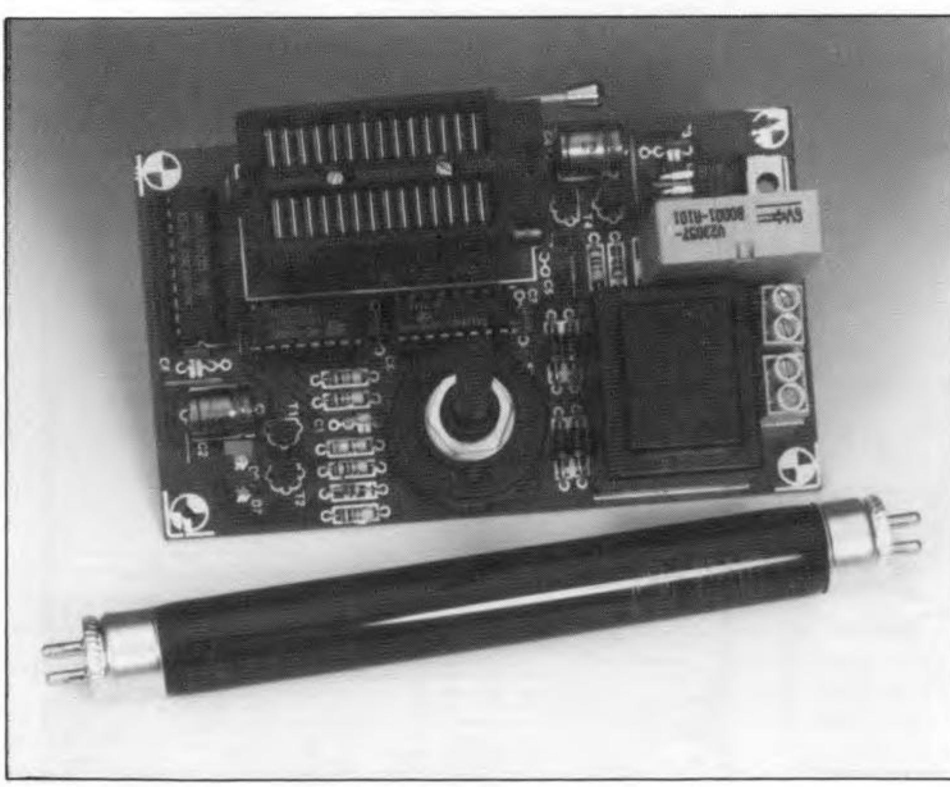
Circuit IC_3 checks the contents of the memory location selected by the address counter. Its $P=\overline{Q}$ output, pin 19, goes low only if the memory location reads back 'FF_H', whereupon the 16-bit counter is allowed to start.

A further safety measure involves the use of a HCMOS type for IC_3. This fixes the logic levels at around half the supply voltage (in this case, 2.5 V). Since EPROMs are traditionally used in a standard TTL logic environment, where '0'<0.7 V, and '1'>2 V, an extra voltage margin is obtained for the level comparison.

The circuit is powered by a conventional supply. The transformer secondary voltage is first rectified by diodes D_1-D_4, then smoothed by reservoir capacitor C_4. Next, the familiar 7805 is used to step the unregulated voltage down to a regulated +5-V rail.

Circuit operation

As soon as the supply is switched on, pin 19 of IC_3 will supply a logic high level — but only if an EPROM is in-



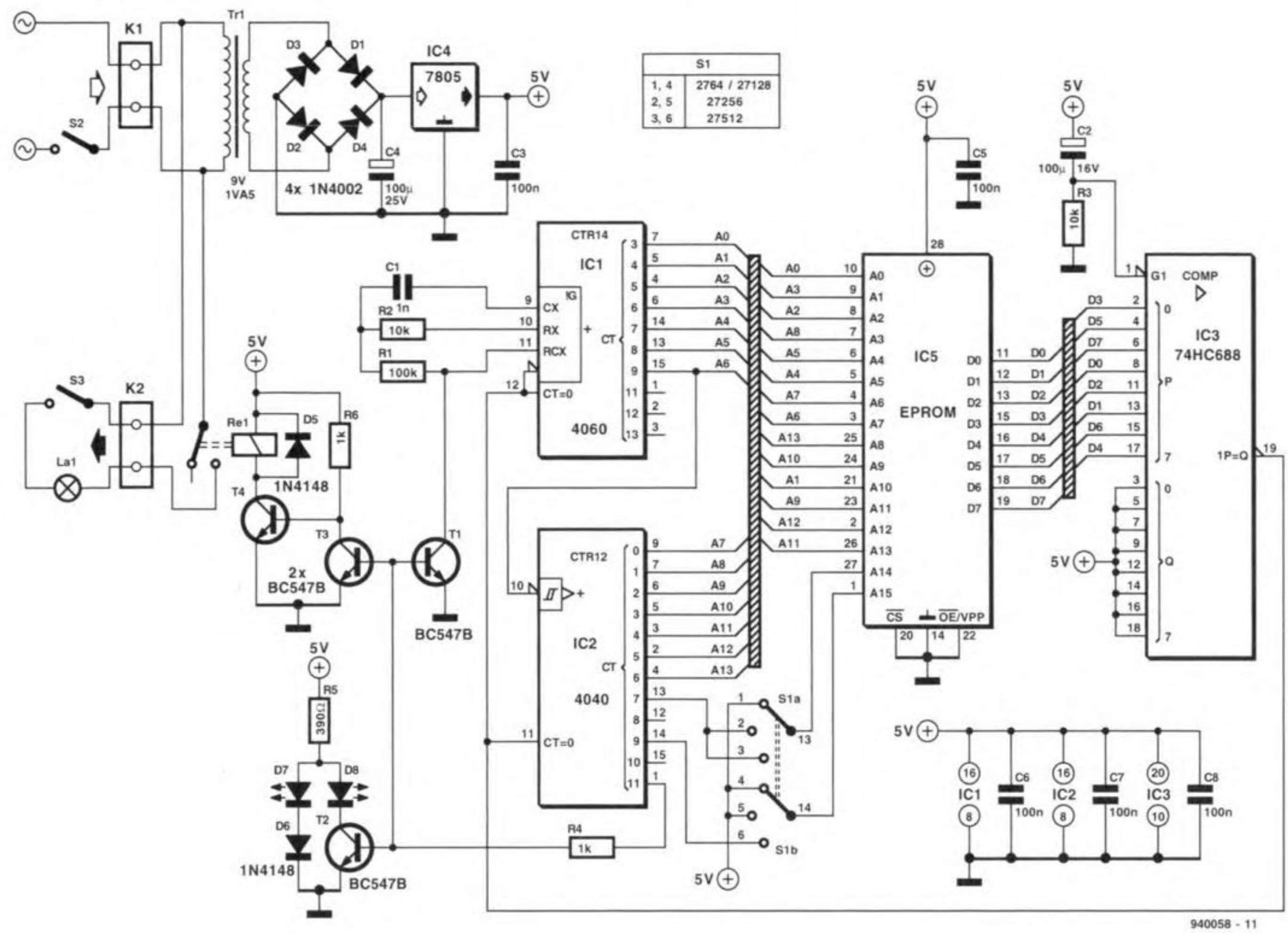


Fig. 1. Circuit diagram of the intelligent EPROM eraser. Build it, and keep EPROM erasing time to an absolute minimum.

serted which is not blank. As a result, IC_1 and IC_2 are reset, and their outputs go low. This causes transistors T_1, T_2 and T_3 to be switched off, so that LED D_7 lights, and the relay is energized via T_4. The UV lamp is switched on, and the erasure process is started.

Once the first memory location has been cleared, the address counter is enabled by IC_3, and counting (i.e., address scanning) may commence. If an address is found which reads a value other than 'FF_H', the counter is reset, and the counting operation starts again. When all cells are erased, the counter may continue to work until Q11 of IC_2 goes high. That happens about one minute after the counter has been started, depending, of course, on the clock frequency used. This provides the previously mentioned extra erasure time. When Q11 goes high, transistor T_1 starts to conduct, causing the oscillator in IC_1 to be disabled. Transistor T_3 then starts to conduct also, which causes the relay to be switched off. Consequently, the UV lamp goes out, and the erasure process is finished. LED D_8 lights to indicate that the circuit may be switched off, and the EPROM removed from the ZIF socket.

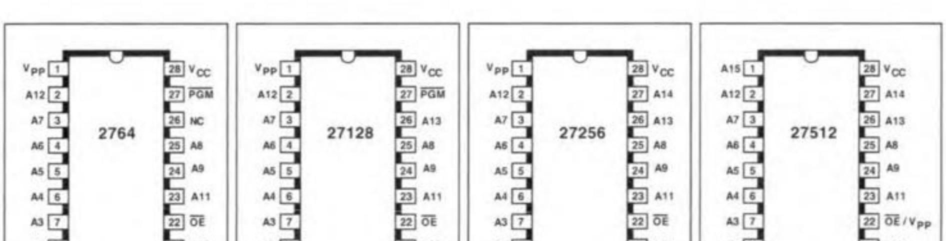


Fig. 2. Pinouts of the EPROMs that can be erased with the circuit.

Construction

The construction of the intelligent EPROM eraser is made relatively easy by a printed circuit board which may be obtained through our Readers Services. The artwork is shown in **Fig. 3**. All parts, except the UV lamp, are accommodated on the board. To make exchanging EPROMs on the

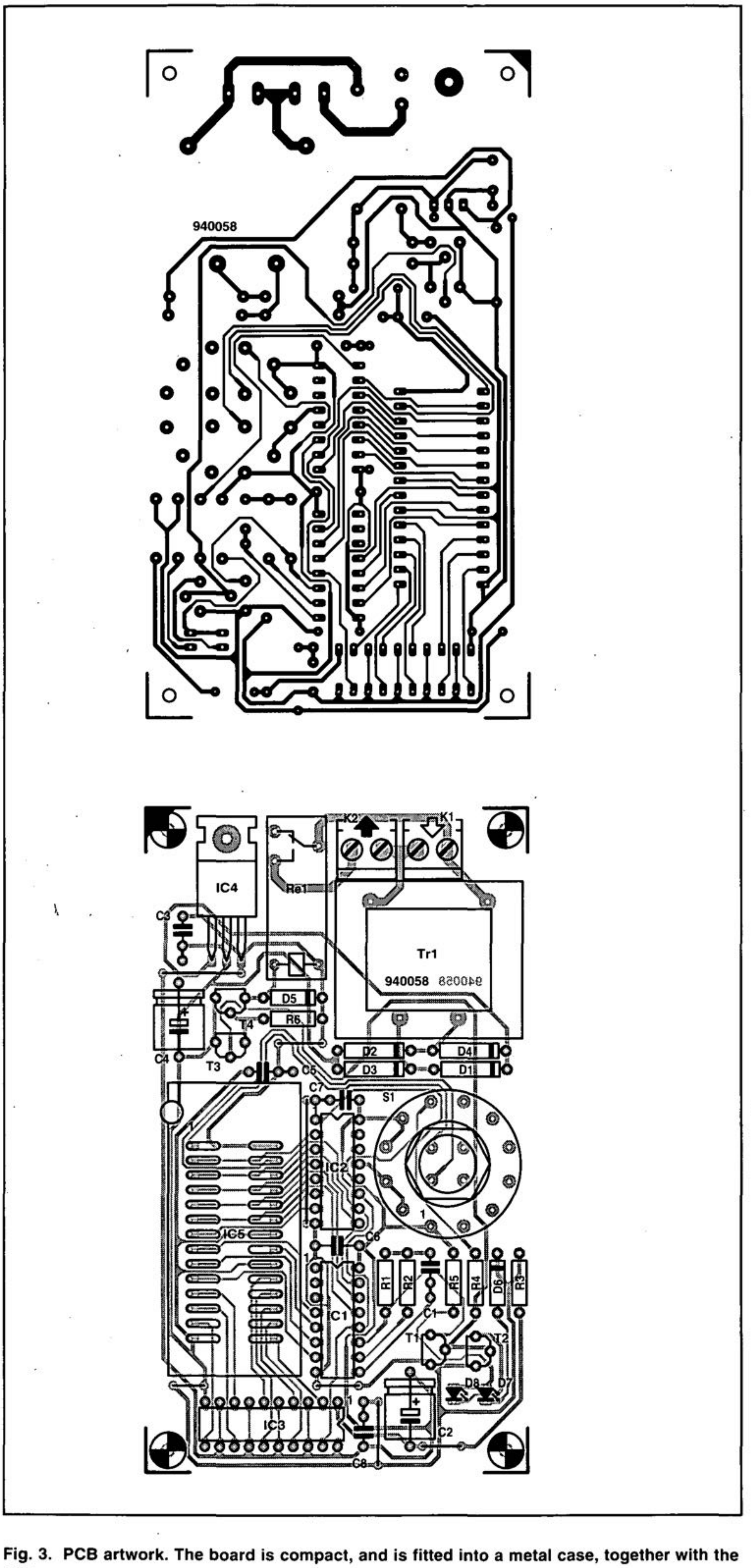


Fig. 3. PCB artwork. The board is compact, and is fitted into a metal case, together with the UV lamp.

eraser as easy as possible, it is recommended to use a 28-way ZIF (zero insertion force) socket.

The completed printed circuit board is fitted into a closed box, together with the UV lamp. The box must be fully closed because the light emitted by the UV lamp is harmful to the eyes and the skin. In all cases, keep exposure to this UV light to an absolute minimum. As an extra safety measure, a microswitch is connected in series with the UV lamp. This switch is mounted such that the lamp goes out if the lid of the box is opened.

Still on the subject of safety, also observe due precautions with regard to the mains voltage which is present on the board.

In view of the heat developed by the UV lamp, it is recommended to house the eraser in a metal box, which is connected to the mains earth.

(940058)

COMPONENTS LIST

Resistors:
R1 = 100kΩ
R2,R3 = 10kΩ
R4,R6 = 1kΩ
R5 = 390Ω

Capacitors:
C1 = 1nF
C2 = 100µF 16V
C3,C5-C8 = 100nF
C4 = 100µF 25V

Semiconductors:
D1-D4 = 1N4002
D5,D6 = 1N4148
D7 = LED, red
D8 = LED, green
T1-T4 = BC547
IC1 = 4060
IC2 = 4040
IC3 = 74HC688
IC4 = 7805
IC5 = EPROM to be erased

Miscellaneous:
K1,K2 = 2-way PCB terminal block, raster 5mm.
S1 = 4-pole, 3-way rotary switch.
S2 = mains switch w. make contact.
S3 = microswitch.
Re1 = E-card-relay (e.g., Siemens V23057 B1-A101).
Tr1 = 9V/1.5VA mains transformer, e.g., Monacor VTR1109.
La1 = UV lamp, e.g., Philips Components UV6.
One 28-way ZIF socket.
Printed circuit board 940058 (see page 70).

FIGURING IT OUT

PART 17 – ALL CHANGE!

By Owen Bishop

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

The rate of change of one quantity with respect to changes in another quantity is known as a **differential**. As explained in Part 5, we represent a differential by a symbol having the form dy/dx. The example given then was $d\Phi/dt$, the rate of change of magnetic flux Φ through an inductor, with respect to time t. If we were to plot a graph of the flux at different times, then $d\Phi/dt$ is the gradient of the graph at any given time.

We are assuming that the dependent variable Φ is a function of the single independent variable t. In symbols:

$$\Phi = f(t).$$

Practically all of the circuits and circuit models we have studied exhibit this one-to-one relationship. In many of them, the voltage at a given point is a function of time. Given an equation that defines how voltage varies with time (for example, $u = 3\sin t$), we differentiate to find the rate of change of voltage at any given instant (in this example, $du/dt = 3\cos t$). But there are other cases in which a quantity depends on **two** (or more) other independent quantities. As an example, consider the familiar Ohm's law equation for the current through a resistor:

$$i = u/r. \qquad \text{[Eq. 126]}$$

As on previous occasions, we are using lower-case letters for quantities that are varying. Ordinarily we would use a capital R for resistance, assuming that its value does not change significantly. But suppose that the resistor is light-dependent or we are using a thermistor. We must allow for resistance to vary just as much as, if not more than, voltage. Current is a function of both voltage **and resistance**:

$$i = f(u, r).$$

We might want to know the rate at which current varies with respect to variations in both voltage and resistance. Think of this in terms of a three-dimensional graph—**Fig. 143**. The curved surface represents the way in which i varies with u and r. We can consider these variations individually. If we cut sections of the surface parallel to the i–u plane, we obtain a set of straight-line graphs. The line AB represents changes in current with respect to voltage, if resistance is constant at 8 Ω. In the other direction, parallel to the i–r plane, sectioning the surface yields a set of hyperbolas. An example is shown by the line CD, which represents changes in current with respect to resistance, if voltage is constant at 4 V.

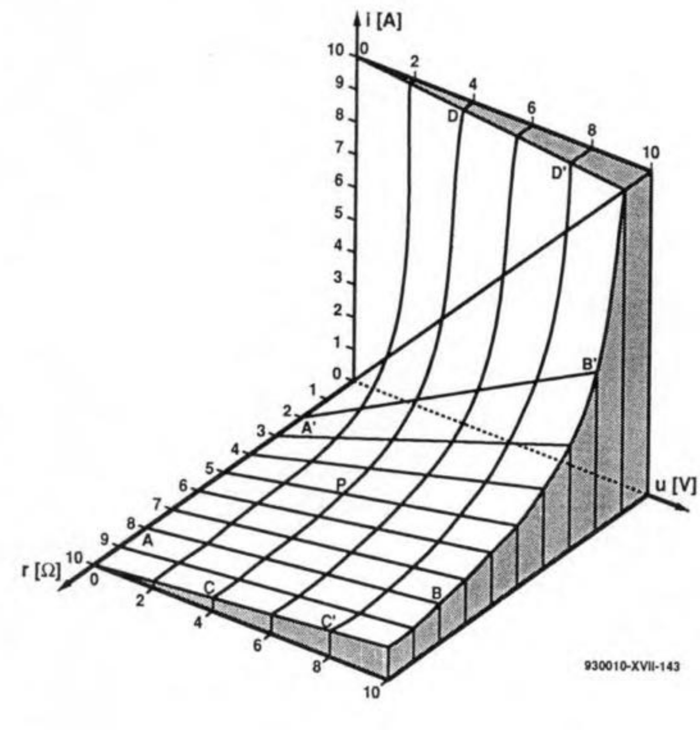


Fig. 143

We calculate the slopes of these curves by using **partial differentials**. The technique corresponds to cutting a slice through the surface, parallel to one or other of the vertical planes. Cutting a slice parallel to the i–u plane means considering r to be constant and then differentiating Eq. 126 with respect to u:

$$\delta i / \delta u = 1/r.$$

The symbol for the differential is written with a special kind of 'd' to indicate that this is a partial differential. We call the δ a 'curly d'. Because we have decided to make r constant, the partial differential (the gradient of a line drawn parallel to the i–u plane) is constant. In other words, we have a straight line, as exemplified by line AB. For the set of lines obtained in this way, the gradient is inversely proportional to r. The smaller the value of r, the steeper the line, as can be seen by comparing line AB with line $A'B'$ in the figure.

Similarly, we find the gradient of lines parallel to the i–r plane. Differentiating Eq. 126 while considering u to be constant:

$$\delta i / \delta r = -u / r^2.$$

The gradient is negatively inversely proportional to the square of the resistance, which is a characteristic of hyperbolas. For a given resistance, it is also directly proportional to the voltage; at double the voltage, for example, gradients at all points along a curve are doubled. This can be seen by comparing curve CD with curve $C'D'$.

Example

Let us put some numbers into these equations and see how they work out. Take a point P on the surface, representing $u = 4$, $i = 0.8$. The third value necessarily follows from the first two. At this point, the gradient in the direction parallel to the i–u plane is

$$1/r = 1/5 = 0.2.$$

The units of this gradient are amperes per volt (A/V). The gradient parallel to the i–u plane is

$$-u/r^2 = -4/25 = -0.16.$$

Units are amperes per ohm (A/Ω). Now suppose that the voltage is increased by 0.1 V and the resistance is decreased by 0.5 Ω. The change in current resulting from a change in voltage is 0.2 A V^{-1}. A change of 0.1 V produces a current increase of $0.2 \times 0.1 = 0.02$ A. Similarly, the change in current resulting from a change in resistance is $-0.16\,A\,\Omega^{-1}$. A change of −0.5 Ω produces a current change of $-0.16 \times -0.5 = 0.08$ A. Both changes are current **increases** and their effects are cumulative. The total change in current is 0.02+0.08 = 0.1 A. The new current is 2+0.1 = 2.1 A.

This result is only approximate, because we have considered the effects of voltage and resistance separately. **Figure 144** is an enlarged view of the surface of **Fig. 143**, in the region of point *P*. Increasing voltage by 0.1 V is equivalent to moving from *P* to *Q*. Decreasing resistance by 0.5 Ω is equivalent to moving from *P* to *R*. Both produce an increase of current (an uphill climb on the sloping surface of the graph). We have simply added these increases. But really, we want to know the change that occurs in going from *P* to *S*. Going from *P* to *Q* and then adding on the current increase that was calculated **on the basis of being at *P*** is not quite the same thing as going directly from *P* to *S*. The change in level between *Q* and *S* is not quite the same as that between *P* and *R*. Having said this, we still find this technique a useful one, provided that we keep to small changes in *u* and *r*. If *PQ* is small, the increase in going from *Q* to *S* is not very different to the increase in going from *P* to *R*, so simply adding the increases is precise enough.

Small increments

Let us summarise this in mathematical terms. The small change in voltage is Δu. The rate of change of current with respect to voltage at a given point is $\delta i / \delta u$. The actual change in current produced by the small change in voltage is

$$(\delta i / du) \times \Delta u.$$

Similarly, the change in current resulting from a small change in resistance is:

$$(\delta i / dr) \times \Delta r.$$

Provided that both Δu and Δr are small, these two may be added (taking account of sign) to find the overall change in current.

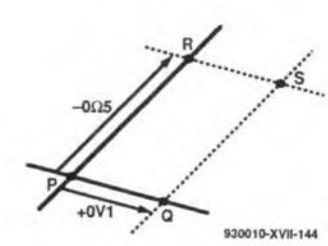


Fig. 144

Another example

We will work this example straight through, without comment, as a practical summary of the method. A 33 Ω resistor has a current of 2.5 mA passing through it. The power dissipated in the resistor is $p = i^2 r$. We are asked to find the change in power when the resistor increases by 1 Ω and the current decreases by 0.1 mA. To begin with, $p = 0.0025^2 \times 33 = 206.25$ µW. The change in power resulting from the change in resistance is:

$$(\delta p / \delta r) \times \Delta r = i^2 \times 1 = 6.25 \times 10^{-6}.$$

The change in power resulting from the change in current is:

$$(\delta p / \delta i) \times \Delta i = 2ir \times -0.001$$
$$= 0.165 \times -0.0001 = -16.5 \times 10^{-6}.$$

The total change in current is $(6.25 - 16.5) \times 10^{-6} = -10.25 \times 10^{-6}$. The new power level is $(206.25 - 10.25) \times 10^{-6} = 196$ µW.

We can easily check this result by working out the power when $i = 2.4$ mA and $r = 34$ Ω. For this pair of values, $p = 0.0024^2 \times 34 = 195.84$ µW. This, of course, is a precise value and it might be wondered why we need to go into the complications of differentiation to obtain only an approximate value. The reason is that the method of partial differentiations is applicable to equations much more complex than the simple one used here to illustrate the method. It can be applied to any combination of resistances and currents, it being necessary only to insert the appropriate values of Δi and Δr into the equations once the initial differentiation has been done.

Percentages

The technique can also be applied when the changes are expressed as percentages. Using the resistor of the previous example, we might specify that the resistance decreases by 0.5% and the current decreases by 2%. Calculating the changes:

$$\Delta r = -0.005r$$

and

$$\Delta i = -0.02i.$$

Total change in power is:

$$i^2 \Delta r + 2ir\Delta i = -0.005 i^2 r - 0.04 i^2 r$$
$$= i^2 r \times -0.045 = p \times -0.045.$$

The percentage change in power is −4.5%.

Three or more

The technique may be extended to include three or even more variables. Partial differentiation is no longer just a matter of cutting slices through a three-dimensional surface, for the 'graph' is a surface in space of four or more dimensions. But the principle is just the same. If *z* is a function of *a, b, c, …,* then

$$\Delta z = \frac{\delta z}{\delta a} \times \Delta a + \frac{\delta z}{\delta b} \times \Delta b + \frac{\delta z}{\delta c} \times \Delta c + \ldots \quad \text{[Eq. 127]}$$

As am example, consider the non-inverting amplifier circuit of **Fig. 145**. The output voltage, *z*, is related to the input voltage, *u*, by the equation:

$$z = u(a/b+1) \quad \text{[Eq. 128]}$$

The values of the resistors are represented by variables *a* and *b*. Here we have three independent variables, and can write three partial differentials by differentiating Eq. 128 appropriately:

$$\delta z / \delta u = a / b + 1;$$

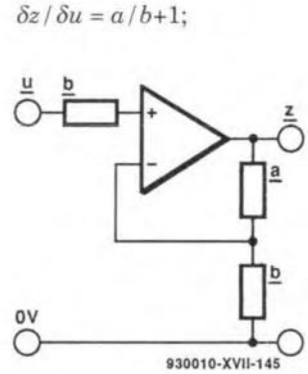


Fig. 145

$$\delta z / \delta a = u / b;$$

$$\delta z / \delta b = -ua / b^2.$$

We will use this example to illustrate the method of estimating errors and work in percentages. The aim is to calculate the maximum positive error in *z* that my be brought about by resistor tolerances and the precision of the input voltage. Factors that will result in an increase in *z* are:

- Increase in *u*; suppose this is precise to ±2%; $\Delta u = 0.02u$.
- Maximal value of *a*; suppose tolerance is 1%; $\Delta a = 0.01a$.
- Minimal value of *b*; suppose tolerance is 1%; $\Delta b = -0.01b$.

Using an equation of the same form as Eq. 127:

$$\Delta z = (a/b+1) \times 0.02u$$
$$+ (u/b) \times 0.01a$$
$$+ (-u/a/b^2) \times -0.01b,$$

so that

$$\Delta z = 0.02u\,(2a/b + 1) \quad \text{[Eq. 129]}$$

Equation 129 is used to calculate the change in *z* for any combination of values of *u, a,* and *b*. For example, if $u = 5$ V, $a = 100$ Ω and $b = 200$ Ω, then

$$\Delta z = 0.02 \times 5(200/200+1) = 0.2.$$

The output voltage may be up to 0.2 V greater than its calculated value. With the exact values given, $z = 7.5$ V, but errors may take it up to a maximum of 7.7 V.

Related rates

In the examples above, the two (or more) independent variables have been independent of each other. But it may happen that they are related to each other because they are both functions of some other quantity. Very often we find that two variables are functions of time. The input to a circuit may be a ramping voltage or a sine wave; in either case, the voltage is a function of time. In the same circuit, there may be photoresistor responding to a periodically changing light level; the resistance of this is also a function of time. If a current in the circuit is affected both by the input voltage and the resistance of the photoresistor, we have a circuit which can be analysed by partial differentials.

For example, suppose that

the input is given by $u = 2\sin 4t$. The resistance of the photoresistor is given by $r = 100+40t^2$ (a light source is approaching the sensor and the inverse square law applies). The current is $i = u/r$. We want to know the rate of change of i.

The standard equation (Eq. 127) tells us that:

$$\Delta i = \frac{\delta i}{\delta u} \times \Delta u + \frac{\delta i}{\delta r} \times \Delta r$$

This equation does not require the change to occur in any specified length of time. Now we will introduce the time element by saying that the changes must take place in a short length of time, Δt. Dividing both sides by Δt:

$$\frac{\Delta i}{\Delta t} = \frac{\delta i}{\delta u} \times \frac{\Delta u}{\Delta t} + \frac{\delta i}{\delta r} \times \frac{\Delta r}{\Delta t}$$

If Δt is made very small, approaching zero, the expressions containing it become differentials:

$$\frac{\mathrm{d}i}{\mathrm{d}t} = \frac{\delta i}{\delta u} \times \frac{\mathrm{d}u}{\mathrm{d}t} + \frac{\delta i}{\delta r} \times \frac{\mathrm{d}r}{\mathrm{d}t} \qquad \text{[Eq. 130]}$$

The expression on the left is the rate of change of current with time. We call it the **total differential** of z with respect to t. Note that Eq. 130 does not contain any finite quantities such as Δt, and so it gives an exact result, not an approximation. Let us apply this equation to our example. Calculating the partial differentials from the equation given (the same as in the first example this month):

$\delta i / \delta u = 1/r$;

$\delta i/\delta r = -u/r^2$.

Calculating the other differentials in the usual way from the equations given:

$\mathrm{d}u/\mathrm{d}t = 8\cos 4t$;

$\mathrm{d}r/\mathrm{d}t = 80t$.

Substituting in Eq. 130:

$$\mathrm{d}i/\mathrm{d}t = (8\cos 4t)/r - 80ut/r^2. \qquad \text{[Eq. 131]}$$

We can use Eq. 131 to find the rate of change of current when one of the quantities has a specified value. For example, when $t = 0.1$ s, $u = 0.7788$ V and $r = 100.4\ \Omega$. Then, substituting in Eq. 131 gives:

$\mathrm{d}i/\mathrm{d}t = 72.77\ \mathrm{mA\ s^{-1}}$.

If instead we need to know the rate of change of current when the voltage has a particular value, we calculate the corresponding values of r and t, then substitute these in Eq. 131.

In this example, the time-related variables are synchronised with each other. This is because the differentials still include t. This need not be the case. If, for example, we state that the voltage is ramping upward at $3\ \mathrm{V\ s^{-1}}$, the equation is: $u = 3t$. Differentiating this gives $\mathrm{d}u/\mathrm{d}t = 3$. The rate of change is constant; it does not depend on how long has elapsed since we began timing. We might also specify that the resistance is increasing at the rate of $4\Omega\ \mathrm{s^{-1}}$, so that $\mathrm{d}r/\mathrm{d}t = 4$. The variables are still both time-dependent, since they are **rates of change**, but they are not synchronised in the time-scale. Substituting these differentials in Eq. 130 gives:

$\mathrm{d}i/\mathrm{d}t = 3/r - 4u/r^2$.

In this case, it is possible to specify any pair of values for u and r and calculate the corresponding rate of change of i. For example, when $u = 2$ V and $r = 4\ \Omega$, then:

$\mathrm{d}i/\mathrm{d}t = 3/4 - 8/16 = 0.25\ \mathrm{A\ s^{-1}}$.

As well as rates of change with respect to time, we often encounter rates of change with respect to temperature. The values of many components, including resistors, capacitors and regulators are all functions of temperature. The **temperature coefficient**, often known as the **tempco** of a resistor, may be of the order of 200 parts per million per degree Celsius (or per kelvin which, in this context, amounts to the same thing). As a differential, and now using the variable t to represent temperature instead of time, we have:

$\mathrm{d}u/\mathrm{d}t = 2r \times 10^{-4}$.

Given the tempcos of resistors and other components in a circuit, the method described above can be used to calculate the rate of change of current or other quantitu with respect to temperature changes.

Test yourself

1. Given that z is a function of both x and y, calculate the partial differentials $\delta z / \delta x$ and $\delta z / \delta y$ when
   (a) $z = 2x+3y$
   (b) $z = 3x^2/4y$

2. A 22 µF capacitor has a p.d. of 9 V across it. The charge q is given by $q = cv$. If the voltage increases by 0.2 V and the capacitance decreases by 0.5 µF, use the method of partial differentials to calculate by how much the charge changes.

3. In a voltage-regulating circuit (**Fig. 146**), a constant voltage $U = 10$ V is applied. The band-gap reference produces a reduced voltage with a tempco of 40 ppm/°C. The resistor has a tempco of 100 ppm/°C. When the temperature is 25 °C, u is exactly 2.5 V and $r = 7.5\ \mathrm{k\Omega}$. The current through the resistor is $i = (U-u)/r$. What is the tempco of the current?

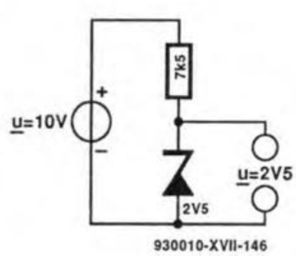


Fig. 146

Answers to Test yourself (Part 16)

1. (a) Odd function; no constant term, no cosine terms; expect to find only sine terms.
   (b) Half-wave repetition, neither even nor odd; expect to find a constant term, even cosine terms and even sine terms.
   (c) Even function; expect to find a constant term and cosine terms.

2, This has half-wave repetition; with the constant term ignored, it is an odd function; expect to find a constant term and even sine terms.
   $A_0 = 3\pi/2$
   $a_n = 0$
   $b_n = -3/n(\cos n\pi+1)$
   When n is even, $\cos n\pi = 1$, and $b_n = -6/n$.
   When n is odd, $\cos n\pi = -1$ and $b_n = 0$. Only even terms. The series is:
   $y = 3\pi/2 - 6[(\sin 2t/2) + (\sin 4t/4) + (\sin 6t/6) + \ldots]$

APPLICATION NOTE

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

OPTOISOLATOR TYPE IL300

AND ITS USE IN ISOLATION AMPLIFIERS

By K. Schönhoff

The electrically isolated transfer of analogue signals is a frequent occurrence in measurement techniques, for instance, in measurements on the mains supply, or on signals of which the parameters are largely unknown. Also, for safety reasons, it is often required to isolate ground connections.

These requirements are usually met by the use of an isolation amplifier. Such an amplifier could be described as an operational amplifier with isolating properties. An isolation amplifier performs the transfer of analogue data (current, voltage) between two points without an electrical connection between these points. Isolation amplifiers prevent earth loops, suppress common-mode signals, ensure separation of potentials and lessen the effect of capacitive and inductive interference.

Isolation amplifiers can be divided into three categories that use capacitive, transformer or optical coupling respectively. Techniques that are based on acoustic, piezo-electric, thermal or mechanical coupling are used only in special applications.

Capacitive coupling

In this, the signal transfer is brought about by two small capacitors, which alternately pass on pulse edges that contain the information—see **Fig. 1**. Use is made of a modulation/demodulation combination in which the modulator forms a control loop via a reference demodulator. This arrangement compensates linearity errors produced in the modulator.

Isolation amplifiers that use capacitive coupling are easy and inexpensive to produce, since the capacitors, together with the semiconductor chips, are normally housed in a modified ceramic DIL enclosure. They have a large bandwidth (about 60–200 kHz), are d.c. stable and have good linearity. The output signal is staircase-shaped owing to the sample-and-hold trap.

Transformer coupling

In this, the oldest isolation transfer technique, the d.c. signal is chopped up, transferred by amplitude modulation and then rectified—see **Fig. 2**. Modern transformer coupling systems use a form of feedback to obtain better linearity.

The system provides good suppression of interference, but its bandwidth is not very large.

Optical coupling

Optoisolators provide an excellent means of converting analogue signals into corresponding light signals. However, precise and linear signal transfer requires careful design. In **Fig. 3**, the light receiver is copied in the control circuit where it provides the nominal values for the light-intensity control loop. This arrangement compensates for ageing effects and non-linearity of the light source. It is, of course, imperative that both light receivers operate in identical conditions and receive exactly the same amount of light.

Optical coupling enables the continuous transfer of analogue signals and is insensitive to interference. Its speed of operation is high (up to 10 MHz) and it produces no spurious signals, since it does not use modulation and demodulation.

Other techniques

Use is often made of frequency-to-voltage converters followed

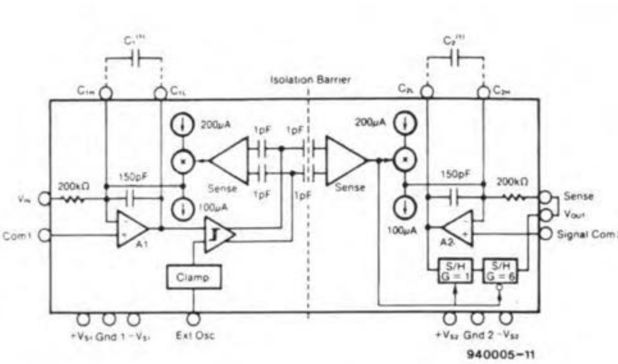


Fig. 1. Isolation amplifier using capacitive coupling.

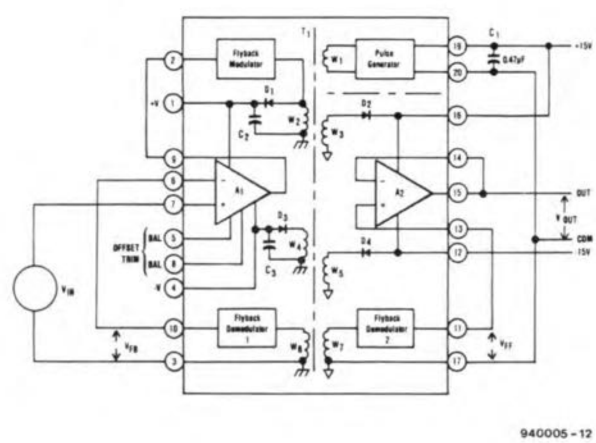


Fig. 2. Isolation amplifier using transformer coupling.

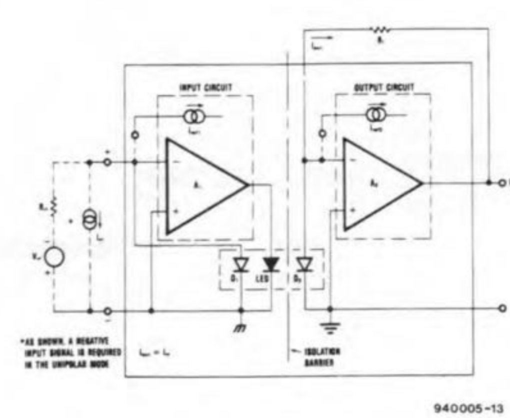


Fig. 3. Isolation amplifier using optical coupling.

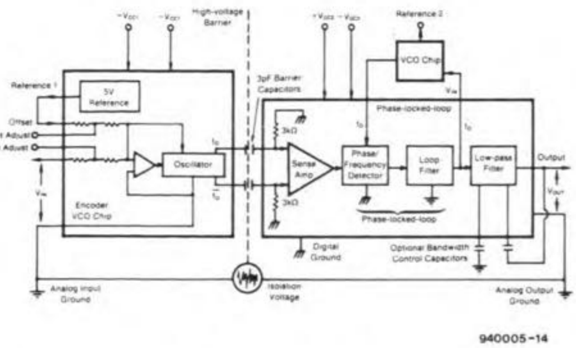


Fig. 4. Isolation amplifier with voltage-frequency converter using capacitive coupling.

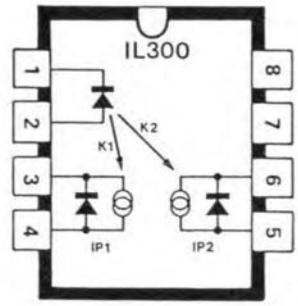


Fig. 5. Pinout of the IL300.

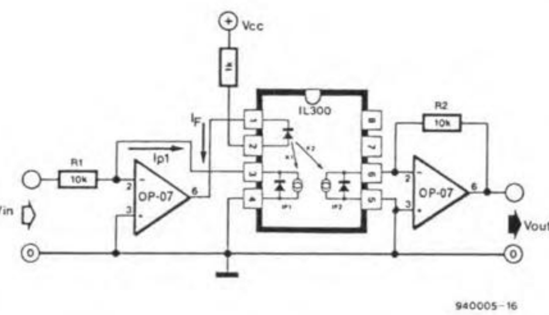


Fig. 6. Isolation amplifier for unipolar signals in a photovoltaic configuration.

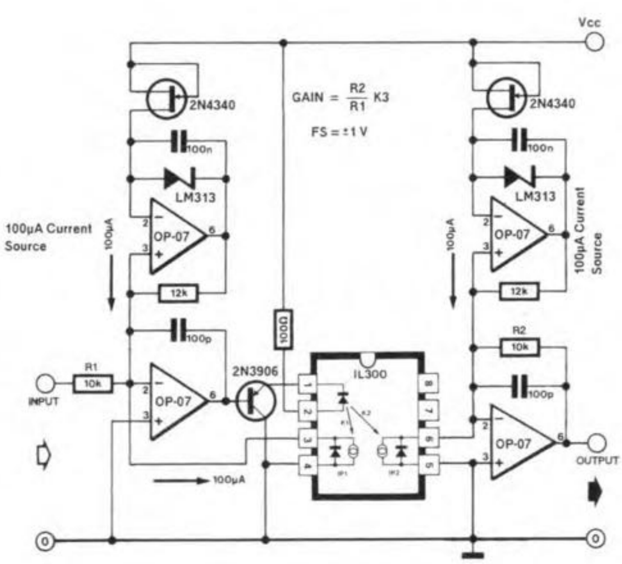


Fig. 7. Isolation amplifier for bipolar signals in a photovoltaic configuration.

by an isolation section using capacitive, transformer or optical coupling—see **Fig. 4**. This results in limited bandwidth but excellent linearity, particularly if optical coupling is used.

There are also analogue-to-digital converters with a serial output and optical signal transfer. These provide very good linearity (16 or more bits), but have a very limited bandwidth and cause signal delay. The latter may be detrimental in, for instance, control circuits.

Discrete designs?

Integrated circuits containing the components for most of the outlined techniques are commercially available. However, for special applications it is often necessary to build an isolation amplifier from discrete components, so that certain parameters can be optimized. The problem with this is the lack of suitable coupling elements.

Easily available optoisolators have large tolerances and suffer from ageing effects, temperature dependence of the transmitter capacitance, and non-linearity of the transmit and receive diodes.

Especially the non-linearity of optoisolators makes their application often difficult. The linearity improves slightly when the drive is reduced, but then the signal-to-noise worsens.

One way of improving the linearity is the use of feedback by means of a second optoisolator. Unfortunately, its application is handicapped by the necessary matching of the optoisolators. Even after the most careful calibration, their parameters may diverge widely because of temperature variations. In extreme cases, the negative feedback may even turn into positive feedback (that is, oscillations).

Dual optoisolators, because of their thermal coupling, would be the answer to the problem were it not for the fact that their transmitter and receiver are at the same side of the isolation barrier.

The IL300

The Type IL300 optoisolator from Siemens is, however, a device that has been designed especially for the transfer of analogue signals. It consists of a combination of an infra-red LED and two photodiodes—see **Fig. 5**. The feedback diode captures part of the light emitted by the LED. From this, a control signal is derived that largely compensates for the non-linearity and the time/temperature dependence of the LED. Since the two diodes are closely coupled, there is no likelihood of differences in parameters. That is, the feedback remains stable and the output signal is a true reflection of the input signal.

The main parameters of the IL300 are:

- non-linearity: ≤0.01% (with feedback).
- temperature stability: ±0.005% $°C^{-1}$.
- Minimal coupling capacitance.
- Low dissipation: <15 mW.
- Isolation voltage: 7500 V_{pp} 1 s.
- Internal isolation barrier: >0.4 mm.

The photodiodes may be used as voltage sources (photovoltaic) or as resistors (photoconductive).

Used photovoltaically, the diodes provide excellent linearity, minimal noise and minimal drift. The transfer properties meet the requirements for 12-bit converters handsomely.

Used photoconductively, the diodes provide a large bandwidth, but linearity and drift are no better than required of 8-bit converters.

The current gain, K_1, between the LED and the reference diode is typically 0.007, that between LED and receive diode, K_2, is typically 0.007, and the transfer gain between the individual diodes, K_3, is typically 1.0.

The significantly higher gain of traditional optoisolators (0.5–1.5) is achieved by the use of phototransistors in the receiver. Such transistors, however, are much slower than the diodes used in the IL300.

Some applications

The versatility of the IL300 is illustrated in the application circuits that follow.

Figure 6 shows the basic circuit of an isolation amplifier with the IL300 used photovoltaically. In the sample calculation below, the following values were used:

- Maximum current of the OP-07: $I_{0\,(max)} = \pm$ 15 mA;
- $K_1 = 0.007$;
- $K_2 = 0.007$;
- $K_3 = 1.0$;
- Input voltage, U_i = 0–1.0 V.

First, the peak current through

the coupling diode is determined:

$$I_{p1\,(max)} = K_1 \times I_{o\,(max)} = 0.007 \times 15 = 105\ \mu A.$$

Input resistor R_1 is calculated from $I_{p1\,(max)}$ and $U_{i\,(max)}$:

$$R_1 = U_{i\,(max)} / I_{p1\,(max)} = 1 / 105 \times 10^{-6} = 9.524\ k\Omega \approx 10\ k\Omega.$$

The value of R_2 depends on the required total amplification, G, and the transfer gain, K_3, of the IL300. Assuming $G = 1$ and $K_3 = 1$:

$$R_2 = R_1\, G / K_3 = 10^3 \times 1 / 1 = 10\ k\Omega.$$

The small-signal bandwidth of the amplifier is 45 kHz.

Figure 7 is an isolation amplifier for bipolar input signals. The 2N4340 transistors provide relief for the output stage of the OP-07 and thus improve the drift behaviour. The current sources determine the operating zero point of the IL300 to make bipolar operation possible. If the requirements are not so stringent, the

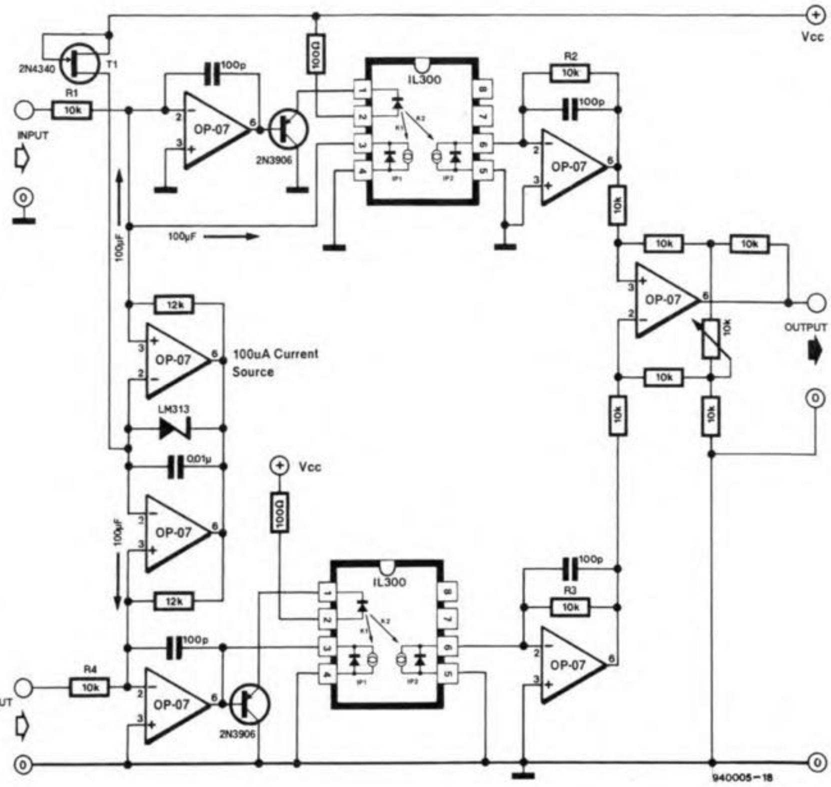


Fig. 8. Isolation amplifier with differential inputs, also for bipolar signals, in a photovoltaic configuration.

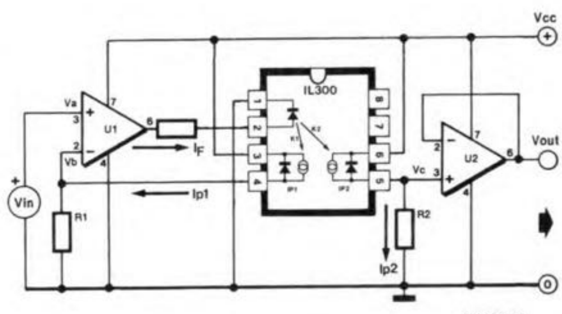


Fig. 9. Isolation amplifier for unipolar signals in a photoconductive configuration.

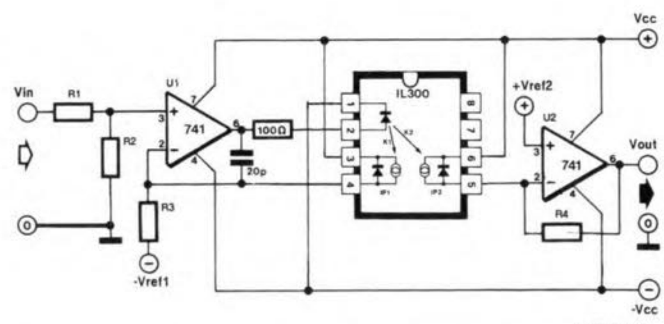


Fig. 10. Isolation amplifier for bipolar signals in a photoconductive configuration.

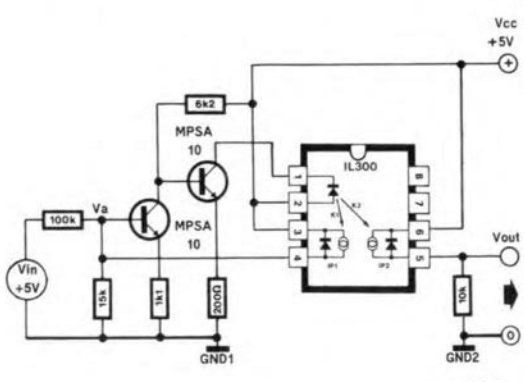


Fig. 11. It does not have to be complicated: just two transistors and an IL300 can provide good electrical isolation of analogue signals.

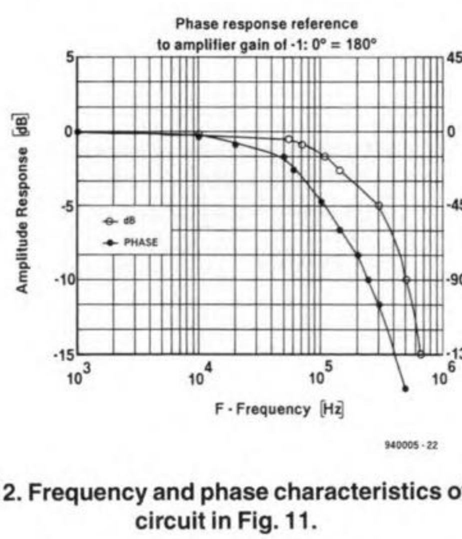


Fig. 12. Frequency and phase characteristics of the circuit in Fig. 11.

Absolute Maximum Ratings	Symbol	Min.	Max.	Unit
Emitter				
Power Dissipation (T_A=25°C)	P_{LED}		160	mW
Derate Linearly from 25°C			2.13	mW/°C
Forward Current	If		60	mA
Surge Current (Pulse width <10µs)	Ipk		250	mA
Reverse Voltage	V_R		5	V
Thermal Resistance	Rth		470	°C/W
Junction Temperature	T_J		100	°C
Detector				
Power Dissipation	P_{DET}		50	mW
Derate linearly from 25°C			0.65	mW/°C
Reverse Voltage	V_R		50	V
Junction Temperature	T_J		100	°C
Thermal Resistance	Rth		1500	°C/W
Coupler				
Total Package Dissipation at 25°C	P_T		210	mW
Derate linearly from 25°C			2.8	mW/°C
Storage Temperature	T_S	-55	150	°C
Operating Temp	T_{op}	-55	100	°C
Withstand Test Voltage 1 min., 60 Hz	WTV		4420	VAC_{RMS}
Withstand Test Voltage 1 min	WTV		6250	VAC_{PEAK}
Withstand Test Voltage 1 sec., 60 Hz	WTV		5300	VAC_{RMS}
Withstand Test Voltage 1 sec	WTV		7500	VAC_{PEAK}
Working Voltage	WV		1700	VAC
Lead Soldering Time at 260°C			10	sec

Fig. 13. Absolute maximum ratings of the IL300.

current sources may be replaced by resistors.

Figure 8 shows the circuit of a differential amplifier that may be used as an instrumentation amplifier with a resolving power of 12 bits. The common-mode rejection is determined by the IL300 and the differential amplifier at the output. Since the output stage is arranged as a differential amplifier, a current source is not required. The circuit provides a bandwidth of 50 kHz and a very high common-mode rejection of 140 dB at 10 kHz.

Figures 9 and **10** show the IL300 in a photoconductive configuration. The circuit in **Fig. 10** has a bandwidth of ≥ 100 kHz.

Figure 11 shows an isolation amplifier designed with discrete components. This circuit is often used in the instantaneous value feedback loop of the control stage in a D-A converter. There is compensation for the ageing of the LED. **Figure 12** gives an idea of the (large) bandwidth of this circuit. Since the transfer of the instantaneous value is carried out in an exact and reliable manner, the control circuit can be at the same side of the barrier as the converter.

Finally

The power supply to the input section must, of course, be electrically isolated from the remainder of the amplifier. The requirements for this are identical to those for the signal transfer. Normally, d.c.-to-d.c. converters are used for this. Note, however, that the transformers used in this may transfer noise and interference. Temporary substitution of a battery supply will quickly indicate whether this is really caused by the transformers or whether there is a fault in the amplifier.

[940005]

Characteristics (T_A=25°C)

	Symbol	Min.	Typ.	Max.	Unit	Test Condition
LED Emitter						
Forward Voltage	V_F		1.25	1.50	V	I_F=10 mA
V_F Temp. Coefficient	$\Delta V_F/\Delta$°C		−2.2		mV/°C	
Reverse Current	I_R		1	10	µA	V_R=5 V
Junction Capacitance	C_J		15		pF	V_F=0 V, f=1 MHz
Dynamic Resistance	$\Delta V_F/\Delta I_F$		6		Ω	I_F = 10 mA
Switching Time	t_R		1		µS	ΔI_F=2 mA, I_{FQ}=10 mA
	t_F		1		µS	ΔI_F=2 mA, I_{FQ}=10 mA
Detector						
Dark Current	I_D		1	25	nA	V_{det}=−15 V, I_F = 0 µA
Open Circuit Voltage	V_D		500		mV	I_F=10 mA
Short Circuit Current	I_{SC}		70		µA	I_F=10 mA
Junction Capacitance	C_J		12		pF	V_F=0 V, F=1 MHz
Noise Equivalent Power	NEP		4 [illegible]		W/√Hz	V_{det}=−15 V
Coupled Characteristics						
K1, Servo Current Gain (I_{P1}/I_F)	K1	0.0050	0.007	0.011		I_F=10 mA, V_{det}=−15 V
Servo Current, see Note 1, 2	I_P1		70		µA	I_F=10 mA, V_{det}=−15 V
K1 Temperature Coefficient	ΔK1/ΔT		-0.5		%/°C	I_F=10 mA, V_{det}=−15 V
K2, Forward Current Gain (I_{P2}/I_F)	K2	0.0036	0.007	0.011		I_F=10 mA, V_{det}=−15 V
Forward Current	I_P2		70		µA	I_F=10 mA, V_{det}=−15 V
K2 Temperature Coefficient	ΔK2/ΔT		-0.5		%/°C	I_F=10 mA, V_{det}=−15 V
K3, Transfer Gain (K2/K1) See Note 1, 2	K3	0.56	1.00	1.65	K2/K1	I_F=10 mA, V_{det}=−15 V
Transfer Gain Linearity	ΔK3		±0.25		%	I_F=1 to 10 mA
Transfer Gain Linearity	ΔK3		±0.5		%	I_F=1 to 10 mA, T_A=0°C to 75°C
K3 Temperature Coefficient	ΔK3/ΔT		±0.005		%/°C	I_F=10 mA, V_{det}=−15 V
Photoconductive Operation						
Frequency Response	BW (-3 db)		200		KHz	I_{FQ}=10 mA, MOD=±4 mA, R_L=50 Ω,
Phase Response at 200 KHz			-45		Deg	V_{det}=−15 V
Rise Time	t_R		1.75		µs	
Fall Time	t_F		1.75		µs	
Insulation – Isolation						
Input-Output Capacitance	C_{IO}		1		pF	V_F=0 V, f=1 MHz
Common Mode Capacitance	C_{cm}		0.5		pF	V_F=0 V, f=1 MHz
Common Mode Rejection Ratio	CMRR		130		dB	f=60 Hz, R_L=2.2 KΩ
Insulation Resistance	R_{IO}		100		GΩ	Vio=500 VDC
Withstand Test Voltage	WTV	4420			VAC_{RMS}	Rel. Humidity ≤ 50% I_{IO}≤10 µA, 1 min
WTV	6250			VAC_{PEAK}	Rel. Humidity≤50% I_{IO} ≤10 µA, 1 min	
WTV	5300			VAC_{RMS}	Rel. Humidity ≤ 50% I_{IO} ≤0 µA, 1 sec	
	WTV	7500			VAC_{PEAK}	Rel. Humidity ≤ 50% I_{IO}≤10 µA, 1 sec

Fig. 14. Parameters of the IL300.

SMALL LOOP ANTENNAS FOR MW AM BCB, LF AND VLF RECEPTION

PART 1 — INTRODUCTION AND THEORETICAL BACKGROUND

By Joseph J. Carr, B.Sc., M.S.E.E.

THE Very Low Frequency (VLF) bands are found between 10 kHz and 30 kHz; the LF frequencies occupy 30 kHz to 300 kHz; and the medium wave (MW) band is from 300 kHz to 3,000 kHz. Within the medium wave region is the AM broadcast band, occupying 540 to 1,700 kHz.[a] Activity in those bands provides some interesting opportunities for DXers. There are AM broadcasters on 145 to 280 kHz, while maritime CW (Morse) traffic is heard on frequencies around 500 kHz.[b] In the 100-kHz region one finds Loran-C navigation stations, and in the 10 to 14-kHz region are Omega navigation stations. Time and frequency standards stations operate at 50 kHz and 60 kHz. A number of beacon and other miscellaneous stations are heard throughout the LF/VLF spectrum. Several navies operate submarine communications stations in the VLF bands (several operate in 20 to 40 kHz region).

An increasingly popular activity is natural terrestrial radio and solar radio astronomy observations in the VLF region. Two different sorts of activity are found. Some people like to listen for 'whistlers'; i.e., natural radio signals created by distant bursts of lightning (Mideke, 1992). They are electromagnetic waves (not sound waves) in the 1 to 10 kHz region, and typically exhibit an exponentially decaying tone. The solar radio astronomy activity is observing the effects of solar events that create **sudden ionospheric disturbances** or SIDs (Taylor and Stokes, 1992). While shortwave propagation drops out due to D-layer absorption during SIDs, VLF propagation is enhanced. It is this enhancement that observers look for when detecting SIDs.

Noise is a big problem in the LF/VLF spectrum, and to some extent in the MW band. Lightning static and other 'hash' is seen. There is also considerable harmonic energy from 50 or 60-Hz power mains. While such frequencies do not normally produce high harmonic levels, when power lines are considered it is seen that even proportionally very weak harmonics have considerable energy content simply because of the high power levels handled by the lines. As a result, sensitive radio receivers see large amounts of power line interference. There is also considerable hash and 'birdies' from television receivers, video cassette recorders and other household and commercial and industrial electronics devices.

The antenna problem

Regardless of the interest of the LF/VLF/AM BCB listener, there is one problem that is common to all of them: **the antenna**. Resonant antennas for LF/VLF are huge, and AM BCB antennas are not far behind. Over the frequency range mentioned above, a quarter wavelength antenna would range from 150 to 25,000 metres in length. Friend, mentor, and past-president of the **American Radio Relay League** (ARRL), the late Vic Clark (W4KFC), once told me of seeing a rhombic for LF submarine communications in Peru in the early 1960s. Such antennas are usually 1.5λ to 7λ on each side. This one was 38.6 km (24 miles) on each of its four sides! A quarter wavelength vertical on 25 kHz is three kilometers high. If small verticals or short random length wires are used instead, performance suffers accordingly. Clearly, such antennas are beyond most of us.

A popular solution to the problem of receiving antennas for LF/VLF/AM BCB

[a] The USA recently extended the AM BCB upper limit from 1,610 kHz to 1,700 kHz.

[b] It is expected that some nations will cease routine manned monitoring of 500 kHz, the international CW calling and distress frequency, in the near future.The US Coast Guard has already discontinued its regular 500 kHz manned radio watch. Monitoring for automatic distress signals will continue however.

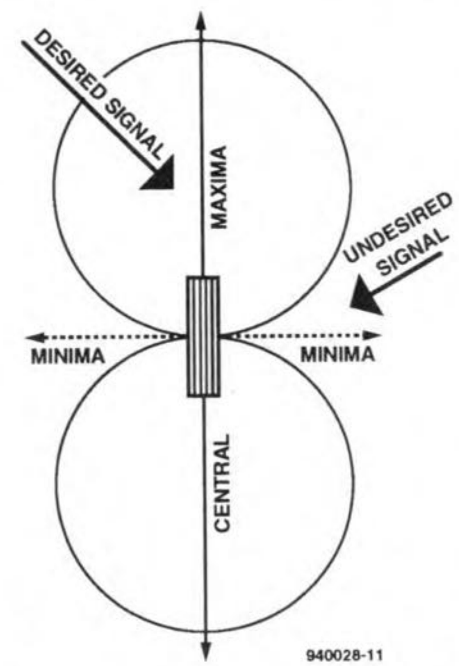


Fig. 1. Directivity pattern of the loop antenna.

reception is the use of small loops. The loop antenna can resonant or non-resonant, although the resonant form will produce a much larger output signal. In this article we will take a look at some of the basic issues in designing a simple loop antenna for the LF/VLF bands.

Small loops vs. large loops

Several different forms of loop antenna are used extensively, but they generally fall into two classes: **large loops** and **small loops**. The large loop variety typically have overall perimeter lengths that range from about 0.5λ (e.g., the miniloop) to 2λ (e.g., the bisquare). Perhaps the most common example of the large loop antenna is the 1λ square, i.e., a square wire loop that is 0.25λ per side (this antenna can be considered a 'one-element quad'). The radiation pattern of this antenna at its resonant frequency will be a bidirectional 'figure-8' pattern, similar to a dipole, with the maxima perpendicular to the plane of the loop and the nulls off the ends. The Delta Loop antenna is another popular example.

Small loop antennas, on the other hand, have perimeter lengths which are considerably less than one wavelength. A naval training manual from World War II era used ≤0.22λ as the maximum perimeter length for radio direction finding (RDF) loops; Jasik (1961) used ≤0.17λ; Kraus (1950) used ≤0.1λ; and **The ARRL Antenna Book** (ARRL 1988) recommends ≤0.085λ.

A difference between large and small loop antennas is seen in the current distribution around the antenna perimeter. In a large loop, the current varies as a function of length, with nodes and antinodes at critical points (location dependent upon polarization and therefore feedpoint). In the small loop, however, the current is uniform across the entire length of the loop perimeter.

A radio signal is a transverse electromagnetic wave, and as such it possesses orthogonal magnetic (H) and electric (E) fields that propagate together along the same line; energy oscillates back and forth between the E and H fields. The polarity of an antenna (horizontal or vertical) depends on the direction of the E-field. The large loop antenna responds primarily to the E-field of the radio signal, while the small loop responds primarily to the H-field. This attribute of the small-loop antenna is one reason why the antenna performs well in the presence of interfering signals from power mains and electrical appliances. Such signals near to the receiver tend to be dominant E-field phenomena, so the small loop is less sensitive to them than to radio signals.

Small loop antennas are built using a number of common symmetrical geometries (circular, square, hexagonal, octagonal, pentagonal and triangular), although the circular and square are probably the most common. For most applications, the square form is probably the easiest to construct, so is used for most of the examples in this article. An interesting aspect to the circular and square loops is that the far fields of the two antennas are very nearly equal when the square of the loop area (A) is greater than, or equal to, 1/100 of the wavelength (i.e. $A^2 \geq \lambda/100$).

Small-loop directivity

The directivity of the small loop antenna is opposite that of the large loop, even though both produce 'figure-8' patterns. Like the dipole, most large loops have maximum sensitivity perpendicular to the plane of the loop. The small loop, however, has maximum sensitivity off the ends, and has the nulls (or 'minima') perpendicular to the plane of the loop (see **Fig. 1**).

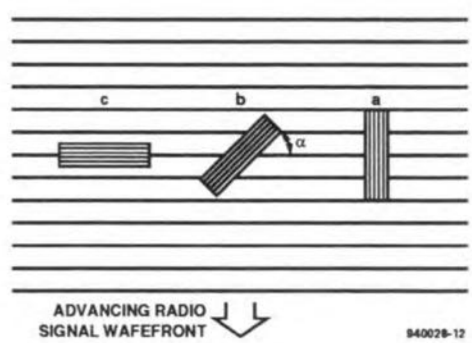


Fig. 2. Three different orientations with respect to advancing radio wave show why the null is broadside to loop plane. In (a) the loop is orthogonal to the wave, so more lines of force cut the conductors (maximum response); At (b) the antenna is oriented at an angle α, so the response is reduced by cos α and (c) the antenna is broadside to the wave, so few lines of force cut the antenna.

Figure 1 also illustrates one reason why the loop is used for RDF, and why it is popular with receiver operators in the lower frequency bands (up to about 7 MHz). The nulls are very sharp compared with the maxima, so can be used to either pin point the direction of the station, or to null it out. Receiver operators can place the null in the direction of an interfering station (or electrical noise source), nulling it out, even when the direction of the maxima is not 'dead on' the direction of the desired station. Even

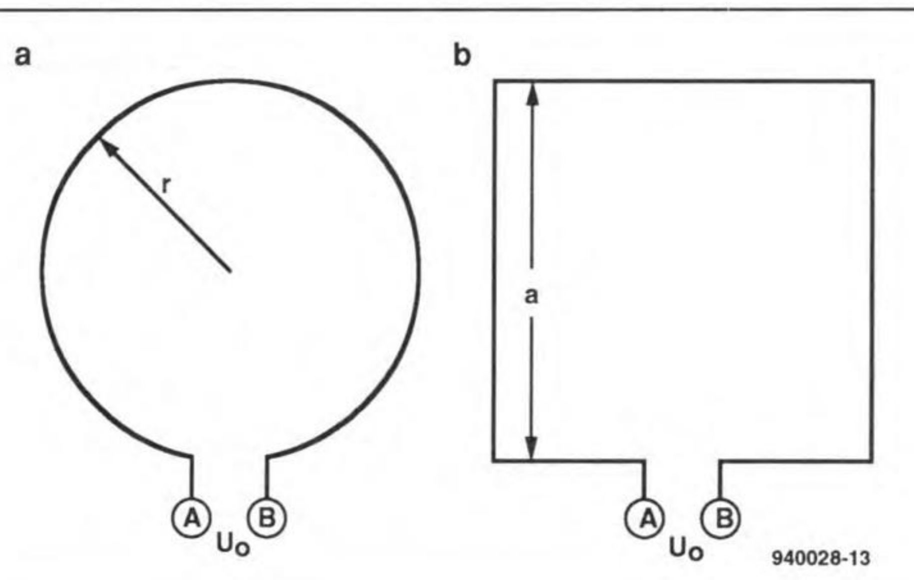


Fig. 3. Dimensions and form of two loop antennas: a) circular and b) square.

though the antenna is not optimized for the desired station, it is optimized for the overall situation because the ratio between the desired and undesired signal strengths is increased. As with signal to noise ratio, the 'desired signal to interfering signal ratio' is critical in eliminating co-channel and adjacent channel interference.

The directivity of the small loop, shown in Fig. 1, might not be immediately intuitive without reference to **Fig. 2**. Here we see three loops at different angles with respect to the advancing wavefront of the radio signal. The loop antenna at 'A' is positioned for maximum sensitivity to the radio signal, i.e., at a angle of 0 degrees. Note that it is cut by several lines of force by the signal, so produces a maximum induced current in the loop conductors. At 'B' the loop is oriented at some angle α with respect to the advancing wave. This loop sees a signal that is less than that of 'A' by the cosine of the angle. Finally, at 'C' we see the broadside situation, i.e., the antenna is oriented 90 degrees with respect to the advancing wave front. In this case, the induced signal is theoretically zero, although some small signal will exist in practical antennas, which accounts for the sharpness and deepness of the null.

Structure of the small loop antenna

The simplest form of loop antenna is the single turn circular (**Fig. 3a**) or square (**Fig. 3b**) form. This type of loop antenna is frequently seen in the shortwave bands, especially above 7 MHz. Amateur radio single-turn RDF 'foxhunting' loops in the 10-metre and 6-metre bands are well known. At the VLF, LF and AM BCB frequencies, however, single turn loops are not too popular because of the small signal pick-up exhibited. Examination of Eq.[1] demonstrates that a small loop would need a fairly large physical size (especially area A) in order to compensate for the lack of turns when the wavelength (λ) gets large. The output voltage, U_o, is found from:

$$U_o = \frac{2\pi A N E_f \cos\alpha}{\lambda} \qquad [1]$$

Where:
U_o is the output signal level in volts;
A is the area of the loop in square meters (m²);
N is the number of turns in the loop;
E_f is the strength of the radio signal in volts per metre (V/m);
α is the angle of arrival of the signal, as defined above, in degrees;
λ is the wavelength of the received signal, in metres (m).

The circular loop antenna is physically a bit harder to construct than the square loop, so for this article we will concentrate largely on the square loop. **Figure 4** shows the basic structure of the multi-turn square loop. Two largely equivalent forms of square loops are the **depth wound** (as shown) or the **planar wound**. In the latter form, the turns are wound all in the same plane, with the square formed by any one turn being slightly different in dimension than the turns adjacent to it. In the depth wound type, the turns form the same size square, but they are, so to speak, stacked on top of each other. In the depth wound square loop, the 'b' dimension is the loop thickness, while in the planar wound form the 'b' dimension is its width. A constraint on the square loop design is that the length of each side be five times (or more) either the width or depth ($a \geq 5b$). Shortly we will discuss methods for winding the square loop.

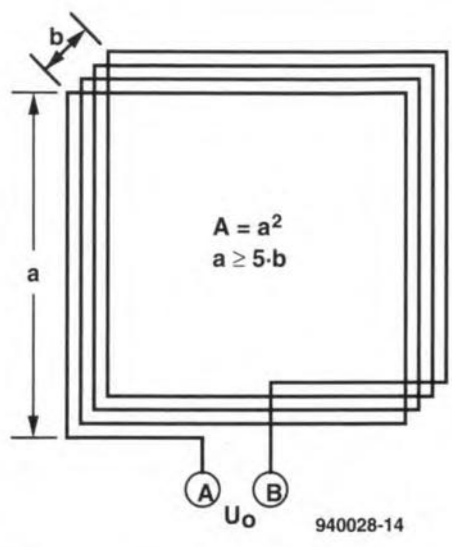


Fig. 4. Form of the standard square loop.

You may sometimes see small loop antennas characterized by their 'effective height' (H_{eff}), which is a theoretical construct that compares the output signal level (U_o) to a vertical piece of the same type of wire of height H_{eff}. The effective height is:

$$H_{eff} = \frac{2\pi N A}{\lambda} \qquad [2]$$

When practical examples are worked, it will be noted that for any easily obtained field strength the output voltage of a small loop antenna is quite small. There are three strategies for increasing the output voltage of the loop without resorting to amplification (which can also be used). First, note that the output voltage in Eq.[1], and the effective height in Eq.[2], are directly proportional to the number of turns (N) in the loop. By adding more turns, the output voltage is increased. However, there is a practical limitation to this approach because adding turns increases resistive losses and also stray capacitance. The stray capacitance can be remarkably high, and results in a relatively low self-resonance point if too large.

The second strategy is to increase the area of the loop. The area rises according to the square of the loop sides ($A=a^2$). Thus, doubling the length of the loop sides quadruples the output voltage. The practical limitation on this approach is the physical difficulties of building and siting a large loop. I have seen 5-metre to the side square VLF loops, but at a rural

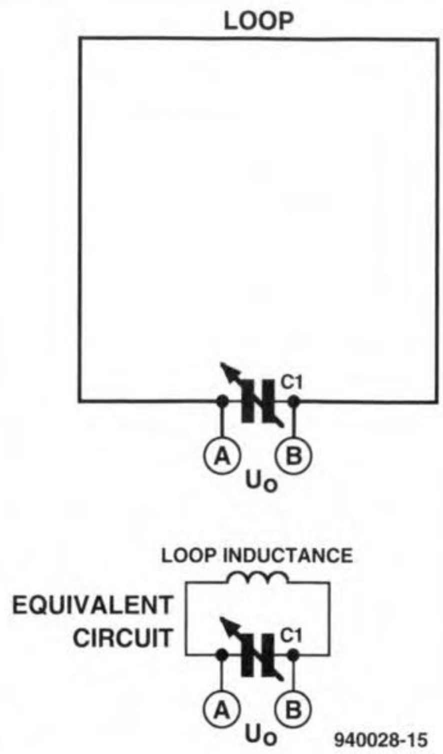


Fig. 5. Resonating the loop (parallel version).

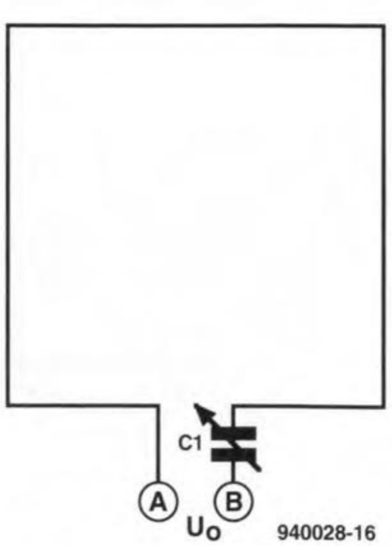


Fig. 6. Resonating the loop (series version).

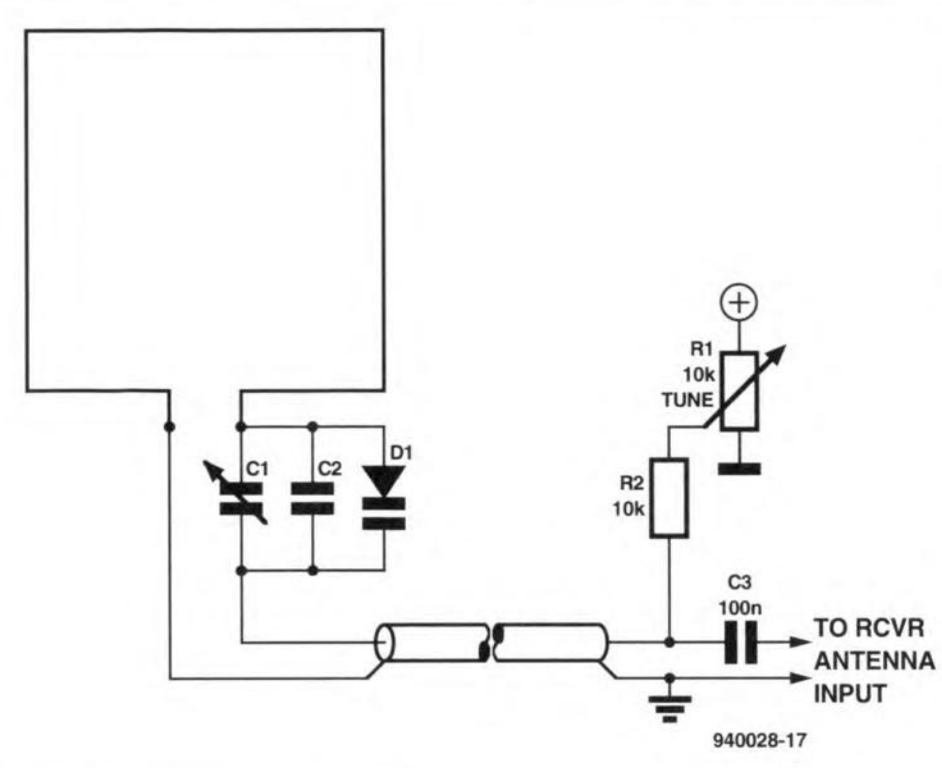


Fig. 7 Using a varactor diode to remotely tune the loop using a direct voltage.

Shape	K_1	K_2	K_3	K_4
Triangle	0.006	1.155	0.655	0.135
Square	0.008	1.414	0.379	0.33
Hexagon	0.012	2.00	0.655	0.135
Octagon	0.016	2.613	0.7514	0.0715

Table 1. Shape co-efficients K_x for Grover's equation.

university site, and not an urban residential site.

The third strategy is to resonate the loop inductance with a capacitance. The output voltage of the resonant vice non-resonant loop is multiplied by the figure of merit (Q) of the resulting *LC* tuned circuit; typical values run from 5 to 200, with values near 100 being quite easy to obtain. The output voltage of the resonant loop is:

$$U_0 = \frac{2\pi A N Q E_f \cos\alpha}{\lambda} \qquad [3]$$

Figure 5 shows the resonant loop and the equivalent *LC* circuit (neglecting resistance). Only one turn is shown for simplicity, but the same diagram also suffices for multiturn loops. A capacitance (C_1) is in parallel with the loop inductance. When resonated, the loop will produce a signal many times larger than the signal of the un-resonated loop. However, the gain in signal output is at the expense of convenience. The resonant loop must somehow be tuned, which makes remote siting of the loop antenna (e.g., outdoors or in the attic) somewhat more difficult. In some cases, the loop is series resonated, as shown in **Fig. 6**. This approach is probably somewhat less common than the parallel approach, but is available to those who wish to try it.

The capacitors used for loop antennas tend to be receiving type 'broadcast variables'. I have used 365-pF single section variables, 380-pF, and 3×365-pF with all three sections in parallel (1,100 pF). Another good prospect is the 500-pF single section units offered by **Maplin Electronics** (P.O. Box 3, Rayleigh, Essex, SS6 8LR, UK).

Figure 7 shows the use of a voltage variable capacitance diode ('varactor') to resonate the antenna. The additional capacitors are a trimmer (C_1) and a parallel capacitor used to increase the overall capacitance. These extra capacitors are needed generally because few varactors are available over 500 pF, so it may be the case that practical sized loops cannot easily be resonated with only a varactor at LF and VLF frequencies.

Loop inductance

There are several different ways to calculate loop inductance, and they produce somewhat different results. Three methods are found in the literature, although one (Patterson, 1987) applies largely to single-turn loops made of large diameter conductors (3 to 20 mm) so is used mostly at high MW and low HF. For AM BCB and below, one might wish to pick from Somerfield's (1952) equation or Grover's (1946) equation.

Somerfield's equation:

$$L = \frac{2a\mu_0 N^2}{\pi}\ln\left(\frac{16a}{b}\right) \qquad [4]$$

Where: $\mu_0 = 4\pi \times 10^{-7}$, and the turns are spaced on width b, a distance of $b/(n-1)$ apart (i.e. close-wound). The Somerfield equations seem to work within an expected range of error, but Grover's equation seems closer to the actual inductance measured in empirical tests performed by me.

Grover's equation:

$$L = K_1 N^2 a\left(\ln\left(\frac{K_2 a n}{(n+1)b}\right) + K_3 + \left(\frac{K_4 (n+1) b}{a n}\right)\right) \qquad [5]$$

Where:
L is the inductance in microhenrys (µH);
a is the length of a loop side in centimeters (cm);
b is the loop width in centimeters (cm);
n is the number of turns in the loop;
K_1 through K_4 are given in **Table 1.**

The equations of Patterson (1967), reported by David (1991):

$$L = (0.00508a)\left(2.303\log\frac{4a}{d} - \phi\right) \qquad [6]$$

Where:
L is the loop inductance in microhenrys (µH);
a is the length of the perimeter of the loop (inches);
d is the conductor diameter (inches);
ϕ is a factor from **Table 2**.

Shape	Factor (ϕ)
Circle	2.451
Octagon	2.561
Hexagon	2.6.6
Pentagon	2.712
Square	2.853
Triangle (equilateral)	3.197

Table 2. Shape factors ϕ for Patterson's equation.

Performing several experiments with VLF loops showed that the Somerfield equation differed from the inductance read by a digital *LC* meter by 20 to 25 percent, while the inductance calculation from Grover was nearer the mark. One loop that I built measured 6.96 mH on the meter, and Grover predicted 6.33 mH. A BASIC program for calculating the loop inductance based on Grover is provided with this article. The version of the program published here will run on any MS-DOS compatible machine, and must be run through BASICA, GW-BASIC or their equivalent.[c]

To resonate the loop requires a specific capacitance determined by the resonance equation. Because there is often a tremendous stray or distributed capacitance (C_s), this value must be accounted in the calculation. In one loop that I built, measuring 61 cm per side with 50 turns of wire, the capacitance was nearly 200 pF. This value was confirmed two ways. First, by finding the self-resonant point, and knowing the inductance, it could be calculated. Second, by measuring the resonant frequency with a known capacitance (330 pF) and noting it was too low for the loop inductance and fixed capacitance. The stray capacitance could be calculated by subtracting the fixed 330-pF capacitance from the calculated capacitance needed to resonate the inductance to the frequency measured. Crude, but it works well. Both methods yielded numbers of the order of 200 pF, with about 15 percent difference between them. The overall capacitance needed can be calculated from knowledge of the loop inductance and the desired resonant frequency:

$$C_1 + C_s = \frac{1}{4\pi^2 f^2 L} \qquad [7]$$

Where:
f is the resonant frequency in Hertz (Hz);
C_1 is the value of the fixed capacitor in farads (F);
C_s is the stray capacitance in farads (F);
L is the inductance in Henrys (H).

Shielded small-loop antennas

The ideal pattern of **Fig. 1** shows uniform main lobes and sharply defined nulls. These attributes are the key to the successful use of the loop. But when the loop is close to the ground or other objects, the pattern will be severely distorted by stray coupling. The nulls tend to fill in, and the maxima lobes tend to become distorted. The answer to this problem is to isolate the loop antenna in a shielded enclosure, as in **Fig. 8**.

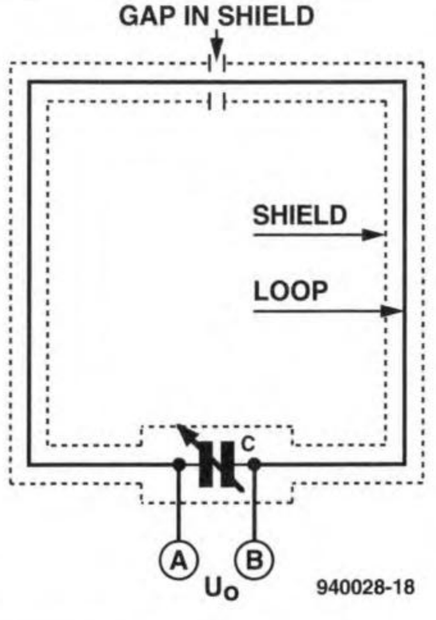


Fig. 8. Shielded resonant loop.

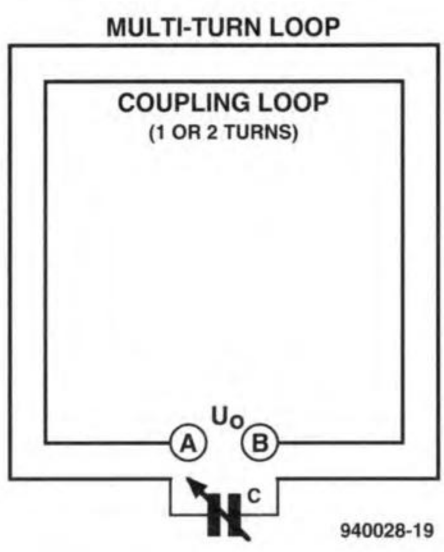


Fig. 9. Loop antenna with a coupling loop.

The gap in the shield is needed in order to allow the antenna to be exposed to the magnetic component of the electromagnetic wave. Because the shielding is effective against the electric field, and the loop responds most favourably to the magnetic field, the shield offers even further protection against stray *E*-fields from power lines and appliances.

Common practice calls for use of a small metal box to hold the variable capacitor and any coupling circuitry, and copper foil to surround the loop windings. Thin gauge copper foil can be found at hobby shops, the kind that model aircraft and ship builders frequent, as well as shops that sell to amateur doll house manufacturers. Generally, 36 to 44 gauge foil is used. Thicker copper material, such as roofing copper, is usable as well, but provides little or no electrical advantage at the cost of more money and considerable effort in forming the shield. While foil can be worked with a cheap pair of scissors, roofing copper must be worked with professional 'tin snips' or equivalent tools.

Coupling the loop antenna to the receiver

Before an antenna can be useful it must be coupled to the receiver. The simplest method for coupling a loop antenna to the receiver is to connect a coaxial transmission line across the output voltage terminals (A-B). While simple and appealing, it is also not a very good method, especially if the loop is tuned (the coax capacitance will detune the resonance point). Other methods are clearly needed.

Figure 9 shows the use of a transformer coupling winding on the loop antenna. The main winding is the multiturn loop antenna, and, in this case, it is resonated to a specific frequency by C_1; non-resonant loops also work using this same style of coupling. The coupling loop is one or two turns of wire wound along with the main loop. Some builders prefer to embed the coupling loop in the same bundle of wires as the main loop. For most applications, one or two turns will suffice for the coupling loop, although at LF frequencies, where the main loop may be 50 to 150 turns of wire, more may be needed.

In some antennas, the coupling loop is resonated with a series connected variable capacitor. Because there are so few turns on the coupling loop the series resonant capacitor must be much larger than the main tuning capacitor. As a result, it is not common to find coupling loops resonant below the MW BCB because of the capacitance required.

The coupling loop can be either grounded at one end, for single-ended output, or used in a balanced configuration. In any event, it is common practice to use a preamplifier to boost the output of the coupling loop. Almost any wideband amplifier that covers the frequency range of interest will suffice, and we will examine a few candidate designs shortly.

Some people use a transformer between the coupling loop and the input of the amplifier. Most small audio output transformers intended for transistor circuits will work well into the VLF region.

[c] A GW-BASIC plus an executable version for modern VGA compatible MS-DOS machines are available on a disk which may be obtained through our Readers Services, order code **1951**. See page 70 for price and ordering information.

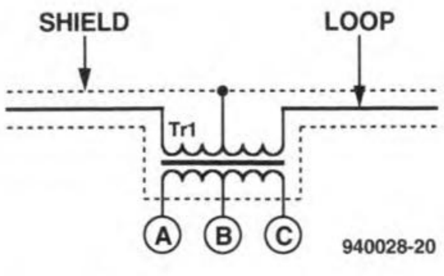


Fig. 10. Using an audio transformer to couple to the receiver.

The most common type used in antenna coupling has an 8-Ω secondary and a 1000-Ω primary. In the antenna circuit, however, the roles of the windings are reversed: the 8-Ω winding is connected to the output of the antenna coupling loop, while the 1,000-Ω winding is connected to the input of a preamplifier.

The frequency response of even the lowest cost audio transformers can be as high as 70 or 80 kHz, although the region below 50 kHz is typically flatter than the higher end of the bandwidth. Commercial grade audio transformers are available with impedance ratios similar to the output transformers cited above, but with ±1 dB bandwidths of 150 kHz. Although the cost is high, they work well in this application. **Mini-Circuits** offers small RF transformers in DIP IC-like packages. Several type numbers are available, with various turns ratios, that offer frequency response down to VLF region; some have ±3-dB points as low as 10 kHz. For loop antennas in the upper end of the VLF region, and lower end of the MW region, a modified 455 kHz interstage transformer can be used; use the type that has a resonant winding and a low-impedance coupling winding. Remove the capacitor that resonates the primary. On most small transistor radio IF transformers, the resonating capacitor is external to the transformer shell, located in a recess in the insulating base of the part. It is a tubular ceramic capacitor and can be removed by crushing.

A slightly different approach is shown in **Fig. 10**. In this case, a non-resonant shielded loop antenna without a coupling loop is used. The transformer is an audio interstage or line-to-line transformer. Devices that have been used successfully include 150:150, 500:500, 600:600 and 1000:1000 Ω units; also used have been step-up transformers. The principal feature required is that the primary winding be centre-tapped. The centre-tap of the transformer primary is connected to the shield surrounding the main loop. The transformer's secondary winding can be connected to the input of a preamplifier, or fed directly to the receiver antenna input.

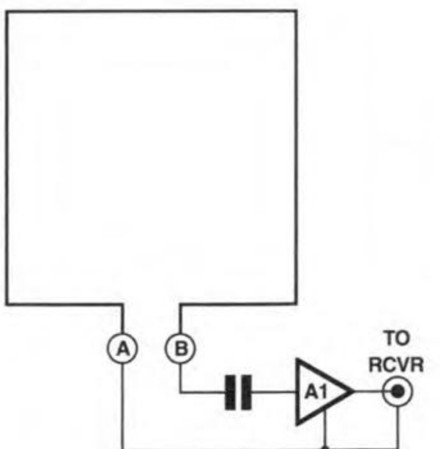


Fig. 11. Using a single-ended preamplifier at the output of the loop.

Figures 11 and 12 show the use of preamplifiers connected to the output of the loop antenna, or a coupling loop. The connection in **Fig. 11** is an amplifier with a single-ended input. One side of the loop output is grounded, and the other is fed to the input of the amplifier. The circuit in Fig. 12 shows the use of a push-pull preamplifier. The outputs of the two active elements are combined in an output transformer; the secondary of the output transformer is connected to either additional amplification or directly to the receiver antenna terminals via coaxial cable.

Next month...

Now that we've learned a bit of the theory about loop antennas, it is time to put it into practice and discuss how one goes about actually making a loop antenna. Next month we will look at practical construction details, loop preamplifier circuits and loop applications that might not immediately spring to mind.

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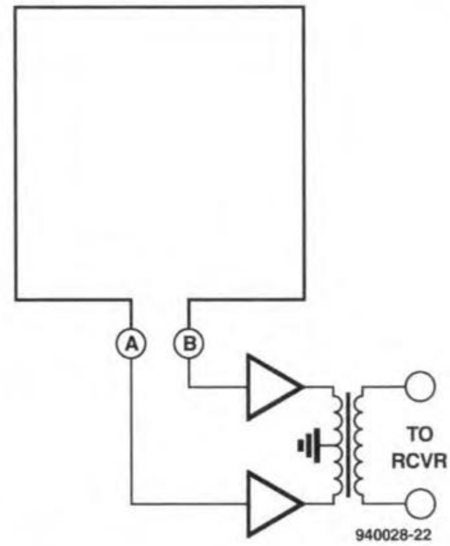

Fig. 12. Using a push-pull preamplifier at the output of the loop.

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NEW BOOKS

HOME SECURITY
By Vivian Capel
ISBN 0 7506 1809 4
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Being burgled is traumatic; it is a personal assault. Yet, in Britain, as in most other European countries, a house is broken into every 10–20 seconds every day, seven days a week, year in, year out. What's more, the rate is increasing, the police are powerless. This, and its direct consequence: higher insurance premiums, is a compelling reason to make our homes more secure. This book provides practical, independent guidance on how to do this. It shows how burglars work and how to thwart them. In a jargon-free way, the selection and installation of alarm system are described in this practical guide for home-owners.

There is a plethora of security devices now on offer. Large sums can be spent, yet weak links may be left unrecognized – except by the burglar. Good independent advice is scarce, as most security firms favour their own products. This book explains the pros and cons of alarm system, including how to avoid faults and find them if they occur. 'Friendly security' is particularly featured, and other security devices are given a critical scrutiny. Further measures to avoid becoming a victim of crime are also discussed.

Descriptions and circuits are given for two simple tested designs, one for the home and one for a foolproof public hall system, for those with some technical and constructional skills.

With recent estimates suggesting that the market for security is growing at 12% a year, this is a timely and straightforward guide to the wide range of technologies and devices used to maintain security in homes and small premises.

This **Butterworth Heinemann** publication is available from **Reed Book Services, PO Box 5, Rushden NN10 9YX**.

ELECTRONIC MUSIC and MIDI PROJECTS
By R.A. Penfold
ISBN 1 870775 24 4
Price £ 9·95 (soft cover)
Home made equipment has been part of the electronic music scene for just about as long as there has been electronic music. In the early days, there was often no alternative to using home constructed equipment since ready made alternative were conspicuous by their absence. Also, the commercial electronic music gear that was available tended to be quite expensive. Today's ready-made electronic instruments are much more affordable, but in many cases, it is still possible to produce home constructed units for significantly less than the cost of broadly comparable ready-made alternatives. It is also still possible to build effects units that have no true commercial counterparts.

Whether you wish to save money, boldly go where no musician has gone before, rekindle the pioneering spirit, or simply have fun building some electronic music gadgets, the designs featured in this book should suit your needs. The projects are all easy to build, and some are so simple that even complete beginners at electronic project construction can tackle them with ease. The basic mixer, MIDI tester, MIDI lead tester, metronome, electronic swell pedal, THRU box, and MIDI automatic switcher are all well suited to beginners. Stripboard layouts are provided for every project, together with a wiring diagram. The mechanical side of construction has largely been left to individual constructors to sort out, simply because the vast majority of project builders prefer to do their own thing in this respect.

None of the designs requires the use of any test equipment in order to get them set up properly. Where any setting is required, the procedures are very straightforward, and they are described in detail.

The projects in this book are primarily aimed at keyboard players. Guitarists are catered for in a separate book, *Electronic Projects for the Guitar*, from the same publisher and author as this publication.

PC Publishing, 4 Brook Street, Tonbridge, England TN9 2PJ.

IN-CIRCUIT TESTING
By Allen Buckroyd
ISBN 0 7506 0930 3
Price £ 35·00 (hard cover)
This book is aimed at potential purchasers and users of in-circuit automatic testers who are attracted to the concept of ICT, but who may need help. This includes test engin-eering managers who need guidance on which equipment to buy for a given application (and how to financially justify the purchase), and ATE programmers, test engineers and technicians who would welcome practical advice on how best to use the chosen ATE.

The work, by Allen Buckroyd, project manager at GEC-Marconi Communications and author of *Computer Integrated Testing*, is the key introduction to ICT for ATE staff and test engineers.

The book gives a description of what ICT can and can not do, and answers many questions on how tests are actually carried out, with benefits and drawbacks of the techniques.

This **Butterworth Heinemann** book is available from **Reed Book Services, PO Box 5, Rushden NN10 9YX**.

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This book, dealing with power amplifiers and preamplifiers for monaural and stereophonic reproduction from microphone, radio, tape and pick-up signals, was first published by Mullard Ltd, a wholly owned subsidiary of Philips Electronics NV, in 1959. Subsequently, in 1988, Mullard changed its name to Philips Components Limited. It is now reprinted by

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LETTERS

Dear Editor—I have read with great interest the 8051/8032 assembler course written by Dr. Ohsmann. I have constructed the BASIC computer and some extensions with success, thanks to the diskette bought from you. I have also constructed two applications for automatic control of hydraulic presses. All the knowledge acquired is thanks to you.

I and my son are designers and makers of photocolorimeters for chemical analysis and we hope to make an instrument controlled by one microcontroller of the 8051 family. Unfortunately, we do not know how to calculate the logarithm of the measure in assembler language which is primordial in this equipment (the absorption of light, the perception of sound, and so on, follow a logarithmic law). The 8052 BASIC does the job, but the setup occupies too much space in the memory and, besides, the 8032 is much cheaper. We would therefore like to work in assembler language (which would also be useful for making a photometer, an audiometer and a deciblemeter, among others). Four decimal places would be sufficient for these purposes.

Can you ask the author to write the necessary routines in assembler? I am, of course, prepared to pay a reasonable price for this. Or can you suggest any books in which I may find the required information?
Lorenzo A. Andali, Rosario, Argentina.

This letter has been copied to the author for a direct reply, but other readers may be interested in replying to Mr. Andali's request. If so, write in the first instance to us.

[Editor]

Dear Editor—I have been a regular subscriber to *Elektor Electronics* for 13 years and have usually found something of direct interest to me in every issue, but because I do not own a PC and have little interest in computers, there have been times when I have hesitated about renewing my subscription. However, when I looked through the March issue, I was glad I had not cancelled my subscription at the end of last year, because I found much of interest in it. I have always made much of my own test gear (in the early days one had to), so designs such as the AF signal tracer would always appeal to me. I have built innumerable audio amplifiers (from 1936 onwards) so I always enjoy articles on these, though I should now never build an amplifier such as the 100 watt design. I skip Owen Bishop's mathematics because I no longer need them, but Joseph Carr's contributions are always helpful and practical.

I shall certainly make both the electronic fuse and the car battery monitor – two very interesting designs; the latter being both ingenious and challenging. I couldn't help wondering, though, why the designer of the electronic fuse used such highly rated semiconductors for a project where the maximum voltage is 25 V and the maximum current 1 A. The BD679 is rated at 4 A and TIC106 at 400 V, 5 A. I suppose the answer is that these components are inexpensive and readily available, and lower rated substitutes would make very little difference to the overall cost.

Any article on building electronic circuits is of interest to me, so the article on p. 62 was initially appealing – but as I had not previously encountered prototyping board, I looked for it in the catalogues and was appalled to find that the cheapest one (from Maplin) was priced at £ 8; the handful shown in your photograph would cost nearly £50. I suppose plating a board with a multitude of separate copper squares is inevitably expensive. However, this material is not, as your heading suggestes, the same as stripboard or Veroboard, which is only about a quarter of the price. I presume your design staff do not have to pay for the materials they use, so the cost would not worry them. The statement that "the tools and materials … will be found even in the poor man's workshop" evoked hollow laughter from me – and I have a very well equipped workshop. It does not run, however, to a Weller soldering station costing over £ 100 (if the RS catalogue is to be believed).

I do not like stripboard and I avoid using it wherever I can, but instead I use plain matrix board and wire up the components on the underside with 0.5 mm tinned copper wire, insulating it only where wires cross. This does involve making loops at the wire ends with a pair of fine round-nosed pliers, which is perhaps more tedious than the technique described in your article, and the method is best suited for simple circuits such as the phase-shift oscillator in the article. But I did use it for the 'first cuckoo in spring' (your April 1985 issue) and that contained six ICs as well as quite a few other components.

No criticisms I have made can alter my opinion that *Elektor Electronics* is a very well produced magazine with excellent coverage of the electronics scene, providing a range of projects that cater for the interests of most serious electronics enthusiasts. Long may it continue to do so.
H.F. Howard, Bristol

Thank you for your interesting letter (only partly reproduced here), which I have copied to various members of our staff. It is most heartening to read that the urge to make things electronic does not wane with advancing years.

Your question as to the rating of the semiconductors used in the electronic fuse has been passed to the designer for comment. I hope that his answer will find a space in one of our forthcoming letter columns.

I have also passed your comments on 'Building electronic circuits' to the relevant designers and intend to have their comments in a future letters column. I think it is not so much that the design staff do not have to pay for the materials they use as the fact that as professional designers they are prepared to spend more money on their own home workshop than many other enthusiasts. I suspect, however, that even they do not pay £100 for a soldering station – prices I have heard mention are about £30.

I hope that you may long continue to be able to read our magazine.

[Editor]

Readers' comments on these letters and on any other subjects published in *Elektor Electronics* are, as always, welcome.