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Simple RF signal generator

Active mini subwoofer

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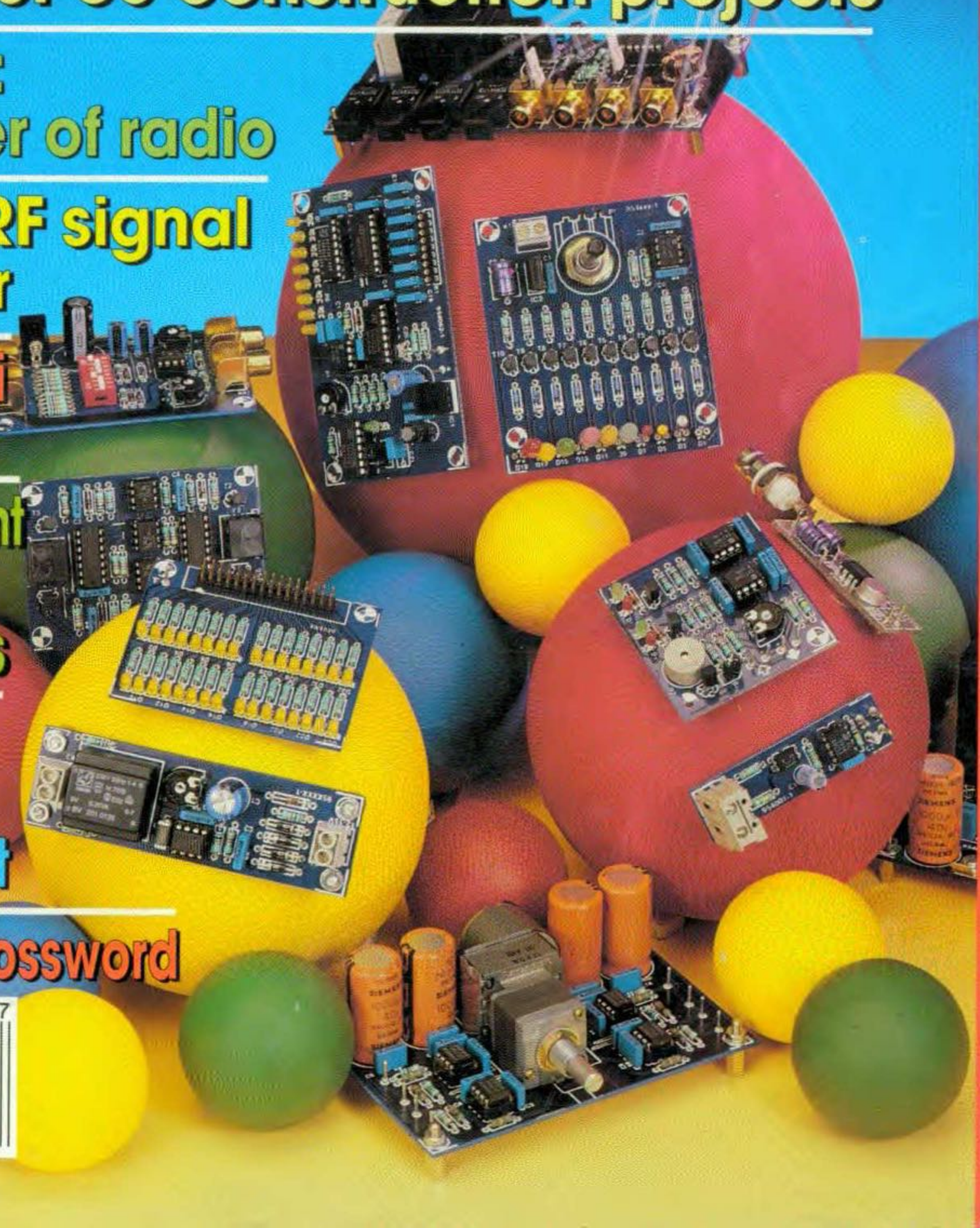
Summer crossword



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From the World of Electronics

THE INFORMATION SUPERHIGHWAY LEADS TO SALFORD

Nynex CableComms, one of the three largest cable operators in the United Kingdom, is teaming up with Salford's City Council and universities to make the area in northwest England a test-bed for exploring the interactive potential of broadband cable networks in the provision of local services.

They are not there yet, but the project is already helping to overturn some of the received wisdom where the first lanes on the information superhighway might lead in Britain.

The United States-owned cable operator – the licencsed cable provider for the region – has spent the past six months agreeing the terms of the experiment with local partners. These are: the City of Salford, the University of Salford, and University College Salford (the latter two will merge into a single university in the middle of 1996).

The three parties plan to collaborate in setting up local experimental broadband services in a range of public-sector settings. "We're not talking about the traditional video-on-demand type stuff," says John Edwards, Nynex's regional director in Salford. "The five areas are crime prevention, community service, health, education and training, and local business development."

The choice of sectors emerged from discussions with public servants from all over the city, held earlier this year. The local police force might sound like an unlikely pathfinder in the information revolution, but the acting Chief Superintendent of Police was at one of the sessions. He contributed several ways in which the force could use the network: everything from sending out high-quality identikit images to transmitting DNA records to other stations.

Pilot projects

It will be several years before some of the more ambitious ideas are translated into reality. However, the partners are hoping to start early pilot projects in most, if not all, of the five areas very soon. "There are a number of opportunities that I think could be exploited pretty readily," says John Willis, chief executive of the City of Salford. "For example, I think we could move fairly quickly towards using the network to supply information about local services more efficiently."

For their part, the two universities are planning to use the project to build on their established interest in developing commercial and educational applications for the new information technologies. "The Open University, as currently conceived, has taken current technology about as far as it's possible to go," says John Squires, principal of University College Salford. "What we're talking about here, among other things, is the possibility for a kind of Open University, Mark 2, that exploits the benefits of interactivity in an accessible and exciting way."

For the cable operator, which expects to invest around £20 million in the project, it is clearly an opportunity to foster goodwill among the local community. This can be in short supply for a company that spends most of its time digging up streets, as other cable operators in Britain have discovered. "The geographical nature of the franchise system means that the operators are very locally focused," says Niall Hickey, press spokesman for the Cable Communications Association. "They want to establish themselves as good corporate citizens."

Discounted prices

This has led several companies to build up links with local schools and other public bodies. In Croydon, South London, for instance, United Artists Communications, has now cabled up around 85% of the local secondary schools at heavily discounted prices. By and large, such projects have tended to look at how schools and others might take advantage of cable services that already exists.

But the commercial – and technological – bottom line to all the civic-worthiness is the need to foster demand for fibre-optic services which have yet to be established. The UK cable industry plans to invest around £10 billion by the end of the decade laying cable networks nationwide.

At present, cable providers can earn a return on this investment offering entertainment, a banned segment of the market for British Telecom, as well as standard telephone services. But the ban will be lifted some time during the next decade. By then, BT should be able to offer popular video-on-demand services over its existing telephone network. All the cable companies would like to find ways to earn a return from services which are only possible on their (higher-capacity) networks.

Comprehensive network

It will be a while before there is a comprehensive network nationwide, so the first experiments will have to be geographically confined. The logic of the Salford project, however, is that finding ways to let people interact, locally, is not merely a stopgap to them interacting with the world outside. Local public agencies could be a fertile source ofpotential fibre-optic applications in their own right, responsible for a wide range of information and activities to which the new technology could usefully be applied.

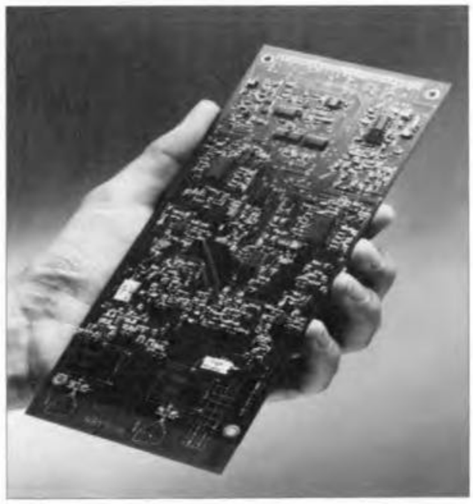
More broadly, the development of these and other applications must necessarily be demand-led, and people are more likely to think of new products to demand if they have already seen the network fill a local niche. "No-one is denying that broadband networks can allow you to be talking to the world," says Carl Grose, executive director of Nynex for the north-west. "But when it comes down to it, you're still talking about people. We're not going to make money offering facilities tat are somehow worldwide, but are actually anonymous at a local level."

Critics bemoan the lack of a 'bigh picture' in the government's piecemeal approach to re-cabling the United Kingdom for the interactive era, since it could mean that some parts of the country get left behind. Yet, the Salford projects indicates that, for all the worries about universal coverage, some individual communities could benefit from the rather limited horizons of the companies laying cable in their area. Indeed, Salford's first trip down the information superhighway may be just around the corner.

"SPREADING" RADIO NETWORKS

Low-cost spread-spectrum radio technology can provide an attractive alternative to existing hard-wired communications, including fibre-optics, for both point-to-point and local-area-network (LAN) applications.

Recent work done by the Instrumentation Group of the Central Research Laboratories Ltd (CRL) of Britain's Thorn EMI Group has resulted in the development of an innovative spread-spectrum system. Already, a major international computer manufacturer has licensed the technology so as to apply it to radio local-area networks (RLANS).



A 2.45 GHz prototype spread-spectrum radio transceiver

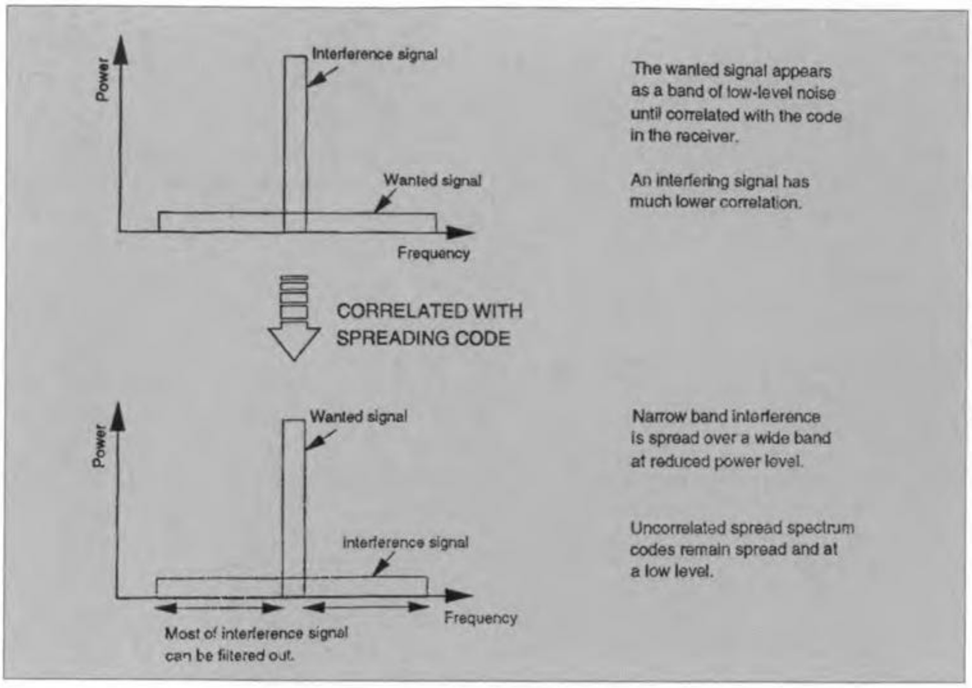


Diagram illustrating the spread-spectrum principle.

Operating at 2.45 GHz, the spread-spectrum radio transceiver (SSRT) provides data rates of up to1 Mbit/s at a cost expected to make the application of such a device viable in a wide range of situations. This data rate is more than adequate for most applications, including the secure transmission of graphics data.

There are many cases where high data rate is not the limiting factor, immunity from interference and ease of installation being of greater importance. Whereas existing UHF narrow-band radio systems used for remote sensing and process control are becoming increasingly prone to interference, especially in large towns and cities, spread-spectrum enables data to be transmitted and received with much more confidence and little or no extra cost.

The technical advantages of spread-spectrum technology have been recognized for many years, especially by the military. The benefits include the ability to suppress interference and deliberate jamming and resistance to locating the source of transmission and to eavesdropping. While deliberate jamming may not be relevant to civilian users, freedom from interference certainly is.

Robust alternative

In the past, the technology involved was too complex and thus too expensive to be transferred to the civilian sector. However, the CRL developments will enable spread-spectrum to be used for both RLANs and point-to-point links in cost-sensitive applications. Overall, the latest technology provides a low-cost ad robust alternative to existing hard-wired communications and removes the need to install cabling, which is costly, time-consuming and inflexible. In addition, the elimination of cabling obviates a major source of overall system unreliability.

As CRL has taken into account the existing and proposed frequency allocations in many regions and individual countries around the world, the outcome has a truly international scope. In particular, operating in the 2.45-GHz band, the laboratories' system satisfies a wide range of allocations, including CFR (USA) and ETS300328 which will supersde existing national standards throughout the European Union.

With spread spectrum, the wanted signal is combined with a spreading code to produce a pseudo-random spread signal which is used to control the transmitter. When received, the resultant signal appears as a band of low-level noise except to a receiver that uses the same spreading code with which it is synchronized. This receiver is thus able to extract the wanted signal by filtering out the majority of the interfering background signals.

In the SSRT, a 31-bit spreading code is used which is long enough to maintain

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data security and yet minimize synchronization time. However, to obviate interference from similar nearby systems, the spreading code generator is programmable so that any one of 16 codes can be switch-selected. They have been chosen to minimize cross-correlation.

Potentially troublesome

The system operates at a clock rate of 16 MHz to generate direct-sequence spread-spectrum signals which are used to control the digital frequency synthesizer. Generating frequencies directly in the 2.45 GHz band avoids the production of potentially troublesome spurious frequencies (mixing products) as well as costs being reduced as no conversion stages or filters are required.

In the initial CRL implementation, the entire transceiver is built on to a single surface-mount printed-circuit board measuring about 220×100 mm. This includes control electronics, frequency synthesizers together with one transmitter and dual receivers, and 'printed' antennas. Should OEMs require external antennas, coaxial connectors may be fitted.

A transmit/receive RF switch selects one antenna, while the second antenna and duplicate receiver channel provide a spatial diversity capability. A diversity switch continuously monitors signals strengths to select the stronger signal from the appropriate antenna/receiver channel. A data recovery section, linked to the spreading code generator in the control section, converts the analogue receiver output to digital levelsand monitors the signal to detect the spread-spectrum synchronization and tracking signal levels.

Maximum flexibility

Fast hardware synchronization provides initial synchronization in 512 microseconds with tracking then being provided under processor control for the duration of the received data packet. The data packet format, error-correction coding, and network protocols, are under the control of the OEM's equipment for maximum flexibility in application.

This is not a case of the CRL designers abrogating responsibility. In practice, SSRTs will often need to operate in conjunction with an existing system where the data protocols have already been established.

Consequently, when a master transmits, all slave stations will receive the data and present it to the user's interface. The user must then determine the destination of the data packet by inspection of the user's address within the data packet. Similarly, when a slave transmits, the data pacxket is expected to be received by the master, while the other slaves receive the same data, but determine that the data packet is not intended for them. This is achieved by inspection of the user's address data within the data packet.

The range of operation is estimated at 2.5 km in line-of-sight and 50 m within a typical office building. These estimates are based on a 400 kbit/s data rate and dipole antennas, and take into account factors such as fading margin, detector signmal/noise rationand permissible transmitter power.

Competitive equipment

The SSRT will be prices at around £300 to £350, which makes the equipment competitive for a wide range of applications in both technical and com mercial environment. However, the key issue is that the availability of a standard commercial product will enable the technology to be adopted at an early stage.

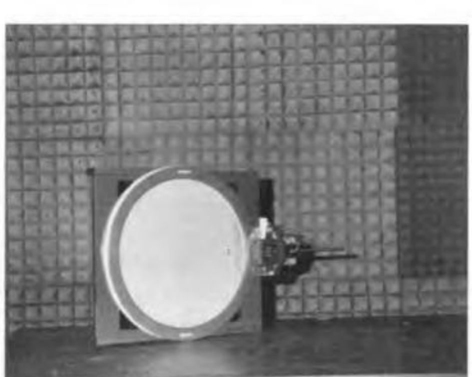
For example, in the retail sector, supermarkets require immediate access to point of sale data. At the same time, they want to minimize cabling to the check-out positions so as to have a flxible store layout. In addition, data and barcode readers for stock control are being used increasingly. In view of the vast amount of lighting and referigeration equipment which can generate electrical noise, stores have always been a difficult environment. Another application is in remote telemetry where it is often necessary to monitor large numbers of outstations. These are just two of the many applications which will emerge for this innovative spread-spectrum technology.

Central Research Laboratories Ltd, Dawley Road, Hayes, England UB3 1HH.

MAKING TUMOURS VISIBLE AND AIRCRAFT INVISIBLE

Question: What can make aircraft invisible but cancer tumours visible?

Amnswer: RFSS, that is, Reconfigurable Frequency Selective Surfaces. RFSS is the brainchild of a team led by Dr John Vardaxoglu at Loughborough University's Department of Electronic and Electrical Engineering in Leicestershire, England. RFSS was first developed as a way to sharpen up a pilot's virw of the world via his radar screen. But is now finding, or expected to find, a wealth of other applications, including the diagnosis and, in the future, the treatment of cancer, and the provision of more channels for satellite broadcasting and telecommunications.



Demonstrator RFSS in the Loughborough laboratory

The secret of RFSS is a copper-coloured cone about the size and shape of a traffic cone. This is a specially-designed radome, one of the streamlined covers used to protect the radar antennas that protrude from the fronts of aircraft. The radar projects beams of microwaves around the aircraft, which are reflected back from other aircraft, targets on the ground or storms looming ahead, all of which show up on the radar screen. The radome protects the antennas agains damage from hail, rain or birdstrikes.

Electromagnetic windows

The radome is made of polyester covered with a thin layer of copper, which forms the frequency-selective surface. The thin layer of copper is covered with tiny circles, formed by removing rings of copper each about one centimetre in diameter, revealing the plastic beneath. These circles acts as electromagnetic windows, allowing only a very narrow band of radar wavelengths to pass through the radome in either direction. This screens out irrelevant wavelengths and prevents them from confusing the picture on the pilot's screen.

Radomes already produced at Loughborough are not reconfigurable: theyallow only band of wavelengths to pass. But the next generation, a prototype of which is now working, will allow the size if the circles to be varied so that different frequencies can be selected and allowed to pass.

Patented technology

This is done by making it possible to alter the diameter of the rings while the radar is in use, thus altering the frequencies that are selected. The technology which makes this possible has been patented, and is now under further development in the Antennas and Microwaves Laboratory at Loughborough.

One use for this could be in advanced military aircraft with 'smart' RFS surfaces. A smart skin made of a material which would absorb or reflect any chosen wavelengths could be tuned to absorb rather than to reflect enemy radar. RFSS may soon be involved in battles of wits with smart radar, in which wavelengths are constantly altered so as to image aircraft with smart skins.

A very different use is being developed with the radiology department of the Leicester Royal Infirmary in Leicester near Loughborough. The aim is to use RFSS as a back-up scanning system to detect deep cancer tumours at an early stage of development in patients who are being scanned with the use of Magnetic Resonance Imaging—MRI.

In MRI, microwaves beamed into a patient's body in a very powerful magnetic field stimulate the emission of radiation that is characteristic of the type of atom which produces it. This radiation is detected and processed in a computer to produce detailed images of soft as well as hard structures, including tumours.

Tissue mass

However, there is often uncertainty as to whether a mass of tissue deep in the body is a malignant tumour or something else. If some technique were available that would help to decide whether such structures were tumours or harmless, normal structures, actually during the MRI scan, this would enable doctors to provide the best possible treatment at the earliest possibe stage.

The Loughborough team are developing detectors equipped with RFSS which can be tuned to pick up only the precise range of frequencies of infrared or other radiation produced by tumours, which as a consequence of their high rate of metabolism act as local 'hot spots' in the body. They give off tell-tale radiation at wavelengths not produced, or produced at much lower levels, by normal tissues.

In a pilot study, this technique has been used alongside MRI. The aim is to use it to confirm during an MRI scan whether or not masses of tissue imaged by MRI are malignant tumours. The next step may be to use microwaves from a transmitter fitted with RFSS to treat the cancer. Selected wavelengths can be focused on tumours to literally cook them to death. This might even be possible during the MRI scan itself.

Commercially important

It is possible that the most commercially important of the many likely applications for RFSS will be in making it feasible to provide more channels for broadcasting and telecommunications, in the first place via satellites. Broadcasting antennas covered with RFSS on satellites could be made effectively invisible to one set of broadcasting frequencies used by other antennas on the satellite, while themselves broadcasting in the same direction using different frequencies.

In this way, several different broadcasters can beam their broadcasts in the same directions. This could allow a network of overlapping radio beams to be used to supplement or replace cable for local telephony and radio networks. Prototypes at Loughborough are already being used to develop this concept.

The development of RFSS at Loughborough is being sponsored by the British Government through the Physical Sciences and REngineering Research Council and by British Aerospace.

Dr John Vardaxoglou, Department of Electronic and Electrical Engineering, Loughborough University of Technology, Loughborough, Leicestershire, England LE11 3TU.

UK MAINS SUPPLY

In spite of misgivings in certain quarters regarding the 'reduction' of the mains voltage in the United Kingdom, the Institution of Electrical Engineers advises as follows. As from 1 January, 1995, the single-phase supply voltage in the UK has been reduced in statutory legislation from 240 V ±6% to 230 V +10%, –6% in order to move towards an alignment with Europe. European countries have moved their nominal voltage up from 220 V to 230 V as from 1 January, 1995.

This is a 'paperwork exercise' and the actual voltage provided at customers' terminals by a UK supply authority will remain at 240 V for many years to come – there is no agreement to change further as 240 V is within the agreed tolerance band. Consequently, there will be no difference in the operation of domestic consumers' appliances.

HOT BIRD 1 IN ORBIT

EUTELSAT's new television satellite, Hot Bird1, was launched in late March and is now in orbit at 13° East where it is co-located with EUTELSAT II-F1.

Hot Bird 1's 16 transponders have gone into service in the second half of April. The satellite is dedicated entirely to television and radio channels for reception by cable, community and direct-to-home antennas throughout Europe and the Mediterranean Basin and is fully optimized for either analogue or digital transmissions.

European Telecommunications Satellite Organization, Tour Maine-Montparnasse, 33, Avenue de Maine, BP 19, 75755 Paris Cedex, France.

NEW BROADCAST STANDARDS AND SYSTEMS

The Institution of Electrical Engineers, IEE, is to hold its Seventh Residential Course on 'New Broadcasts and Systems' at the University of Durham from 3rd to 7th July, 1995.

The course will provide delegates with an in-depth understanding of the new broadcast standards, the principles behind them and the background to their development.

It is aimed at engineers already working in the broadcast industry, whether in equipment design, manufacture and installation, or in programme production and transmission.

Topics, which will be covered by acknowledged experts in the field, will range from digital audio and video fundamentals through current studio and systems practice, to the use of low bit rate systems such asMPEG and DVB for the delivery systems of the future.

Further details from *Janet McCready, Electronics Division, IEE, Savoy Place, London WC2R 0BL. Telephone 0171 344 5421. Fax 0171 497 3633.*

38TH LONGLEAT AMATEUR RADIO RALLY

The 38th Longleat radio rally will be held on Sunday 25 June, 1995, from 09:30 to 17:00. Participants should follow the brown tourist signs for "Longleat House" (not the Safari Park) from the A36. There will be 150 trade stands and 20 club stands. Parking is free, and there are catering and bar facilities on site as well as camping and caravanning facilities. There is also a RSGB bookstall and membership services stand.

Further information from *Gordon Lindsay, G0KGL, 66 Jubilee Crescent, Mangotsfield, Bristol BS17 3AZ. Telephone/fax/answerphone 0117 940 2950.*

RESULTS OF IEEIE SALARY SURVEY

Results of the Salary Survey conducted last January among members of The Institution of Electroncis and Electrical Incorporated Engineers (IEEIE) show that highest paid Fellows and Members now earn £45,300 per annum, and more than 55% of Fellows and Members earn over £25,000 per year with median earnings at £ 25,600 against £ 24,000 in 1993. Over the two years since the last survey, Graduate Members have moved ahead markedly from £ 17,100 p.a. to £ 20,000 p.a. in the public sector and from £17,000 p.a. to £18,900 p.a. in the private sector.

Turning to academic qualifications, almost 25% of Fellows, Members and Graduates now have a Higher National Diploma (HND) or UK degree and over 27% of Technician Members and Associate Members have a Higher National Certificate (HNC).

Copies of the survey, which covers not only remuneration – categorized in age and level of responsibility – but also areas of employment, pension provisions, bonus payments and an analysis of benefits received by type of work are available from The Secretary, IEEIE, Savoy Hill House, Savoy Hill, London WC2R 0BS. Telephone 0171 836 3357.

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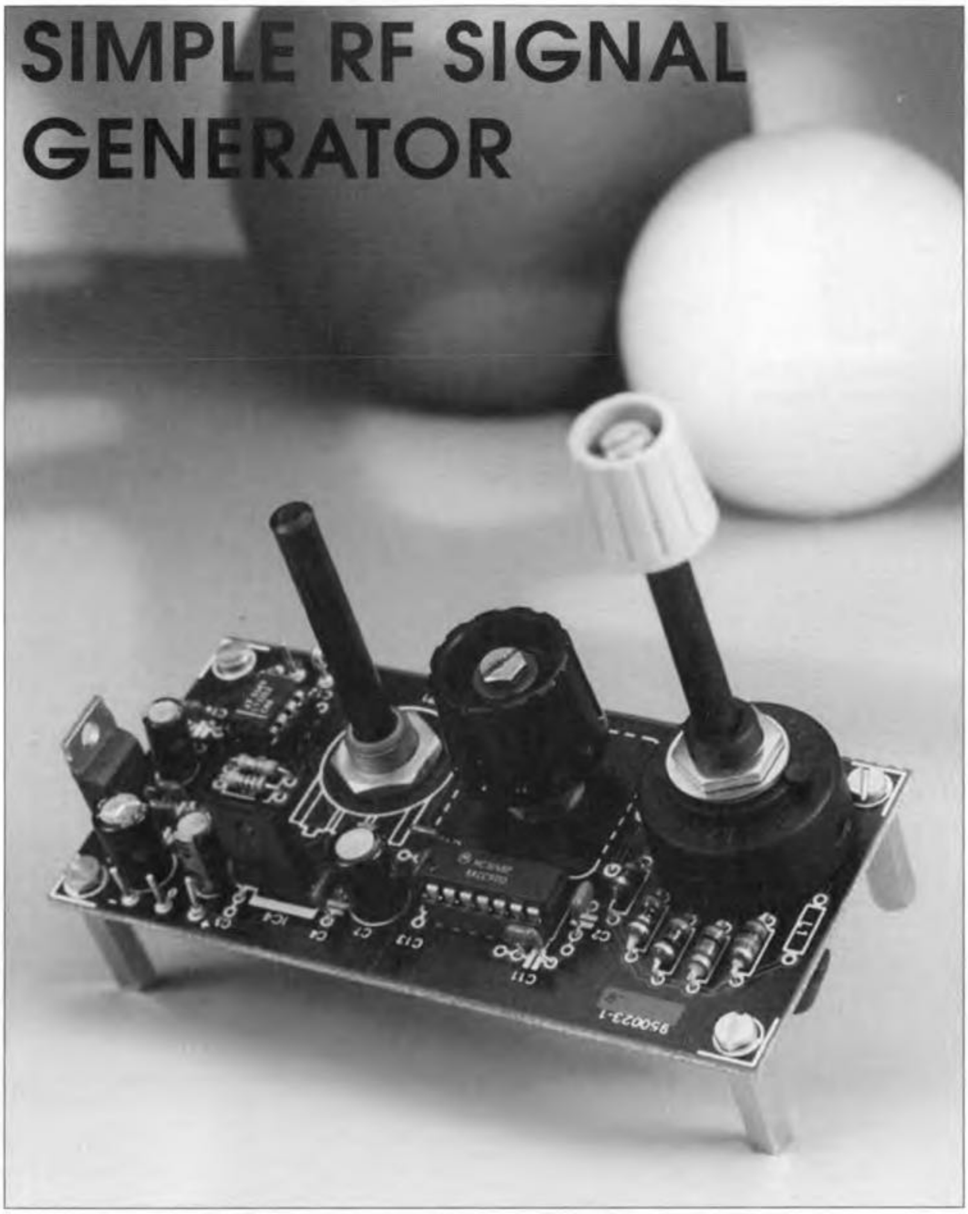
SIMPLE RF SIGNAL GENERATOR



This article proves that building a no-frills RF signal generator has nothing to do with sorcery as many d.c. oriented constructors seem to assume. The design is based on two ICs only, and uses ready-made inductors in the frequency determining sections.

Design by L. Lemmens

LET's face it: you can't test a radio or, for that matter, any piece of RF circuitry without an appropriate signal generator. Fortunately, it is not always necessary to know the generator's exact output signal level, as long as you are sure that you are on the right frequency, and that a reasonably strong signal is being injected into the circuit under test. This will bring nine out of ten circuit faults to light. An oscilloscope is then an invaluable tool, as the types used by hobbyists will usually work quite well up to about 20 MHz. Armed with the present RF signal generator and such an oscilloscope you are ready to tackle, say, an FM radio whose 10.7 MHz IF (intermediate frequency) section is at fault, or find the resonance frequency of an inductor-capacitor combination, or ...

Circuit description

The signal generator is based on Motorola's MC1648 voltage-controlled oscillator (VCO) IC. Here, the device is used as an *L-C* tuned oscillator rather than as the VCO in a PLL system for which it was originally designed by the manufacturer. The internal circuit diagram of the MC1648 is shown in **Fig. 1**. A low-distortion oscillator is created by connecting an external inductor-capacitor combination (parallel tuned circuit) between the TANK and Bias Pt. terminals. The oscillator incorporates positive feedback by coupling the base of transistor Q_6 to the collector of Q_7. An automatic gain control (AGC) is incorporated to limit the current through the emitter-coupled pair of transistors (Q_6 and Q_7), and allow optimum frequency response of the oscillator. Transistors Q_9 and Q_{11} provide the bias drive for the oscillator and output buffer Q_2-Q_3. By setting a suitable bias level for the buffer, the MC1648 can be made to supply square wave output signals, a feature which is not used here.

A glance at the circuit diagram, **Fig. 2**, reveals that the internal buffer of the MC1648 is not used at all. The oscillator signal is taken directly from the TANK terminal, and fed to opamp IC_2 via coupling capacitor C_{15} and level control P_1. The reason for using an external buffer opamp instead of the buffer transistors contained in the MC1468 is (1) that the LT1252 gives much better performance and (2) allows 50-Ω loads to be driven directly. Also, the AGC of the MC1468 can then be set for pure sinewave operation. That is achieved by connecting the AGC pin to ground via a capacitor, C_{11}.

The oscillator frequency is determined by the position of rotary switch S_1 and the setting of tuning capacitor C_1. The generator has five ranges:

1	200 kHz to 800 kHz
2	660 KHz to 2.7 MHz
3	2.1 MHz to 8.1 MHz
4	6.5 MHz to 29 MHz
5	18 MHz to 56 MHz

All frequencies are approximate, depending on the tolerance and *Q* (quality) factor of the inductors used (L_1 through L_5).

The maximum output level in ranges 1, 2 and 3 is of the order of 1.4 V_{pp}. In range 4, the maximum output level drops from about 1.3 V_{pp} at the lowest frequency to about 600 mV_{pp} at 29 MHz. Similarly, the level in range 5 drops from about 1 V_{pp} at 18 MHz to 300 mV_{pp} at 56 MHz. All levels were measured with a 50-Ω load connected to the output. Although range 5 may extend up to about 90 MHz, the output level will not be stable above 60 MHz or so, and there is a fair risk of the oscillator quitting suddenly at these frequencies.

The LT1252 is a current feedback opamp from Linear Technology which is marked by a fairly constant signal

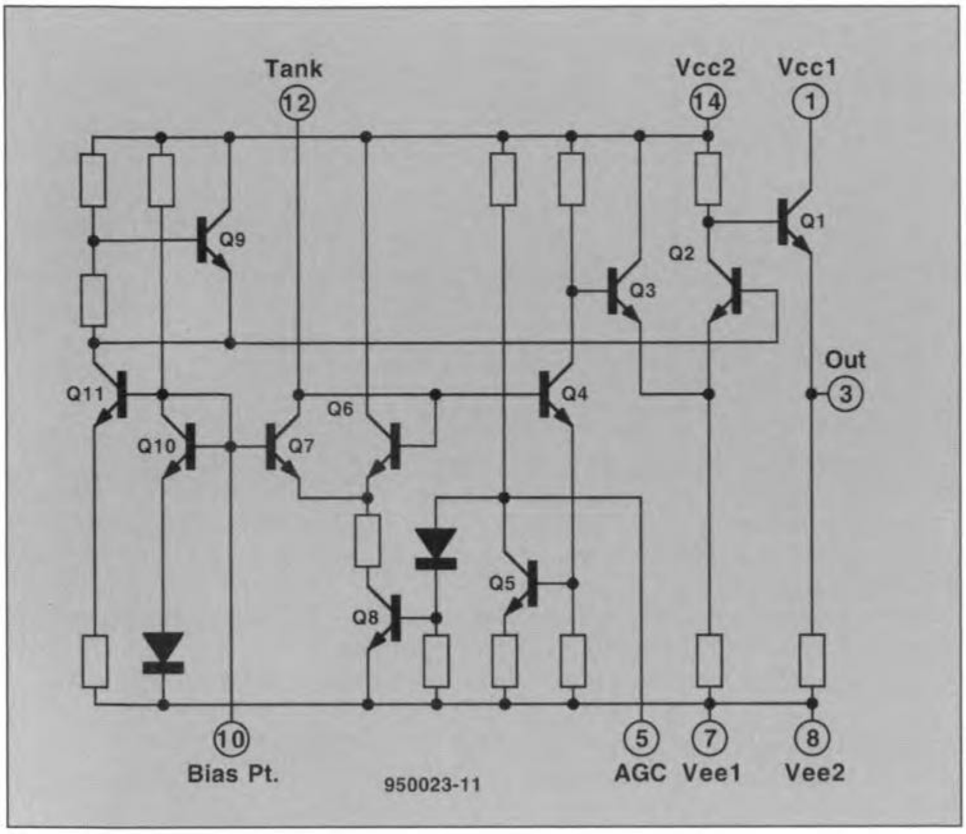


Fig. 1. Internal schematic of Motorola's MC1648 voltage-controlled oscillator (VCO).

bandwidth at all gain settings. The device is actually a low cost amplifier for video applications. The linearity of the LT1252 is outstanding. As with all current feedback amplifiers, the impedance at the inverting (negative) input determines the bandwidth. For maximum stability of the amplifier, this impedance should be resistive, not capacitive. Here, the gain of IC_2 is about ×8 (actually, $1+(R_2/R_3)$). Resistor R_1 matches the output of the opamp to 50-Ω loads as usual in RF technology.

The symmetrical power supply is entirely conventional, and based on the familiar 7805/7905 three-pin voltage regulators. The unregulated input voltage to the supply should remain below ±15 V to avoid high dissipation in the voltage regulators.

Construction

The RF signal generator is best built on the printed circuit board shown in **Fig. 3**. For an RF project, construction is remarkably simple. The tuning capacitor, C_1, and the RF output level control, P_1, are **mounted at the solder side of the board**. The spindles of these parts, however, are at the component side of the board. While the potentiometer terminals may be soldered directly to the relevant copper spots at the solder side of the board, the terminals of the tuning capacitor are connected via two pieces of component wire with a length of approximately 12 mm. The rest of the construction is all plain sailing. Although not strictly necessary, sockets may be used for the two ICs. Switch S_1 is a 12-way PCB mount rotary switch which is limited to five positions with the 'stop' ring provided.

COMPONENTS LIST

Resistors:

R_1 = 47Ω
R_2 = 680Ω
R_3 = 100Ω
P_1 = 22kΩ linear potentiometer

Capacitors:

C_1 = 500pF tuning capacitor (Conrad or Maplin)
C_2-C_6;C_{11};C_{15} = 100nF
C_7;C_9 = 100µF 25V radial
C_8;C_{10} = 47µF 16V radial
C_{12};C_{13};C_{14} = 47nF

Inductors:

L_1 = 1mH
L_2 = 100µH
L_3 = 10µH
L_5 = 100nH

Semiconductors:

IC_1 = MC1648P (Motorola)
IC_2 = LT1252 (Linear Technology)*
IC_3 = 7905
IC_4 = 7805

Miscellaneous:

K_1 = BNC socket.
S_1 = 12-way 1-pole rotary switch, PCB mount.
Printed circuit board, order code 950023-1, see page 70.

* MicroCall (01840) 261678

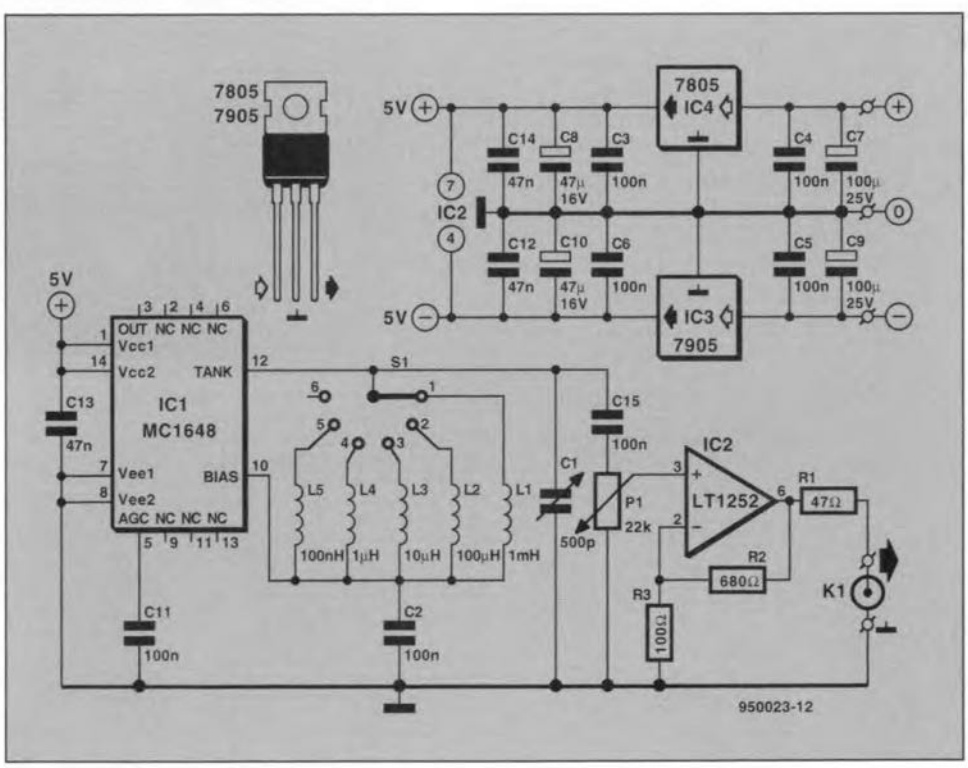


Fig. 2. Circuit diagram of the simple RF generator. Note that the RF signal is taken directly from the MC1648's TANK pin, rather than from the OUT pin. A current feedback opamp is used to buffer the RF signal, and step it down to 50-Ω impedance.

The voltage regulators do not require heatsinks. The RF output pins on the board are connected to the RF output socket via a short length of 50-Ω coax cable (RG58/U or RG174/U).

The completed board may be fitted into any reasonably sized all metal enclosure, such that the spindles of S_1, C_1 and P_1 are accessible from the front panel. In nearly all cases, the spindle of C_1 will have to be extended. The power supply is best incorporated into the case.

Calibration

The tuning scale of the signal generator is best calibrated with the aid of a frequency meter or a general coverage receiver. The output level control may also be fitted with a 'mV$_{pp}$' scale made with the aid of an oscilloscope.

(950023)

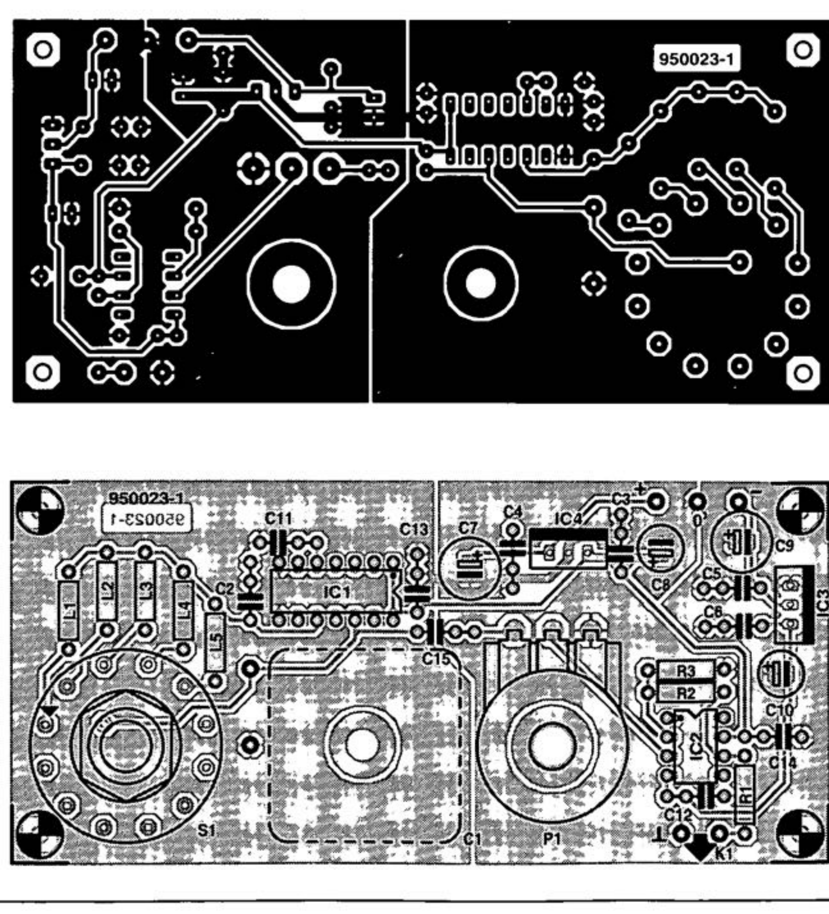


Fig. 3. Track layout and component mounting plan of the single-sided printed circuit board for the RF signal generator (board available ready-made, see page 70).

ACTIVE MINI SUBWOOFER

Design by T. Giesberts

Most proprietary subwoofers require you to be well-off and strong: the first to be able to afford them and the second to be able to lift them. The active unit presentedhere does not cost the earth and can be moved readily even if you are not a trained weightlifter. Even so, its performance stands comparison with many a commercial unit.

A loudspeaker system that has become very popular over the past few years is a combination of a subwoofer and two medium/high speakers. Since human hearing is not (or hardly) directional as far as low frequencies are concerned, ONE subwoofer suffices in most cases and this may be situated below a settee or a table without its performance being affected. Since the medium/high frequency units are also fairly small, the entire speaker system is ideal where there is not much space or if you don't want your living room taken up by the spaker system.

The drive unit

Ihe drive unit used in the prototype is Type SPP-110/8 from Monacor. The data sheet of this unit gives cut-off frequencies (–3 dB points) of 50 Hz and 8000 Hz. The frequency characteristic is shown in **Fig. 1**.

A low cut-off frequency of 50 Hz is fine for good bass performance. However, this requires an enclosure with a (specified) volume of not less than 5 litres, which is what we do not want to use for a MINI subwoofer. The present design uses a volume of 2 litres, which shifts the low cut-off frequency to 120 Hz. To bring this back to about 50 Hz, a correction network with a frequency characteristic shown in **Fig. 2** is used. This characteristic, combined with that of a 2 litre enclosure gives an overall characteristic as shown in **Fig. 3**.

In order to raise the low frequencies, a 55 W power amplifier is provided. The amplifier also corrects the frequency sensitivity of the human ear. Human hearing is much less sensitive to low frequencies than to medium and high frequencies. Thus, more sound pressure (read: amplifier power) is needed to ensure that low frequencies sound as loud as other, higher, frequencies. This reduced sensitivity of the human ear to low frequencies is also taken into account in recording studios.

Circuit description

The circuit diagram of the electronics part of the active subwoofers is shown in **Fig. 4**.

The correction filter is formed by IC_{1d}, IC_{1c}, and IC_{1b}. It is sub-divided into two low-pass sections. The first, a third-order section consisting of IC_{1d} and IC_{1c}, has a fixed cut-off frequency. The second, based on IC_{1b}, has a cut-off frequency that can be varied with P_1. This arrangement makes it possible to vary the frequency characteristic (that is, the amount of bass) to a small extent: this is shown for the two extreme positions of the potentiometer in **Fig. 5**. The output of the corection filter is taken from pin 7 of IC_1 and applied to the power amplifier via C_8.

The power amplifier consists of an integrated input stage, IC_{1a}, driver stages T_2 and T_3, and output stages T_4 and T_5.

At the onset of a (sinusoidal) signal from IC_{1a}, the bases of both T_2 and T_3 areat zero volts. During the positive half of the signal, T_3 is off and T_2 begins to conduct when its base-emitter potential rises above +0.6 V. During the negative half, the roles are reversed: T_2 is off and T_3 begins to conduct when its base-emitter voltage drops below –0.6 V. The waveforms of the input signal and the potential developed across R_{21} are shown in **Fig. 6**. The kinks in the voltage across R_{21} show that signal is distorted to an appreciable extent.

This cross-over distortion is eliminated by the use of T_1. Since the degree to which this transistor conducts is determined by the setting of P_2, the transistor may be considered as a potentiometer—see **Fig. 7**. This potentiometer enables the bases of T_2 and T_3 to be biased with +0.6 V and –0.6 V respectively. This means that even in quiescent operation, that is, when there is no input signal

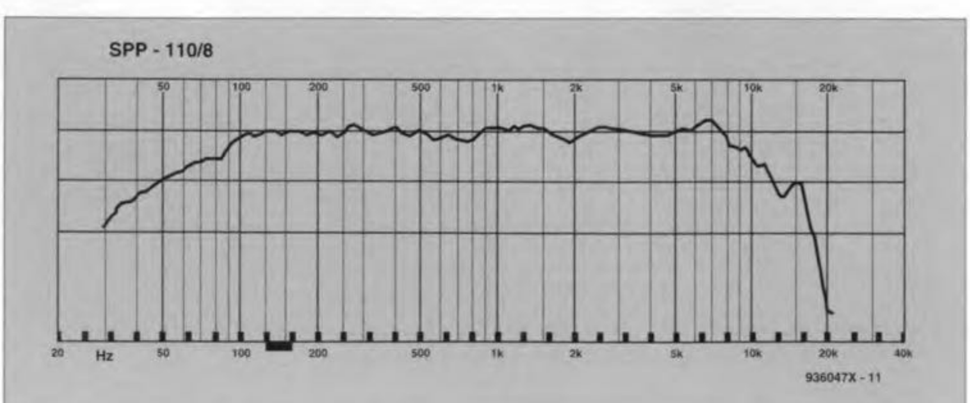


Fig. 1. Frequency characteristic of the drive unit.

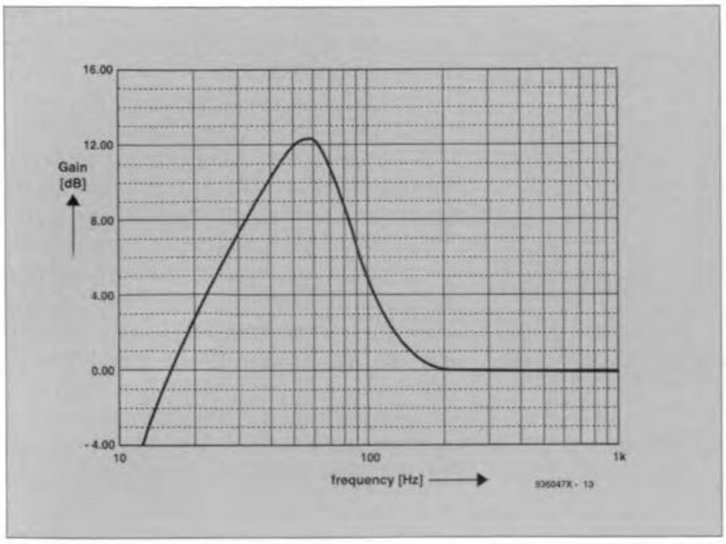


Fig. 2. Because the drive unit is fitted in too small an enclosure, the lower cut-off frequency shifts to about 120 Hz.

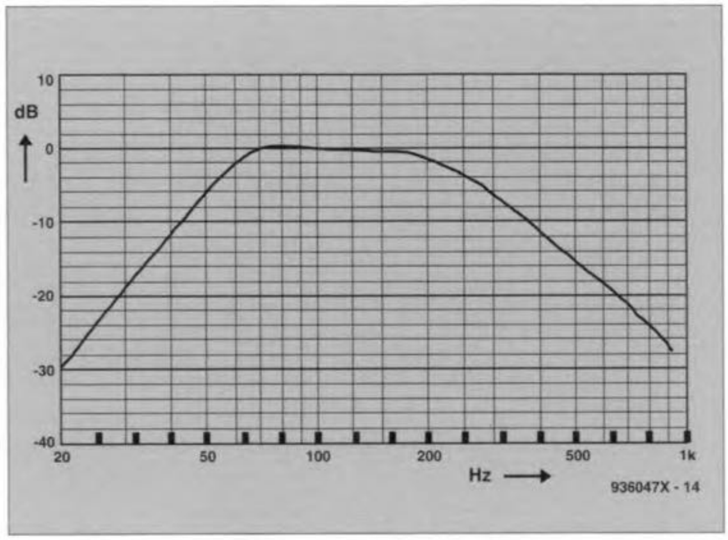


Fig. 3. The lower frequencies attenuated by the box are amplified by a correction filter.

(U_{in} = 0 V), a small current flows through T_2 and T_3. If then a signal is input, the relevant transistor will switch on immediately, since it was already on the verge of conduction. This arrangement ensures that the output signal shows no cross-over distortion.

Transistor T_1 is used instead of a potentiometer so as to compensate for the temperature dependence of T_2–T_5. Since T_1 is also affected by the ambient temperature, the transistor varies the bias to T_2 and T_3 in accordance with temperature variations. This in turn ensures that the quiescent current through the output transistors is kept constant.

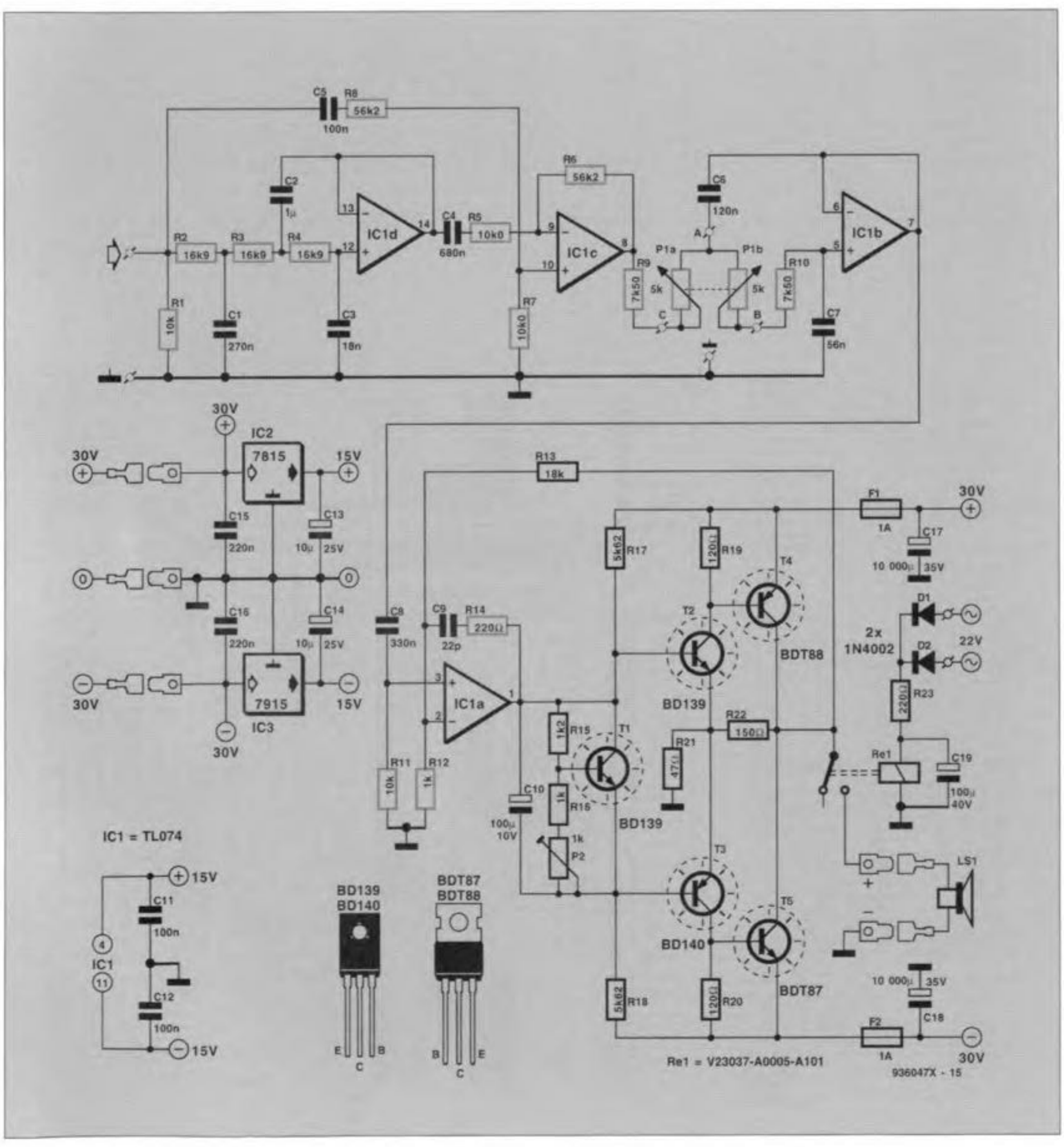


Fig. 4. Circuit diagram of the active filter and power amplifier.

The output of the power stages is applied to the loudspeaker via a relay. This relay is energized, after a short delay when the mains supply to the unit is switched on, and denergized immediately the supply is switched off. This prevents annoying clicks to be emitted by the speaker.

The power supply for the electronics is straightforward and traditional (see **Fig. 8**). The supply lines to the op amps must be stabilized and this is effected by regulators IC_2 and IC_3 in Fig. 4.

Building the circuit

The circuit is intended to be built on the printed-circuit board in **Fig. 9**. Populate the board in the traditional way, starting with the wire bridges and finishing with capacitors C_{17} and C_{18}.

Use solder lips for connections that carry large currents as shown in the close-up in **Fig. 10**.

It is also advisable to fit suitable solder tags to the (heavy-duty) wires to the power supply and loudspeaker.

The connections to P_1 must be in screened cable to prevent hum caused by stray fields. Solder the screen of this cable to the earth point on the board.

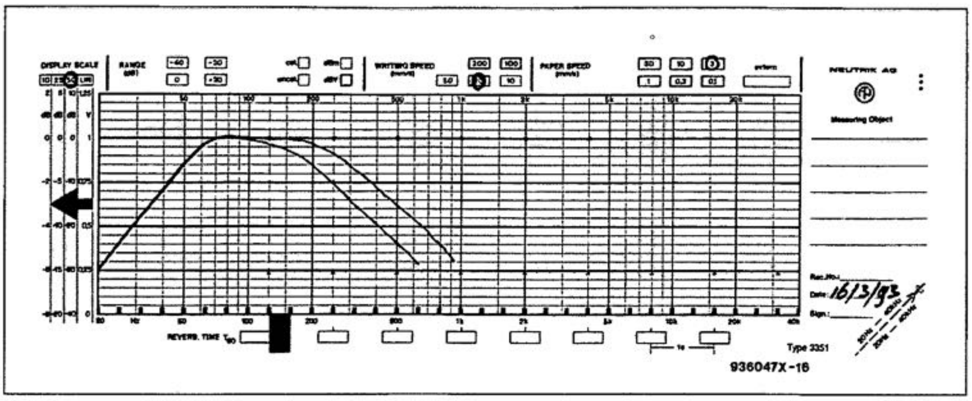


Fig. 5. Overall frequency characteristics for the two extreme positions of P_1.

Use screened cable also for the link from the input to the board and keep it as short as feasible.

When the board is finished (**Fig. 11**), it must be mounted on to the aluminium back cover of the enclosure as shown in **Fig. 11** and **Fig. 13**. The heat sink must be screwed on to this cover on the outside after a rectangular opening has been made in the cover through which the transistors protrude (since these must, of course, be screwed to the heat sink. Note that they must be insulated from the heat sink by means of ceramic washers and heat conducting paste). A wiring diagram for the back cover assembly is given in **Fig. 12**. The design is based on the use

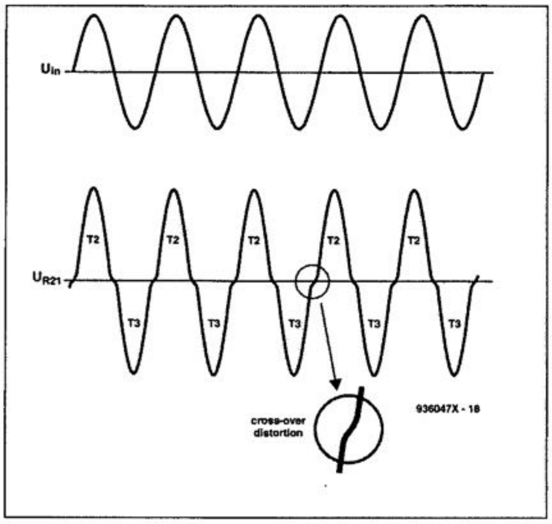


Fig. 6. Cross-over distortion is caused by the base-emitter voltage threshold having to exceed a certain value.

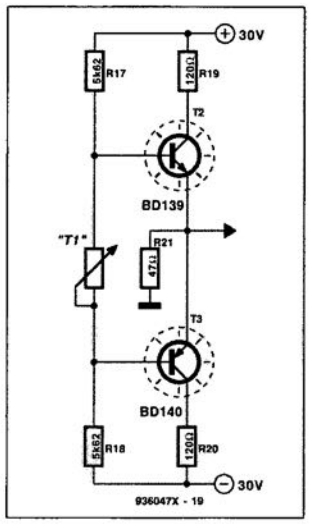


Fig. 7. For clarity's sake, T_1 is replaced by a potentiometer.

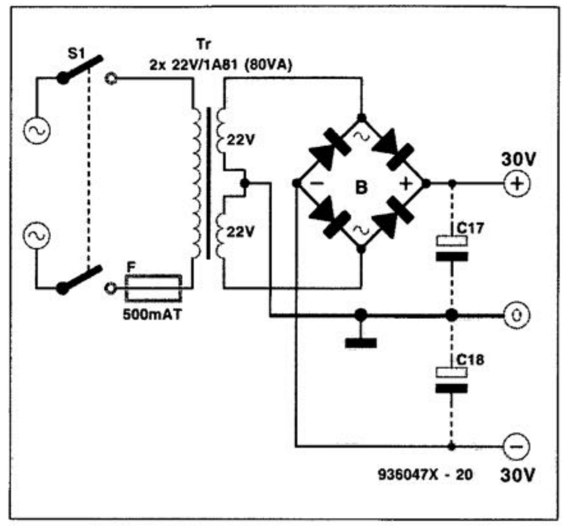


Fig. 8. The power supply must be built separately since the board does not allow for it.

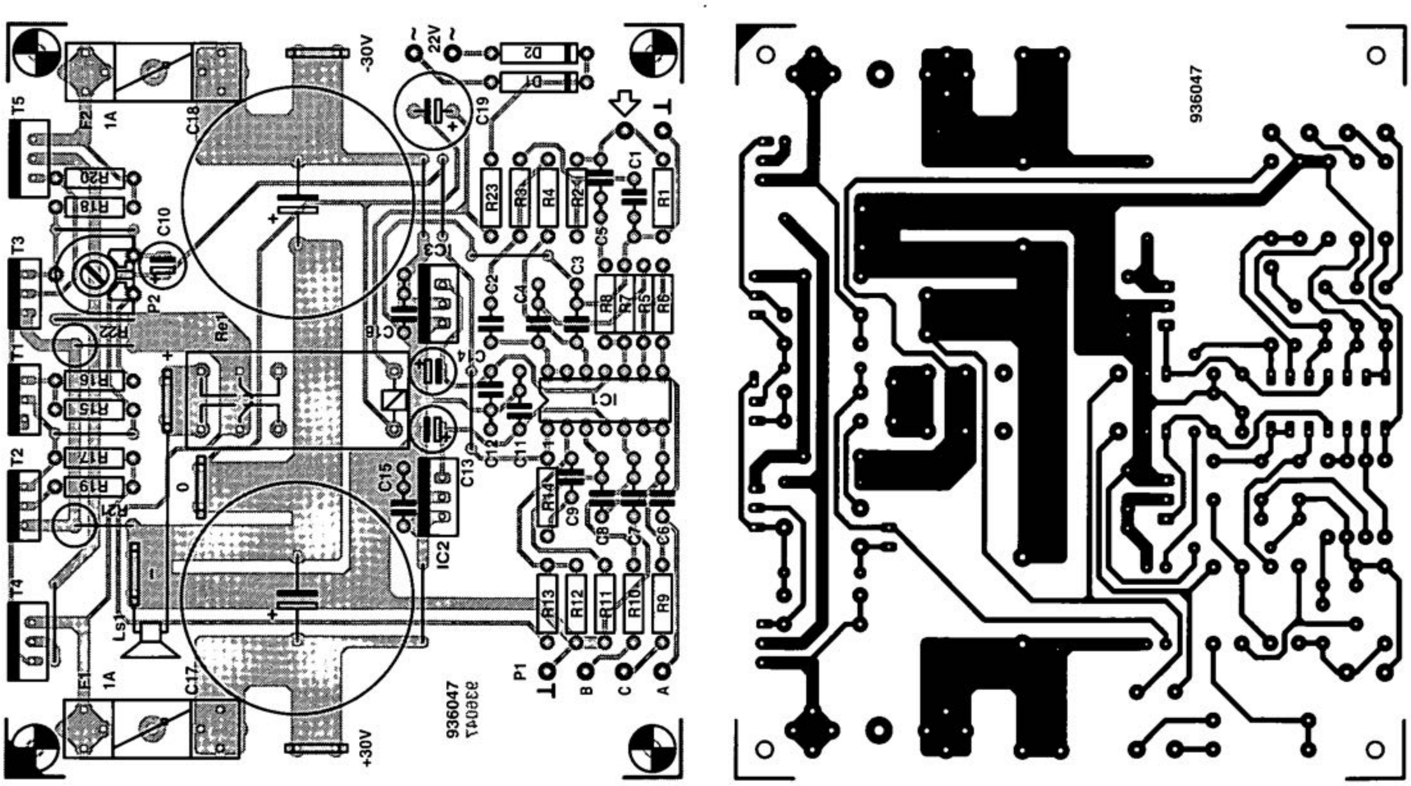


Fig. 9. Printed-circuit board for the active filter and power amplifier.

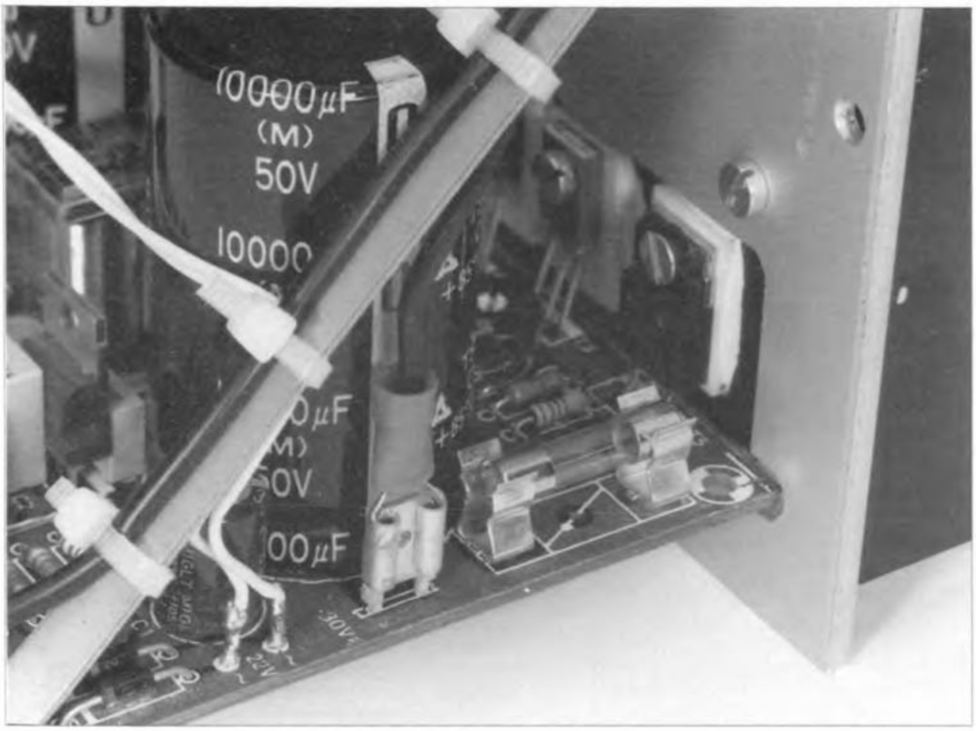


Fig. 10. Car-type flat connectors are ideal for use with large currents.

of a toroidal mains transformer, which has the advantage of producing only a weak stray field. It is fitted on to the bottom panel of the enclosure once this has been completed.

Enclosure

The prototype enclosure is made from 10 mm thick chipboard, which is strong and easy to work with. However, other types of wood or woodboard may be used. The cutting diagram for the various panels is shown in**Fig. 15**. The finished box, ready for varnishing or painting, is shown in **Fig. 16**. When the drive unit, finished amplifier and mains transformer have been fitted, fill the box with with suitable sound-damping material, fit the aluminium rear panel, and make sure that the box is made airtight (use a proprietary sealing compound).

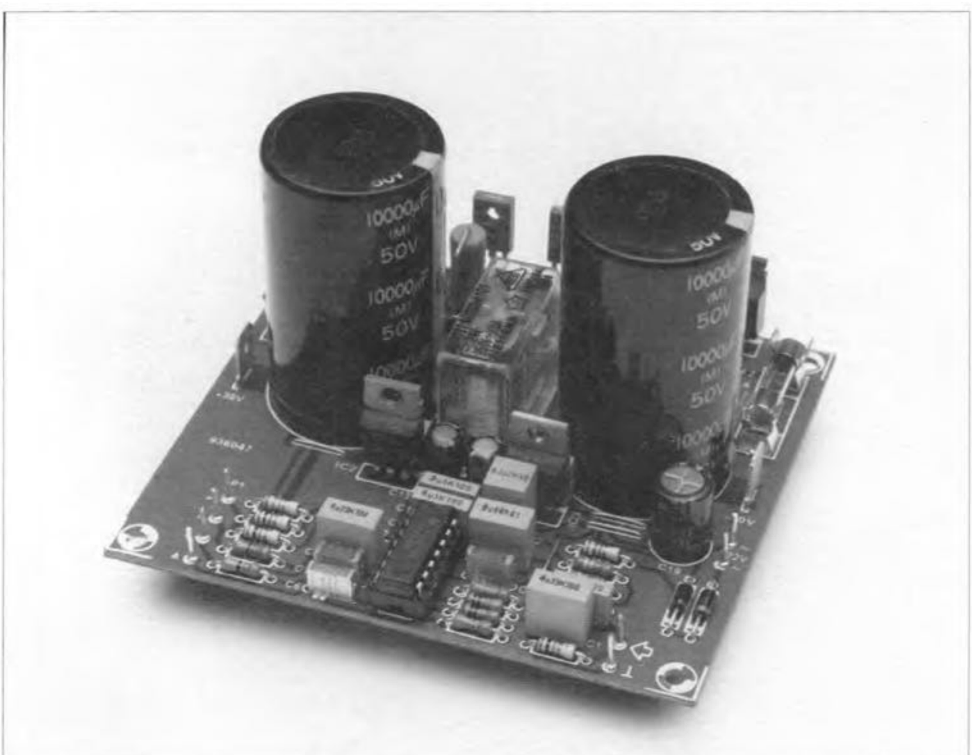


Fig. 11. Finished prototype board.

Setting up

The setting up consists merely of adjusting the quiescent current with P_1. This done by replacing fise F_1 with an ammeter, set to the mA range. Short-circuit the input of the amplifier and switch on the mains. Note the meter reading, say, x mA. Vary P_1 until the meter reads $(x + 5)$ mA.

Usage

The subwoofer can be connected to the left-hand or to the right-hand channel output of the power amplifier, since recordings are invariably made so that each channel has the same bass frequencies available. Anyone not entirely happy with this can use the auxiliary circuit in **Fig. 17**. The subwoofer must, of course, be supplemented by two medium/high frequency speakers as shown in **Fig. 18**.

Parts list

R_1, R_{11} = 10 kΩ
R_2–R_4 = 16.9 kΩ, 1%
R_5, R_7 = 10.0 kΩ, 1%
R_6, R_8 = 56.2 kΩ, 1%
R_9, R_{10} = 7.50 kΩ, 1%
R_{12}, R_{16} = 1 kΩ
R_{13} = 18 kΩ
R_{14}, R_{23} = 220 Ω
R_{15} = 1.2 kΩ
R_{17}, R_{18} = 5.62 kΩ, 1%
R_{19}, R_{20} = 120 Ω
R_{21} = 47 Ω, 5 W
R_{22} = 150 Ω, 5 W
P_1 = 5 kΩ, linear, stereo potentiometer
P_2 = 1 kΩ preset

Capacitors:
C_1 = 270 nF
C_2 = 1 µF, 63 V, polypropylene
C_3 = 18 nF
C_4 = 680 nF
C_5, C_{11}, C_{12} = 100 nF
C_6 = 120 nF
C_7 = 56 nF
C_8 = 330 nF
C_9 = 22 pF
C_{10} = 100 µF, 10 V, radial
C_{13}, C_{14} = 10 µF, 25 V, radial
C_{15}, C_{16} = 220 nF
C_{17}, C_{18} = 10,000 µF, 35 V, radial for board mounting
C_{19} = 100 µF, 40 V, radial

Semiconductors:
D_1, D_2 = 1N4002
T_1, T_2 = BD139
T_3 = BD140
T_4 = BDT88
T_5 = BDT87

Integrated circuits:
IC_1 = TL074
IC_2 = 7815
IC_3 = 7915

Miscellaneous:

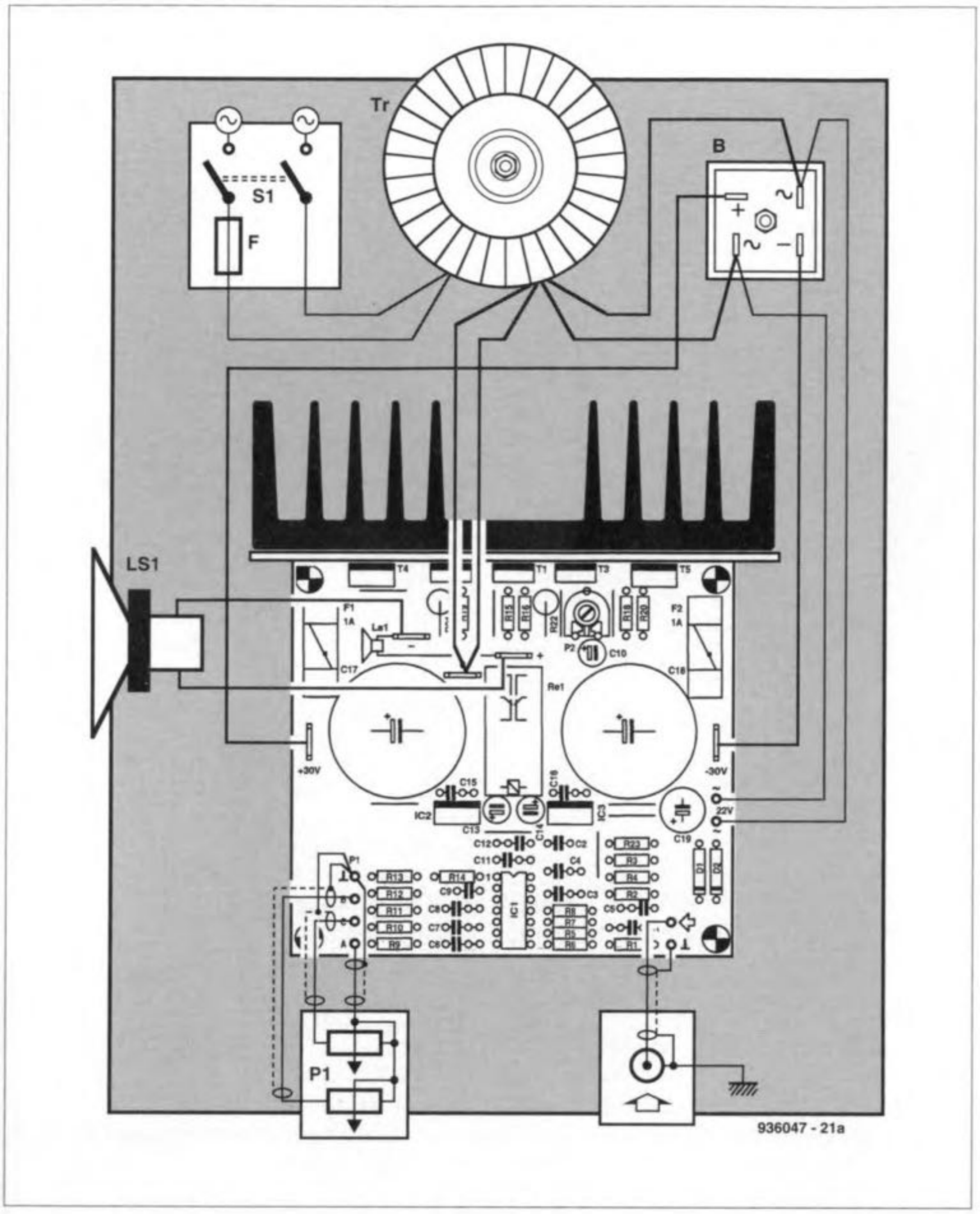


Fig. 12. Wiring diagram of the mini subwoofer.

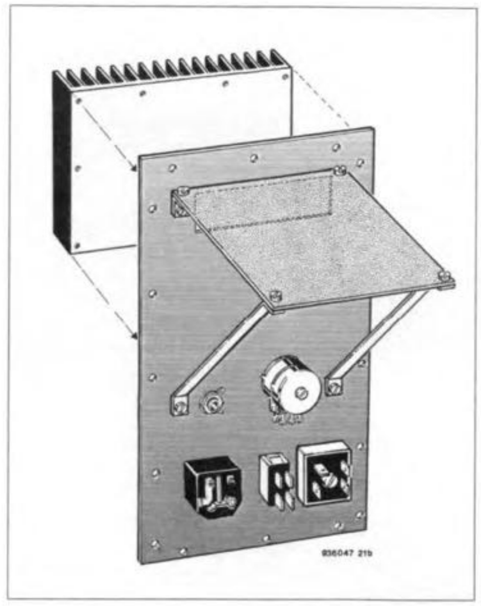


Fig. 13. One stage in the construction of the subwoofer ...

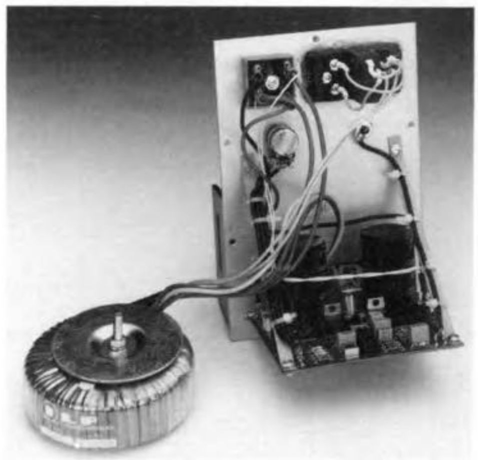
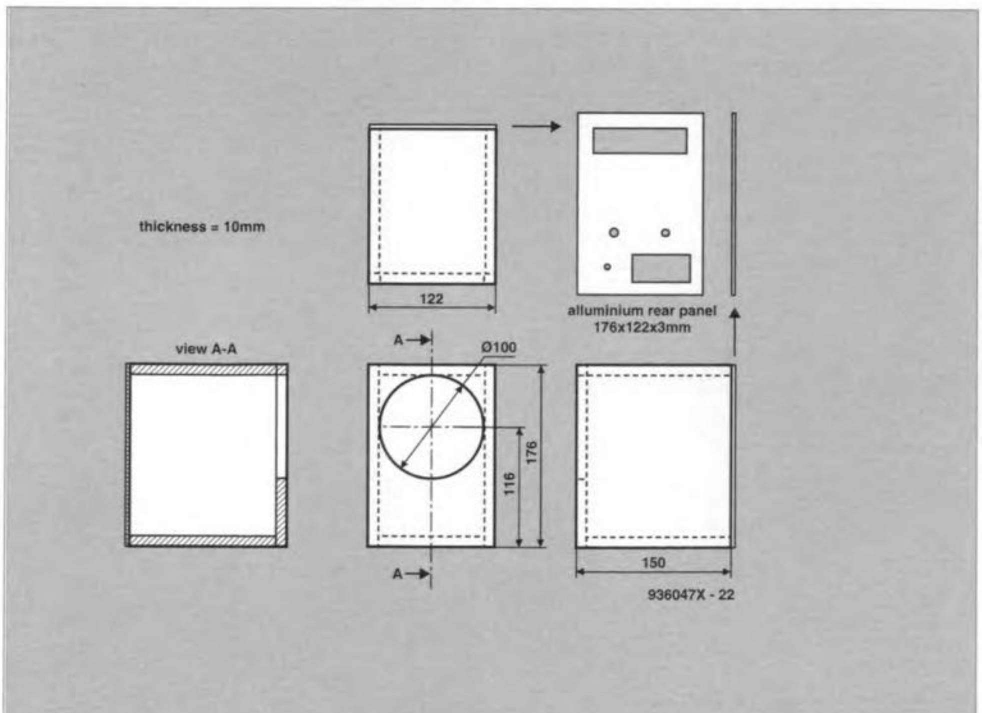


Fig. 14. The completed electronics part of the subwoofers.



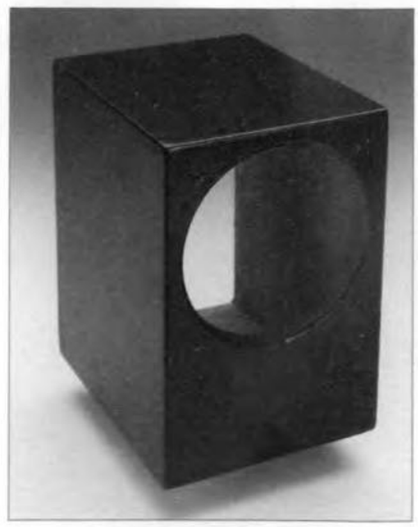


Fig. 15. Construction diagram of the enclosure for the subwoofer.

Fig. 16. The 'bare' enclosure.

Re_1 = relay 24 V, 1100 Ω, 1 make contact
Ls_1 = drive unit, Monacor SPP110/8 or equivalent
Heat sink (T_1–T_5), <1.5 K W^{-1} e.g. SK71/75 SA*
F_1, F_2 = fuse 1 A
Mains transformer, e.g. ILP 31015; 2×22 V/1.81 A
Mains entry plug with integral fuseholder and switch
Soldering pins and lugs as needed
Five ceramic washers (TO-220)
Heat conducting paste as needed
Audio input plug

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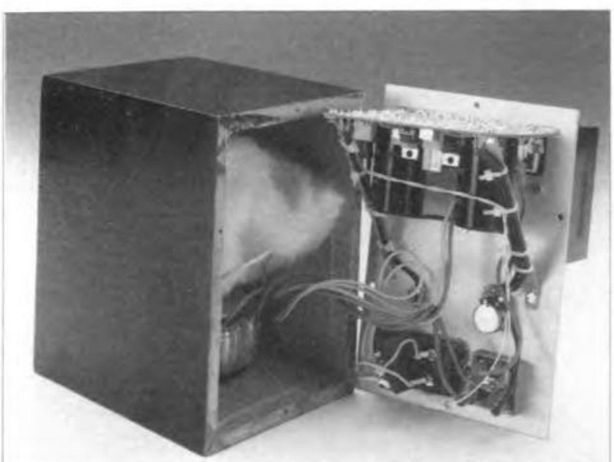


Fig. 17. The completed box with electronics fitted.

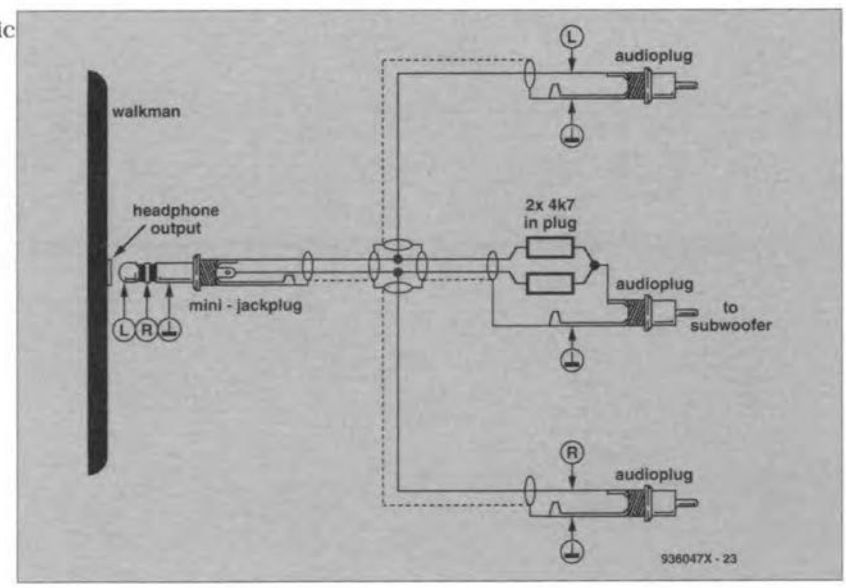


Fig. 18. How to connect the subwoofer to the sound output.

OUTDOOR SATELLITE RECEIVING EQUIPMENT

BACKGROUND TO LNB AND ACTIVE SPLITTER TECHNOLOGY

At the time of writing this article, the history of receiving equipment for communication satellites can be traced over a period of about 10 years. Apart from a series of extended features for satellite TV receivers, developments in the area of outdoor equipment, that is, low-noise converters (LNCs), splitter systems and ancillaries, have moved at a terrific pace.

By R. Badenhausen

YOU may remember those first LNCs (or LNBs as they are sometimes called) used in the Eutelsat era. Indeed, you may still be using such an LNB to view satellite TV programmes. The effective length of these units is determined also by the feed horn system fitted in front of the LNC. The waveguide flange at the input of these LNCs usually allows either a mechanical (motor-driven) or a magnetic polarizer to be fitted. Although their operation was fairly easy to understand, mechanical polarizers often developed, well, mechanical, problems. Not surprisingly, they were relatively short-lived, and generally replaced by the more reliable and much faster magnetic counterparts. The photograph in **Fig. 1** shows such a polarizer fitted on to a typical HEMT LNB.

A direct current sent through a coil sets up a field which operates on a small ferrite rod positioned in the centre of the waveguide. The drawing in **Fig. 2** shows this arrangement. A virtually linear relationship exists between the strength of the induced magnetic field and the rotation of a linearly polarized wave received from the dish. To obtain a 90-degrees rotation of a received wave, the coil should have a number of turns which produces a magnetic field strength of about 500 A/m, depending on the exact specification of the ferrite material used. A full discussion of the operation of the Faraday or polarisation rotator is, unfortunately, beyond the scope of this article. The basics are, however, explained with reference to **Fig. 3** in the inset on the next page.

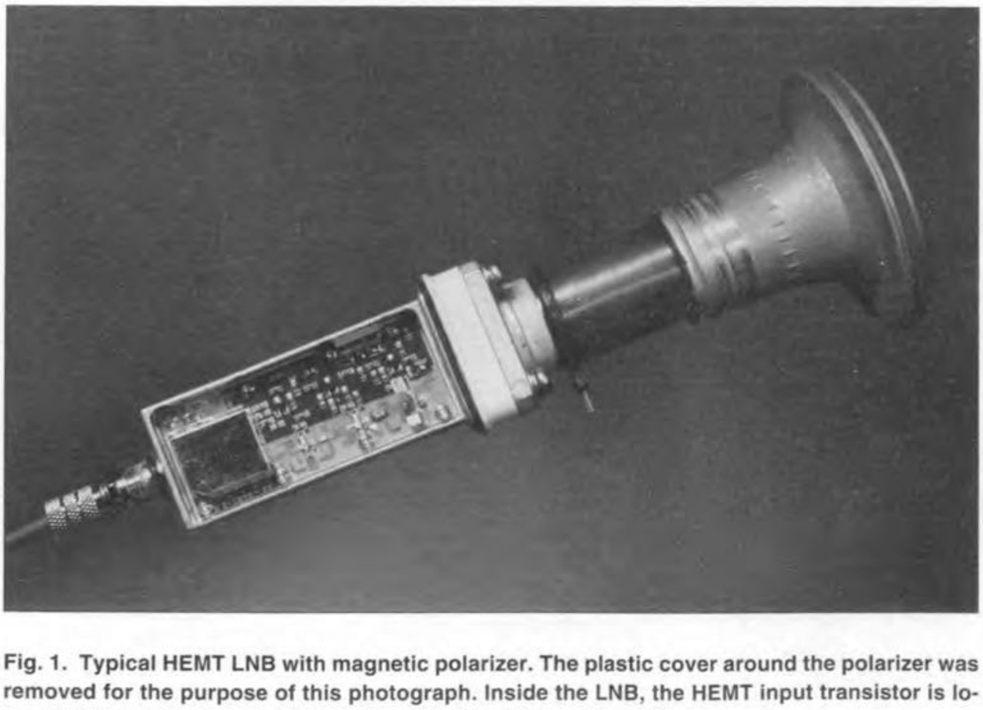


Fig. 1. Typical HEMT LNB with magnetic polarizer. The plastic cover around the polarizer was removed for the purpose of this photograph. Inside the LNB, the HEMT input transistor is located underneath the small screening fitted upright on the PCB.

Figure 4 clarifies the way taken by the received signal after passing through feed horn, and shows the waveguide-to-PCB transition. The waveguide is rectangular, type R120/IEC153. The probe at the electrical input of the LNA (low-noise amplifier) only picks up co-oriented waves from a H/V (horizontal/vertical) mixture. The input amplifier and mixer in the LNA are usually based on HEMT (high electron mobility transistor) technology. The relative ease with which HEMTs can be produced in very large volumes has certainly contributed to the fact that satellite TV reception is now within reach of the masses. As prices of LNCs tumbled, the electrical specifications of the LNA section in particular improved dramatically to a level where a noise figure of about 0.6 dB was reached coupled to a gain of 13 dB. That was about five years ago, when the HEMT finally took over from its Ga-As MESFET predecessor. At about the same time, SMA (surface mount assembly) technology had evolved to a level where new ways of producing an LNB could be attempted. The result of all these improvements was the now widely familiar Marconi LNB. This LNB was one of the first types which did not require an external polarizer to be fitted in front of its waveguide input, horizontal/vertical selection being accomplished with the aid of two direct voltage levels applied to the LNB via the downlead coax cable. The internal H/V selector results in a more compact LNB on to which a conical shaped feed horn is fitted. The size of this feed horn is determined mainly by the construction of the feed holder. These types of LNB are also marked by several, independently operating, LNA circuits, as illustrated by **Fig. 5**.

If you want to exchange an older converter with a large feed horn against a Marconi-type LNB, you may run into problems because the existing LNB holder may not keep the new LNB at the right focal position. In certain cases, a new LNB holder structure will have to be made to make sure that the Marconi LNB is correctly focussed. While most magnetic polarizers require a holder with an inside diameter of about 30 mm, the Marconi LNB is usually fitted with one of two standardized feed horns having an outside diameter of 23 mm or 40 mm. In a few cases, 23-mm Marconi LNBs come with an adaptor for use with 40-mm holders.

State of the art converter technology

When designing HEMT based input circuits, a compromise must be found between highest achievable amplification and lowest noise figure. An important factor in this design strategy is the source impedance of the input circuit, in other words, the matching between the waveguide-to-PCB transition and the input of the HEMT. Thanks to the good dielectrical characteristics of Teflon® circuit boards, inductors, capacitors as well

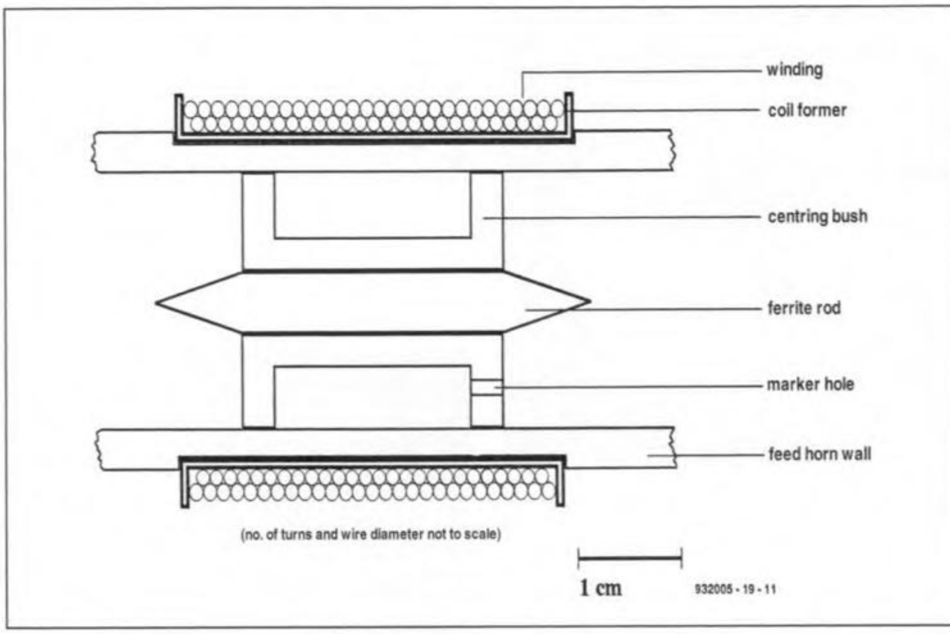


Fig. 2. Typical construction of a magnetic polarizer. The ends of the ferrite rod are tapered to avoid undesirable wave reflections. The marker hole only serves for production purposes.

BASICS: magnetic polarizers

The plane of a linearly polarized wave may be rotated by modifying the permeability of ferrite materials with the aid of a surrounding magnetic field.
As illustrated in Fig. 3, the ferrite then has a complex permeability, which requires two real components to be taken into account along with two other, imaginary, quantities, which are not discussed here. The real components are μ'_+ (<1) and μ' (>1). Here, the point of operation is assumed to be in the non-saturated regions caused by a magnetic field.
Another analytical assumption is that each linearly polarized wave consists of two circularly polarized, counter-rotating components (positive and negative circular waves), each having half the amplitude.
In a medium made from ferrite material, the phase rotation angle, ϕ, of two counter-rotating circular waves is calculated from

$$\phi'_\pm = 2\pi f k l \sqrt{\varepsilon_0\, \varepsilon'\, \mu_0\, \mu'_\pm}$$

where
f = operating frequency (MHz);
k = correction factor (<1, e.g., attenuation);
l = effective length of medium in direction of wave propagation;
$\varepsilon_0 = 8.85 \cdot 10^{-12}$ F m^{-1};
ε' = dielectric value of medium;
μ_0 = permeability constant; $1.26 \cdot 10^{-6}$ H m^{-1};
$\mu'_\pm$ = real permeability values of medium;
ϕ = in radians.

By applying the magnetic field, the positive circular wave is rotated by an amount

$$\phi'_+ = 2\pi f k l \sqrt{\varepsilon_0\, \varepsilon'\, \mu_0\, \mu'_+}$$

similarly, the negative circular wave is rotated by an amount

$$\phi'_- = 2\pi f k l \sqrt{\varepsilon_0\, \varepsilon'\, \mu_0\, \mu'_-}$$

The resulting difference between the angles of the two circular polarized waves which leave the medium then becomes

$$\phi' = \phi'_- - \phi'_+$$

or, converted to the linear polarized wave,

$$\phi = 0.5\, \phi'$$

In practice, owing to technical restrictions, most waveguide is only partially filled with ferrite material. Consequently, some deviations will occur in the above (simplified) calculation models.
Ferrite-ceramics in general have dielectric values greater than 10, and initial permeability values of up to 1,000. Such values are necessary to be able to use relatively low magnetic bias power levels, say, below 100 mW. As regards insertion loss, most manufacturers of magnetic (ferrite) polarizers indicate a value ≤0.2 dB, and state a value of at least 20 dB for the cross polarization isolation. As an aside, raising the magnetic field strength by an amount H_R, causes absorption of the received wave, converting its useful RF energy almost entirely into heat in the ferrite material.

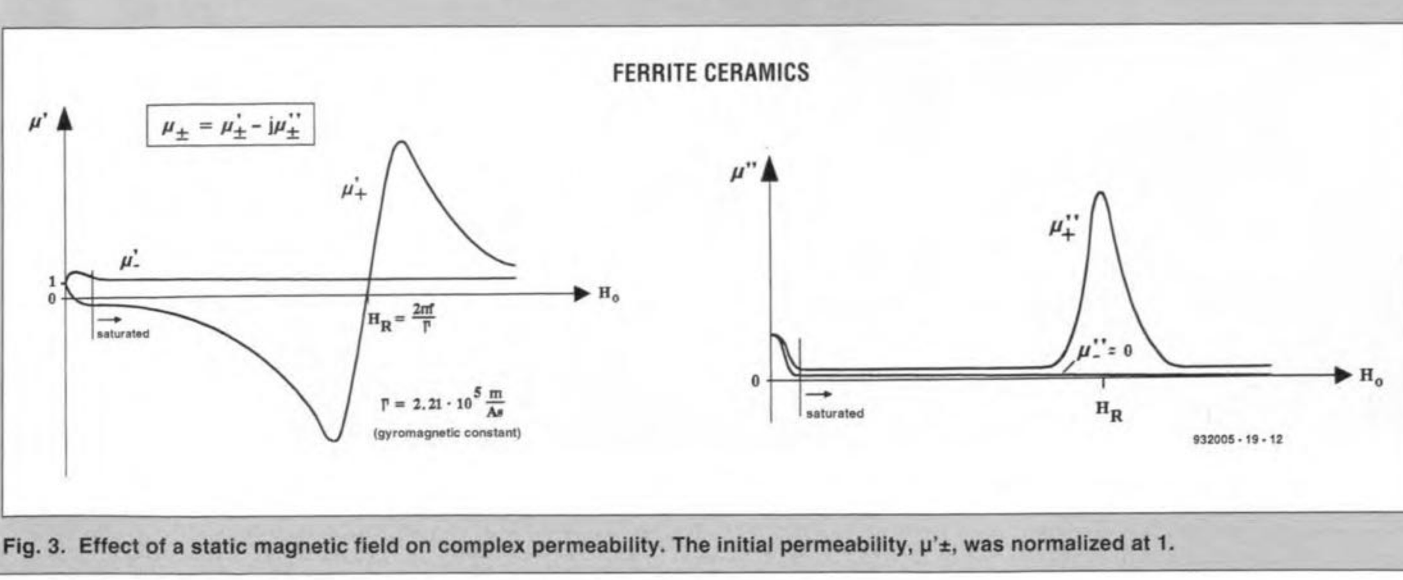


Fig. 3. Effect of a static magnetic field on complex permeability. The initial permeability, $\mu'_\pm$, was normalized at 1.

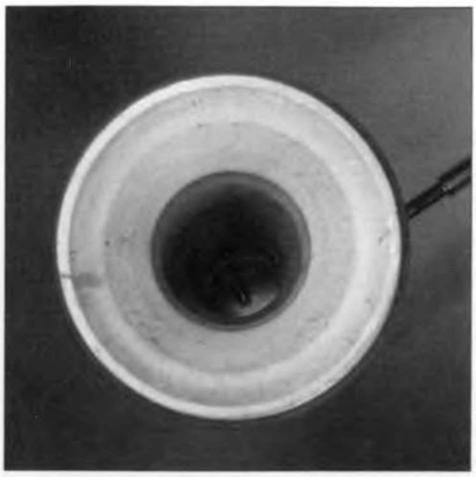


Fig. 4. A quarter-wave probe is used to couple the RF signal from the waveguide to the input of the LNA. Obviously, the polarizer was removed from the LNB input for the purpose of this photograph.

as filters may be realized virtually lossfree with the aid of microstrip technology. Examples of microstrip sections are visible in a number of areas at the board side shown in **Fig. 6**. At the same time, the logic and control circuit for the H/V switching is quite extensive, too, as shown in **Fig. 7**.

Repair work on faulty LNBs is usually limited to voltage measurements on the logic parts and the voltage regulator. Type numbers being rare on microwave semiconductors, it is nearly impossible to replace these parts. In most cases, attempts at repair are futile and a waste of time, and a defective LNB is best replaced by a new one.

Special attention should be given to the waterproof sealing of the LNB case. Marconi-type LNBs with two outputs (H and V) are either 'twin' or 'dual' versions. Although a number of manufacturers of these LNBs do state the type and approval numbers on an adhesive, the indication 'twin' or 'dual' is usually missing. A 'dual' version supplies only one polarisation plane on each of its outputs, while a 'twin' version contains, in principle, two completely independent V/H converters. The selection is usually made with the aid of a few jumpers in a shared circuit.

Active splitter systems

The *Multiswitch* signal distribution system allows the use of 'dual' type LNBs. **Figure 8** shows such an active splitter, which allows up to four satellite TV receivers to be connected. It should be noted, however, that this splitter system

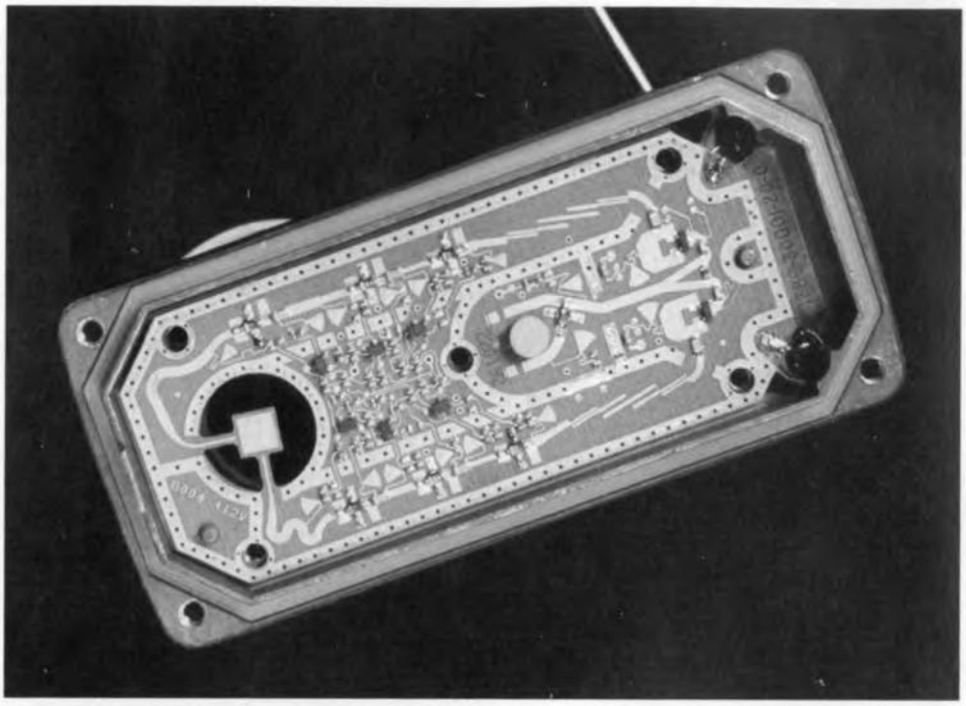


Fig. 6. The almost completely symmetrical microwave and IF signal paths of the twin LNB are clearly visible in this picture. Also seen are the two quarter-wave 'antennas' for the two polarization planes. The square pad which joins the probes at the centre of the waveguide serves mainly to compensate electromagnetic reflections caused by the LNA input circuitry.

Fig. 5. A 'twin' LNB: two mutually independent V/H converters with a shared microwave fed system in a single case. Ideal for two satellite TV receivers on a single dish.

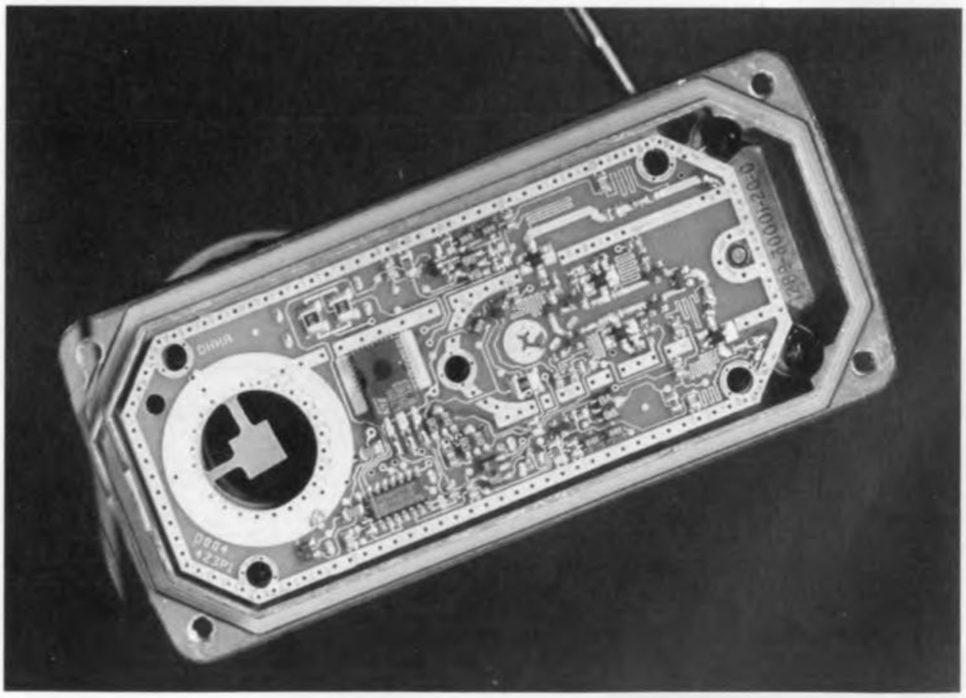


Fig. 7. For the purpose of this photograph, the PCB was flipped and put back into the LNB case. Most of the components at this side of the board go into the power supply and polarization selection logic.

Fig. 8. An active four-way switch with optional insertion of terrestrial (VHF/UHF) TV signals. LNB signals are applied separately for the two polarization planes, for which the Multiswitch supplies the appropriate supply voltages.

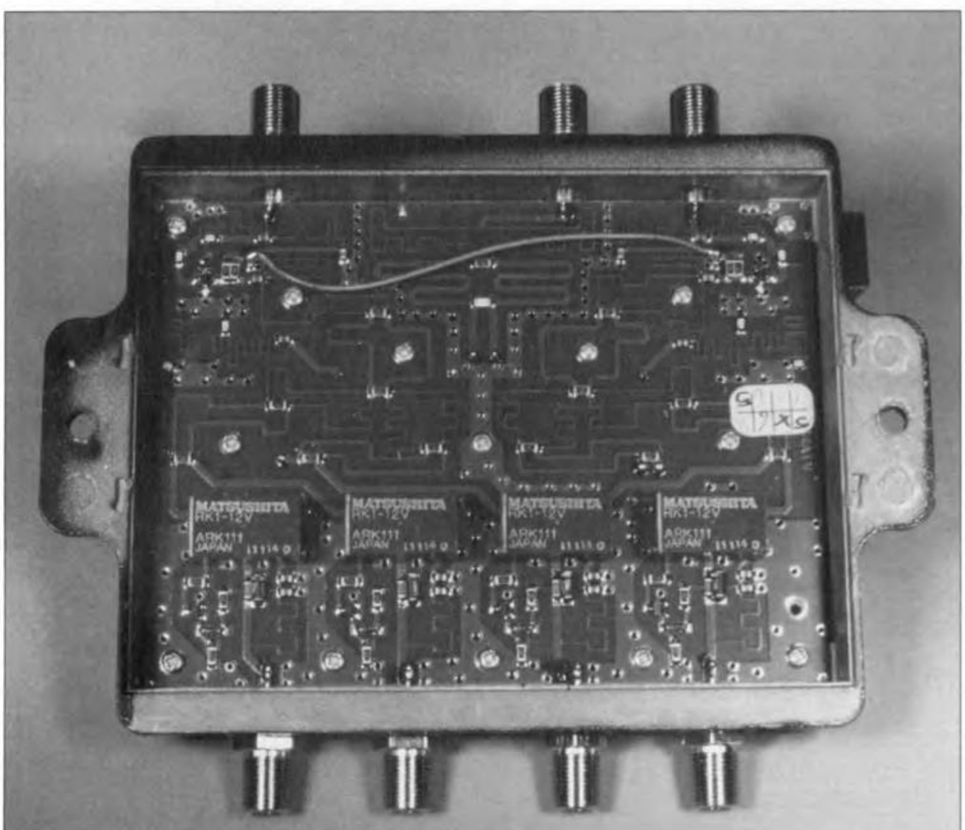


Fig. 9. A look inside the Multiswitch from Lynk. The Matsushita RF relays switch the LNA IF signal for the H or V plane to the individual outputs (1-4), under the control of the polarization selection voltage sent up by the satellite TV receiver.

is, generally speaking, a passive circuit as far as the connection of terrestrial TV signals is concerned. In particular, when more than two participants are connected, it may be necessary to use an external antenna booster to compensate the attenuation of about 15 dB in this frequency range (40-960 MHz). Also, as regards installation of a Multiswitch, it should be noted that these units are fitted in a metal enclosure which may not be fully watertight, although it affords sufficient RF screening.

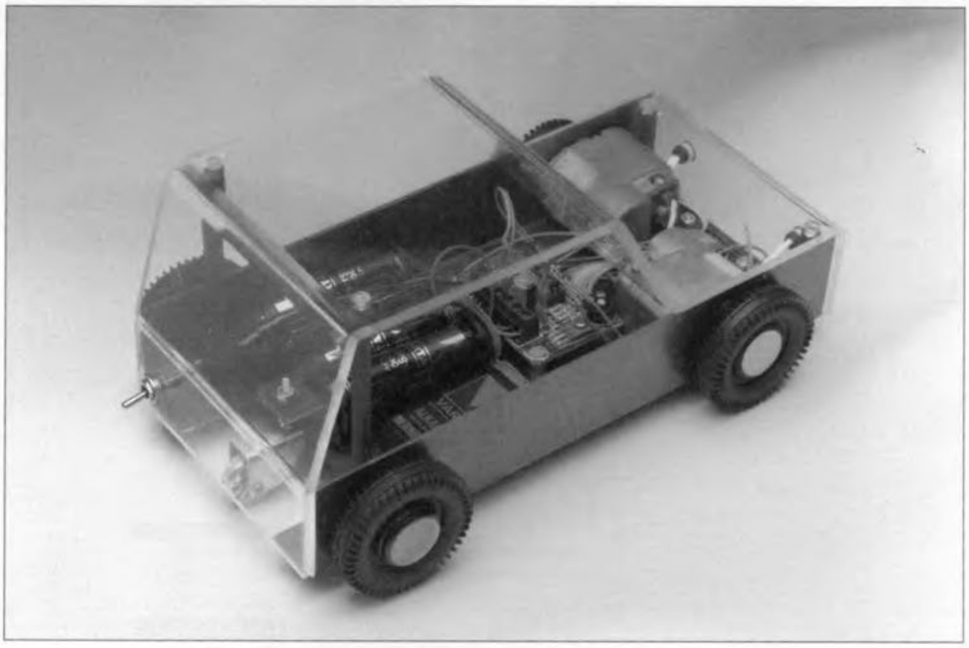
The Multiswitch system is often an ideal — and certainly cost-effective — alternative to multi-user satellite TV systems in which cables to the individual users are already present, or when where re-arranging some of the cables is not problematic. Easy to install, the Multiswitch is still a fairly complex unit from a point of view of electronics, as shown in **Fig. 9**.

A 'twin' type LNB may also be used in conjunction with a Multiswitch splitter, but only if the latter provides the normalized LNB supply voltages at its inputs. Although this is usually the case, most manufacturers of active splitters recommend the use of 'dual' type LNBs, simply because these are less expensive and better available.

H/V converters have input and output markings which resemble those of Multiswitches. These units supply the horizontally and vertically polarized signals in separate bands within the range 950 MHz to 2,000 MHz. This approach, which is not as cost effective as the Multiswitch, has an advantage when is not possible to go round a signal feed to several participants on a single (existing) cable. (932005)

MINI ROBOT CAR

This playful project demonstrates the operation of a small electric vehicle which is capable of tracking a line. Inexpensive and easy to build, the project forms an excellent introduction to the fascinating world of robotics.



Design by L. Pijpers

ROBOT cars in large factory halls find their way with the aid of sensors and a track in or on the floor. The track may consist of a white line painted on a dark floor, or another reflective substance. Such tracks can de detected with optical sensors. Other variants of track types include metal strips which can be followed with the aid of a metal detector, or a slot in the floor which can be followed by using a probe or another mechanical device.

To avoid collisions with personnel and objects, most robot cars have additional sensors, for instance, infrared detectors, cameras, or a kind of radar based on ultra-sonic waves, laser light or radio waves. To enable them to be stopped in the (unlikely) event of a collision or malfunction, these robot vehicles usually have a number of easily accessible switches.

The miniature robot car described in this article is capable of following a black line drawn on a light surface, or a white line drawn on a dark surface (as illustrated by the photograph in **Fig. 1**). The description of the operation of the electronics is based on the assumption that a black line is followed on a light surface.

The sensors used are reflection types, based on a combination of an infra-red LED and a phototransistor. The sensors, shown in **Fig. 2**, drive two small electric motors via some electronics. The motors are used to power the front wheels of the car.

Drive system

The circuit diagram, shown in **Fig. 3**, indicates a remarkably low component count. The symmetry of the circuit is quite obvious and not surprising because the front wheels of the car are powered separately by motors M_1 and M_2. Because the speed of a normal electric motor is far too high for the present application, motors with a built-in reduction gear are used. Also, because the motors must turn in opposite directions to move the car straight on, they receive opposite supply voltages.

The current through the motors is switched on and off by darlington transistors, T_1 and T_2. LEDs D_1 and D_3 are provided to indicate activity of the relevant motor. Diodes D_2 and D_4 are connected in parallel with the respective motor to suppress back e.m.f. surges which are generated when the motor coil is switched off.

Electrical power is furnished by a set of batteries, the size and type of which will depend mainly on the size of the car. NiCd, lead-acid and ordinary dry batteries are all suitable, although the rechargeable types are, of course, preferred because of the environmental aspect. The battery voltage is not critical: depending on the type of motors and the desired 'top speed' of the car, voltages between 6 V and 15 V may be used. If you use NiCd cells, simply put a number of these in series until you have the desired voltage.

On the right track

In the present measurement and control circuit, the two reflection sensors, IC_2 and IC_3, 'check' if they are looking at a reflecting surface, for instance, at either side of a track made from black tape. The detectors are mounted such that one of them is always to the right of the track, and the other, to the left of the track. Each of the infra-red LEDs in the detectors then illuminates a section of the surface underneath. The light reflected by the (white) floor at either side of the track is detected by the phototransistors, which enable the relevant motor to be powered. When the vehicle swerves from the track, one of the phototransistors will detect less reflected light, or no light at all, and the relevant motor will be slowed down. Because of the electrical characteristics of the sensors used, this is a fairly gradual process, i.e, not an abrupt action, which serves to correct the direction of the vehicle.

The circuit diagram indicates that the phototransistors in IC_2 and IC_3 are n-p-n types. The collectors are connected directly to the positive supply voltage. The base of the phototransistor is not bonded out to a pin because the device is driven by light rather than voltage. Assuming that a fixed emitter resistor is used, the amount of collector-emitter current which flows through the device depends on the amount of light detected by the phototransistor. The emitters of the two phototransistors are connected to ground via resistors R_2 and R_3, and a section of preset P_1. As long as there is sufficient light on the phototransistors, they will keep conducting, and the emitter voltage will be 'high' (nearly the supply voltage). When the vehicle diverges from the track, however, one

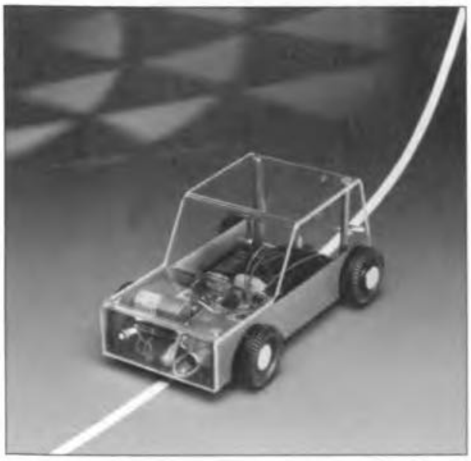


Fig. 1. The mini robot car is capable of following a white or a black track on a smooth surface such as cardboard.

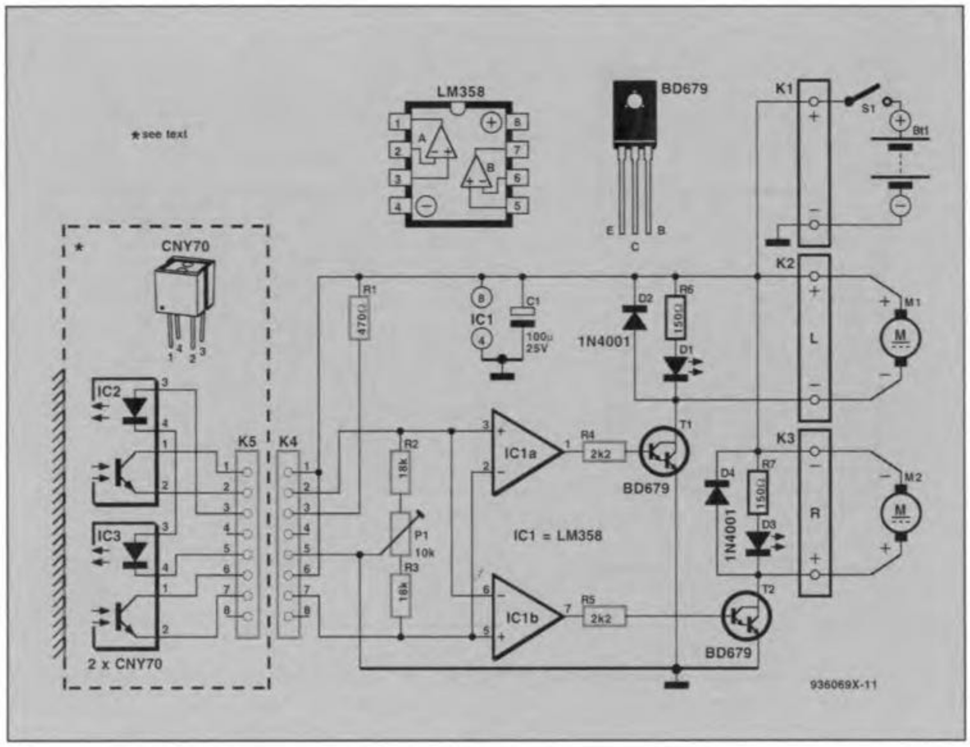


Fig. 3. When reflection sensor IC2 detects more 'white' than IC3, motor M1 runs and powers the front left wheel. The little vehicle then makes a right turn.

of the phototransistors will switch off, causing its emitter voltage to drop considerably (down to almost 0 V). When the sensor is exactly above the border between the dark track and the white surface, the emitter voltage will be roughly half the supply voltage.
The sensor voltages are combined by two opamps, IC_{1a} and IC_{1b}, and then fed to the motor driver transistors, T_1 and T_2. The motor associated with the sensor which sees a 'white' surface (or more white than the other sensor, see further on) is energized.

Both phototransistor emitters are connected to an inverting and a non-inverting input of one of the opamps. Consequently, when the emitter voltage changes, the output of one opamp will go low, while that of the other will go high. An example: assuming that IC_2 is suddenly unable to detect a white surface, its emitter voltage drops low. This low level also reaches pin 3 of IC_{1a} (non-inverting input) and pin 6 of IC_{1b} (inverting input). The result is that the output of IC_{1a} (pin 1) goes low, while the output of IC_{1b} (pin 7) goes high. Transistor T_1 is then switched off because its base current is removed. T_2, on the other hand, does receive base current (via R_5), and starts to conduct. Motor M_1 stops, and M_2 starts to turn. Assuming that M_1 is fitted at the left-hand side of the vehicle, and M_2 at the right-hand side, the vehicle will turn to the left.

As already discussed, each of phototransistor is connected to two opamps. However, the opposite is also true, i.e., each opamp is connected to two phototransistors. Looking at IC_{1a}, for instance, it is seen that both inputs of this opamp are actually connected to IC_2 **and** IC_3. This causes the opamp to behave like an ordinary difference amplifier, whose output can be made to go high by pin 2 dropping low or pin 3 going high. In practice, that means that motor M_1 is switched on either as a result of IC_2 detecting more light, or IC_3 detecting less light. Remember, by 'light' we mean infra-red light reflected by the white floor. The story is the same for opamp IC_{1b} and motor M_2. Summarizing, the operation of each motor is governed by the light difference detected by the pair of sensors, rather than the absolute output level of the sensor it belongs with.

Because of the electrical coupling between the two symmetrical halves of the circuit, a kind of 'electrical equilibrium' is created. This balance occurs, theoretically, when both sensors detect an equal amount of light, and P_1 is exactly at the centre of its travel. Only in that (hypothetical) case, M_1 receives just as much current as M_2, and the mini car would drive straight on. In practice, that will never happen because of the relatively high gain of IC_{1a} and IC_{1b}, and the fact that the vehicle is constantly busy correcting its course, which can only be achieved by switching M_1 and M_2 on and off all the time. In fact, the vehicle tracks the line along a slightly zigzagging course. If the vehicle has a constant tendency to swerve to one direction, that can be corrected by adjusting the preset.

Construction and test

As far as the electronics are concerned, everything fits neatly on the printed circuit board shown in **Fig. 4**. This board is available ready-made through the Readers Services. Before you start fitting the parts, cut the board in two sections. The small section is for the detector, and the large section, for the motor driver electronics. The completed driver board is shown in **Fig. 5**. Once all parts are mounted, run a thorough visual inspection on your solder work and the values and orientations of all components.

Assuming that everything is to your satisfaction so far, you may interconnect the boards, and connect the supply voltage. The motors are not connected as yet. Their activity is indicated by LEDs D_1 and D_3. Stick a piece of black adhesive tape on a piece of white cardboard, and move the sensor

Fig. 2. The two reflectors mounted at the underside of the car ensure that the vehicle stays on the track.

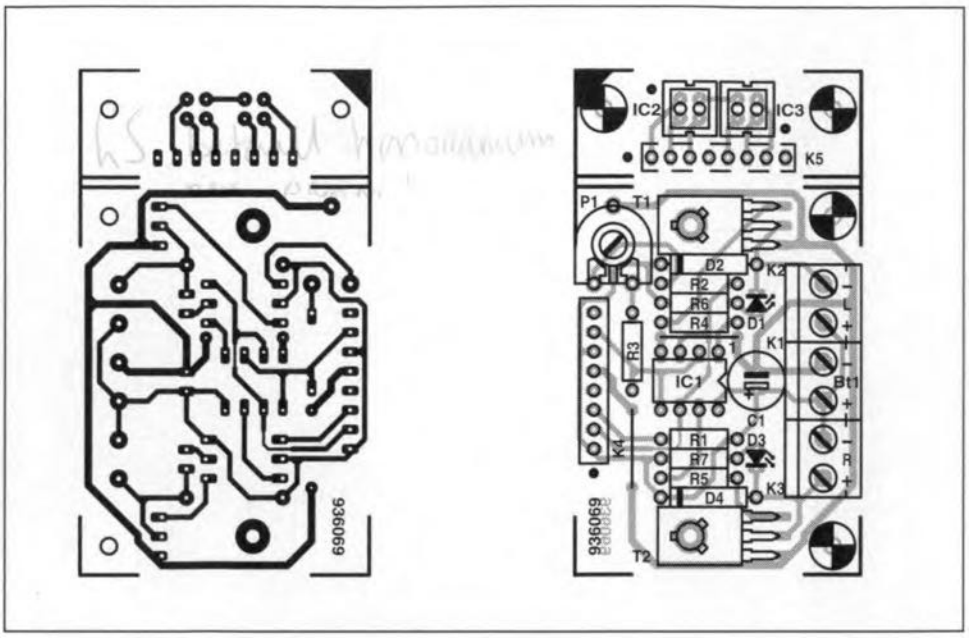


Fig. 4. Track layout and component mounting plan of the printed circuit board designed for the mini robot car (board available ready-made, see page 70).

COMPONENTS LIST

Resistors:
R_1 = 470Ω
R_2;R_3 = 18kΩ
R_4;R_5 = 2kΩ2
R_6;R_7 = 150Ω
P_1 = 10kΩ preset H

Capacitors:
C_1 = 100µF 25V radial

Semiconductors:
D_1;D_3 = LED, red, 5mm dia.
D_2;D_4 = 1N4001
T_1;T_2 = BD679
IC_1 = LM358
IC_2;IC_3 = CNY70

Miscellaneous:
K_1;K_2;K_3 = 2-way PCB terminal block, pitch 5mm.
K_4 = 8-way SIL pin header.
K_5 = 8-way SIL socket
M_1;M_2 = 12VDC model car motor w. reduction gear.
S_1 = on/off switch
Bt_1 = 9V to 15V battery set, with holder.
Printed circuit board, order code 936069 (see page 70).

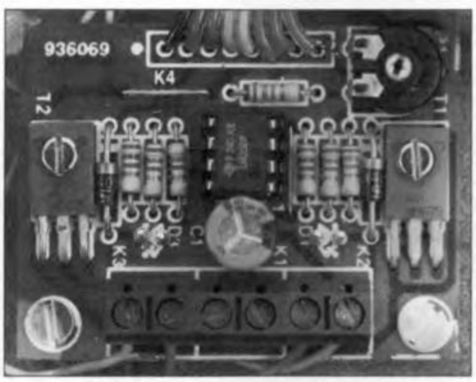


Fig. 5. The large board accommodates the darlington transistors which switch the motor current. The two reflection sensors are fitted on a separate little board.

board to either side of this track. The LEDs should come on and extinguish as you follow this track and simulate diversions. If this checks out, you may run the same test with the motors connected to the board (mind the polarity!). One motor will run at a time. Adjust P_1 if there appears to be a divergence to one side.

Building the vehicle

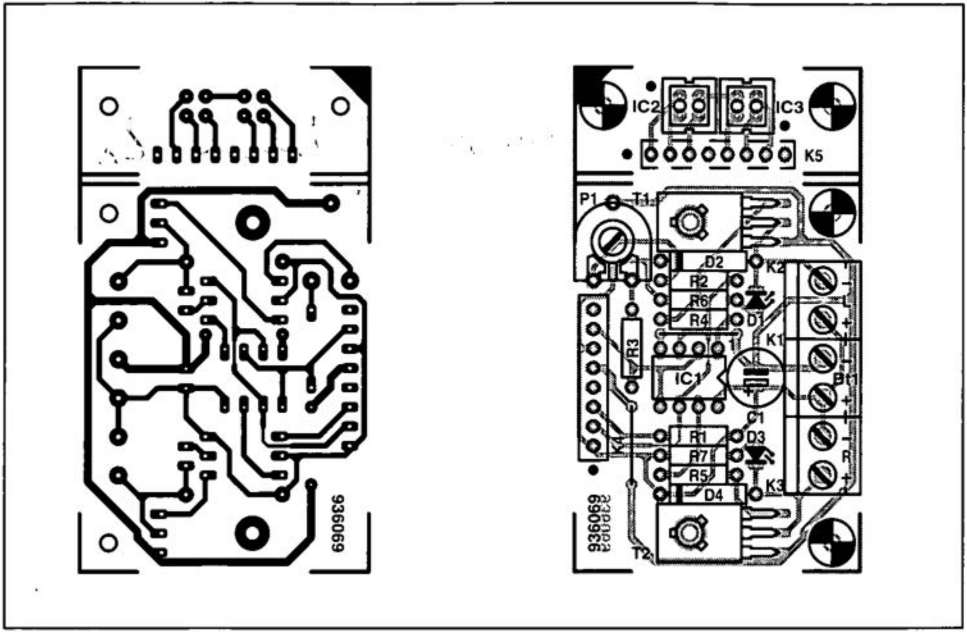
Although you are perfectly free to make a model car in the latest Italian style, the emphasis here is on a simple little vehicle to demonstrate elementary robotics. The prototype is actually a three-wheeler built from pieces of perspex cut and bent to form a basic chassis. Perspex is easily bent into the desired shape with the aid of a hot-air gun or an electric paint stripper. If you do not fancy building your own chassis, pay a visit to a modellers' shop. **Figure 6** shows the bottom side of the three-wheeler. The third wheel, which is not used to power the car, is fixed on a vertical spindle, and serves to improve the steering characteristics. The two rear wheels are dummies which are not in contact with the floor.

Once the mini car works, you may want to experiment with different types of track. It is best to start with a narrow black track (which fits between the sensors) on a light surface. Alternatively, you may want to use a wider, white, track (underneath the sensors) on a dark surface. In the latter case, be sure to swap the wire pairs on the motors. In both cases, P_1 will have to be set roughly to its centre position. Another interesting experiment is to make the car drive along a single black/white border. That requires P_1 to be adjusted almost fully clockwise or anti-clockwise. Arriving from the left, for instance, the car will faithfully track the border. Coming from the right, however, it will not even be able to detect the border!

In case the car is unable to 'see' the track, experiment a little with the distance between the sensors and the track. This distance must be between 1 mm and about 5 mm.

Finally, a remark about the power supply. Although the prototype was powered by a 9-V battery in series with a couple of 1.5-V 'mono' cells, it is , of course, better to use batteries of the same type to arrive at the desired voltage. The reason for this choice should be clear: only then will all the batteries be drained simultaneously, so that they can also be charged simultaneously. (936069)

Fig. 6. Prototype of the car viewed from underneath. The third wheel at the rear of the car improves the steering. The two rear wheels are dummies which do not touch the floor.


936069
IC2
IC3
K5
P1
T1
D2
K2
R2
R6
R4
D1
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IC1
Bt1
C1
R1
D3
R7
R5
K4
D4
K3
T2
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MARCONI - THE FATHER OF RADIO

GS betaalt honorarium per pagina!

By Ian Poole

This year, it is a hundred years ago that Marconi gave the first demonstration of a usable radio system near Bologna in Northern Italy. From these beginnings, he went on to become one of the greatest figures in the history of the development of radio. There is little doubt that, without his drive and innnovation, radio would not be as advanced as it is today.

Photograph by courtesy of
The Institution of Electrical Engineers

Upbringing

Guglielmo Marconi was born in Bologna in 1874, the son of a wealthy business man named Guiseppe. In his early years, he travelled widely, accompanying his mother on her trips abroad. During his periods at home, he was given private tuition. Unfortunately, when he started to attend a college, it was found that his previous tuition had been inadequate and he did not show any signs of his future brilliance.Eventually he abandoned his formal education without any qualifications.

Despite this, his last years in education had left him with a keen interest in physics. Seeing this, his mother arranged some tuition with a family friend, Professor Righi, who was a lecturer at Bologna University. It was a a result of reading an article by his new tutor that Marconi became very interested in the newly discovered Hertzian waves.

First experiments

Quickly he started to repeat some of Hertz's original experiments. In these, Hertz had made a spark jump across a spark gap in the transmitter circuit. This made a more feeble spark jump across a much smaller spark gap in a circuit nearby. Using this set-up, Marconi was able to achieve distances of a few metres between the two circuits.

Marconi realized that the method of detecting the transmitted spark was very insensitive. To improve this, he made a coherer. This was a device invented by a Frenchman, Edouard Branly, and later made into a form that could be used with Hertzian or radio waves by Sir Oliver Lodge, an English researcher.

Basically, a coherer consists of a glass tube filled with metal filins. At either end, electrodes are inserted as shown in **Fig. 1**. Normally, the resistance between the two electrodes is very high, but when a spark signal is detected, it causes the filings to cohere (stick together) and so make a low resistance path between the electrodes. This can be used, for example, to sound a bell. A small tap is then given to the tube to de-cohere (separate) the filings, and often this was provided by a tapper attached to the bell circuit as shown in **Fig. 2**. In this way, the bell would sound repeatedly whilst a long signal was present.

Not satisfied with the performance of the coherer in this form, Marconi set about improving it. As he did not have a sound mathematical background, he started experimenting with different compositions of metal and soon found a far better combination (than used by Branly). In fact, it was Marconi's practical, experimental approach that enabled him to make so many of his discoveries.

Soon, Marconi was able to span much greater distances with his system using a transmitter like that shown in **Fig. 3**. For many of his tests, he would operate the transmitter in the house and one of his father's employees would be with the receiver in a field nearby. When a signal was received, this would be signalled by waving a flag, although when they went out of sight over a hill, a shotgun had to be fired so that its report could be heard.

Once a range of about 2 km could be spanned, the Marconi family decided that it was time to approach the authorities to see if they were interested in exploiting the idea. The first demonstration took place in 1895 to the Italian Ministry of Posts and Telegraphs. They declined the offer, however, because they could not see any advantages over the existing wire telegraph systems they already used.

Move to England

Marconi was naturally disappointed with the refusal by the Italian authgorities. However, he was not deterred and the

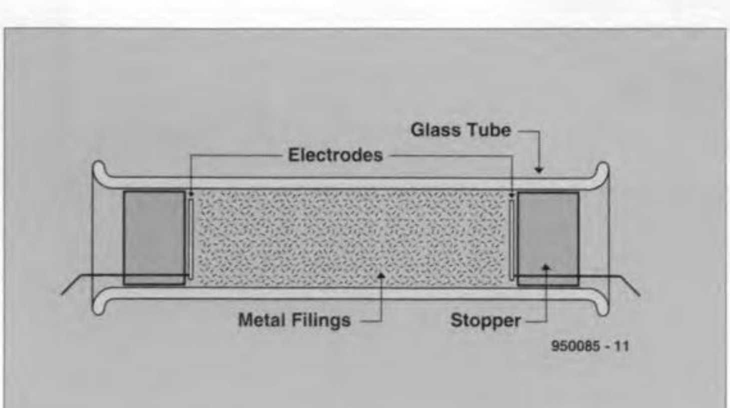


Fig. 1. Diagram of a coherer.

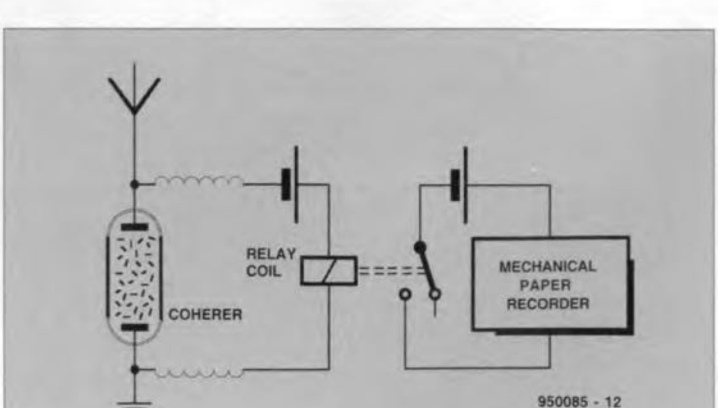


Fig. 2. Circuit diagram of an early receiver.

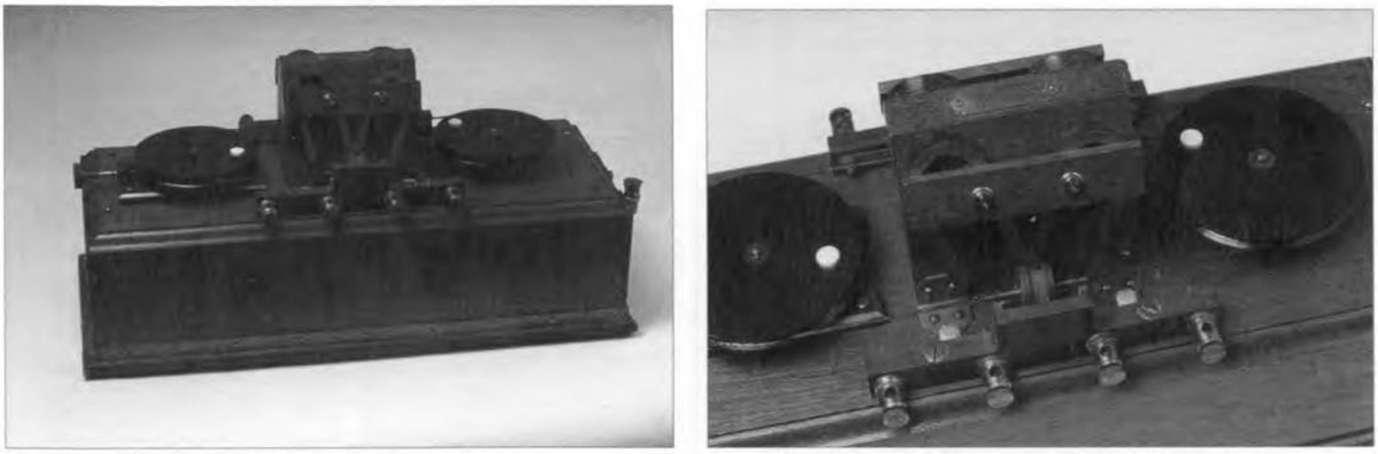


Fig. 3. General view and close-up of an early magnetic detector. (From the Journeaux Historic Wireless Section). Photographs by courtesy of G.C. Arnold Partners.

family decided that there might be more opportunities in England where Marconi's mother had many roots.

Arriving in London in 1896, Marconi and his mother were met by his cousing who was himself an engineer. He gave Marconi an introduction to A.A. Campbell Swinton, who saw possibilities in the idea and arranged a demonstration to Williaam Preece, the then Chief Engineer for the Post Office. Preece was very impressed with the demonstration, but bureaucracy meant that the Post Office were very slow in coming forward with any offers.

By this time, Marconi had given several demonstrations, including some to the Press, and the response was very favourable. As a result, he decided to set up his own company. This was formed in July 1897 and called 'Wireless Telegraph and Signal Company Limited'. The formation of the company enabled Marconi to borrow money to further his development.

It did not take him long to start on further work. By the end of 1897, he had erected some masts over 40 metres high at The Needles Hotel on the Isle of Wight so that he could perform tests with a boat steaming up and down the Solent. From this site, he was able to achieve distances of over 30 kilometres.

Next, Marconi made the first international link. In 1899 he set up stations at Wimereux near Boulogne in France and South Foreland in Kent, England. Apart from the test receiving a large amount of press coverage, a major discovery was made when the transmissions were picked up at his factory in Chelmsford. Up until then, it had been thought that radio waves could travel only over line-of-sight paths.

The challenge of the Atlantic

With this new discovery, Marconi started to think of communications over even greater distances. By proving that communication could be made to ships in the Atlantic, a whole new market for his equipment could be opened up. Accordingly, Marconi decided to try to make communication across the Atlantic. However, this was a major challenge both in terms of the technical achievement and the finance needed to carry out the project.

He set up the first of his stations in Poldhu in Cornwall, England. Here he erected a ring of twenty masts over sixty metres high to support a cone of wires which formed the aerial. A similar aerial was erected at the station on the other side of the Atlantic at Cape Cod in Massachusetts, USA. Unfortunately, the design of the aerial was poor and the one at Poldhu was totally destroyed during a gale. The one at Cape Cod was also found to be unsafe, distorting in a strong breenze.

With typical resilience, Marconi did not let this disaster destroy his dreams. He set about rebuilding the station at Poldhu, but this time he used a smaller, more robust design for the aerial. He also decided to relocate the American station to Newfoundland, Canada, to make the transmission path shorter. It was also necessary to use simpler wire aerials which would be supported by kites or balloons.

Tests commenced in December, 1901, with transmissions from Poldhu.

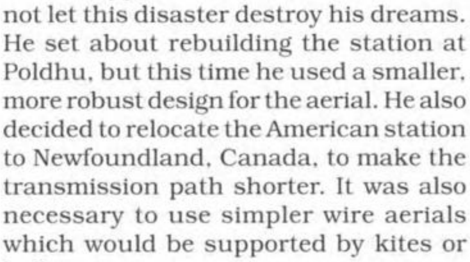


Fig. 4. The circuit of one of Marconi's first transmitters.

Unfortunately, weather in Newfoundland was bad, and the kite supporting the aerial moved rapidly in the wind making the receiver difficult to tune. In addition to this, the signals were very weak and so Marconi resorted to the use of a 'self-restoring coherer' and a sensitive pair of headphones. This type of coherer was an early form of rectifier which enabled the signals to be detected much as they are in many modern sets. Using this, Marconi and his assistant were just able to detect the letter 'S' being transmitted from the other side of the Atlantic. News of this was enthusiastically received by the press and Marconi was hailed a hero.

Further developments

Whilst Marconi had proved it was possible to span the Atlantic, he had not yet succeeded in proving that a reliable service could be maintained. To achieve this, he set up a station on Cape Breton Island, Canada. Here again, an aerial collapsed and this proved that more work was neede to improve the aerial designs.

Shortly afterwards, when he was back in England testing out a new aerial, he noticed that a wire on the ground pointing in the direction of the transmitter picked up a better signal. After further development, the inverted L or Marconi aerial was conceived—see **Fig. 4**. This became the mainstay of Marconi installations for many years and is in use even today.

As it became clear that useful and reliable communications could be provided between ships and the shore, the number of installations started to grow. The first commercial installation on a merchant ship was completed in 1900, and by 1902 a total of seventy ships were using Marconi apparatus; by 1910, this figure had risen to over 250. Many of these installations used a magnetic detector. This was another of Marconi's developments and was more sensitive than previous types. This gave Marconi installations a competitive edge over his rivals.

Shorter waves

The first long-distance communication systems used very long wavelengths, often greater than 2000 metres. However, in the 1920s, experiments were undertaken into the possibility of using short waves. Marconi, too, started some experiments. In 1923, he set up a transmitter at Poldhu and monitored the signal strength as he sailed away. He noticed that the strength fell at first, but at a distance of around 4000 km he noted that the short-wave transmissions were stronger than those from the very-high-power long-wave stations. This was despite the fact that the short-wave transmitter was much less powerful.

With proof of the viability of the short-wave bands, many organizations started to use them. The British Government required a network of stations to link various locations in the British Empire. The Marconi company took the contract and very quickly installed the network. Even though much of the technology was very new, the system proved to be very reliable and was in service for over forty years.

In the following years, Marconi continued the trend to the use of shorter wavelengths, exploiting wavelengths of about one metre. This was made easier by the fact that valves for these frequencies were becoming available. However, to prove the viability of these new frequencies, Marconi installed a link between the Vatican and Castel Gandolfo, the Pope's summer residence.

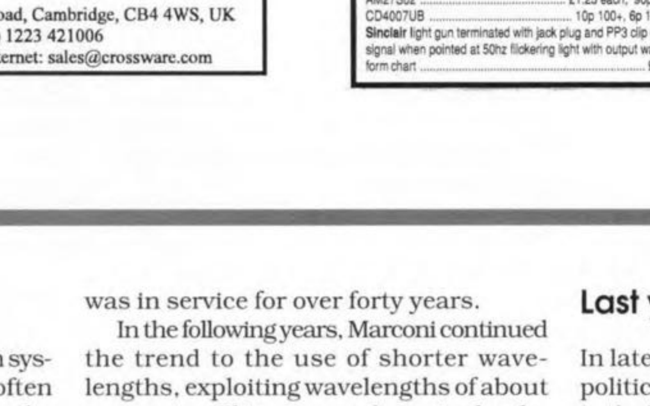


Fig. 5. The inverted 'L' or Marconi aerial.

Last years

In later life, Marconi became involved in politics in his native Italy. He was elected to the Italian Senate and undertook many diplomatic missions for his country. However, his last years were troubled by ill health and he died in July 1937 at the comparatively early age of 63.

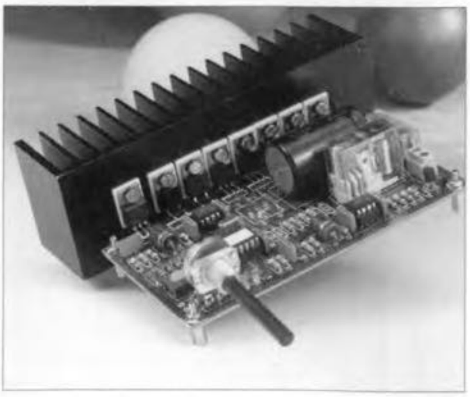
Such was the impression Marconi had made on the technology of radio that on the day after his death, all radio transmissions were silenced for two minutes in his honour.

[950085]

A history of Marconi's Wireless Telegraph Company can be found in *Setmakers*, by Keith Geddes in collaboration with Gordon Bussey, ISBN 0 9517042 0 6 (British Radio and Electronic Equipment Manufacturers' Association, Landseer House, 19 Charing Cross Road, London WC2H 0ES).

FAST NiCd BATTERY CHARGER

Design by A. Rietjens



A microprocessor-controlled charger is described that surely meets the requirements of most users of NiCd batteries. It is fast, suitable for all types of NiCd battery and affords protection against overcharging

The present charger takes into account the many comments on previously published chargers and requests for more facilities from a variety of readers during the recent past.

The charger is fast and offers a choice of two fast-charging periods: 1 h 15 min and 2 h 30 min. It is suitable for charging batteries with a capacity of up to 4 Ah. Up to 10 batteries in series can be charged simultaneously. Overcharging is prevented by the batteries being discharged before charging can take place and by the provision of delta-peak protection. After the charging cycle has ended, trickle-charging keeps the battery fully charged .

The operation of the charger is wholly controlled by a Type ST62T20 microcontroller, which also provides an indication of the modes of operation and any error conditions.

A moving coil meter may be added to give a constant indication of the charging current.

Operation

The charging principle is simple and efficient and is based on the fact that when NiCd batteries are being charged, the danger zone is reached only when the batteries are nearly fully charged. All sorts of temperature effect and pressure effect may then happen, depending on whether the batteries were charged rightly or wrongly.

In the danger zone, the charger ensures that the charging current is brought back to a low level, so that overcharging is prevented.

At the onset the charging current is fairly high, but this is gradually reduced to keep the process within safe limits. Moreover, during the entire charging process, the battery voltage is monitored to show up any drop. Such a drop indicates that the battery is fully charged and that charging should cease. This is the earlier mentioned delta-peak protection.

Basically, the charger supplies a charge for a given fixed time to the batteries. After that times has elapsed, trickle-charging is commenced. Owing to this fixed time, it must be assumed at the onset of charging that the batteries are flat. This is why all batteries are first discharged with a current of 1C until their voltage has dropped to 60% of its initial value.

After the batteries have been discharged, they may be charged in 1 h 15 min or in 2 h 30 min. In the first case, the charging time is divided into 15 min periods and the initial charging current is 2C—see **Fig. 2a**. After each 15 min period, the charging current is halved, except that the current during the fifth period is the same as during the fourth. After the five periods have elapsed, trickle-charging is begun. The total charge is thus

$$2C\cdot{}^1/_4 + 1C\cdot{}^1/_4 + {}^1/_2C\cdot{}^1/_4 + {}^1/_4 C\cdot{}^1/_4 + {}^1/_4C\cdot{}^1/_4 = 1C$$

Those who doubt the robustness of their batteries and therefore prefer to play it safe can choose the longer

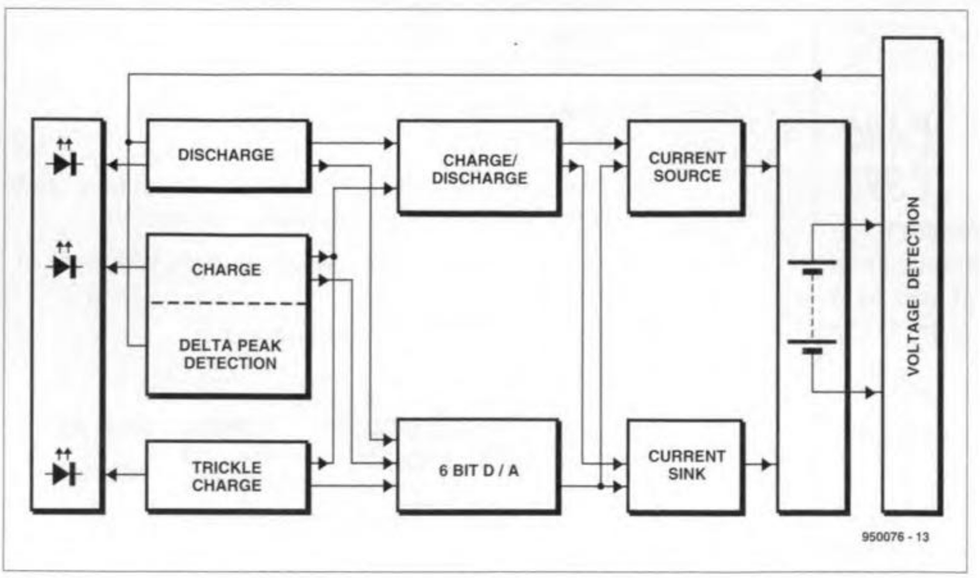


Fig. 1. Block diagram of the charger.

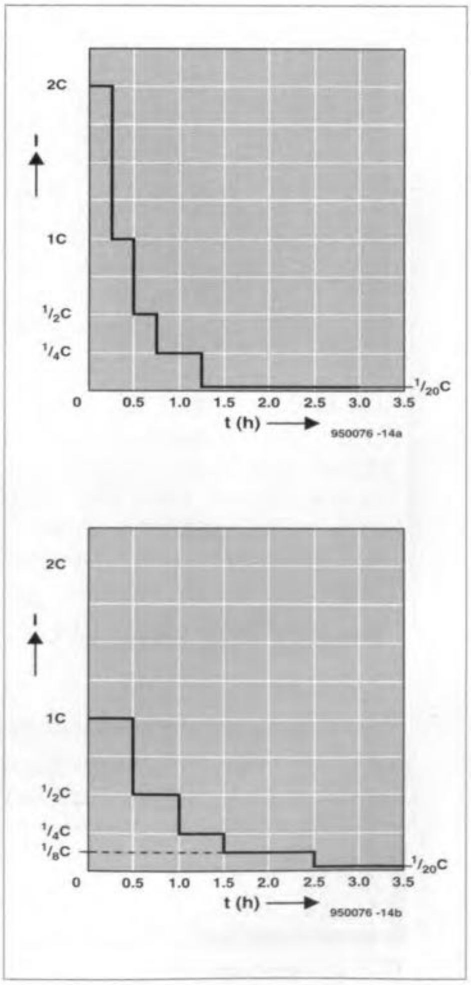


Fig. 2. Charging time vs current characteristic for (top) 1 h15 min charging process and (bottom) 2 h 30 min charging process.

charging period of 2 h 30 min. Again, the time is divided into five periods, but now of 30 min, while the initial charging current is 1C—see **Fig. 2b**. The total charge remains, of course, the same:

$$1C\cdot{}^1/_2 + {}^1/_2C\cdot{}^1/_2 + {}^1/_4C\cdot{}^1/_2 + {}^1/_8C\cdot{}^1/_2 + {}^1/_8C\cdot{}^1/_2 = 1C.$$

Charging at the 2.5 h rate is recommended for batteries that can not cope with heavy charging currents. During the final period, charging is at ${}^1/_8$C, which is very close to the ${}^1/_{10}$C rate which all NiCd batteries can handle.

The block diagram in **Fig. 1** gives a fairly complete picture of the principle of operation. The charger goes through three successive steps: discharging, charging and trickle-charging. For the discharging, a current sink is required and for charging, a current source. In the present charger, both functions are fulfilled by one current source, which is reversed by a relay. Since there is only one current source, a single digital to analogue converter, DAC, suffices to define the current.

During the charging and discharge processes, the voltage of the batteries is monitored constantly. Each step in the process has its own LED indicator to show at all times what is happening. Optionally, a panel meter can be added to give a constant reading of the current.

Circuit description

In **Fig. 3**, the microcontroller, IC_2, controls the relay via T_3, the LEDs indicating discharging, charging and trickle-charging, and the DAC.

The charging current is set in accordance with the battery capacity by the current source (this will be reverted to), and P_1 provides fine adjustment. The charging period is set with S_2. The battery voltage is monitored by IC_4.

The current source consists of IC_3, T_1, T_2 and R_{14}–R_{21}. Depending on the control voltage at pin 3, IC_3 will drive T_1 and T_2 to a smaller or greater degree until the current through R_{14}–R_{21} reaches a level at which the potentials at pins 2 and 3 are equal. The control voltage is generated by the DAC, which here consists simply of a number of resistors. The MSB is determined by PB_2 and the LSB by PB_7. When only the MSB is high, the voltage at pin 3 of IC_3 will be about 0.5 V. When all the bits are high, there will be potential of 1 V. The current through each of resistors R_{14}–R_{21} is then 0.5 A or 1 A respectively. Changing the number of resistors will thus enable the maximum current through the source to be varied.

Apart from with R_{14}–R_{21}, the current may also be varied with P_1. The potential at the wiper of this potentiometer may be varied from 0 to 2.5 V or from 2.5 V to 5.0 V, depending on the position of S_2. Turning the potentiometer allows variation of 50% in the capacity of the battery to be charged, which is independent of S_2. This switch only determines whether charging will be for 75 min or 150 min. If only one resistor of R_{14}–R_{21} is fitted, the current can be varied with P_1 such that batteries with capacities of 250–500 mAh can be charged. With each added resistor, the range of capacities is extended by the same value, that is, three resistors give a range of 3× 250 = 750 mAh to 3×500 = 1500 mAh. See **Table 1**.

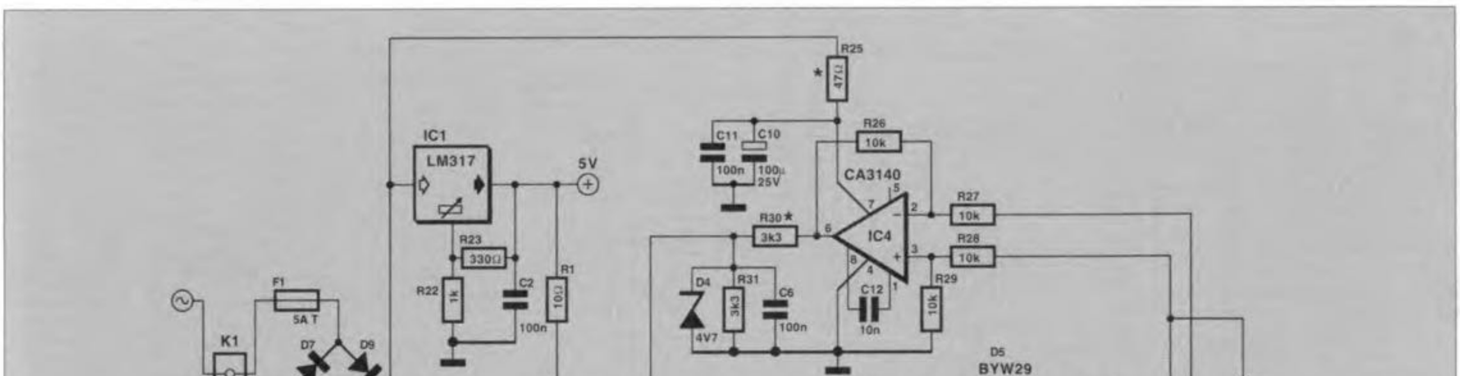


Fig. 3. Circuit diagram of the battery charger.

The optional 100 µA panel meter, M_1, enables the current to be monitored continuously. It should, of course, be borne in mind that the current is determined by the number of resistors. When a battery capacity is selected with P_1, the meter can be set P_2 to half full-scale deflection (f.s.d.) during discharging of the battery. This makes it possible to follow the charging process on the meter. The meter may be given a new scale extending from $^1/_8$C to 2C.

The battery voltage is measured by IC_4, a differential amplifier with unity gain. Thus, the output of IC_4 carries the battery voltage, plus or minus the drops across the parasitic resistances in the sockets. These parasitic resistance can cause difficulties with complex charging processes. However, in the present charger, it is sufficient if the voltage applied to IC_2 via potential divider R_{30}-R_{31} remains below 4.5 V during the entire process. This is because the software does not consider absolute voltage levels, but percentage differences ($^5/_8$ of the initial voltage during discharging and $^1/_{40}$ of the peak voltage in the delta-peak detection). This is elaborated on in **Table 2**.

Software program

An overview of the software program is given in the flow diagram in **Fig. 5**. This shows the entire process from start to finish.

It starts with the question whether there are batteries in the holders and whether these are inserted properly (BATTERY?) If so, discharging is commenced (RELAY ON/SET DISCHARGE) and D_1 is switched on.

Subsequently, a small charginf current flows to enable a deeply discharged battery to recover.

When during discharging the voltage measurement shows that the battery voltage has dropped to $^5/_8$ of the initial value, the relay is changed over and charging begins (RELAY OFF/SET CHARGE), whereupon D_2 lights. The manner of charging is then selected in accordance with the chosen setting: 1 h 15 min or 2 h 30 min. That choice immediately determines the situation during the first period: 30 min at a current of 1C or 15 min at 2C. The current is halved ($I = I/2$) until the fifth period is reached. A delta peak check is carried out during the entire charging process. At the end of period 5 ($I = {}^1/_{20}$C), trickle-charge indicator D_3 lights.

Note that each of the three LEDs has two functions. Their main functions are indicating the discharge, charging or trickle-charging mode,

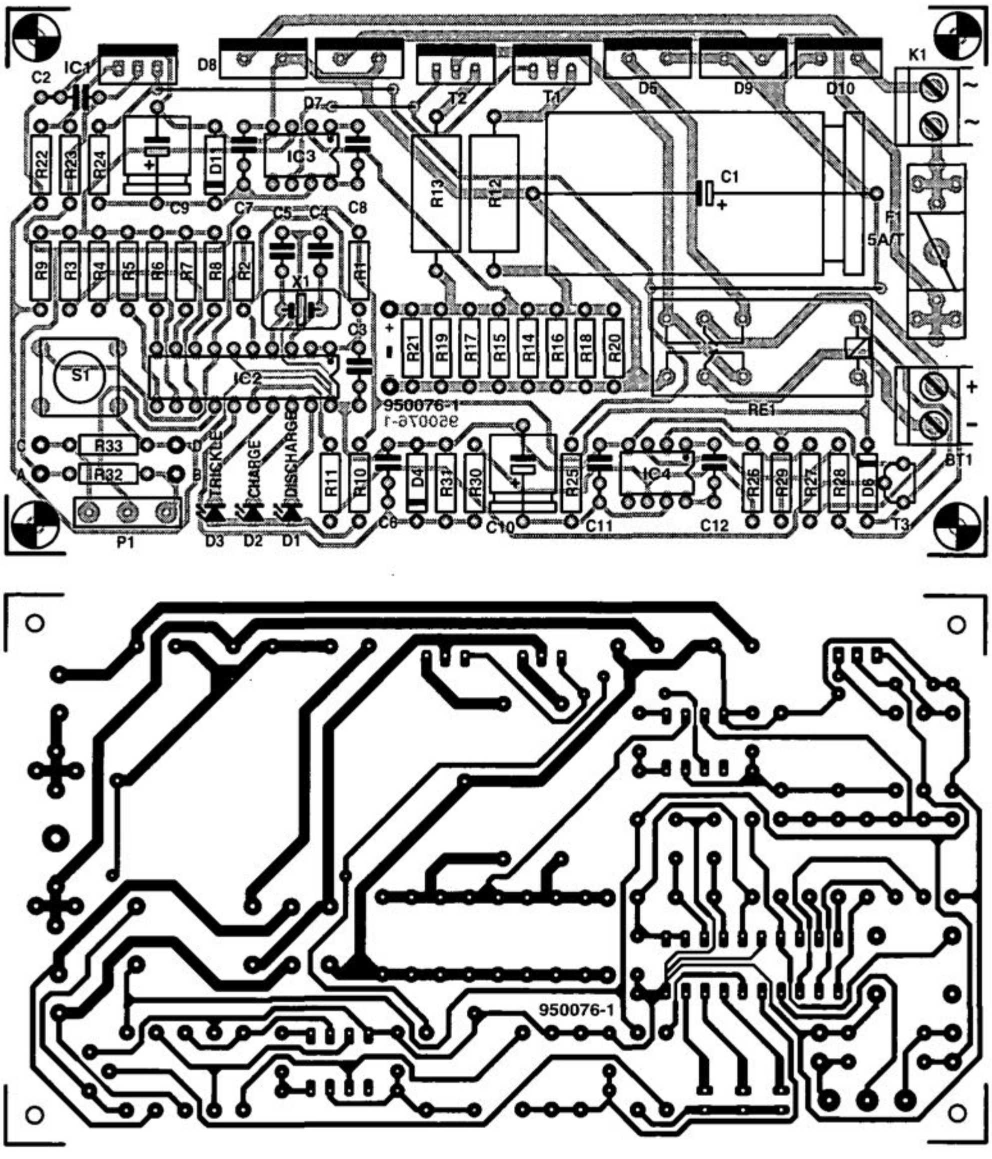


Fig. 4. Printed-circuit board for the battery charger.

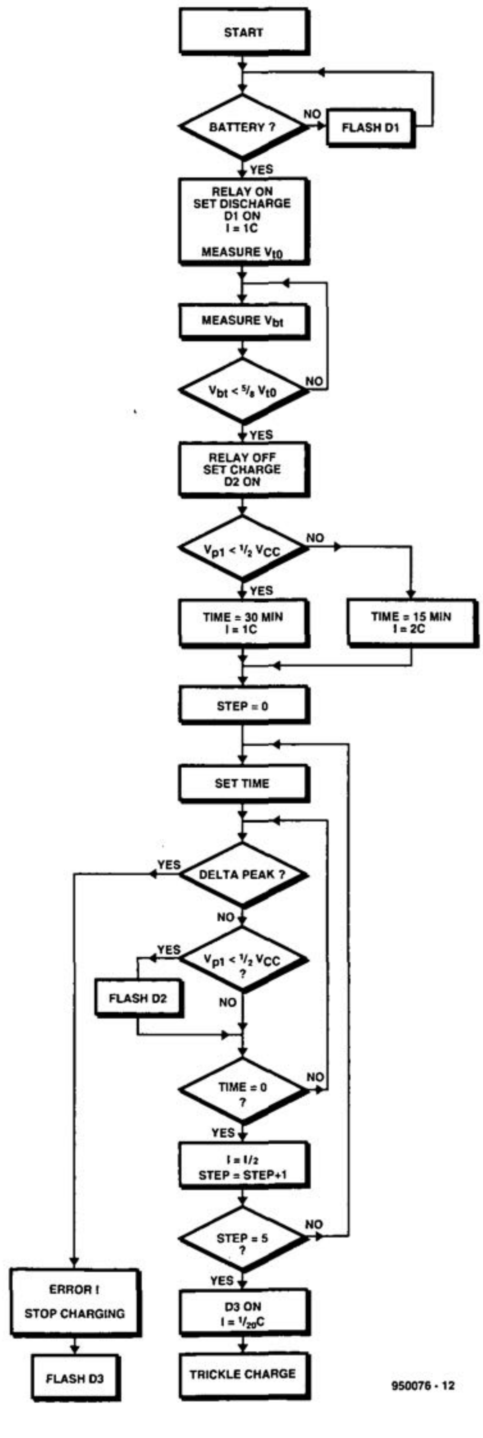


Fig. 5. Flow diagram of the software.

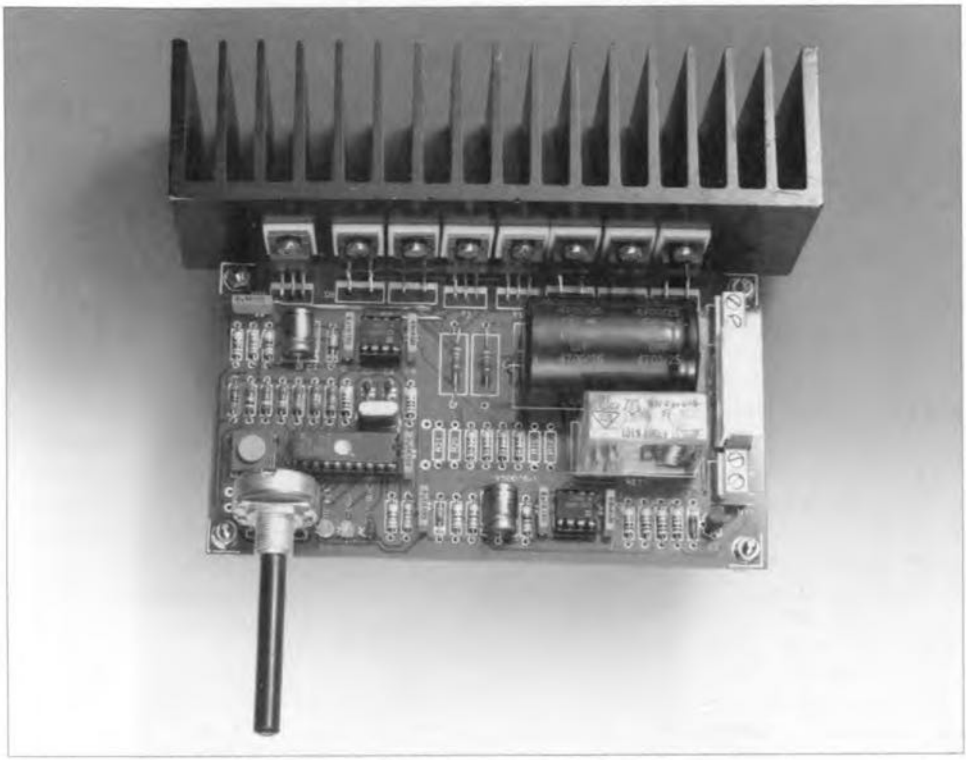


Fig. 6. The completed prototype board with heat sink.

when they light continuously. When they flash, something is happening. When the answer to question "BATTERY?" is negative, D_1 flashes. When a charging time of 2 h 30 min is chosen, D_2 flashes. When the delta-peak protection detects a voltage drop, charging is stopped and D_3 flashes.

Construction

The charger circuit is intended to be built on the printed-circuit board in **Fig. 4**. Populating the board is straightforward, although one of the long sides needs some extra attention. This side houses all the components that dissipate more than the usual heat: T_1, T_2, D_5, D_7–D_{10} and IC_1. Cooling of these components when large capacity batteries are charged at the 1 h 15 min rate is no luxury. Since they are at the side of the board, it is a fairly simple matter of fitting them all one large heat sink (see **Fig. 6**). Note that insulating washers should be used in all cases.

The secondary of the mains transformer is connected to terminal block K_1. See Table 2 for various secondary voltages. A correct choice of transformer is important so as to minimize the dissipation. It is advisable to take a value equal to the maximum supply voltage divided by 1.4. The current drain from the transformer is shown in Table 1: it is advisable to round off the values shown upward. For example, for charging four batteries with a capacity of 500–1000 mAh, the tables show that a supply voltage of 12–14 V and a current of 2 A are required. This requires a secondary voltage of 10 V and a current of 2.5 A. If, according to the table, a peak current of 4 A is required, it is advisable to rate the secondary at 5 A.

Whether to use two, four or eight 1 Ω resistors in the R_{14}–R_{21} positions is entirely dependent on the capacity of the batteries—see Table 1.

The values of resistors R_{24} and R_{30} depend on the supply voltage, that is, the number of batteries-see Table 2.

If varying numbers of battery are to be charged, the values for the largest expected number in Table 2 should be used. The minimum voltage should be used, however. For instance, for 4, 5 or 6 batteries, the supply voltage should be 14 V, R_{30} becomes 6.8 kΩ and R_{24} becomes 49 Ω.

The holder for the batteries to be charged is connected to terminal block BT_1.

It may be a good idea to give the indicator diodes different colours (red, green and yellow, for instance).

Potentiometer P_1 can be fitted directly on to the board. The connecting points for charging time selector S_3 are located directly behind P_1.

If the optional moving-coil meter is used, this should be connrected, together with R_{34} and P_2 to '+' and '–' between C_3 and R_{21}.

Assembly and usage

The choice of an enclosure will depend to a large extent on the shape and dimensions of the required mains transformer. Fortunately, there is a wide choice of suitable enclosures. Whatever, it is important to ensure that the heat sink can radiate heat freely.

The holder for the batteries is best mounted on top of the enclosure, but it may also be kept separate from the enclosure and be connected to it by a short cable (and plug and socket).

Light-emitting diodes D_1–D_3 must,

Resistor(s) fitted	Battery capacity (mAh)	Charging time (h)	Maximum current (A)
R_{14}	250–500	1.25 or 2.5	1
R_{14}, R_{15}	500–1000	1.25 or 2.5	2
R_{14}–R_{17}	1000–2000	1.25 or 2.5	4
R_{14}–R_{21}	2000–4000	2.5	4

Table 1. The charging current depends on the capacity of the batteries and is decided by the number or parallel-connected resistors R_{14}–R_{21}.

Number of batteries	R_{30}	R_{24}	Supply voltage (V) Minimum	Maximum
1	100 Ω	49 Ω	7	8
2	100 Ω	49 Ω	9	10
3	1.5 kΩ	49 Ω	11	12
4	3.3 kΩ	49 Ω	12	14
5	4.9 kΩ	49 Ω	12	16
6	6.8 kΩ	49 Ω	14	18
7	8.2 kΩ	49 Ω	16	20
8	10 kΩ	150 Ω	18	22
9	12 kΩ	270 Ω	20	24
10	15 kΩ	390 Ω	22	26

Table 2. The number of batteries in series dictates the supply voltage.

of course, be fitted in such a way that they are freely visible from the outside. This means that in most cases they can not be soldered on to the PCB, but must be connected to it via flexible insulated circuit wire. The same applies, of course, to P_1, S_1, and S_2. As regards S_1, note that although in the prototype this was a PCB type switch, it may be advisable to use a standard push-button type and fit this on the front panel.

Use of the charger is straightforward: insert the battery or batteries into the holder, switch on the mains and leave well alone for the selected charging time. To eliminate any likelihood of error, it is good practice to start the discharge/charging cycle by pressing reset switch S_1. If everything is in order, the red LED will light. If the batteries to be charged are really flat, the discharge period will be fairly short, which means that the red LED will go out quite quickly, whereupon the yellow LED will light.

Finally, since the charging current range set by R_{14}–R_{21} can be varied by a factor 2 with P_1, it is necessary to set P_1 to the exactly required level on the (optional) front panel meter. If this is not used, measure the current with an ammeter in series with the batteries or measure the voltage across R_{14}–R_{21}.

Battery capacity

The capacity of a rechargeable battery is normally stated on its case in mAh or Ah. Charging and discharge currents are relative to this capacity. A current that discharges a battery in one hour is, therefore, indicated by 1C(apacity). A current of 2C thus discharges a fully charged battery in half an hour.

A current of $^1/_{10}$C is normally taken as a safe charging current. This level of current needs to flow into a discharged battery for 14–16 hours to recharge the battery fully. Even if this current flows for a much longer time, the battery does not become overcharged. A current of $^1/_{20}$C is normally used for trickle-charging. Thus, for a battery of (1000 mAh), C = 1 A; 2C = 2 A; $^1/_2$C = 500 mA; and $^1/_{10}$C = 100 mA.

Fast charging is carried out with an appreciably higher current and can only be used with batteries that are suitable (not all are: see data sheet). Also, some battery holders have a resistance that interferes with the charging process, especially when the charging current is a pulse train.

Finally, keep an eye on the dissipation, particularly when the number of batteries to be charged is small, or the number of series-connected batteries being discharged is large.

Parts list

Resistors:
R_1 = 10 Ω
R_2, R_{26}–R_{29}, R_{32}, R_{33} = 10 kΩ
R_3 = 9.09 kΩ, 1%
R_4 = 19.1 kΩ, 1%
R_5 = 39.2 kΩ, 1%
R_6 = 78.7 kΩ, 1%
R_7 = 158 kΩ, 1%
R_8 = 316 kΩ, 1%
R_9 = 1.0 kΩ, 1%
R_{10} = 470 Ω
R_{11} = 680 Ω
R_{12}, R_{13} = 0.33 Ω, 2 W
R_{14}–R_{21} = see text and Table 1
R_{22} = 1 kΩ
R_{23} = 330 Ω
R_{24} = see text and Table 2
R_{25} = 47 Ω
R_{30} = see text and Table 2
R_{31} = 3.3 kΩ
R_{34} = 2.7 kΩ (optional)

Potentiometers:
P_1 = 10 kΩ
P_2 = 5 kΩ preset (optional)

Capacitors:
C_1 = 4700 µF, 25 V
C_2, C_3, C_6, C_7, C_{11} = 100 nF
C_4, C_5 = 22 pF
C_8, C_{12} = 10 nF
C_9, C_{10} = 100 µF, 25 V

Semiconductors:
D_1 = LED, 3 mm, red
D_2 = LED, 3 mm, yellow
D_3 = LED, 3 mm, green
D_4 = zener, 4.7 V, 500 mW
D_5, D_7–D_{10} = BYW29
D_6 = 1N4148
D_{11} = zener, 15 V, 1 W
T_1, T_2 = BUZ10
T_3 = BC547B

Integrated circuits:
IC_1 = LM317
IC_2 = ST62T20 (Order no. 956509 see p. 70)
IC_3 = CA3160
IC_4 = CA3140

Miscellaneous:
K_1, BT_1 = 2-way terminal block, pitch 5 mm
S_1 = push-button switch
S_2 = double-pole change-over switch
X_1 = crystal 8 MHz
Re_1 = relay, 6 V, 5 A with two change-over contacts
F_1 = fuse, 5 A, slow with holder
M_1 (optional) = moving =-coil meter, 100 µA
Heat sink, SK411SA/50 or SK85SA/50 (Dau)
Mains transformer = see Tables 1, 2
PCB Order no. 950076 (see p. 70)

[950076]

Parameters

- Fast charging in 1 h 15 min or 2 h 30 min.
- Automatic discharge before charging can commence.
- Charging monitored by delta-peak protection circuit.
- Suitable for batteries with a total capacity of up to 4 Ah.
- Operation affected minimally by parasitic resistance of battery holder.
- Circuit adjusted to accord with battery capacity with potentiometer.
- Simple, single-sided printed-circuit board.
- When charging completed, automatic transfer to trickle-charging.
- Modes of operation (charging, discharging, trickle-charging) indicated by LEDs.
- Bad contact indicated by LED.
- Error indication by LED.

APPLICATION NOTE

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

Fault-tolerant CMOS multiplexers

A Maxim Application

All analogue multiplexers act as the solid-state equivalents of digitally controlled rotary switches. But multiplexers at the interface between an electronic system and its environment have an additional function—they act as insurance policies against malfunction and damage.

Designers may control the system, but input multiplexers contend with everything outside the system—poor installation, careless operators, and electrical noise. They must operate correctly in the presence of ground loops, electrical interference from motors and engines, and unintentional inputs such as 240 V a.c.

In the following discussion, the serial-MOSFET switch emerges as the most economical choice in overcoming these problems. Other designs offer overvoltage protection alone, but only the serial-MOSFET approach combines overvoltage and fault tolerance without the need for external components.

Multiplexer construction

The switches in common multiplexers have been designed the same way for nearly twenty years: each consists of an n-channel and p-channel MOSFET connected in parallel on a silicon substrate, and driven with opposite-polarity gate-drive voltages (**Fig. 1a**). This connection provides a symmetrical signal path through the parallel source-to-drain resistances, producing a characteristic double hump in the curve of on-resistance vs input voltage (**Fig. 1b**). (Many designs minimize this effect by driving the body connection of the n-channel MOSFET with signal voltage).

The presence of each device polarity guarantees that at least one of the two MOSFETs will conduct for any input voltage between the supply rails. Thus, the multiplexers can handle any signal level that falls between the rails.

A multiplxer switch ceases to be a switch, however, when signal voltage exceeds either supply rail. Each switch includes two parasitic diodes, intrinsic to the MOSFET source and drain structures, which provide current paths to the rails (**Fig. 2**). Both diodes are reverse-biased during normal operation, but any signal excursion beyond the rails applies forward bias to one of the diodes, clamping the signal at 600 mV beyond the rail. Because the diodes are present when power is removed, they also clamp (at ±600 mV) when the rails are at zero volts.

Parasitic diodes provide a useful clamping function, but they also introduce problems. Excessive current in the diodes can cause overheating and damage in the signal souce as well as the multiplexer (**Fig. 3**). Somewhat lower levels of current (below that of overheating and damage) can still cause latchup in the multiplexer. And once it crosses a diode junction, the fault current becomes a flow of injected minority carriers that 'spray' into the silicon substrate. Collected by other switching devices, this current can induce an error voltage in every channel.

Turning on a parasitic diode clamps the multiplexer output to one supply

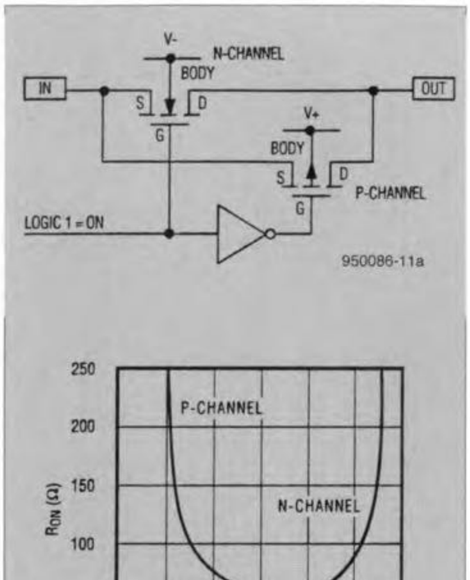


Fig. 1. The traditional CMOS switch is a transmission gate (a), whose on-resistance vs signal voltage characteristic exhibits a double hump as shown in (b)

rail—an action that can damage external circuits connected to that output. The cause of damage may not be obvious, but an output transient (to the rail) caused by momentary overvoltage at the multiplexer can destroy an analogue-to-digital converter's input, or cause differential overload and long settling times in an op amp.

Protecting the multiplexer

Several design measures offer protection for a CMOS multiplexer and its associated external circuits. These measures include connecting a resistor in series with each channel input, connecting diode-resistor networks to control the fault effects, and choosing a multiplexer whose architecture and process technology provide fault-tolerant properties.

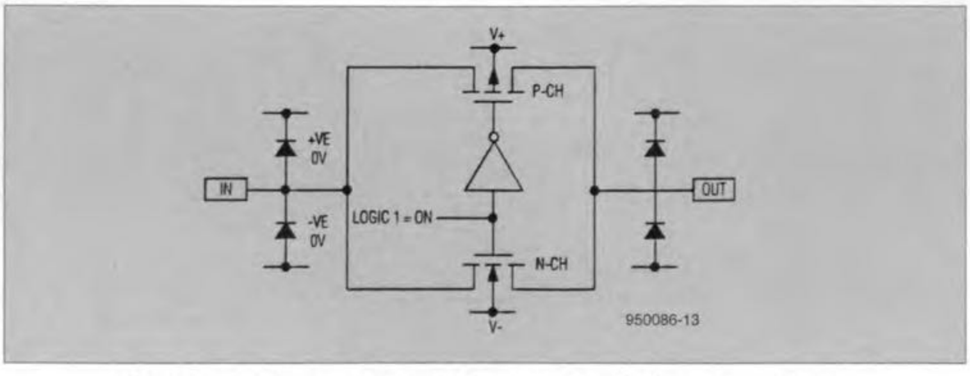


Fig. 2. Parasitic diodes provide a path for fault current when a conventional analogue switch is exposed to overvoltage.

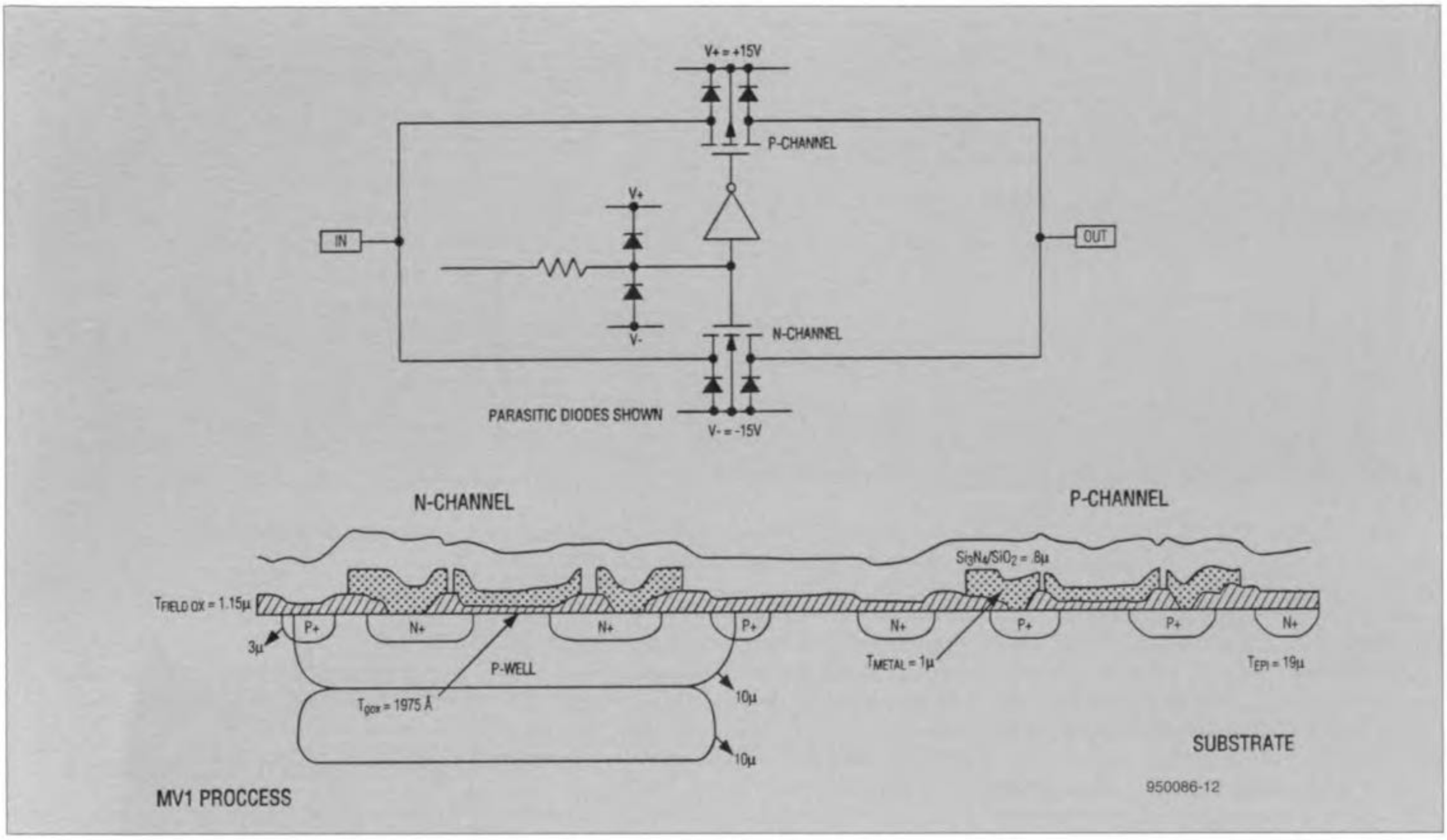


Fig. 3. A closer look at the transmission-gate MOSFETs shows parasitic diodes tied to the supply rails.

The simplest form of protection adds series resistors that work in conjunction with the internal protection diodes (**Fig. 4**). The resistor values are critical because they present a tradeoff: larger values give more protection, but introduce greater signal errors.

Leakage current into the multiplexer also flows through the series resistors, causing an error voltage that worsens with temperature (the leakage doubles for each 8 °C increase above ambient). Lowering the resistor values can reduce this error to an acceptable level, but the lower value may allow too much diode current, threatening latchup in the multiplexer. As a rule, unless otherwise specified in the data sheet section on absolute maximum rating, you should limit the diode current to 20 mA continuous or 40 mA peak.

Low leakage currents can offset this drawback of large protection resistors. New, ultra-low-leakage multiplexers like Maxim's MAX328 and MAX329 have extended the design limits for series-resistor protection over those for earlier-generation multiplexers. The new devices' low leakage (±1 pA at 25 °C; ±20 nA at 125 °C) allows bvery high-valued protection resistors. Resistors of 150 kΩ, for instance,

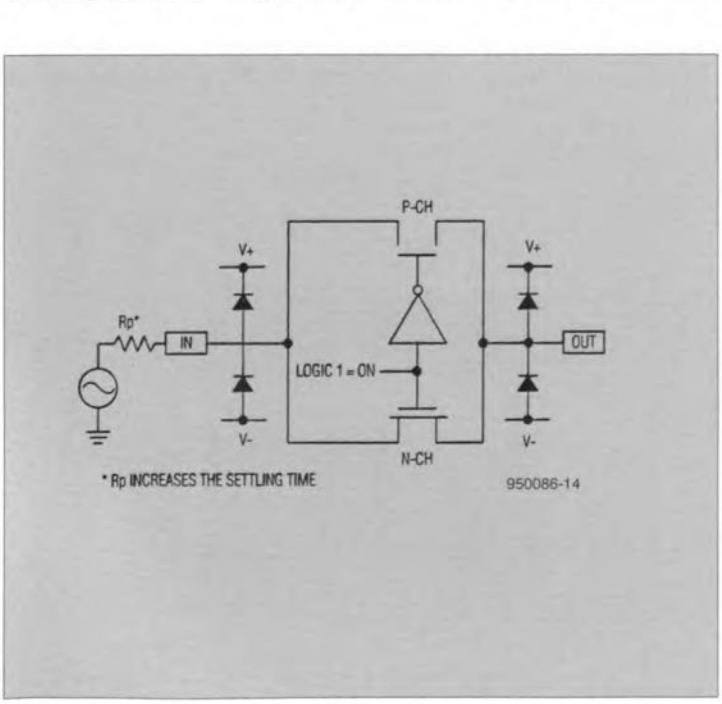


Fig. 4. Adding a series resistor to the switch of Fig. 2 limits fault current, but also adds to the switch resistance and lengthens settling time.

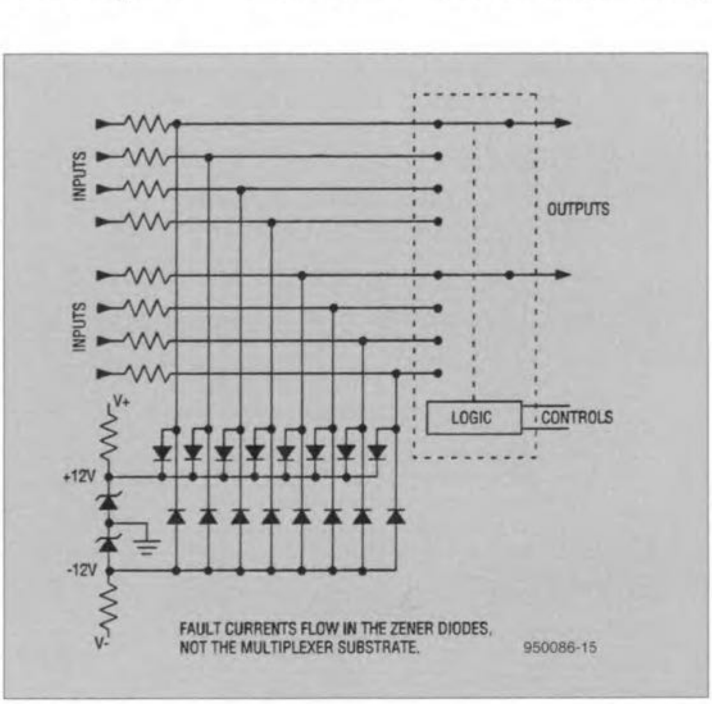
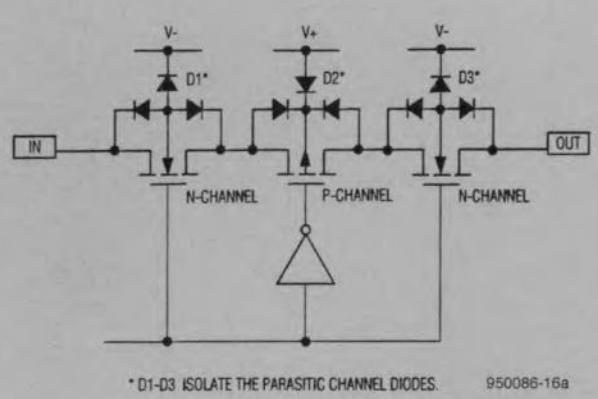


Fig. 5. Fault protection for a conventional multiplexer entails current-limiting resistors, two zener diodes for a bipolar clamp-voltage network, and dual clamp diodes for each channel.



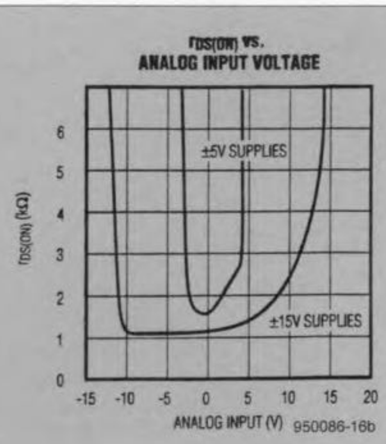


Fig. 6. The 3-MOSFET switch element of a fault-tolerant multiplexer (a) has an on-resistance characteristic shaped like a bathtub (b)

admit fault currents of only 1 mA while withstanding ±150 V inputs. At ±1500 V, they admit fault currents of only ±10 mA. The resistors produce only ±3 mV of additional error at 125 °C.

Note that ±1500 V protection resistors require 15 W ratings for continuous duty. But, in most applications, you can scale this thermal rating considerably, because the overvoltage has a much lower duty factor. External resistors thus offer flexibility—you can choose different resistor values for different channels in the same device, and scale their power ratings as required. Integrated resistors, on the other hand, are constrained by their package power rating; this rating may limit the number of channels that can withstand overvoltage at the same time.

The series-resistor approach protects the multiplexer, but it doesn't prevent corruption of signals in the selected channel. Thesesignals are at the mercy of overvoltage in any of the unselected channels. But the direct cause isn't overvoltage; it's fault current (the inkjected minority carrier mentioned earlier) flowing into the substrate via one or more protection diodes. Eliminate that substrate current and you eliminate the gross signal errors.

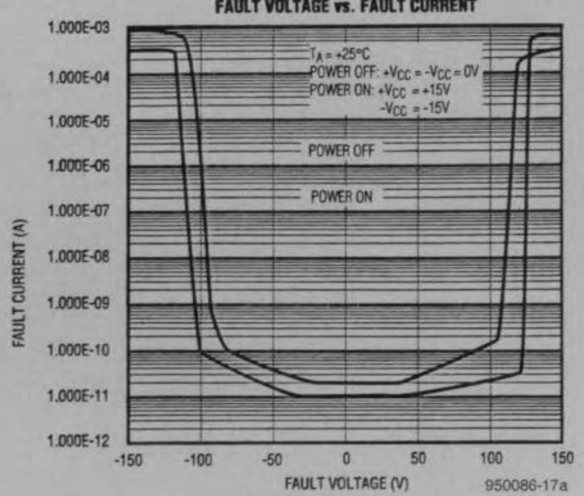
One way to handle the fault current is to divert it into an external network (**Fig. 5**). Two zener diodes produce ±12 V clamp levels, centred within the multiplexer's ±15 V supply rails. Then, instead of flowing through an internal protection diode, the fault current owing to overvoltage in any channel flows through one of the two external protection diodes for that channel.

Though it offers excellent protection, this technique requires alarge number of external components. Moreover, the external diodes produce additional leakage current that precludes use of the high-valued series resistors discussed earlier. The external components represent extra board space, not to mention the cost of purchase, test, through-hole aseembly, and inventory. A better solution is to integrate this protection with the multiplexer, on a single chip.

Fault-tolerant multiplexers

Fault-tolerant multiplexers require no external components, yet are capable of withstanding high levels of overvoltage without corresponding high levels of fault current. They achieve this protection with an internal design that is quite different from that of conventional multiplexers.

Each switch in a fault-tolerant multiplexer is actually a series connection of three MOSFETS, in the order n-channel/p-channel/n-channel (**Fig. 6a**). Internally generated drive voltages turn the switch on by simultaneously driving the n-channel gates to the positive rail and the p-channel gate to the negative rail. The switch then remains on for as long as the analogue signal (which modulates the gate-source



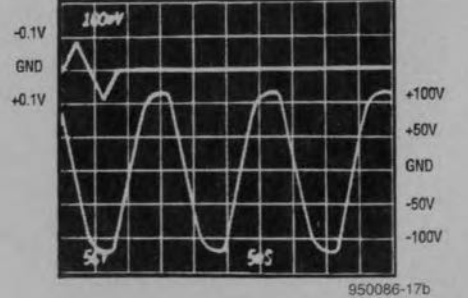


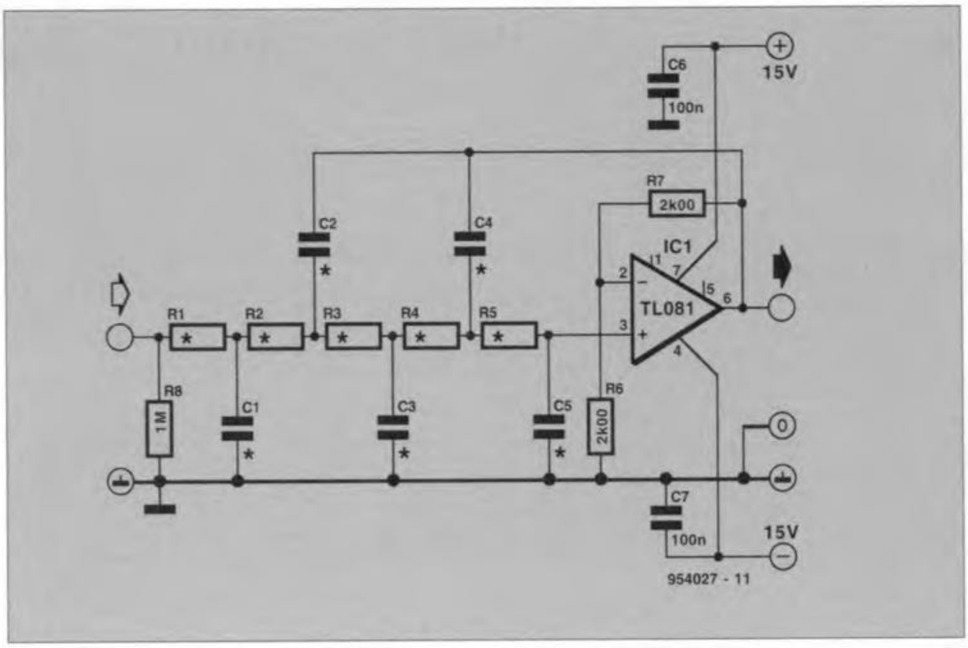
Fig. 7. The onset of avalanche (fault) current in a MAX388 multiplexer defines a fault-tolerant region of approximately ±100 V (a). A 0 V signal in the selected channel (b, top trace) is unaffected by ±100 V applied to an off channel.

5TH-ORDER BESSEL FILTER

Bessel filters have a fairly constant transit time (delay) over the pass band and their characteristic thus shows no ringing. This means, however, that the characteristic starts falling off much earlier than in, say, a Butterworth section. Also, it takes much longer for the slope of the characteristic to reach the theoretical value of 30 dB/octave; in a Butterworth filter this occurs almost immediately after the cut-off point.

The table gives various values for the capacitors and resistors. The cut-off point for each combination is 1 kHz. The last line of the table gives theoretical values for the capacitors for equal values of the resistors – these are required for conversion to a high-pass filter.

Design by T. Giesberts
[954027]



	C_1 (nF)	C_2 (nF)	C_3 (nF)	C_4 (nF)	C_5 (nF)	R_1 (kΩ)	R_2 (kΩ)	R_3 (kΩ)	R_4 (kΩ)	R_5 (kΩ)
1	10	4.7	6.8	3.3	1.5	25.036	20.688	25.239	20.291	21.724
2	12	5.6	8.2	3.9	1.8	20.375	17.658	20.727	17.333	18.214
3	15	6.8	10	4.7	2.2	16.036	14.552	17.015	14.495	15.003
4	18	8.2	12	5.6	2.7	13.288	12.168	13.990	12.372	12.151
5	22	19	15	6.8	3.3	10.486	10.313	11.060	10.245	10.036
6	27	12	18	8.2	3.9	8.587	8.456	9.359	8.395	8.560
7	33	15	22	10	4.7	7.223	6.668	7.621	6.836	7.092
8	39	18	27	12	5.6	5.955	5.736	6.140	5.686	5.996
9	47	22	33	15	6.8	5.030	4.644	5.084	4.477	4.922
10	56	27	39	18	8.2	4.416	3.689	4.274	3.734	4.023
11	68	33	47	22	10	3.709	2.976	3.558	3.049	3.277
12	82	39	56	27	12	3.093	2.478	3.063	2.451	2.727
13	22.671	10.690	16.137	6.9535	3.3489	10.0	10.0	10.0	10.0	10.0

voltages) remains within limits set by the n- and p-channel gate-source thresholds.

Typical gate-source thresholds are 1.5 V for n-channel devices and 3 V for p-channel devices. Therefore, with ±15 V supplies, the thresholds confine a multiplexer's input signals to the range –12 V to +13.5 V. Because one of the three MOSFETs in a switch begins to turn off as the signal exceeds either limit, the switch-on resistance versus input voltage assumes a characteristic 'bathtub-shaped' curve (**Fig. 6b**).

The resulting high impedance in the off state is very convenient: the switch is off, blocking the overvoltage, and fault current is virtually zero. Substrate (fault) current flows only as a resultmof avalanche, which occurs when the overvoltage exceeds a limit set by the MOSFET's geometry and doping levels.

Below the avalanche limit, signals in the selected channel remain unaffected because the overvoltage produces no substrate current (**Fig. 7**). The series-connected switch also turns off when power is removed. This behaviour simpl;ifies the design of redundant systems, because multiplexers connected to common signal lines can be powered down without loading the lines.

The Maxim range of series-structure multiplexers includes the MAX358 and MAX359 devices (1-of-8 and dual 1-of-4) which withstand overvoltages to ±35 V, abd the similar MAX378 and MAX379 that withstand overvoltages to ±75 V. The MAX368 and MAX369 add latched address inputs to the basic 35 V-tolerant models, and the new MAX388 and MAX389 are latched models that withstand ±100 V. The non-latched devices are pin-compatible with industry-standard multiplexers DG508/509; the latched devices are pin-compatible with the latched models DG528/529.

Reference: *Maxim Engineering Journal, Vol. 12*. Maxim Integrated Products (UK) Ltd., 21C Horseshoe Park, Pangbourne, Reading RG8 7JW, Great Britain. Telephone (01734) 845 255. Fax (01734) 843 863.

Maxim products are available, among others, from 2001 Electronic Components (see advert on p. 67).

[950086]

5TH-ORDER BESSEL FILTER

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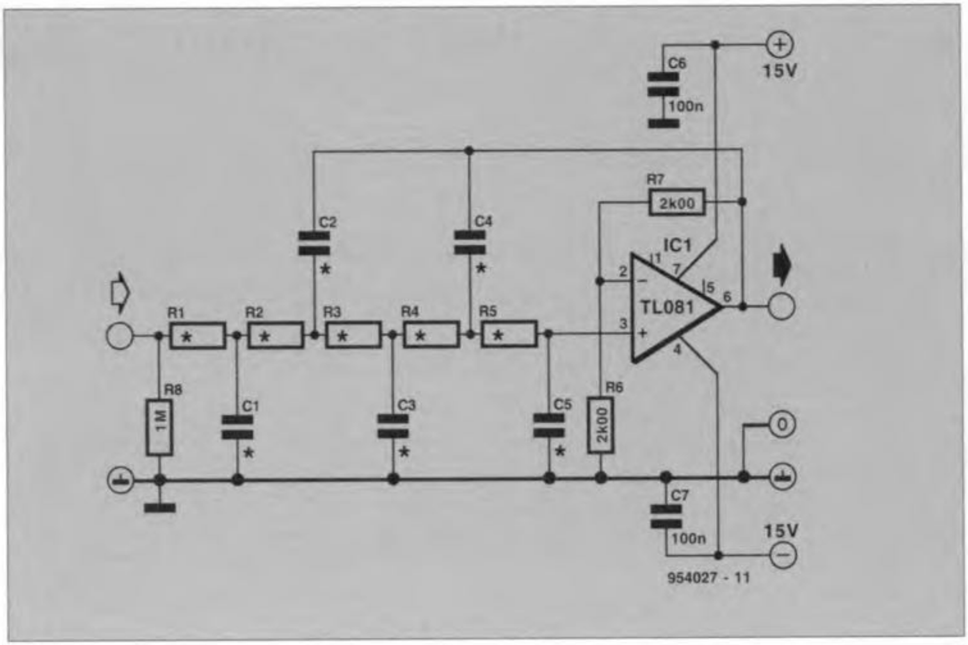
Design by T. Giesberts
[954027]



	C_1 (nF)	C_2 (nF)	C_3 (nF)	C_4 (nF)	C_5 (nF)	R_1 (kΩ)	R_2 (kΩ)	R_3 (kΩ)	R_4 (kΩ)	R_5 (kΩ)
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4	18	8.2	12	5.6	2.7	13.288	12.168	13.990	12.372	12.151
5	22	19	15	6.8	3.3	10.486	10.313	11.060	10.245	10.036
6	27	12	18	8.2	3.9	8.587	8.456	9.359	8.395	8.560
7	33	15	22	10	4.7	7.223	6.668	7.621	6.836	7.092
8	39	18	27	12	5.6	5.955	5.736	6.140	5.686	5.996
9	47	22	33	15	6.8	5.030	4.644	5.084	4.477	4.922
10	56	27	39	18	8.2	4.416	3.689	4.274	3.734	4.023
11	68	33	47	22	10	3.709	2.976	3.558	3.049	3.277
12	82	39	56	27	12	3.093	2.478	3.063	2.451	2.727
13	22.671	10.690	16.137	6.9535	3.3489	10.0	10.0	10.0	10.0	10.0

THE DIGITAL SOLUTION

Part 8 – Moving data

By Owen Bishop

In this series we look closely at digital electronics, what it is, what it does, how it works, and its promise for the future.

A recent example of the benefits of the digital solution is the catalogue with data sheets published by RS Components as a CD-ROM. At present, customers receive both the new-style CD-ROM and the old-style catalogue of nearly 2000 pages weighing nearly 2.8 kg, but we might assume that eventually the book-form catalogue will be phased out with a considerable saving in production costs, postage, shelf-space, and woodpulp (an environmental bonus). The CD-ROM version also has the benefit that it can print orders and, if the computer has a modem, fax the order directly to the RS Components local depot.

Taking the word 'computer' to include mobile telephones, fax machines and any other 'smart' device which sends or receives digital data, this month's article describes how we move data along a communication channel from one computer to another. Moving data in binary digital form brings advantages and raises problems. The fact is that communication by digital means is rapidly expanding relative to analogue communication so presumably the advantages outweigh the problems.

Transmitting data

There are three main channels of communication: copper wire, radio and optical cable. Copper wires are almost invariably used for short distances, such as between a computer and a printer, or between a telephone in the home and the local telephone exchange. In BTs experimental 'Video on demand' in Ipswich, England, which also offers services such as shopping, banking, and games, it has been found that the existing telephone wires are suitable for digital signalling over typical distances betweena subscriber's house and the local telephone exchange. For longer distances, optical fibre has largely replaced the wires. For international communications, both radio (especially satellite radio) and optical fibre are dominant. We will look at each of these channels in turn.

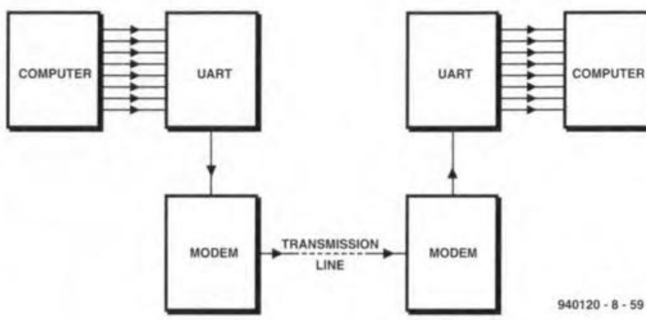


Fig. 60

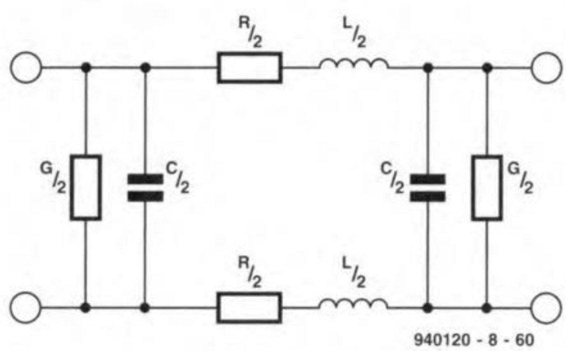


Fig. 61

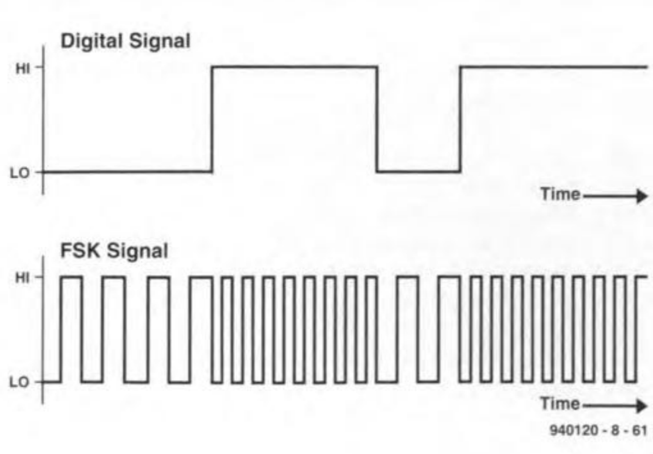


Fig. 62

One way of sending a digital signal along a pair of wires is to transmit a succession of high and low voltages, representing 1s and 0s. In a cable connecting a printer to a personal computer, the high level is 5 V and the low level is 0 V. In order to speed up the transmission of data so that minimum time is required to load the printer's buffer memory, data is sent a byte at a time through eight parallel wires, each wire carrying one bit. The printer cable also has wires for sending control pulses between the computer and printer, a procedure known as *handshaking*. This *parallel* transmission of high-low data is limited to short distances. One reason is that it is four times as expensive to run eight lines over a given distance than it is to run two lines. Also, in the case of transmission over the telephone network, twin wiring already exists and it is uneconomic not to use it. For this reason, the first stage in transmitting data to a communication channel is to convert it from parallel to serial form. Instead of the eight bits being sent simultaneously on eight separate lines, they are sent one after the other on a single line. Serial transmission runs at approximately one eighth of the speed of parallel transmission, but the economy of using only two wires prevails.

One frequently-used plan for transmitting data serially from computer to computer is illustrated in **Fig. 60**. Thenunit that converts parallel data to serial data or to reverse the process at the receiving end is known as a *universal asynchronous receiver/transmitter*, abbreviated to *UART*. Essentially, it consists of a register into which data is loaded eight bits at a time and unloaded bit by bit. Or, in the reverse operation at the receiver, serial data arriving a bit at a time is accumulated in a register until the register is full, whereupon the bits are fed out in parallel along an eight-line cable. The UART also contains circuitry for dealing with timing and other control operations.

Voltage levels on a line of any appreciable length are subject to attenuation and other changes (see box), so that sending signals for long distances along telephone lines or other wired connections as a series of high and low voltages is not the preferred method. They are sent as a continuous tone, the frequency of which is altered to indicate 1s and 0s. This is known as *frequency shift keying*, *FSK*, and is performed by a unit know as a *modem*. This receives a serial train of pulses from a UART and emits one of two tones. While it is receiving a low voltage (0 V = logic 0), it produces a 1070 Hz signal—see **Fig. 62**. While it is receiving a high voltage (5 V = logic 1), it emits a 1270 Hz signal. We say that the signal is 1170 Hz *modulated* by ±100 Hz. At the receiver, the modem determines the frequency of the received signal and produces a high or low output accordingly. The receiving modem *demodulates* the signal, converting it to high and low voltages. The fact that the unit both *modulates* and *demodulates* gives it the long name *mo*dulator-*dem*odulator, which is contracted to *modem*. At the re eiving modem, the demodulated serial data is fed to a UART and converted to bytes of parallel data.

Communication between two computers alonga single channel may be *simplex* (one way), *half-duplex* (two-way, but only one station can transmit at any one time), or *full-duplex* (two-way, both stations able to transmit simultaneously). For full-duplex communication, it is necessary for the two modems to operate at different frequencies. Otherwise, modems would be unable to distinguish between each other's signals. By convention, the modem originating the data transfer transmits at 1170±100 Hz. The answering modem replies with signals at 2125±100 Hz.

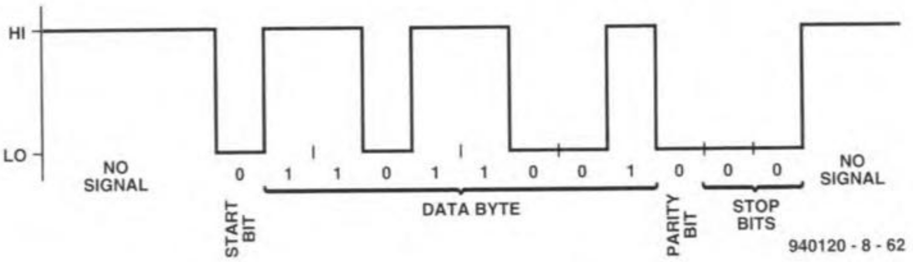


Fig. 63

Asynchronous transmission

The UART is described as 'asynchronous', which means that it may begin to receive a byte at any instant of time. In a synchronous system, the transmission of bytes is set on a regular time-scale under the control of a system clock. For this to be effective, the receiver must keep in step with the transmitter and needs access to the same clock. Otherwise, the receiver will not know when one byte ends and the next one begins. Sending clock pulses along with data pulses is feasible, or the clock pulses can be sent along a separate line, but the most practicable approach is to work asynchronously.

When the UART has loaded a data byte, it adds a few more bits to it before sending it out in serial form—**Fig. 63**. During the intervals between transmission of data, the line is held high or, in the case of transmission by way of a modem, the modem transmits its higher frequency. To begin its transmission, the UART sends a low *start bit*. The arrival of this low bit (or lower frequency) alerts the receiving UART that data is about to arrive. Next come the eight bits of the original data byte, followed by a *parity bit*. The purpose of this is the detection of errors in transmission. Before sending the parity bit, the UART counts the number of 1s in the data byte and adds either a 0 or a 1 so as to make the total number of 1-bits odd. Thus, if the data byte is 1101000, the number of 1s is already odd and the parity bit is 0. But if the data byte is 1101100, which has an even number of 1s, the parity bit is 1. It is also possible to the UART to work with even parity. Whether odd or even parity is in use is part of the specification of the system and, like the rate at which bits are transmitted, is decided before transmission begins. The transmission finishes with one or two *stop bits* to signal the end of the group, after which the signal returns to 'high'.

At the receiver, the UART strips off the start and stop bits. Then it counts the number of 1s and checks that the parity bit is correct. If it is correct, it strips it off and sends the remaining eight bits to the computer as a data byte. If the parity bit is incorrect, it sends a message to the transmitter to ask for the data to be retransmitted. The parity bit is an elementary though effective way of detecting an error in a single byte. There are numerous other ways of error detection that are employed in various circumstances to detect and possibly correct multiple-bit errors.

Optical fibre

Optical fibre transmits a modulated light-beam in the same way that copper wire transmitsa modulated voltage signal, but optical fibre has significant advantages. A fibre in a typical fibre-optic cable is about 0.1 mm in diameter. It is surrounded by a layer of glass or plastic, the *cladding*, which ensures that light striking the surface of the fibre is totally internally reflected. Several fibres are enclosed in a plastic sheath to make a cable only a few millimetres in diameter. This makes optical fibre lighter than copper-wire cable and in addition, it is cheaper to produce. These two features partly account for the popularity of optical fibre for long distances. Total internal reflection means that light passing along one fibre does not pass through to adjacent fibres. Each fibre is isolated from its neighbours and acts as separate channel without interference. A light-emitting diode, LED, or laser diode at the transmitter produces the modulated light-beam under computer control. The light passes along the fibre and is detcted by a photodiode at the receiving end. In a long link, there are repeaters to maintain signal strength. As is usual with digital signals, it is necessary only to discriminate between highs and low, so the recovered signal is identical with the original signal. There is no loss of data.

Each fibre carries several modulated signals simultaneously, each modulated with a different frequency. At the receiver, signals can be separated out and routed to different computers. Because light itself is of extremely high frequencies, the modulation frequencies can also be very high, allowing more of them to be 'stacked' in a given channel, so increasing the quantity of data that can be transmitted in a given time. This is another reason that optical fibre has supplanted wire cables as a communication medium. In addition, optical fibre is not subject to electromagnetic interference.

In certain circumstances, single waves in water travel long distance, regaining their

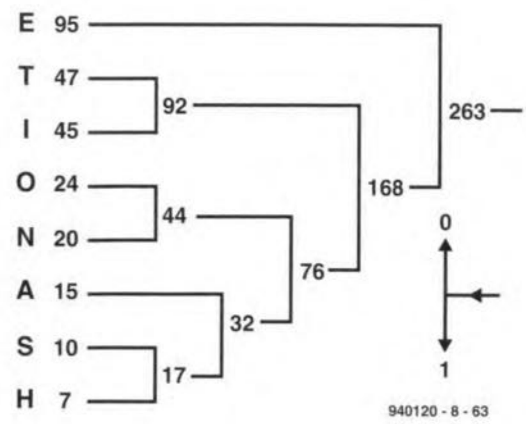


Fig. 64. Huffman tree.

shape after they have collided with other waves. This effect can sometimes be seen in the wake of a large boat. These waves are called *solitons*. A similar situation is found with light waves of a wavelength of 1.3 µm, traveling in optical fibre. A short (a few picoseconds) pulse forms a soliton which travels long distances in the fibre without significant reduction in amplitude, change in frequency or change in shape. On a long cable run (for example, a transatlantic cable), it is usual to employ repeater stations every 100 km to detect the optical signal, convert it to an electrical signal, and use a laser to retransmit the 'cleaned-up' light signal. The use of solitons makes another type of amplification possible. The signal is passed through a special length of optical fibre a few metres long, doped with erbium. This is 'pumped' by a low-power external semiconductor laser, supplying energy to the fibre and so regenerating the soliton. This is amuch simpler system to instal over long distances, particularly in trans-oceanic cables and allows signals of high frequency to be used. This technique is in the experimental stage at present, but initial results are very promising.

Radio

Radio transmission is the alternative to communication by wire or optical fibre. Digital signals may be modulated on to the carrier wave in one of three ways: amplitude shift keying, ASK, (high amplitude for 1, low or zero amplitude for 0); frequency shift keying, FSK (as explained earlier); and phase shift keying, PSK. In PSK, the phase of the carier is shifted by 180° at each transition between a 1 and a 0. There is a more effiient variation of PSK known as *quadriphase* PSK (QPSK), in which the bits are grouped inpairs and phase is shifted by 45°, 90°, 135°, or 180°, depending on whether the bits are 00, 01, 10, or 11. Combinations of methods such as QPSK plus ASK are also used to increase data transmission rates still further. The description above applies to a given channel but, as with optical fibre, several hundred channels at different carrier frequencies can be combined into one transmission. This is known as *frequency division multiplexing*. It requires a multiplexer at the transmitter to collect together signals from many sources, allocating each of them to a different channel. At the receiver, a demultiplexer separates the signals in each channel and relays them to their destinations. An alternative, known as *time division multiplexing*, is often used in satellite transmissions. Here, the user may occupy the whole of the bandwidth of the system, but the transmission time is divided into time slots. Only one user is allowed access at each time slot, transmitting frames of data each time their turn comes round. Data may be transmitted from several earth stations provided that each transmitter is synchronized to send its frame at the correct time. The satellite receives frames from each ground station in turn and relays them as a succession of frames to the receiving earth station where a time demultiplexer splits them up and routes them to their destinations.

Data compression

The need to compress data for storage has already been discussed, but the topic arises here in the context of data transmission. Even with the methods for rapid and multiple transition already outlined, the amount of data to be transmitted is increasing relentlessly. The more we can compress data before moving it, the more data can be sent in a given time, or the more data can be fed into a single channel. The first choice is between lossy and lossless data compression. Lossy compression deliberately discards some of the information (for example, by approximating analogue values with fewer than the ideal number of bits), so that it is not possible to reconstruct the original data entirely at the receiving end. For many classes of data, this is perfectly acceptable. For example, images that do not accurately reproduce the fine details and the exact colours of the original are good enough for video-phones. In the audio field, it is necessary to sample at 44.1 kHz and with 16-bit accuracy for CD recording, but a slower sampling rate and fewer bits are perfectly adequate for voice communication by telephone. In both examples, the human ability to recognize objects, faces, voices, and words, fills in the 'gaps' caused by loss of data. This is similar to the case of movie films and videos, in which a succession of still frames is interpreted in the human brain as a moving picture. For other types of data such as computer programs, data files, and control data, it is essential that data compression is lossless. Lossless compression generally relies on a system of coding at the transmitter, and reversing (decoding) and recovering the original data at the receiving end.

One of the simplest methods of encoding is *run-length coding*. In transmitting a picture, for example, in which most of the document consists of white paper, it is extremely wasteful to encode every pixel. If one particular value is repeated many times, it is sufficient to code it by quoting its value and the length of the run. For instance, a portion of a scan lasting 50 bytes (assuming 1 byte per character as in ASCII coding), and if B is black, G is grey and W is white, the full code might be:

BWGWBBBWWWWWWWWW-WWBGBGWGGGGGGGGGGGGG-BBBBBWWWWWBG

Run-length coding is preceded by a control character such as Escape (27 in ASCII code), which we represent by E in the coded version below. Each E is followed by two bytes coding the repeated character and the length of the run:

BGWEB3EW13BGBGWEG12EB6 EW5BG

This requires only 25 bytes, a reduction of 50%. At the receiver, this code allows the original string to be reconstructed without loss.

Run-length coding is not so useful with text because it is rarely that long runs occur (only one occurrence in the sentence before these brackets and only two within the brackets!). The *Huffman code* relies on the frequency with which tdifferent letters normally occur in text. The commoner the letter, the shorter its code. Coding depends on a Huffman Tree, see **Fig. 64**, the exact configuration of which depends on the relative frequency of letters. Here we use an order based on the letters ETIONASH, being the most frequent in English text, and assume that the relative frequencies are as shown in the figure. For text in other languages, or perhaps for particular disciplines in English, frequencies and the structure of the tree would be different. Figure 64 is only part of a tree; a full tree covers all letters of the alphabet.

To form the tree, the letters are listed in order of frequency, 95 for E down to 7 for H. Then, the two least common letters are grouped together: S+H has frequency 17.

Transmission lines

When a wire is of any length greater than a few metres, it begins to have an appreciable effect on a signal as this passes along it. This applies to both analogue and digital signals, but we shall discuss this with special reference to digital transmissions. A length of copper wire has resistance, R ohms per unit length. Another resistive property is the leakage between the two conductors, G ohms per unit length. The capacitance between the conductors, C farads per unit length, introduces a reactive component, dependent on frequency. Finally, there is another reactive component, the inductance of the wires, contributing L henries per unit length. R, G, C, and L are continuous along the line but, in calculations, a unit length of the cable can be considered as a number of separate (lumped) components as in **Fig. 61**. The total input impedance of the line to a signal frequency f is referred to as Z_0, the *characteristic impedance*, and is given by the equation:

$Z_0 = \sqrt{[(R + j\omega L)/(G + j\omega C)]}$, where $\omega = 2\pi f$. If f is large, as it usually is for the transmission of data, the equation simplifies to:

$$Z_0 = \sqrt{(L/C)}.$$

This approaches a fixed value at high frequencies, typically 50 Ω or 75 Ω. The line impedance brings about attenuation of the signal. For line transmissions over long distances, it is necessary to place repeater stations along the route. These receive the signal, remove noise, amplify it and retransmit it. Since Z_0 is dependent on frequency and since square waves of the type found in digital signals contain a rich spectrum of harmonics, passage along a copper line results in distortion of the original square wave. Further, the velocity of a wave is dependent on frequency, a phenomenon known as *group delay*. In a mixed-frequency signal such as a square wave, the various components of the square wave travel at different velocities; the lower the frequency, the faster they travel. Lower frequencies arrive first, so distorting the waveform. These are two of the disadvantages of using copper wire as a transmission channel. However, different types of transmission line vary in these respects. The local telephone lines have a constant group delay over a wide range of frequencies, which means that they can carry high-frequency signals without distortion. This has important implications for local transmission of data, including video signals, as in the case of the Video on Demand service mentioned earlier.

In order to maximize the power passed from the transmitter to the line. the output impedance of the transmitter should equal Z_0. The same consideration applies at the receiving end. If the input impedance of the receiving equipment is greater or less than Z_0, the energy transfer to the receiver is less than the maximum. Transferring the maximum power into the receiver is not the only reason for matching impedances. The non-transferred energy must go somewhere: when the wave energy reaches the end of the line, part of the wave energy is *reflected back*. As a train of pulses arrives at the receiver, a train of pulses of lower amplitude begins its journey back to the transmitter. If impedances are badly matched at the transmitter end, too, there is a second reflection back in the original direction. The multiple echoes create havoc in a digital signal.

The least frequenct group is formed by A and S+H. Group these together to make 32. This leaves the two least common letters as O and N, so group these to make 44. Group the two least common groups, O+N and A+S+H. Continue in this way to produce Fig. 64. A letter is coded by starting on the right and moving along the tree to the letter. An upward direction at each fork is 0, a downward drection is 1. Thus, the letter N is coded as 1101. The most common letter, E, requires only a single bit, 0. The least common letter, H, requires five bits. Note that no letters are coded by 1, 11, 111, or 1111, and no letter has more than five bits. These facts make it easier to distinguish the beginning of a new group when decompressing. For instance, the word ESTATE is compressed to:

01111010011101000

This is 17 *bits* long. In the ASCII code, in which each letters requires eight bits, the word would be 48 bits long.

The Huffman Tree is but one of several compression techniques. Like many other techniques, it is based on probability. In the Huffman Tree, the letters most likely to occur are coded with the smallest number of bits.

This idea can be extended to groups of letters and to words. Words such as THE, ARE, BIT, and groups forming parts of words, such as ING, MENT, ION, and QU, are more likely to occur than other combinations of letters. Some compression methods scan the text and match it to groups or words that have occurred previously. Instead of repeating the groups or words, the code refers back to the previous occurrence, indicating its length and how far back it occurs.

The Information Superhighway

We are living in an age in which more information is available than ever before. Not only is it available, but it is rapidly transferable from almost any part of the world to almost every other part. Costs are low and data transfer is virtually immediate. This is the beginning of what has come to be called the Information Superhighway. It is yet one more product of the digital solution.

Further reading

Daniel, Christine, and Owen Bishop: *An Introduction to Networks for PC and Mac Users*, Bernard Babani Press, 1995.

Test yourself

1. A modem operating on the standard frequencies is responding to a received message by sending a 0. What frequency is its tone?
2. The ASCII code for capital A is 65 in decimal. Convert this to binary, then add the bits to produce the signal that would be sent by a UART with odd parity.
3. Using the Huffman Tree of Fig. 64, decode this word:

   11101111010011001101101
   1111011111.

   By what percentage has it been compressed in comparison with an ASCII file?

Answer to Test yourself (7)

1. 2; 4.4; 9.68; 14.096; 7.5712; 9.0886; 10.3556; 7.2208; 6.5938; 6.0245; 5.2866; 4.7859. The output obviously dies away to zero but very slowly.

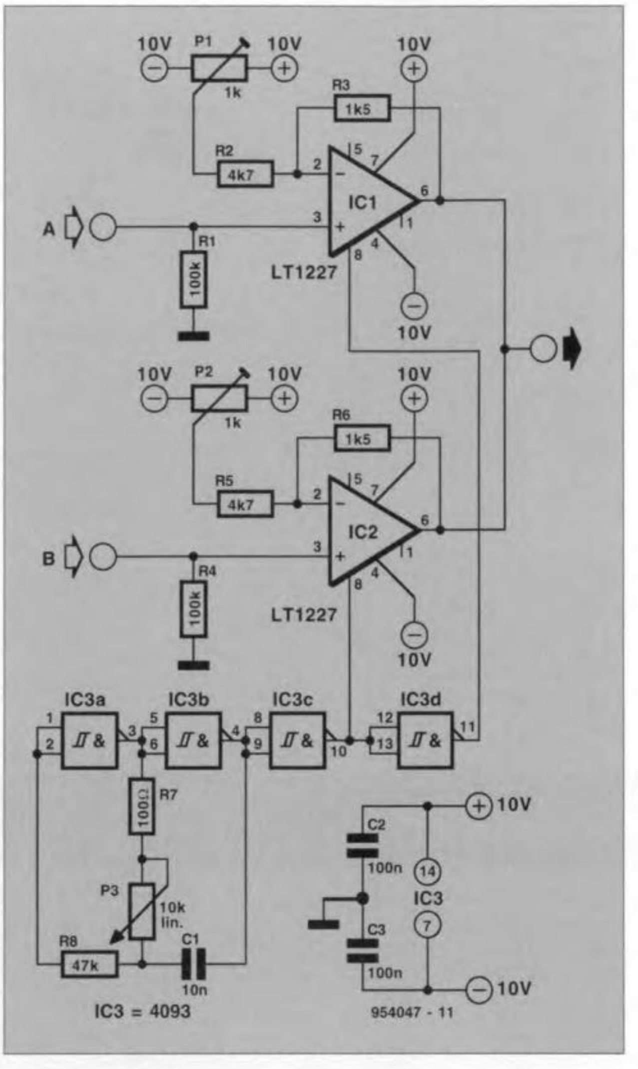
[940120-VIII]

VIDEO/SCOPE MULTIPLEXER

Most of the new, fast current-feedback op amps have an enable input via which it is possible to switch the device on and off. Because of this, the output 'floats', so that it is permissible to couple several outputs, provided that only one op amp is active at any one time. This is a property that makes it possible to construct a multiplexer from a number of such op amps which, because of the speed of the devices, can switch signals up to 20–30 MHz. This is ideal for video applications or, as in the present design, to multiply the number of oscilloscope inputs, for instance, to make a two-channel scope from a single-channel model.

Each of amplifiers IC_1 and IC_2 processes one of the two input signals and offset voltage from P_1 and P_2 respectively. Which of the two channels becomes available at the output is determined by the level provided by rectangular-wave generator IC_3. This level is constantly inverted by gate IC_{3d}, so that at any one time only one of the op amps is active.

The frequency can be set over a wide range with P_3. Experimenting with the value of C_1 is recommended.



Although the supply is specified as ±10 V, it may be any value between ±9 V and ±12 V. Make sure, however, that the 4093 is supplied from the positive line.

The rectangular-wave oscillator is not synchronized with the horizontal deflection of the oscilloscope. It is, therefore, best to use a high frequency with a long time base and a low one with a short time base. This obviates the stream of on/off switching voltages being displayed as a series of interruptions in the signals.

Mind the triggering of the scope, since it may happen that, owing to the offset, the time base is triggered by the oscillator signal instead of by one of the measurement signals. Because of this, the use of an external trigger input is recommended.

Bear in mind that the design is not an elaborate extension of the oscilloscope. For example, the input impedance is only 100 kΩ, whereas that of most scopes is 1 MΩ. However, for video applications, even 100 kΩ is already on the high side. Also, most op amps take about 1.5 µs to switch over. The annoying thing about this is that the switching causes a peak on the output voltage that is sometimes displayed vaguely, like a veil. It may also happen in scopes with a limited brightness control that the display is on the dark side.

Finally, it is advisable to limit the level of input signals to 1–2 V_{pp} to make sure that the amplifiers are not overdriven.

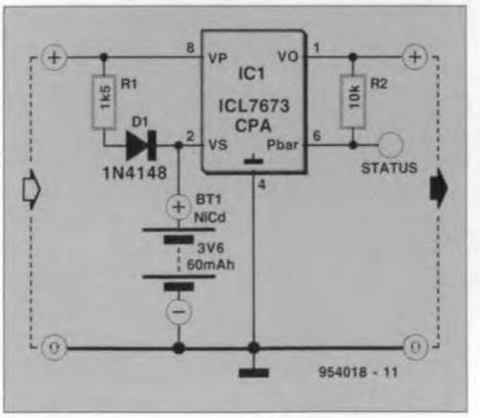
The circuit draws a current of about 15 mA.

Design by K. Walraven
[954047]

BATTERY BACKUP FOR RAM

Systems with RAM must always be connected to a power supply otherwise the data in the memory get lost. Any break in the supply can, therefore, be disastrous. A simple addition to the system, whereby a battery automatically takes over when the normal suply voltage is interrupted can prevent such a disaster from happening.

There are several ways of inserting a battery into the system, one of which is shown in the diagram. The IC arranges the switch-over in such a way that the mains-derived supply line remains unbroken. The circuit also provides a status indication which can be used for signalling or for switching off current-draining parts.



The IC has two inputs: one for the mains-derived power supply and one for the battery. The status output, pin 6 goes high the moment there is switch-over to the battery. The unbroken mains-derived power supply is output at pin 1.

Resistor R_1 and diode D_1 are optional. They ensure that the battery remains charged. The resistor has a value that gives a current equal to the trickle-charging current of the battery. These components must, of course, NOT be fitted when a primary (non-rechargeable) battery is used.

The IC can handle a current of up to 38 mA at pin 8 and up to 30 mA at pin 2.

Design by H. Bonekamp
[954018]

FSK GENERATOR

The generator may be of interest to those who wish to record FSK (frequency-shift keying) signals on their tape recorder. Such signals consist of pulses at two frequencies, which either represent a 0 (low) or a 1 (high). In the generation of these pulses, care must be taken to ensure that no phase errors occur.

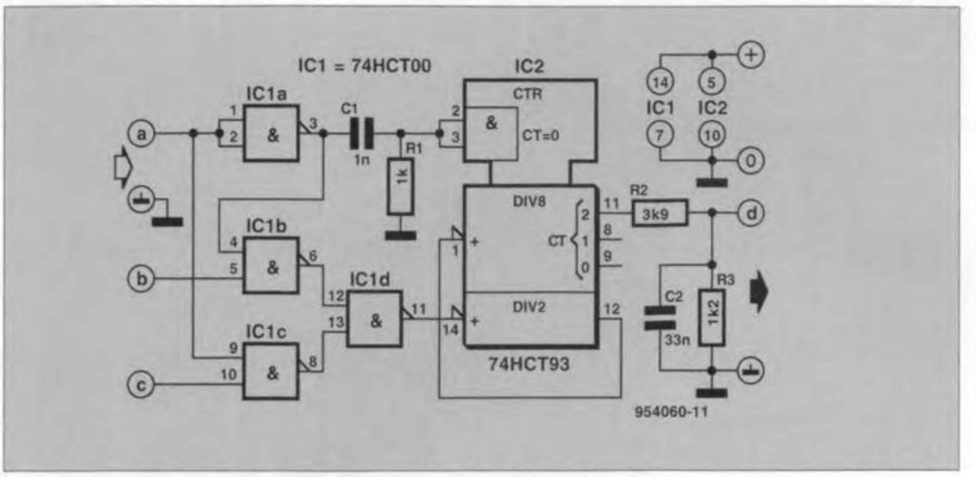
In the circuit, IC_1 chooses between the high and low frequency, represented at the input by a 1 or 0 respectively. A 1 causes f_2*16, and a 0, f_1*16, to be connected to the input of IC_2. This makes it clear why these inputs are designated *16: the :16 scaler divides the input frequency by 16, so that the wanted FSK frequencies are made f_1 and f_2.

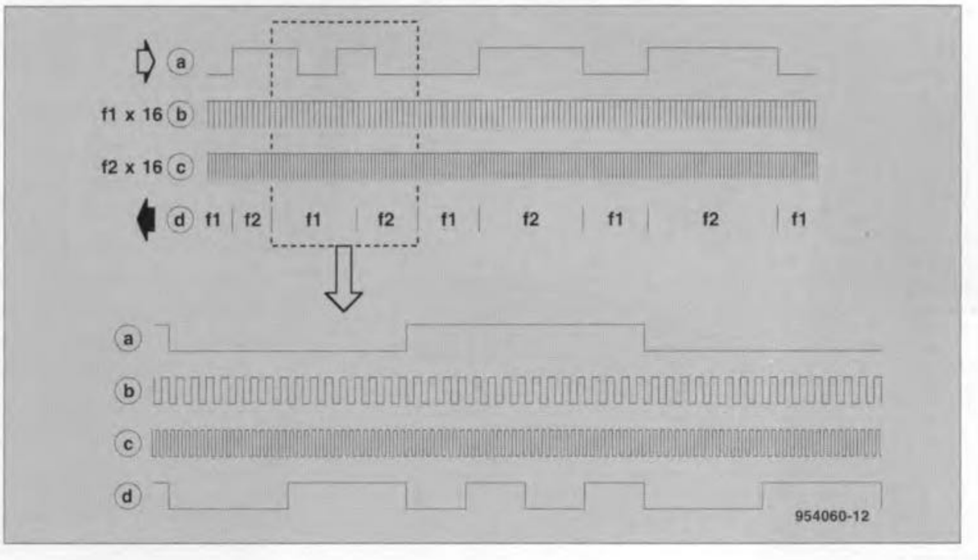
Network R_2-R_3-C_2 is a low-pass filter with upper cut-off frequency f_2, and also provides level matching to 1 V_{pp}, which corresponds to 0 dB in a tape recorder.

Raising the frequency by a factor 16 ensures that no significant phase errors can arise. At the same time, the divider is reset at each trailing edge, which prevents half or other submultiple pulses being generated.

The circuit draws an operating current of about 10 mA.

Design by A. Rietjens
[954060]



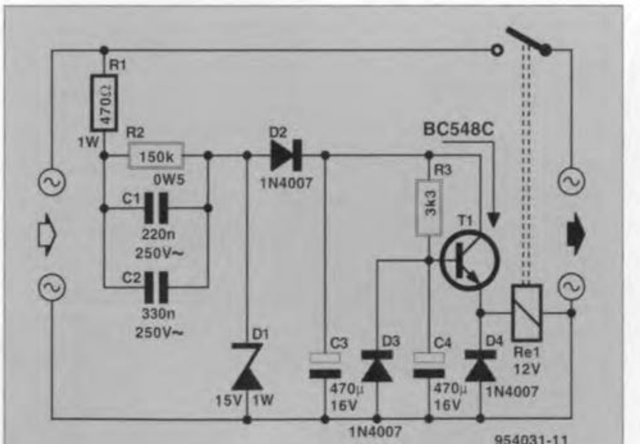


DELAYED MONITOR POWER-ON

In many computer systems, the mains for the monitor is derived from the computer, so that when the computer is switched on, the monitor is switched on also. This can give rise to such a high current, that the domestic mains fuse may blow. This is not just annoying, but it can give rise to data loss and even damage to the hard disk. This difficulty may be remedied by a slow-blow fuse, a separate mains switch for the monitor or the power-on delay in the circuit diagram shown.

The power-on delay unit is connected in the mains lead of the monitor. It provides a time delay in the switching of power to the monitor and thus reduces the initial peak current and ensures that the mains fuse is not overloaded.

Resistor R_1 and capacitors C_1, C_2 draw a current of about 40 mA from the mains. The positive half wave of the mains voltage is limited to 15 V by zener diode D_1. These clipped half waves are rectified by D_2 and smoothed by C_3. From this, an average current of 20 mA is derived which charges C_4 via R_3. When the potential across C_4 has reached a certain level, transistor T_1 is switched on, whereupon relay Re_1 is energized. The relay contact then connects the mains to the monitor.



Diode D_3 and resistor R_2 ensure that C_3 and C_4 are discharged as soon as the mains is switched off.

Diode D_4 is a freewheeling element for the protection of T_1.

The circuit is best built on a small piece of prototyping board and fitted inside the monitor enclosure. If that is not possible owing to lack of space, it must be fitted in a small plastic box of its own.

Design by J. Kirchner
[954031]

FSK DECODER

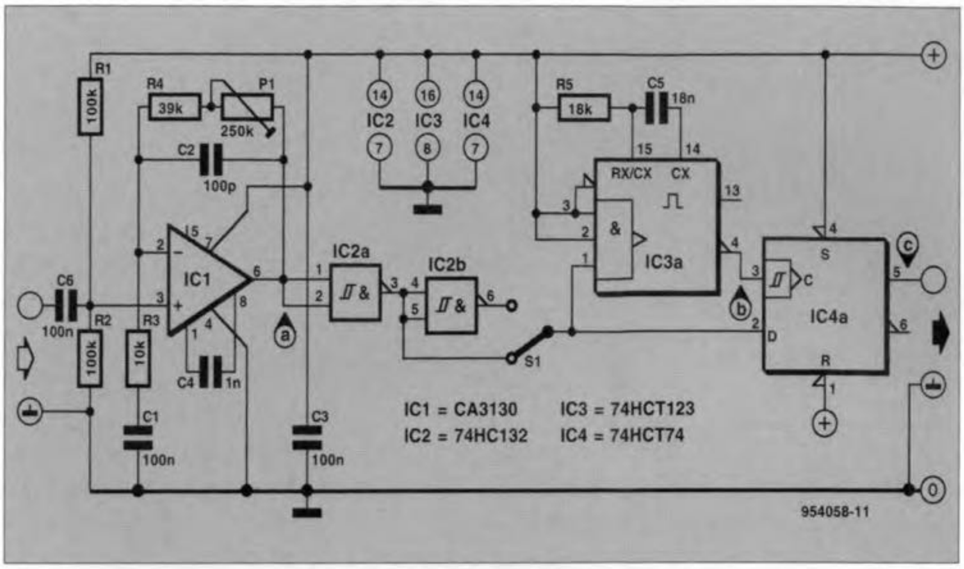
The decoder enables FSK signals, for example, those from a tape recorder, to be converted into digital signals.

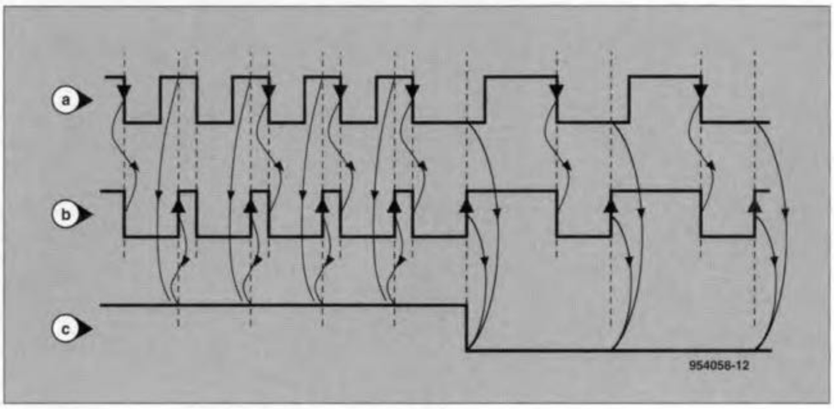
Input amplifier IC_1 raises the analogue signal to a level that is symmetrical around half the supply voltage. This is converted into a digital signal by an HCMOS gate. The switching levels of this gate are at exactly half the supply voltage (wherefore it must be a Type 74HC132).

Since in some tape recorders the input and output signals are in antiphase, S_1 is provided to change the phase of the detected signal. Each trailing edge of the resulting signal triggers monostable IC_{3a}. The mono time is about $^3/_4$ of the period of the highest detected frequency. This means that the clock applied to IC_{4a} generates a clock at the low frequency when this is still in the '0' state, while the high frequency is in the '1' state. Since the digital signal is applied to the D-input, a 0 will be clocked in for the low frequency and a 1 for the high frequency.

The detected 0s and 1s are available at the output, from where they can be processed further. The present circuit is designed for input frequencies of 2 kHz and 4 kHz.

Design by A. Rietjens [954058]



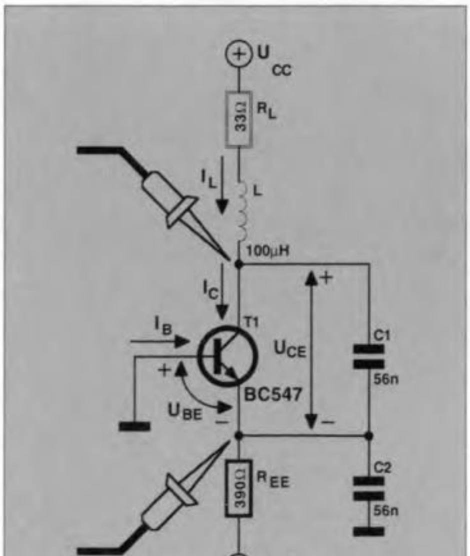


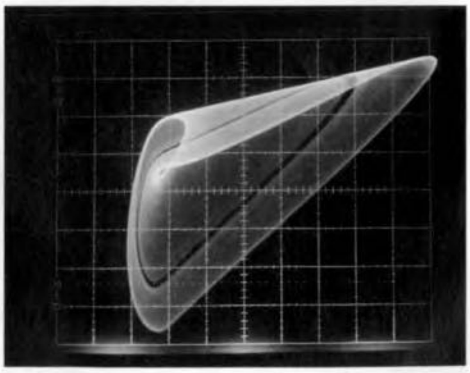
CHAOS

Chaos is all around us: on the roads, in nature, in the weather, and in everyday life. Electronic systems may also show chaotic behaviour, such as the Colpitts oscillator in the diagram. This suffers from what has been called the *multi-oscillation phenomenon*. In this, two or more oscillations exist simultaneously in the steady state. In other words, there are parasitic or unwanted oscillations existing together with the main oscillation. The resultant steady-state signal is severely distorted, which limits its application in communication systems.

The circuit in the diagram is easily built on a piece of prototyping or similar board. It needs a supply of ±5 V. When the probes of an oscilloscope in the *X-Y* mode are connected as shown, its display will show something like the photograph.

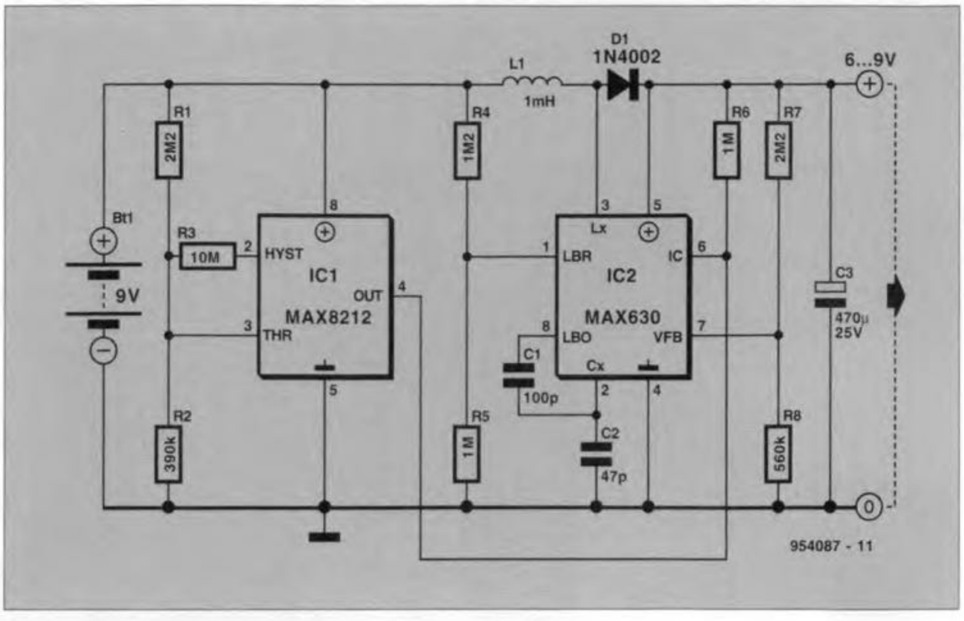
Design by L. Lemmens [954021]





9 V BATTERY LIFE EXTENDER

Often an equipment is supplied by a battery whose nominal voltage is only slightly higher than the required supply voltage. The result is that when the battery is only partly discharged, the equipment already gives a 'low battery' indication and switches itself off. Only half of the battery capacity, if that, has then been used, which is not only wasteful, but also costly.



The circuit in the diagram provides a minimum of 7 V until the 9 V battery voltage drops to less than 2 V. It uses a MAX8212 programmable voltage detector and a MAX630 micropower step-up switching regulator.

The MAX8212 contains a comparator, a 1.15 V band-gap reference, and an open-drain n-channel output driver. Two external resistors are used in conjunction with the internal reference to set the trip point voltage to the desired level. A hysteresis output is also included to allow positive feedback to be applied for noise-free output switching.

The MAX630 is a low-power step-up switching regulator that can handle powers from 5 mW to 5 W. All necessary functions are contained in a compact 8-pin case: a 1.31 V band-gap reference, an oscillator, a comparator and a MOSFET output stage that can provide a current of up to 375 mA. The chip draws a current of only 70 µA, which is virtually independent of the output current and of the duty factor. Moreover, a special circuit ensures that the quiescent-current drain is limited to 1 µA. Consequently, a high efficiency is guaranteed even in low-power applications. The chip operates satisfactorily with (battery) supply voltages from 2 V to 16.5 V.

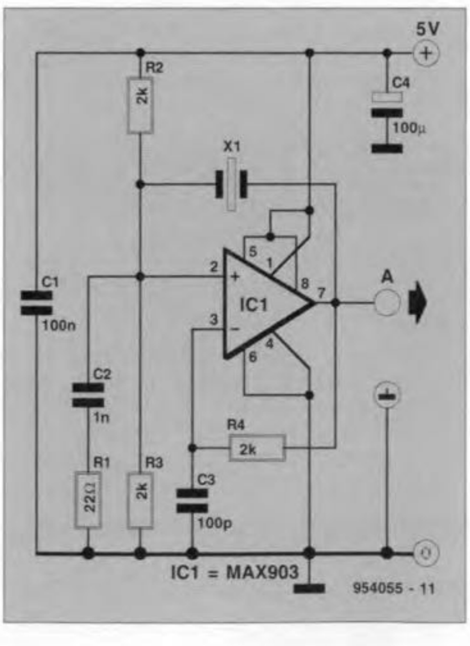
Input I_C, pin 6, of IC_2 is low as long as the battery voltage is higher than 7 V. The chip is then held in the shutdown mode and draws a current of only 10 nA. When the battery voltage falls below 7 V, the output of voltage detector IC_1 goes high, whereupon IC_2 is enabled. This circuit holds the output voltage at 7 V, even when the battery voltage falls further. The 'low battery' detector, LBD (pin 8), is used to lower the oscillator frequency when the battery voltage falls to 3 V in order to increase the permissible output current at this low battery voltage.

The circuit may also be used, with or without the MAX8212, to obtain a guaranteed voltage of 5 V from four 1.5 V batteries even when the total voltage drops below 2 V.

A Maxim application

LOW-POWER QUARTZ OSCILLATOR

Ready-made oscillators are available with standard frequencies only and usually require a buffer when they are to provide a large circuit with a clock. An interesting alternative is shown in the diagram. It is based on Maxim's analogue comparator Type MAX903. This IC draws a current of only 1.3 mA and the entire circuit only 13 mA. The IC has a typical transfer delay of 8 ns and provides a high open-loop voltage gain.



The oscillator uses a quartz crystal in the range 10–20 MHz and requires a load of about 500 Ω. Network R_1-C_1 lowers the open-loop gain at high frequencies to reduce the number of spurious harmonics: this enhances the symmetry of the oscillator signal. If a crystal in the range 2.5–10 MHz is used, the *RC* network may be omitted, but the value of C_3 should then be raised slightly.

Network R_4-C_3 lowers the gain at low frequencies to prevent overtone crystals oscillating at their fundamental frequency.

Capacitor C_4 decouples the supply voltage line.

A Maxim application
[954055]

PROGRAMMABLE FREQUENCY DIVIDER

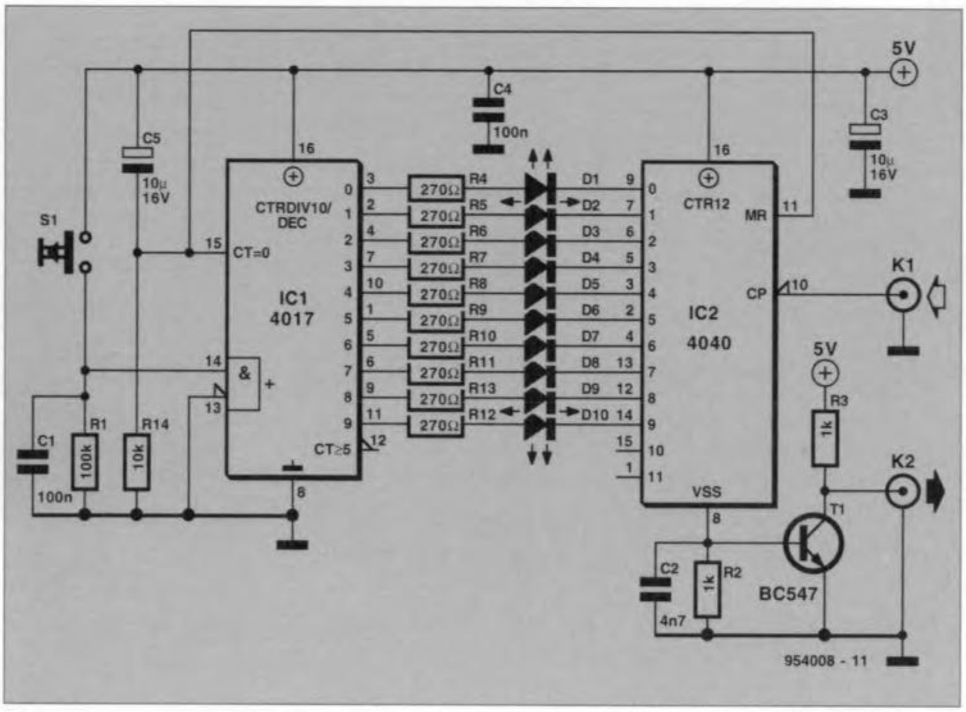
The circuit in the diagram is based on IC_2, which contains a row of series-connected binary scalers. The clock input to the IC via K_1 will be available divided by 2 at output Q_0, by 4 at Q_1, by 8 at Q_2, and so on. However, a closer look shows that Q_0–Q_9 are not used as outputs but as inputs for the output from IC_1.

Circuit IC_1 is a decade scaler with the special property that only one output can be high at any one time, depending on the state of the counter.

Because of the coupling of the two ICs, a current will flow from IC_1 via one of the outputs of IC_2 (for as long as this is low), and the associated resistor and LED to ground. (The LEDs thus flash, visibly or otherwise, at a frequency associated with the selected output of IC_2).

The current flowing to ground will cause a potential drop across the resistor in series with the supply terminal, which is converted by T_1 into a standard logic level. Depending on the state of counter IC_1, the current through the supply connection will have a certain pulse rate, which is determined partly by the signal at K_1.

Depending on the supply voltage and the type of IC_1 used, the maximum clock for IC_2 may be 7 MHz at V_{dd} = 5 V, or 24 MHz at V_{dd} = 15 V.



The state of IC_1 is set with S_1. Network R_{14}-C_5 ensures a defined starting state (power-up reset), while combination R_1-C_1 debounces the switch.

Design by M.S. Nagaraj
[954008]

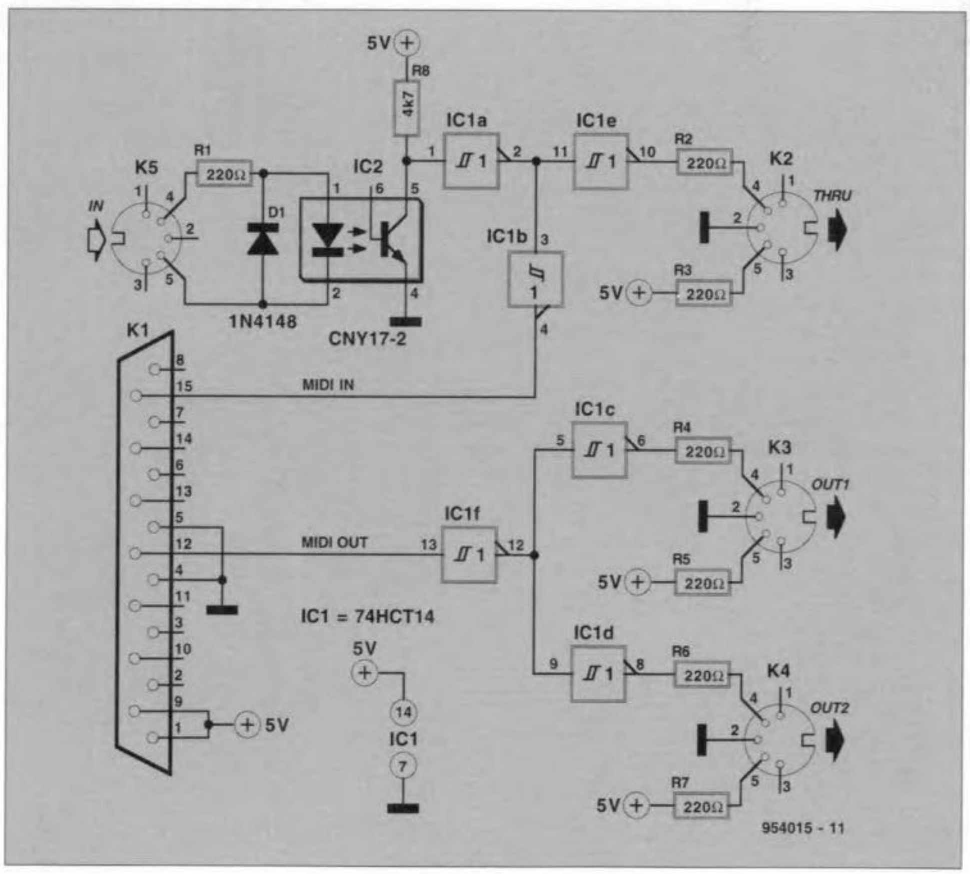
MIDI FOR PC SOUNDCARD

On many PC soundcards, the MIDI (Musical Instrument Digital Interface) is not fully implemented. This is because MIDI signals are not really TTL signals. The MIDI is strictly a current loop which in the receiver is electrically separated from the remainder of the circuitry by an optoisolator.

Fortunately it is not difficult or costly to provide a soundcard, such as the Soundblaster Pro 2 used for this design, with a complete MIDI as shown in the diagram.

The circuit is based on a Type CNY17/II optoisolator and a 74LS14 TTL driver, and may easily be built on a small piece of prototyping board. It is connected to the joystick/MIDI port of the card via a 15-way plug. It is important to note that in the MIDI IN connector ground pin 2 (central between pins 4 and 5) must NOT be connected.

Design by C. Weterings
[954015]



MICROCONTROLLER DEVELOPMENT/TEST TOOL

Anyone involved in the development and application of microcontroller systems is aware of the complexities that can arise during test and faultfinding. Of course, these do not bother those who have the use of a 32-channel logic analyser, but many readers have no such equipment. For them, the present tool may provide a very useful alternative.

It is assumed that the ICs in the microcontroller system are fitted in suitable sockets. These sockets are connected to single LEDs via an appropriate plug-in connector, a length of flatcable and a double-row pin header. The LEDs are low-current types that light at full brightness with currents as low as 2 mA.

The pin header, in combination with individual plug connectors, makes it possible to cater for a variety of IC pinouts and pin numbers.

During test or faultfinding, a separate ground lead is required between the circuit on test and the tool. All LEDs that are linked to an IC pin that is logic high will light. The current of 2 mA per LED can normally be drawn from the microcontroller equipment on test, provided its power supply has the necessary reserve.

Since our eye can not follow fast variations in brightness, there is an adaptation of the software to be tested in the form of relevant time delays for each change in signal. These are, for instance, functions STIME or LTIME on diskette Type 1811 (Assembler Course EASM52 – see *Elektor Electronics* March–June 1994). If one-second stops are inserted into the software, signal variations can be followed well. Even better, but more complex to program, are stops that can only be disabled by the pressing of a switch.

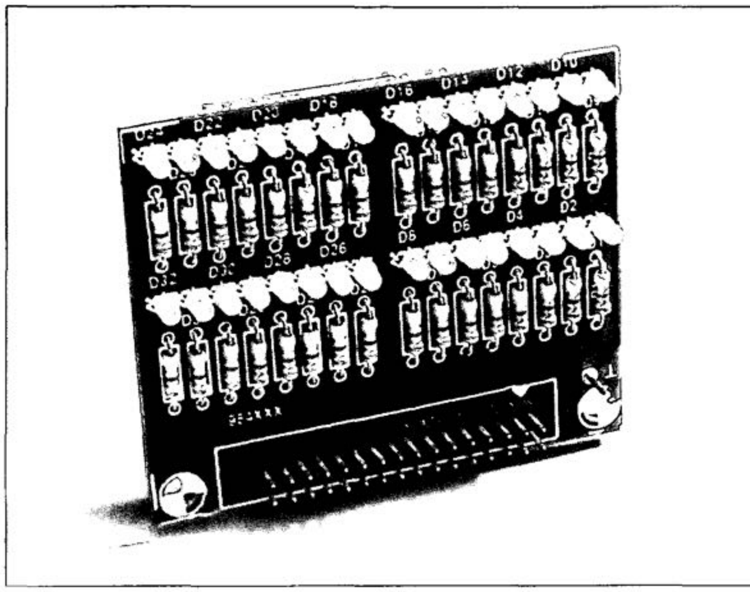
The tool has an undoubted advantage over a logic analyser in that no presettings are required that increase the likelihood of error. Variations in the software can be tested immediately.

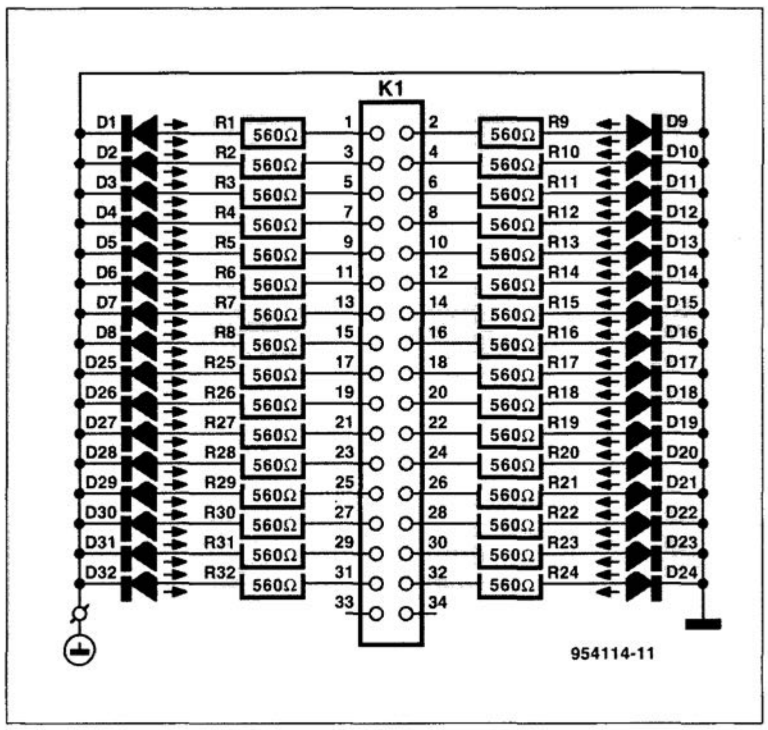
The designer has used the prototype tool for more than a year with his 80C535 development system without any problem whatsoever.

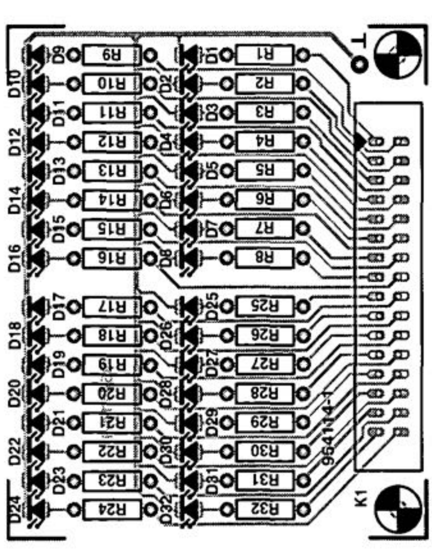
Parts list

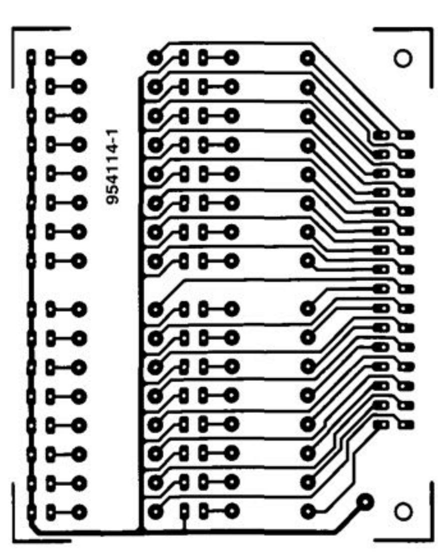
R_1–R_{32} = 560 Ω
D_1–D_{32} = LED, low-current
Double-row pin header, 2×16 pins
PCB 954114 – not available ready made

Design by U. Hinke
[954114]

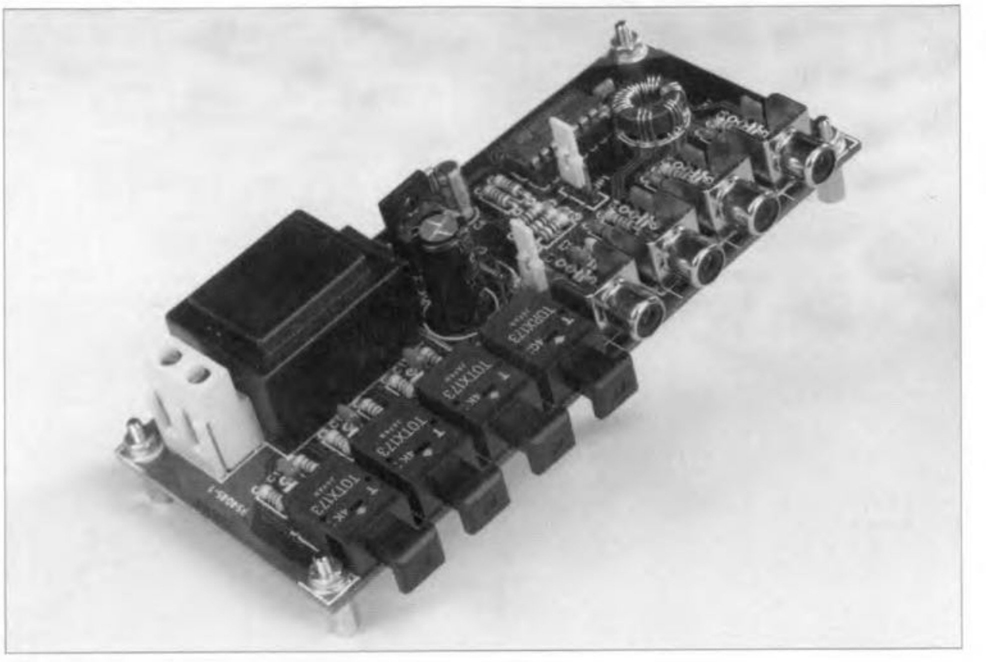








SPLITTER FOR S/PDIF COAX/OPTICAL OUTPUT



THIS circuit was originally designed to enable one S/PDIF output on a CD player to drive several inputs. The circuit acts as a 3-way splitter or format converter for S/PDIF ('digital') or optical signals. The function of the circuit is determined with the aid of two jumpers, as follows:

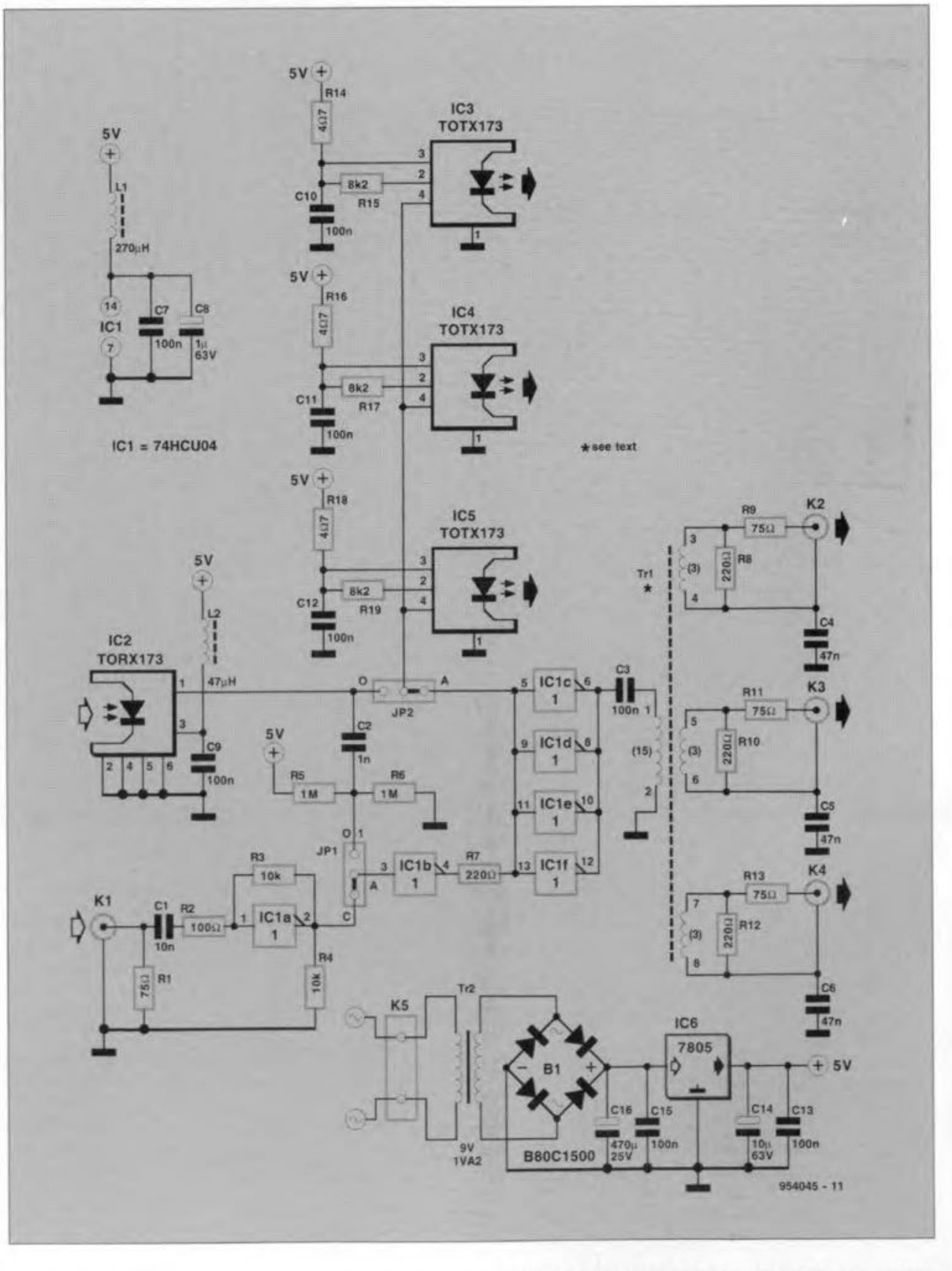
JP_1	JP_2	Function
O	O	optical input to optical and coax output
O	A	optical input to optical (buffered) and coax output
C	O	optical input to optical output, coax input to coax output
C	A	coax input to optical and coax output

In the selections available, JP_1 has priority. The third option indicates that the circuit is capable of splitting two sources separately, while retaining their formats. It is not possible to make cross connections because it is not usually required to convert from coax to optical. The other way around, though, is far more usual, and available with the other three options.

The optical inputs and outputs consist of Toshiba 'Toslink' modules, which are TTL compatible. The coax inputs and outputs are formed by RCA-style 'phono' sockets. Both the Toslink modules and the phono sockets are accommodated on the printed circuit board.

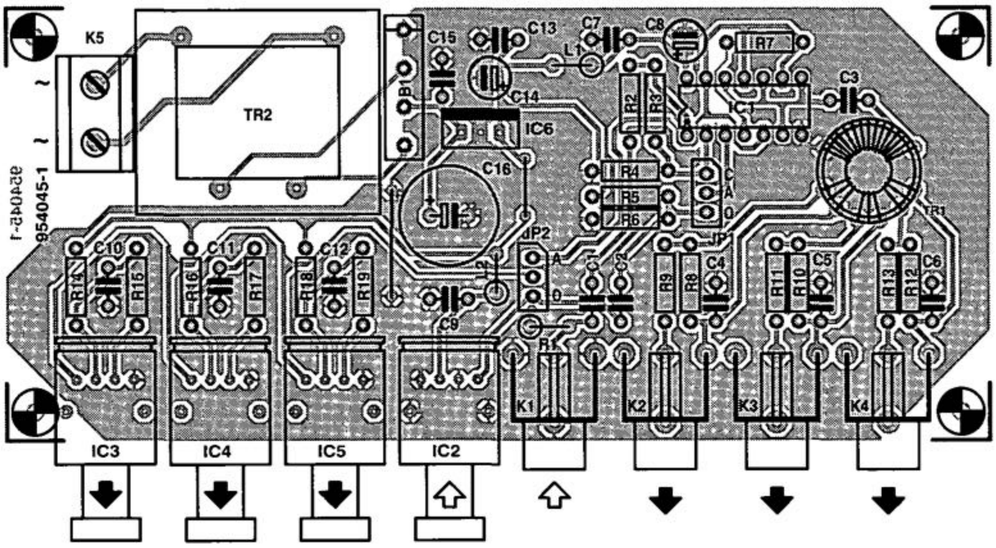
A digital (coax) input signal is first amplified by IC_{1a}. Resistor R_4 lightly loads the gate to prevent it from oscillating if no input signal is connected. In some cases, a resistor with a slightly lower value may be required in this position. Next, jumper JP_1 allows the output of the amplifier (position 'C'), or that of Toslink receiver IC_2 (position 'O'), to be selected. The Toslink signal is superimposed on a half-supply bias created by R_5 and R_6, and then applied to IC_{1b}. The half-supply bias corresponds to the switching threshold of IC_{1b}. This allows Toslink transmitters IC_3, IC_4 and IC_5, to be driven from the output of receiver IC_2, or from the output of IC_{1b}. Remember, the signal supplied by the latter is cleaned and amplified. Simply select the option which gives the best results.

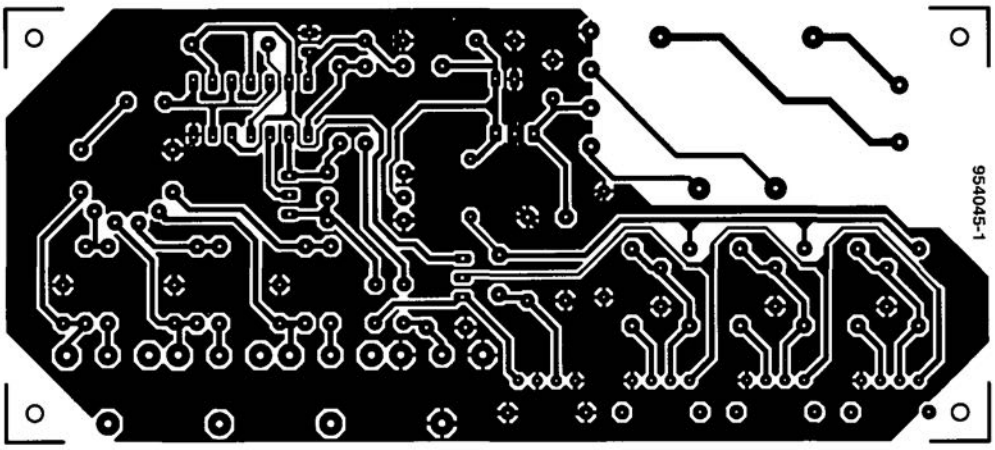
The buffer formed by the four parallel-connected inverters has sufficient drive capacity to match the low impedance of the primary of transformer Tr_1. The transformer actually forms the heart of the circuit. The three digital outputs are electrically isolated from the input, although the r.f. signals 'see' the ground level through capacitors C_4, C_5 and C_6. Matching resistors R_9, R_{11} and R_{13} prevent reflections at the relevant output from exceeding the ×1 level. Resistors R_8, R_{10} and R_{12} similarly damp the ringing effects



which occur on the non-used secondary windings. The drawing shows the practical construction of the transformer, which is based on a type G2.3-FT12 ferrite ring core. This type was chosen because of its high bandwidth and coupling factor, factors which allow the primary winding of 15 turns to occupy only about half the core, while the secondary windings of three turns each are distributed over the remainder. All windings are made from 0.5-mm dia (24 SWG) enamelled copper wire.

The power supply is conventional and 'on board', consisting mainly of a 1.2-VA mains transformer, a bridge rectifier and a 7805 three-pin voltage regulator. All ICs on the board have individual supply decoupling parts. Current consumption of the circuit is of the order of 70 mA. We regret that the printed circuit board shown is not available ready-made.





Parts list

Resistors:

R_1;R_9;R_{11};R_{13} = 75Ω
R_2 = 100Ω
R_3;R_4 = 10kΩ
R_5;R_6 = 1MΩ
R_7;R_8;R_{10};R_{12} = 220Ω
R_{14};R_{16};R_{18} = 4Ω7
R_{15};R_{17};R_{19} = 8kΩ2

Capacitors:

C_1 = 10nF ceramic
C_2 = 1nF ceramic
C_3;C_7;C_9-C_{13};C_{15} = 100nF ceramic
C_4;C_5;C_6 = 47nF ceramic
C_8 = 1µF 63V radial
C_{14} = 10µF 63V radial
C_{16} = 470µF 25V radial

Inductors:

L_1 = 270µH choke
L_2 = 47µH choke

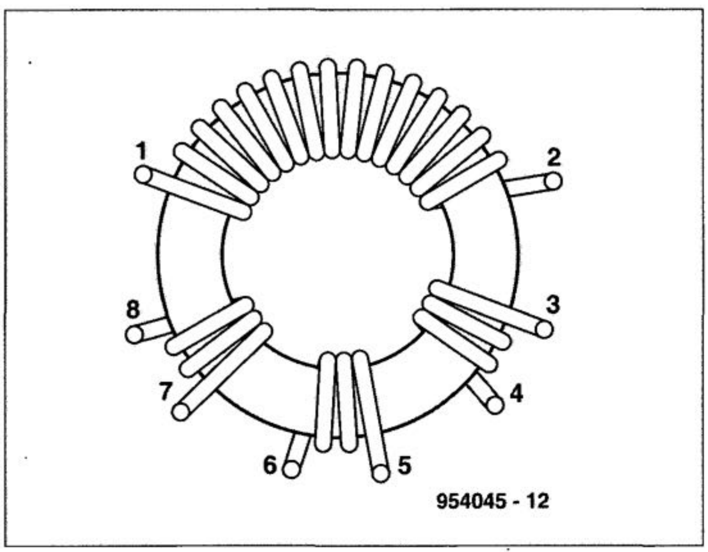
Semiconductors:

IC_1 = 74HCU04
IC_2 = TORX173 (Toshiba)
IC_3;IC_4;IC_5 = TOTX173 (Toshiba)
IC_6 = 7805

Miscellaneous:

JP_1;JP_2 = 3-way pin header, w. jumper.
K_1-K_4 = RCA style PCB mount socket, Monacor T709G.
K_5 = 2-way PCB terminal block, pitch 7.5mm.
B_1 = B80C1500
Tr_1 = G2.3-FT12 ferrite ring core; primary 15 turns 0.5mm dia. ecw, secondary 3×3 turns 0.5mm dia. ecw.
Tr_2 = 9V/1.2VA transformer, e.g, Hahn BV EI 302 0376; Velleman 1090012M; Monacor VTR1109 (1.5VA); Block VR1109 (1.5VA).

Design by T. Giesberts
[954045]



CPU VENTILATOR CHECK

Some of he fastest PCs around these days, particularly 60-MHz+ Pentium and 486-based machines, have a small ventilator fitted on the CPU to help it stay cool. These ventilators are also available as add-on units for PCs which suffer from CPU overheating problems. The circuit shown indicates whether the CPU ventilator is actually running. In fact, the circuit is capable of detecting two fault conditions: (1) ventilator not running because of a wiring fault or a burnt motor winding (i.e., current interruption); or (2) ventilator not running because it is jammed.

Both conditions can be detected by measuring the current drawn by the ventilator motor. No current means condition (1), and too much current means condition (2). These faults are signalled by two red LEDs and a buzzer to alert the user that there is an acute overheating risk for the expensive CPU. For peace of mind there are also two green 'OK' LEDs to indicate that all is well.

The motor current flows through R_7; the resulting voltage drop across the resistor is smoothed by R_{11}-C_5. Resistor R_7 should have a value which causes a voltage of about 0.5 V across C_5 when the ventilator is running normally.

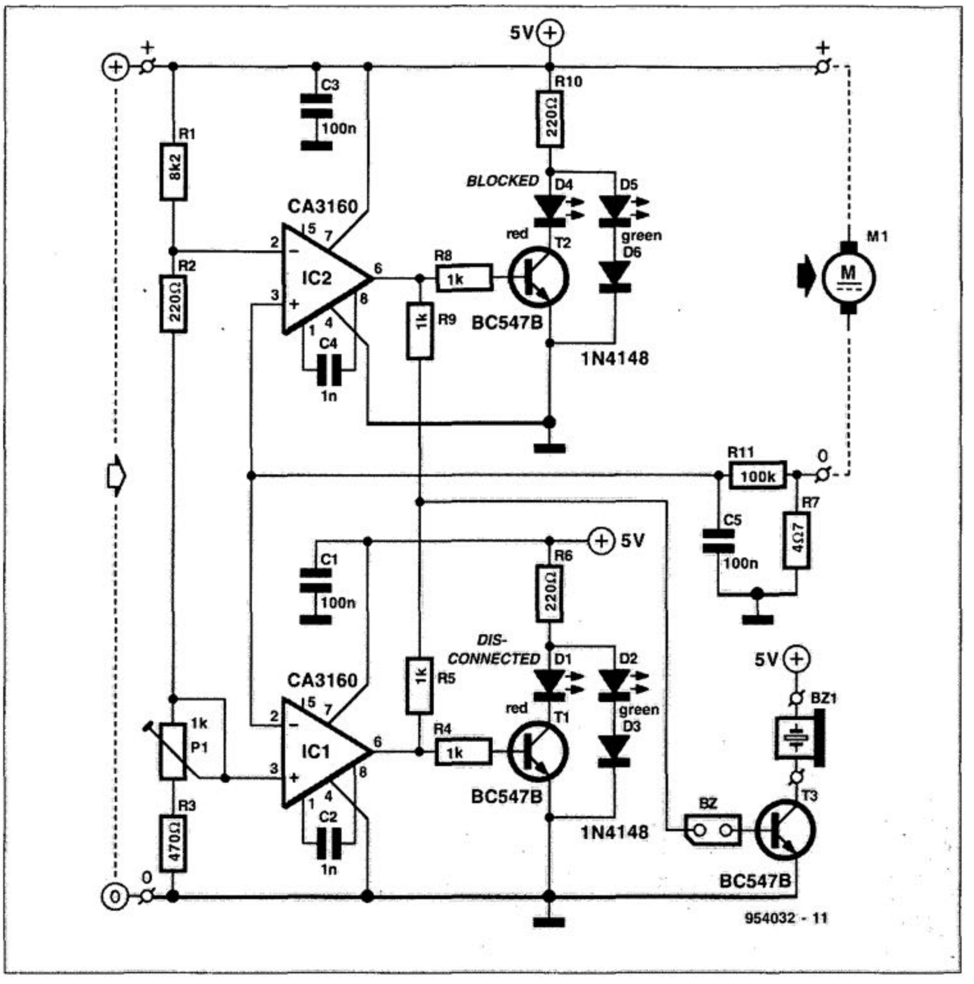
When the ventilator current is interrupted, the voltage at the +input of IC_1 rises above that at the –input. Consequently, the comparator output swings high and switches on transistor T_1. Red LED D_1 then lights and the green one, D_2, goes out.

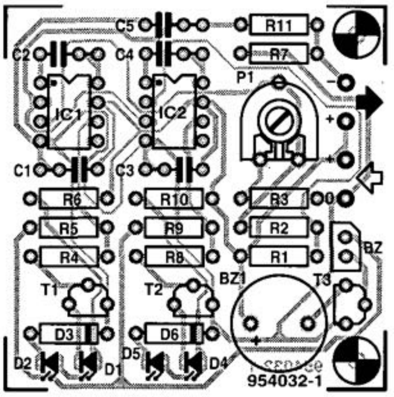
When the ventilator is jammed, the output of IC_2 goes high, and diode D_4 lights, while D_5 goes out. Provided the jumper marked 'Bz' is fitted, the buzzer will sound to signal either of the two fault conditions.

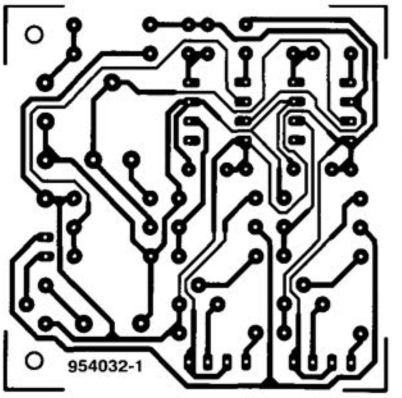
Preset P_1 should be adjusted such that both red LEDs stay out when the motor is running normally.

The circuit is designed for 5-V ventilator units. When a 12-V ventilator is used, all that is necessary is for R_1 to be increased to 15 kΩ, and R_6 and R_{10}, to 1kΩ. The buzzer should then obviously be a 12-V type.

The printed circuit board shown is not available ready-made.







Parts list

Resistors:
R_1 = 8kΩ2
R_2;R_6;R_{10} = 220Ω
R_3 = 470Ω
R_4;R_5;R_8;R_9 = 1kΩ
R_7 = 4.7Ω (see text)
R_{11} = 100kΩ
P_1 = 1kΩ preset H

Capacitors:
C_1;C_3;C_5 = 100nF
C_2;C_4 = 1nF

Semiconductors:
D_1;D_4 = LED, red, 3 mm
D_2;D_5 = LED, green, 3 mm
D_3;D_6 = 1N4148
T_1;T_2;T_3 = BC547B

Integrated circuits:
IC_1;IC_2 = CA3160

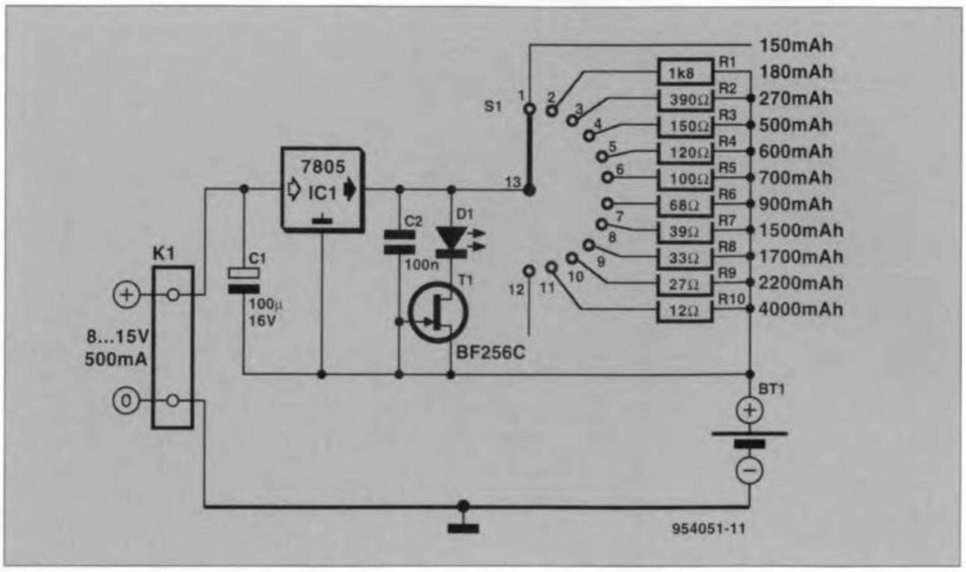
Miscellaneous:
Bz_1 = 5V active buzzer
PCB 954032 – not available ready made

Design by A. Rietjens
[954032]

SWITCHED NiCd BATTERY CHARGER

The charger is designed to charge NiCd batteries with capacities of 150–4000 mAh. A potential 5 V higher than the nominal voltage of the battery to be charged is derived from an 8–15 V mains adaptor by IC_1. The supply voltage is decoupled by C_1 and C_2. The nominal voltage of the battery to be charged, whether this is 1.5 V or 9 V, has no effect on the circuit. It is therefore possible to charge *n* series-connected batteries, provided that the supply voltage is $(n{\cdot}1.5 + 5)$ V.

Charging current can reach the battery in three different ways. Firstly, a current of 5 mA flows from the reference terminal of IC_1 to the + terminal of the battery. Secondly, since the gate-source voltage of T_1 is 0 V, the LED current of 10 mA flows into the battery. Thirdly, the charging current proper, which is provided by the current source formed by the regulator and the resistor specific to each different type of battery. The total charging current is $^1/_{10}$C, so that a flat battery is full charged again within 14–16 hours.



Note that in position 1 of the selector switch no 'proper' charging current flows, because the LED current and the current from the REF terminal of the regulator already make up $^1/_{10}$C for 150 mAh batteries.

The mains adaptor must be able to provide a maximum operating current of not less than 400 mA.

Design by A. Rietjens
[954051]

FLASH KEY ADD-ON FOR TELEPHONE SETS

Although most modern telephones, in particular, those with a number memory, already have a flash key, there are many excellent sets around which have not. This circuit makes these older sets compatible with a domestic telephone exchange which requires a flash key on each extension.

The function of a flash key is simply to interrupt the line current for about ⅒th of a second. In principle, you could mimic this by pressing the receiver switch a few times, but the chances that you achieve the right pulse length are pretty small. This circuit does a much better job.

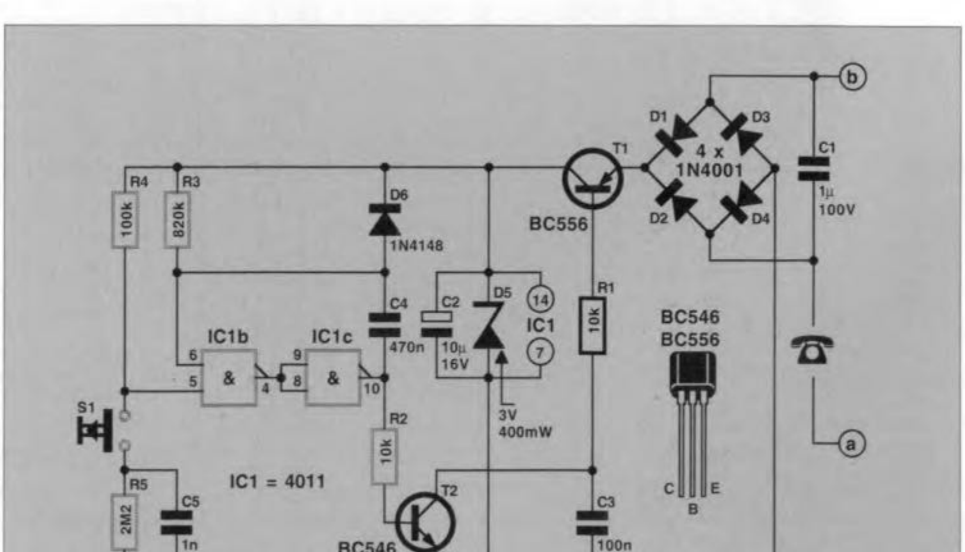
The circuit is connected in series with the telephone line. When the receiver is lifted, or when the extension is called, the voltage across the circuit is the line voltage or the bell voltage respectively. Transistors T_1 and T_2 start to conduct via C_3. This creates a minimum supply voltage of about 3 V for the 4011 CMOS IC which is wired as a monostable multivibrator (MMV). The output of the MMV is normally high, keeping the transistors switched on. By pressing the 'flash' key, the MMV is started via C_5. During the monotime, T_1 and T_2 are pinched off for about 100 ms, the exact time being determined by C_4 and R_3. If the pulse is too long, the circuit may occasionally be disconnected by the exchange. If that happens, simply make R_3 a little smaller to reduce the monotime. If the pulse is too short, you will hear a click, but the extension will occasionally not switch properly. The remedy is then to increase the value of R_3.

The circuit reduces the line voltage by 4–4.5 V, which is equivalent to the effect of a really long line. Fortunately, the reactance of C_1 is almost nought for the speech signal, which is thus not attenuated.

It makes no sense to press the flash key while the set is ringing. Do not do it because the high ring voltage (65 V and more) will arrive at T_1 and T_2 via C_1 and C_3, putting the the transistors at risk.

The circuit is **not** type-approved for connection to the public switched telephone network.

Design by R. Jansweyer
[954069]



DIFFERENTIAL VOLTAGE MEASUREMENTS

Sometimes you wish you had an electronic magnifying glass to do a really accurate voltage measurement, or watch voltage fluctuations within a small range. This article shows how that can be achieved with a simple circuit.

Design by F. Hueber

HOWEVER accurate they are, ordinary voltmeters have deficiencies for some applications. For instance, when dealing with a voltage of 9 V which is to be kept as constant as possible, the check should really concentrate on a small range on either side of the target voltage, say, from 8 V to 10 V. Should the measured voltage fall outside this 'window', its exact value is of no importance because the deviation is too large anyway. In other words, what we are after is: being able to see small voltage variations.

An accurate read-out of such small voltage variations is not possible with the aid of an 'ordinary' voltmeter. For instance, selecting the '10-V' position on the meter gives you a range of 0 V to 10 V, in which the desired range, 8 V to 10 V, is simply too small to enable accurate measurements. In fact, what we need is a voltmeter with a range from 8 V to 10 V.

Although some modern DMMs do offer the above 'voltage window' function, it is, obviously, not practicable to incorporate the (expensive) instrument into existing equipment just for the purpose of monitoring fluctuations of a certain voltage. Time to start designing a 'voltage magnifying glass' tailored for the function.

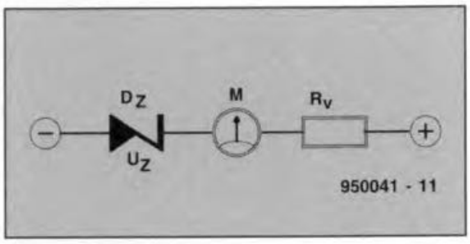


Fig. 1. A voltage magnifying glass can be made with very few components. Unfortunately, this little circuit suffers from poor accuracy.

Voltage subtraction with a zener diode

There are several ways to realize a voltmeter which limits its activities to a relatively small range. One of the simplest options available is shown in **Fig. 1**. A moving coil meter, M, is connected in series with a zener diode, Dz, and a resistor, R_v. The value of the resistor is such that the maximum voltage (in this case, 10 V), produces full-scale deflection (f.s.d.) on the meter. The value of the zener diode is that of the 'low' limit of the desired voltage window, i.e., 8 V.

The effect of the zener diode is that no current flows through the meter coil until the applied voltage exceeds 8 V. In other words, the meter's needle starts to deflect at 8 V, and reaches the f.s.d. point at 10 V. This is exactly what we were after. Resistor R_v takes a value of

$$R_v = (U_{max}-U_{zener})/I_{meter}$$

Unfortunately, the simple circuit of Fig. 1 has an important disadvantage. Because the zener diode starts to conduct a little just before the actual zener voltage, and does no act as an ideal switch when the zener voltage is reached, the start of the scale is badly defined. Another problem is that the zener's temperature dependency degrades the accuracy of this measurement circuit. The upshot is that the circuit of Fig. 1 is not really suitable unless there is a large difference between U_{min} and U_{max}, and that these voltages are not too small. The latter condition is related to the fact that the higher voltage zener diodes have a better defined 'knee' in their U/I curve than lower-voltage ones. Still, be prepared to see highly inaccurate behaviour at the start of the range you wish to monitor.

Bridge circuit

The schematic in **Fig. 2** shows a circuit which does far better in respect of accuracy. What you are looking at is a bridge circuit of which one branch is a reference voltage source consisting of a zener diode which is being supplied with a current I_z via resistor R_z. The other branch is formed by voltage divider R_1-R_2, and also passes a defined current, I_d. A moving coil meter, M, is connected between the junctions of the two branches, A and B, via a series resistor, R_v.

The values of R_1 and R_2 in the voltage divider are selected so that the voltage drop across R_2 equals the reference voltage, U_z. If that is the case, the potential difference between junctions A and B is nought, and no current flows through the meter.

When the applied voltage rises to the maximum value, U_{max}, the potential at junction A remains constant, while that at junction B rises, depending on the ratio R_1/R_2. The resulting voltage difference between A and B causes a current, I_i, to flow through R_v and the meter coil. As before, the value of R_v is selected to give f.s.d. at U_{max}.

That completes the theoretical discussion of the circuit. In practice, things are a little more complex. When the applied voltage rises, I_d and I_z rise. Unfortunately, I_i then also rises because it flows through R_1 and the zener. The effect causes two errors. Firstly, the reference voltage is

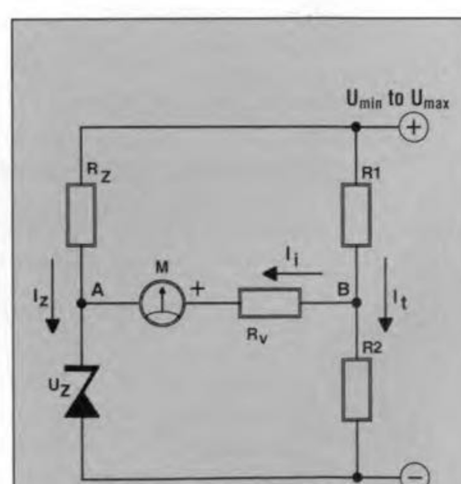


Fig. 2. A circuit with much better accuracy than the one in Fig. 1. Unfortunately, problems arise if the meter has a relatively high current consumption.

changed a little, because the exact zener voltage depends on the current. Secondly, it is hard to avoid the current I_i having an effect on the voltage at junction B. In other words, the accurately defined ratio between R_1 and R_2 is disturbed. The voltage divider is actually loaded, and its behaviour is then difficult to compute.

The problems caused by the measurement bridge are better controlled when I_d, and possibly I_z too, is made much greater than I_i. If the current through the voltage divider is, say, ten times greater than that through the meter, the latter has hardly any effect on the voltage at junction B. Problems occur, however, when a fairly insensitive meter, for instance, a 1 mA or 10 mA type, is to be applied. The total current drain of the circuit then becomes so large that it will form a too heavy load on the voltage source to be monitored.

There are yet other problems. Some of the possible candidates for the function of reference voltage device may not behave as they are expected to. The graphs in **Fig. 3** tell the whole story. Especially with low-voltage zener diodes, the supplied voltage depends strongly on the forward current. This is illustrated by curve 'a' which shows the response of a 3-V zener diode. If you want an accurate low-voltage reference, a green LED (curve 'b') gives much better results, particularly, if you limit the LED current to about 1 mA. By the way, the LED does not light, a condition which never fails to cause confusion with inexperienced constructors about alleged non-functioning of a such a circuit. Curve 'c' shows that zener diodes with a relatively high value are well suited to the application. This curve belongs with a 5.6 V zener diode (read the value on the right-hand scale).

Accuracy, by the way, is always relative. Curve 'd' shows the current dependency of a 'real' voltage reference, in this case, an LM385 bandgap diode. The green LED certainly takes a back seat to the shape of this curve, which remains a straight line over three decades of the current range. In fact, the reference voltage rises only 2 mV against a current rise from 20 µA to 1 mA! Moreover, the temperature coefficient of the bandgap device is negligible at only 20 ppm/K. Although the price of the LM385-2V5 is about ten times that of an ordinary green LED or a zener diode, the device represents excellent value for money if you are after really accurate measurements.

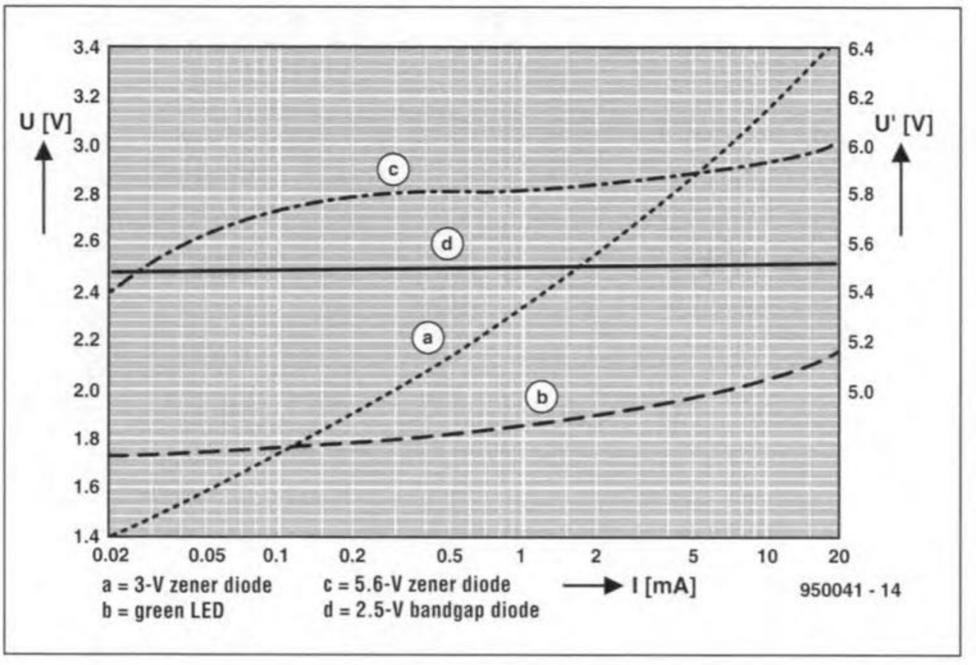


Fig. 3. Current-voltage response of a few reference voltage sources. The green LED and the two zener diodes take a back seat to the extremely stable response of the bandgap diode.

An active measurement circuit

The best way to eradicate all problems with cross-effects between components in the differential voltage monitor is by making the circuit 'active', in other words, by inserting a buffer between the voltage divider and the meter. As illustrated in **Fig. 4**, that is easily accomplished by adding one opamp.

The current through the moving coil meter is now supplied by an opamp, IC_1, whose inputs have such a high impedance that they hardly load voltage divider R_4-R_3-P_1. This eliminates the problems caused by variations of the zener voltage and the voltage supplied by the voltage divider, as discussed above. All resistors are simple to calculate using Ohm's law. The opamp does not need external parts, and almost any ordinary type may be used.

The various component values are computed as follows, based on these assumptions:

Lower voltage, U_{min} = 6V;
Upper voltage, U_{max} = 12V;
Reference voltage, U_{D1} = 2.5 V;
Meter f.s.d. current, I_i = 0.5 mA.

The reference voltage is obtained from an LM385-2V5 bandgap reference. The current which flows through voltage divider R_4-R_3-P_1 is made large with respect to the input current of IC_1. To remain on the safe side, a factor 10 is used. Resistor R_4 then takes a value of 68 kΩ. At the lower voltage, U_{min}, the current through the voltage divider equals

$$I_d = (U_{min} - U_{D1})/R_4 = 0.0514 \text{ mA}$$

The total value of R_3 and P_1 should cause these two resistors to drop the same voltage as D_1 at the previously calculated divider current. So,

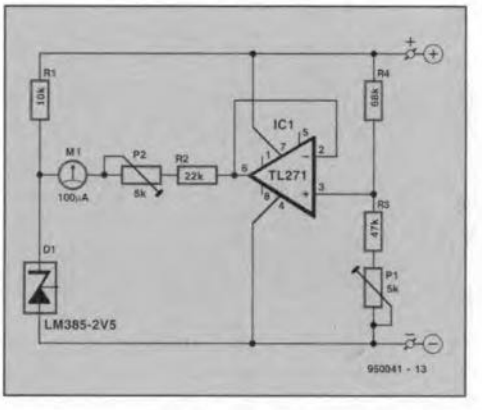


Fig. 4. Most, if not all, accuracy problems with the basic bridge of Fig. 3 may be solved by adding a buffer opamp to drive the meter.

$$R_3 + P_1 = 2.5 \text{ V}/0.0514 \text{ mA} = 48.64 \text{ k}\Omega,$$

which is conveniently arranged by choosing a 5-kΩ preset for P_1, and a 47-kΩ resistor for R_3. The inevitable tolerances may then be compensated by adjusting the preset.

At U_{min} = 6 V, the voltage difference between D_1 and the output of the opamp is nought. As the input voltage rises to the maximum level, U_{max} (12 V), the input voltage of IC_1 also rises to

$$(U_{max}/U_{min}) \times U_{D1},$$

or

$$(12/6) \times 2.5 \text{ V} = 5 \text{ V}$$

A potential difference of 2.5 V then exists between D_1 and the opamp output. Consequently, the meter's series resistor, R_2+P_2, should take a value which causes f.s.d. on the meter at that voltage. The internal resistance of the meter should also be taken into account. As a rule of thumb, this will be about 1 kΩ for a 100-μA meter, 100 Ω for a 1-mA meter, or 10 Ω for a 10-mA meter. With the meter used here, an internal resistance of 1 kΩ should be observed. The series resistor then becomes:

$$
\begin{aligned}
R_v &= (\Delta U/I_i)-R_i \\
&= (2.5\ \text{V}/100\ \mu\text{A})-1\ \text{k}\Omega \\
&= 23\ \text{k}\Omega
\end{aligned}
$$

Here, too, a combination of a fixed resistor and a preset is used for accurate adjustment. The actual values of these components become: R_2 = 22 kΩ and P_2 = 5kΩ.

Adjustment of the 'voltage magnifying glass' is fairly simple. First, apply the lower voltage, U_{min}, and adjust P_1 for zero deflection on the meter. Next, apply U_{max}, and adjust P_2 for f.s.d. on the meter.

Special cases

When U_{max} is very small, you should use a small reference voltage also, for instance, an LED or an LM385-1V25. The output of the opamp used must be capable of handling this lower voltage, so exit the ubiquitous 741. At U_{min}<5 V, it is better to change to a 'rail-to-rail' opamp such as the CA3130.

When U_{max} is very large, and exceeds the maximum supply voltage of the opamp (usually 36 V or ±18 V), the excess voltage may be dispelled by connecting a zener diode in series with the positive supply. In extreme cases, it is better to connect the zener and its current limiting resistor in parallel with the supply.

Problems may also arise if the measurement window is very small. In that case, the voltage rise at the input may not be sufficient to enable the meter to reach its f.s.d. The problem may be remedied by wiring the opamp as a non-inverting amplifier. Using a 10-kΩ resistor between pin 2 and ground, and another 10-kΩ resistor in the feedback path, a gain of ×2 is obtained. Mind you, the output voltage of most ordinary opamps should remain at least 2 V below the positive supply level.

Finally, there may be a problem if the current consumption of the meter is on the high side. The output current of most opamps is limited to 10 to 15 mA. Moving coil meters requiring more current are, therefore, unsuitable for use in the present circuit, unless, of course, an opamp is used which is capable of driving such a large load.

Final notes

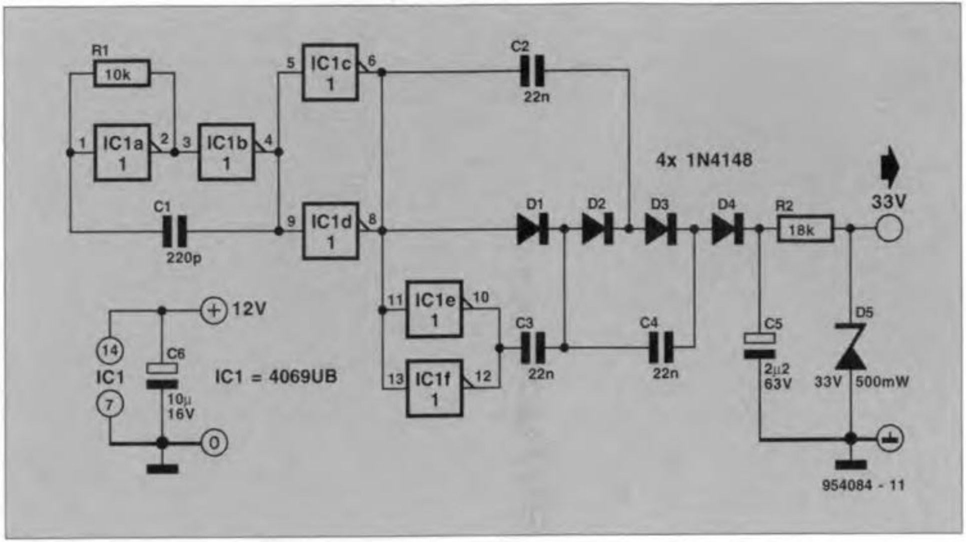
The circuit is, in principle, also suitable for small alternating voltages — just add a rectifier consisting of, say, four 1N4148 diodes and a 220-μF electrolytic capacitor. Do take into account, however, that the voltage across this capacitor is about 1.4 times (minus two diode voltages) higher than the applied alternating voltage. It is also possible to omit the electrolytic capacitor altogether — the rectified voltage is then roughly equal to the average value of the sine-wave, or about $0.9{\times}U_{eff}$. Because of the threshold voltage of the rectifier diodes, the circuit design is a bit more complicated than with direct voltage measurements.

The eight or so components which make up the voltage monitor should not be difficult to fit on a piece of veroboard or stripboard. With some dexterity, this board may be secured to the connections of the moving coil meter, which results in a fairly compact module. (950041)

DC-DC CONVERTER

There are frequent occasions in which a circuit needs a low-current supply at a voltage appreciably higher than the 12 V provided by a standard mains adaptor. A good example of this is the tuning voltage in a v.h.f. receiver, which is normally of the order of 24–33 V.

Oscillator IC_{1a}-IC_{1b} generates a rectangular signal at a frequency of about 200 kHz. This signal is applied to a multiplier which raises the signal level ×4. When pin 10 of IC_1 is low and pin 6 is high, C_3 is charged to about 12 V. When just afterwards the levels at these two pins are reversed, the potential at junction C_3-D_1-D_2 rises to almost double the supply voltage. This voltage is used to charge C_2. When pin 6 becomes high again shortly afterwards, the potential at junction C_2-D_2-D_3 is three times the supply voltage, that is, 36 V with respect to earth. This voltage is used to charge C_4 to 24 V during the next half cycle, so that during the following half cycle the potential across C_4 is 48 V with respect to earth.



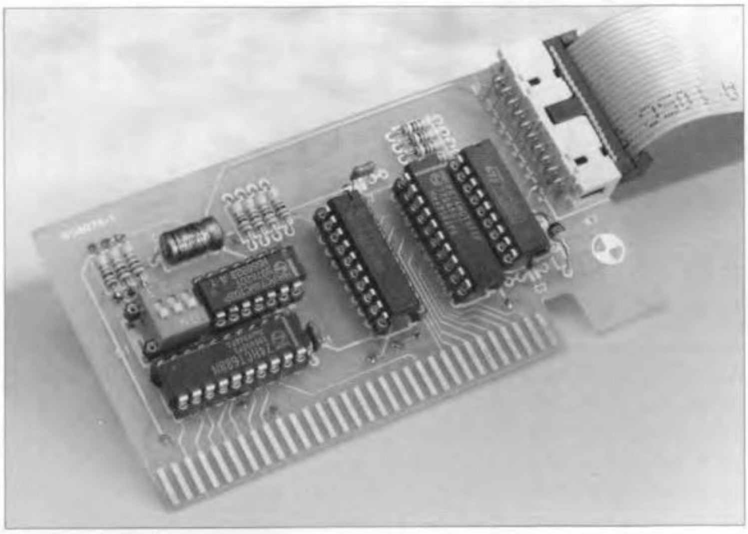
Owing to unavoidable losses, for example in the diodes, the practical output voltage is rather lower than the theoretical value. Nevertheless, in the prototype, with D_5 removed, the output voltage was 46 V. In the practical circuit shown, the output is held at 33 V by D_5-R_2. The maximum output current is 1 mA.

The total current drawn by the converter is about 18 mA.

Design by H. Bonekamp

[954084]

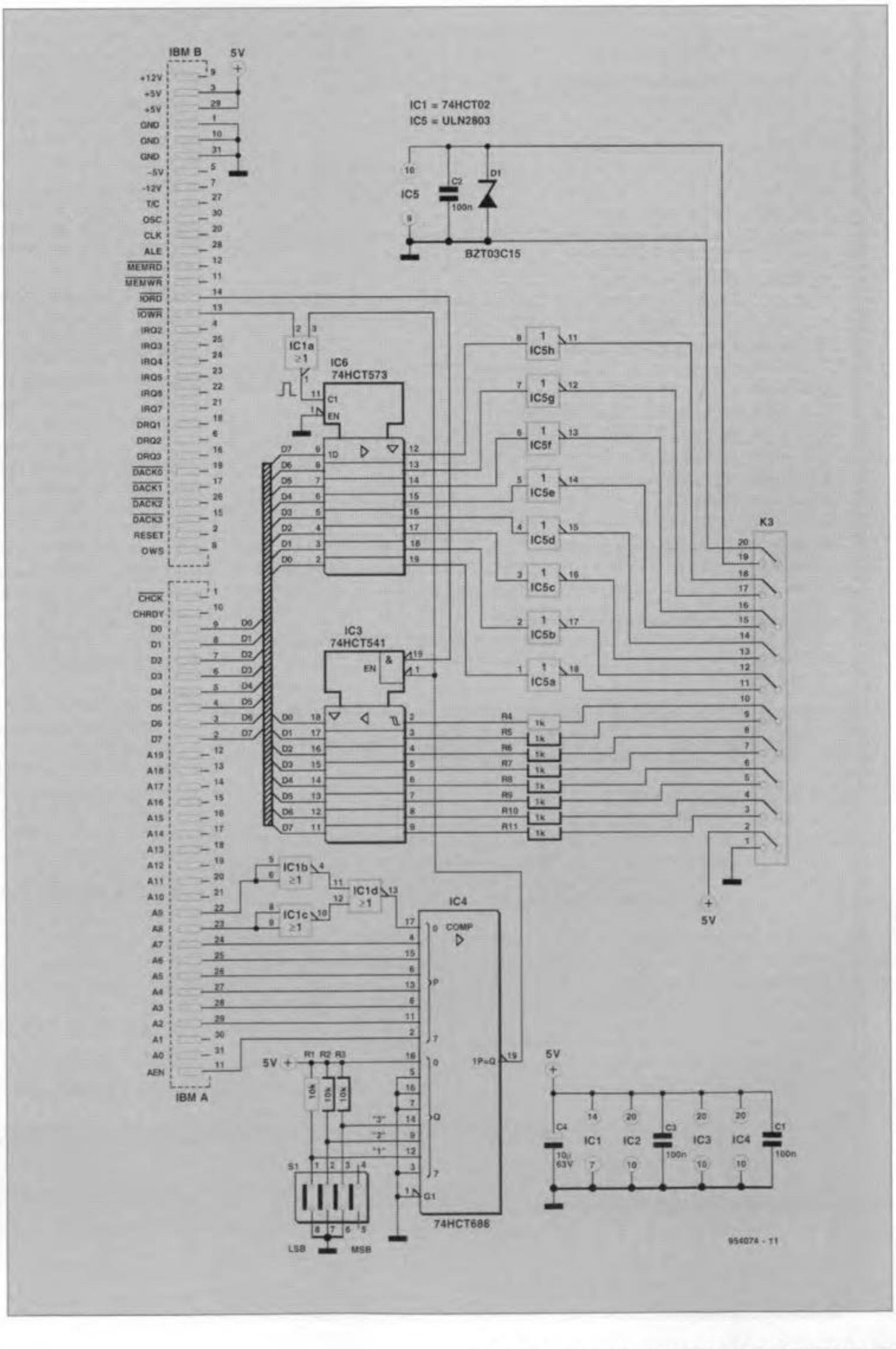
SIMPLE I/O CARD



The I/O card provides
- 8 inputs at TTL level;
- 8 outputs with data memory and power buffer;
- 8 preset I/O addresses.

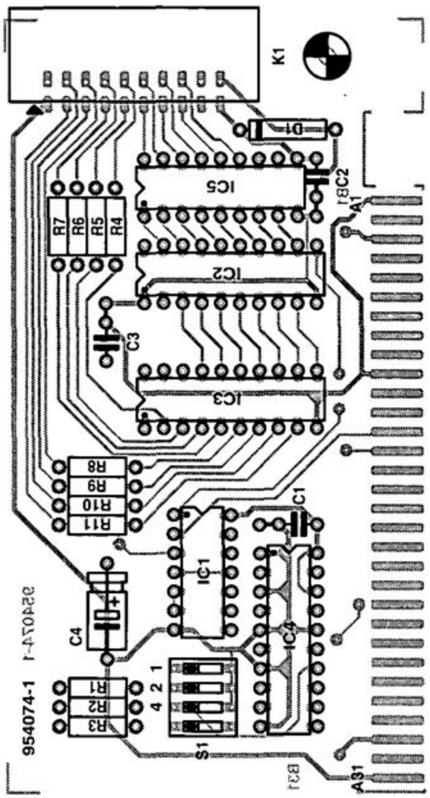
Three sections of dipswitch S_1 are used for selecting an I/O address. When all switch sections are closed (on), the address is 300H. Other addresses are given in the table.

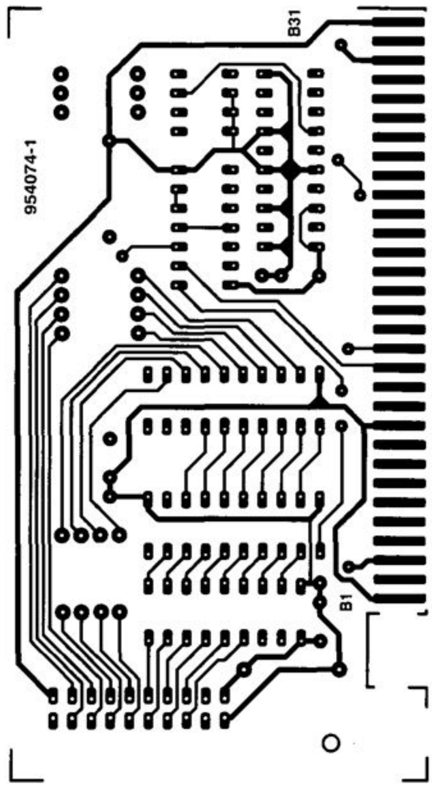
Address (hex)	'4'	'2'	'1'
300	on	on	on
304	on	on	
308	on		on
30C	on		
310		on	on
314		on	
318			on
31C			

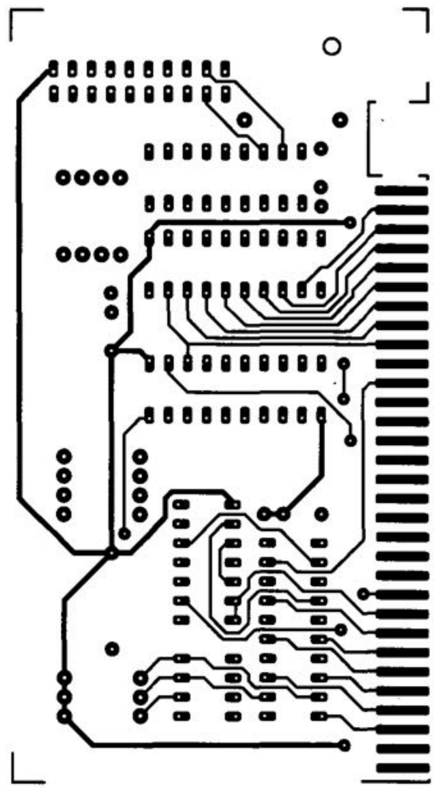


This address may be used to write to, and read from, the card. Written data are stored in IC_2 and are available via buffer IC_5. The properties of this buffer are: open collector, maximum 50 V, maximum 500 mA. Its outputs are collectors of darlington transistors that switch to earth. This arrangement has several consequences. It makes it possible for output potentials higher or lower than 5 V to be switched. Such potentials must be applied externally, that is, the external loads or pull-up resistors must be supplied from an additional voltage source with respect to earth: pin 20 of K_1. The chip has internal diodes that protect the transistors against overvoltage in case of inductive loads. These diodes are all connected to pin 10, from which a fast zener diode, D_1, is linked to earth. The rating of D_1 must be higher than the level of the external voltage source. The specified rating of 15 V is all right for an external voltage of 12 V. If the external potential is 24 V, the zener diode must be rated at 30 V. Apart from this, it is good practice to shunt an inductive load with a freewheeling diode.

The inputs at connector K_1 are applied to buffer IC_3 via additional protection resistors. Note that this buffer should preferably be an HC and not an HCT type, because, when the inputs are used to measure the level at the darlington outputs, a potential difficulty arises. Since the output of a darlington transistor is never lower than 0.7–1.0 V, it may happen that an HCT type does not detect a logic low.







According to manufacturers' specifications, HCT types detect a logic low only if this is below 0.8 V. An HC type has a slightly higher switching threshold.

The inputs are at TTL level, which means that input signals must not exceed 5 V. If necessary, a potential divider can be used at the inputs. Resistors R_4–R_{11} provide additional protection, but it is doubtful whether IC_3 would survive long if permanent 12 V inputs were applied to it.

The I/O card is intended to be constructed on the printed-circuit board shown. The introductory photograph shows the completed board.

Parts list

Resistors:
R_1, R_2, R_3 = 10 kΩ
R_4–R_{11} = 1 kΩ

Capacitors:
C_1, C_2, C_3 = 100 nF
C_4 = 10 µF, 63 V

Semiconductors:
D_1 = BZT03C15

Integrated circuits:
IC_1 = 74HCT02
IC_2 = 74HCT573
IC_3 = 74HC541
IC_4 = 74HCT688
IC_5 = ULN2803

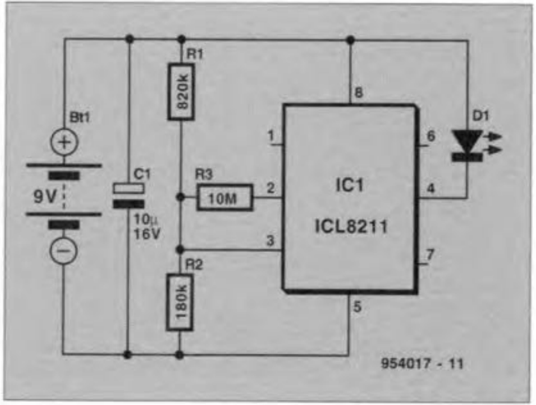
Miscellaneous:
K_1 = 20-way right-angle box header
S_1 = 3- or 4-section DIP switch
PCB Order No. 954074 (see p. 70)

Design by K. Walraven
[954074]

'BATTERY LOW' INDICATOR

Batteries often go flat at the most inopportune moments. What's more, frequently, this is not detected until it is too late, but this can be prevented by the simple circuit shown in the diagram. It enables the battery voltage (here, 9 V) to be monitored and give an indication when the potential drops below a certain level.

When the battery voltage drops below 5.9 V, the LED lights: it will go out only when the voltage rises above 6.3 V. The circuit draws acurrent of about 40 µA when the LED is out, but increases to about 10 mA when the diode lights.



The thresholds are determined with resistors. For the lower limit (at which the LED lights):

$$U_{t1} = (R_1R_3/(R_1+R_3)+R_2)(1.115/R_2) \quad [V]$$

and for the lower limit (at which the diode goes out):

$$U_{t2} = \{(R_2+R_1)/R_2\}\times 1.115 \quad [V].$$

These two equations enable the circuit to be adapted to any given battery.

The Type ICL8211 is a chip from Intersil, which is now part of Harris.

Design by H. Bonekamp [954017]

SMART SCART BOX

A problem occurs when you have a comfortable chair, a satellite TV tuner with only one SCART connector, and two decoders, also with only one SCART (loop-through) connection. Such a situation exists, for instance, in a setup consisting of a Pace 800rd Sat-TV receiver, a Luxcrypt RTL4/5 decoder, and a D2MAC decoder (for TVPlus). Although small SCART boxes are available commercially to connect three SCART apparatuses, manual switching is then required, in this case, between the Luxcrypt and the D2MAC decoder. Obviously, anything to do with manual switching is a drag in this day and age of channel hopping and remote controls, whence the present unit, which selects the actuated decoder automatically, without the need to get out of your chair and flick a switch.

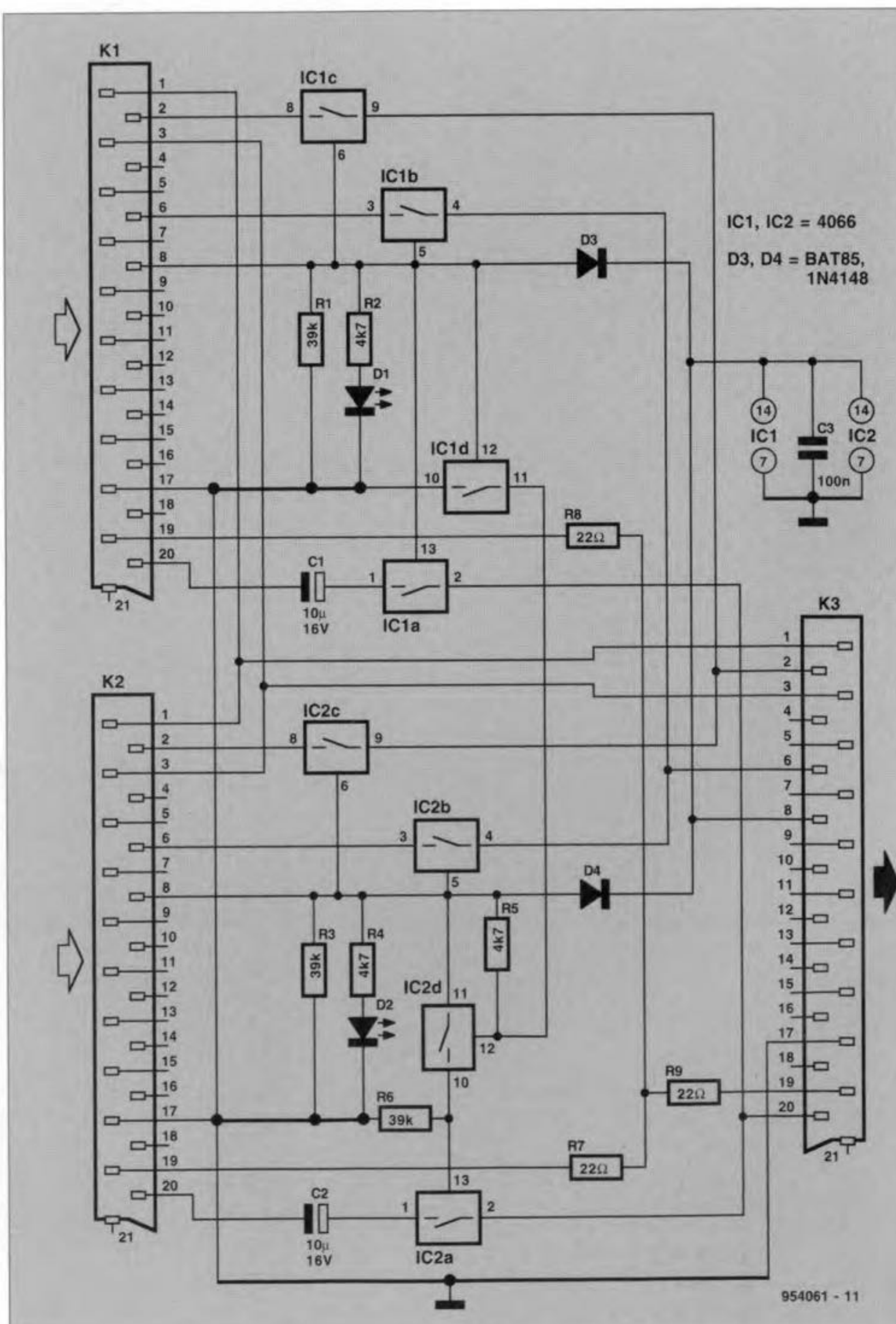
The circuit shown here offers automatic switching between two video sources which, when actuated, supply a switching voltage at pin 8 of their SCART connector. Most satellite TV decoders, as well as video recorders, have such a function. The smart SCART box does not require an external supply voltage, or a battery — the internal supply voltage is stolen from the previously mentioned SCART switching voltage.

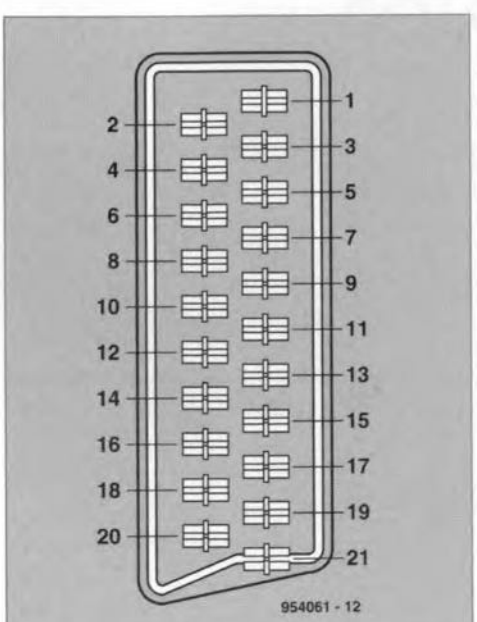
The circuit diagram shows a straightforward two-channel switching arrangement based on the familiar, low-cost, Type 4066 electronic bilateral switches. The 39-kΩ resistors on the switching voltage lines prevent undefined levels at the control inputs of the 4066. Two high-efficiency LEDs show the status of each decoder. The LED current is kept to a minimum to keep the load on the switching voltages as small as possible. The (tantalum) coupling capacitors in the video line, C_1 and C_2, cancel the effect of any offset voltages on the video signal supplied by the decoder. In particular, the author's Luxcrypt decoder was found to have a d.c. offset in its output signal, which caused signal leakage through the relevant 4066 switch when a D2MAC signal with high brightness was present. The tantalum capacitors should prevent this effect.

If the smart SCART box is used with two VCRs and a TV set, the VCR at input I has priority.

The circuit has one shortcoming which is caused by the absence of a power supply, and should not be left unmentioned. The video signal supplied by the satellite TV tuner is distributed between two loads without amplification. This obviously causes signal loss because of the too low load impedance which then occurs at the receiver output. To prevent overloading, you may want to use resistors R_x (start with approx. 82 Ω) between the video output of the receiver and each decoder video input. In some cases, however, these resistors may not be necessary because the lower signal level may either be compensated by setting a slightly higher gain in the decoder (Luxcrypt), or it may not cause problems at all (D2MAC decoder).

It is suggested to build the circuit into





an existing 'dumb' SCART box, which is available cheaply, and saves you the trouble of drilling and hunting for three panel-mount SCART sockets. The component count being quite low, it should not be a problem to construct the circuit on a small piece of veroboard, and use wires to make the connections to the relevant SCART sockets pins.

Design by F.M. Bemelman

[954061]

PIN DESIGNATIONS FOR SCART CONNECTORS

Pin	Decoder SCART	TV SCART	VCR SCART
1	R Audio Out	R Audio Out	R Audio Out
2	R Audio In	n.c.	R Audio In
3	L Audio Out	L Audio Out	L Audio Out
4	Audio Ground	Audio Ground	Audio Ground
5	Blue Ground	Blue Ground	Ground
6	L Audio In	n.c.	L Audio In
7	Blue In	Blue Out	n.c.
8	Ext. Decoder Status In	AV Status Out	AV Status In
9	Green Ground	Green Ground	Ground
10	n.c.	n.c.	n.c.
11	Green In	Green Out	n.c.
12	Serial data I/O Port	n.c.	n.c.
13	Red Ground	Red Ground	Ground
14	n.c.	n.c.	n.c.
15	Red In	Red Out/SVHS Chroma	SVHS Chroma
16	RGB Status In	RGB Status Out	n.c.
17	Video Ground	Video Ground	Video Ground
18	Ground	Ground	Ground
19	Baseband Video Out	Video Out	Video Out
20	Video in	n.c.	Video in
21	Ground (casing)	Ground (casing)	Ground (casing)

6 V MOTIVE-BATTERY CHARGER

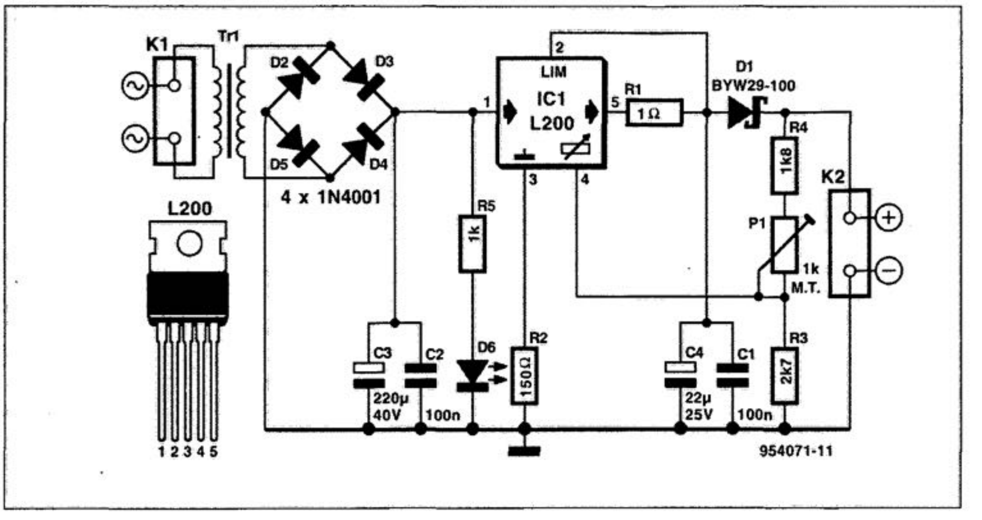
The motive-battery charger described in Ref. 1 has met with great interest, particulary among owners of motor cycles. In response to many requests from owners of vintage cars and motorcycles, a modification is described which allows the charger to keep 6-V batteries in shape during the winter season, when the vehicle or machine is not used.

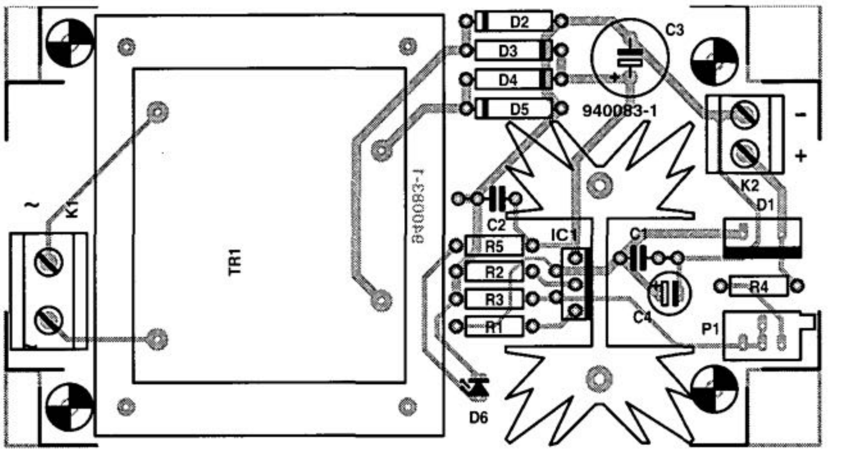
The nominal output voltage of the L200 regulator is calculated from

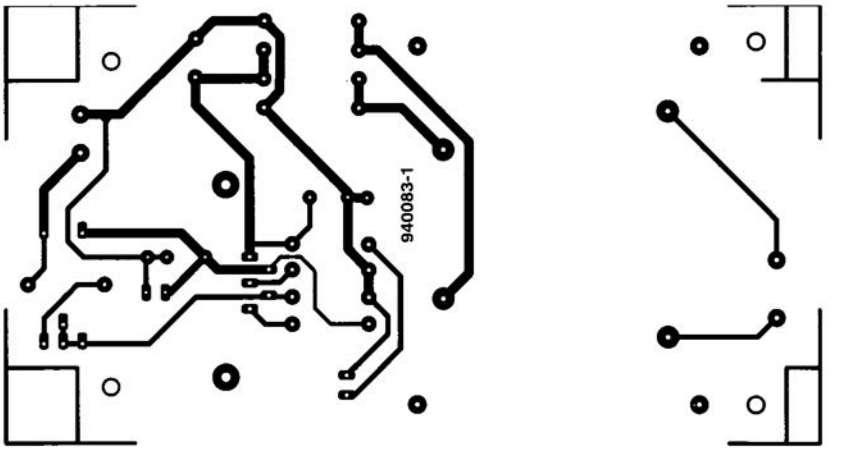
$$U_{\text{out}} = U_{\text{ref}} \times \left(1 + \frac{R_4 + P_1}{R_3}\right)$$

where U_{ref} is 2.77 V (typ.). So, change R_3 to 2.7 kΩ, and R_4 to 1.8 kΩ, and Bob's your uncle. These values allow the charger's output voltage to be adjusted between 6 V and about 7.7 V. The LED series resistor, R_5, is lowered to 1 kΩ.

Because of the lower output voltage (6 V instead of 12 V), the dissipation in the L200 may be on the high side if the original transformer is used. If you need the present 6-V version only, the transformer is, therefore, best changed into a type with a 9-V secondary. Suitable types are the Block VR13/1/9 (9V/1.44 A), the Monacor (Monarch) VTR12109 (12 VA), or the Block PT13/1/9. The latter is a short-circuit protected type (more expensive than the others, unfortunately). All three transformers are rated at 230 V primary voltage, and fit on the printed circuit board. Also do not forget to change the fuse to a 63-mA type. As with the 12-V version of the charger, the current limiter is actuated at a level of 0.5 A.







Parts list

Resistors:
R_1 = 1 Ω
R_2 = 150 Ω
R_3 = 2.7 kΩ
R_4 = 1.8 kΩ
R_5 = 1 kΩ
P_1 = 1 kΩ multiturn preset

Capacitors:
C_1, C_2 = 100 nF
C_3 = 220 µF, 40 V, radial
C_4 = 22 µF, 25 V, radial

Semiconductors:
D_1 = BYW29-100
D_2–D_5 = 1N4001
D_6 = LED, red, 5 mm

Integrated circuits:
IC_1 = L200CV (5-pin)

Miscellaneous:
K_1 = 2-way terminal block, pitch 7.5mm
K_2 = 2-way terminal block, pitch 5 mm
Tr_1 = short-circuit-proof mains transformer, 9 V, 12 VA – see text
Enclosure, synthetic fibre 120×65×65 mm ($4^3/_4$×$2^1/_2$×$2^1/_2$ in)
Heat sink 5 K W^{-1}, complete with fitting/insulating kit
PCB Order no. 940083 (see p.70)

Design by L. Lemmens
[954071]

Reference:

1. Motive-battery charger, *Elektor Electronics* October 1994.

MODIFIED SCART SWITCHING BOX

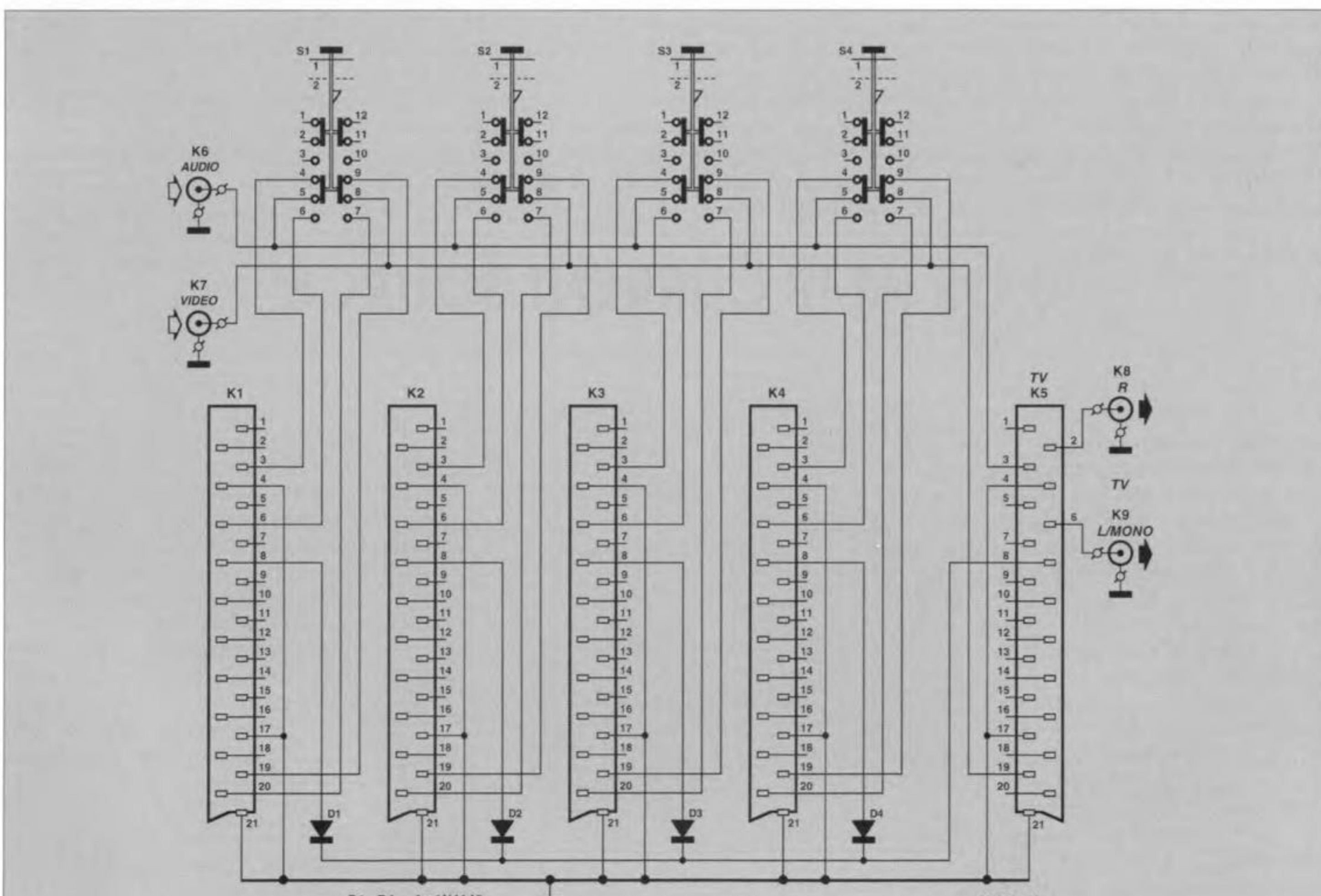
The SCART switching box described in Ref. 1 was carefully designed to make sure that any input is connected to only one output at a time. This was done to prevent signal loss and mismatching effects on the video and audio signals. A rather simpler design is presented here which lets go of the 1-signal-to-1-output principle. The result is a considerable simplification of, in particular, the switches and the way they are wired.

In practice, the present version of the SCART box was tested with a setup consisting of two video recorders and a TV set. Although some signal loss was noted, no serious problems occurred owing to mismatching which is inherent in driving two 75-Ω loads from one 75-Ω source. The advantage of the design shown here is that it is readily extended with one more SCART channel. The switches used should be mechanically coupled (interlocking), which means that only one switch can be depressed at a time.

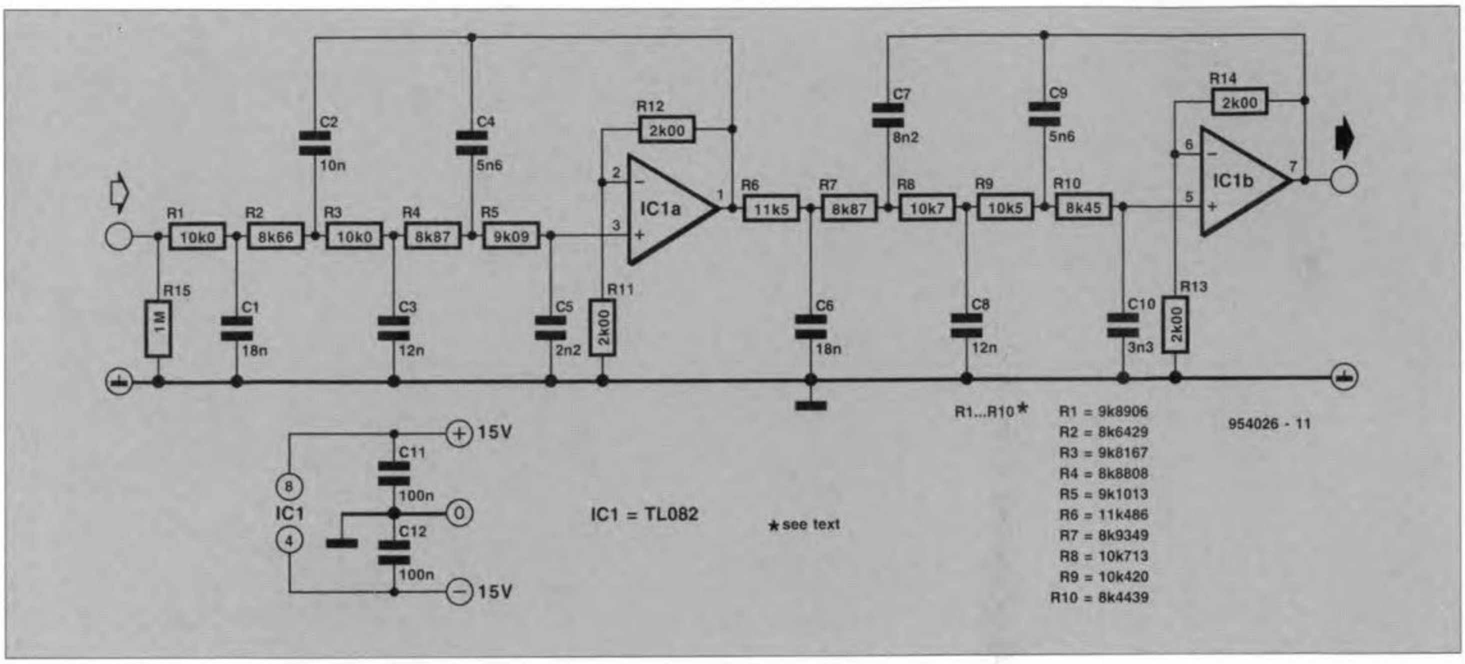
Design by C. Romaldini
[954057]

Reference:

1. SCART switching box, *Elektor Electronics* December 1993.



10TH-ORDER BESSEL FILTER



A Bessel filter has a fairly constant transit time (delay) over the pass band, but its amplitude characteristic starts to fall off rather earlier than in a Butterworth filter. Moreover, the skirts of the amplitude characteristic are not as steep as those of a Butterworth filter. For instance, the –0.1 dB point in a Bessel filter occurs at 184 Hz, and the –1 dB point at 581 Hz; the same points in a Butterworth filter occur at 827 Hz and 934 Hz respectively.

To make the skirts of a Bessel filter steeper, a 10th-order type may be used and this can be constructed by cascading two 5th-order sections. The two sections can not be taken at random, however: they must be attuned to one another. As the diagram shows, the resistor and capacitor values in the second section are different from those in the first section. This is because the first section has a fairly high gain at frequencies beyond the cut-off point, whereas the second has to arrange an early onset of the falling off of the characteristic. Although the cascade may be reversed, it is better for the signal-to-noise ratio to place the higher-gain section first, as shown in the diagram.

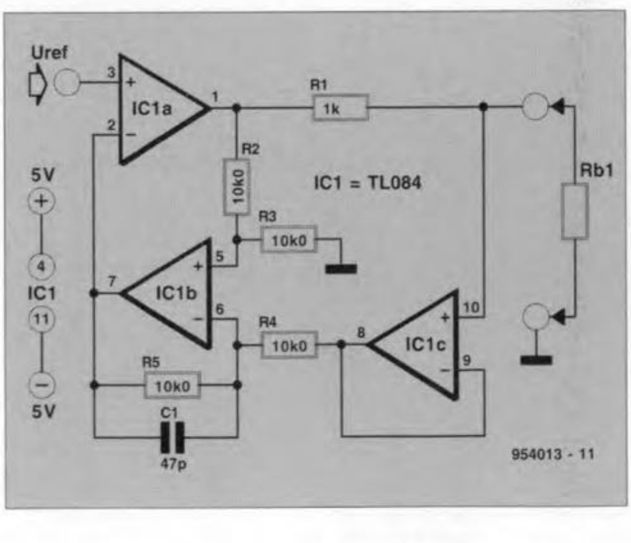
Although not optimal, the specified values of the capacitors are rounded off to E-12 values and those of the resistors to E-96 values. These values give a small variation in both the delay (≤ 0.25%) and the cut-off frequency (990 Hz instead of 1 kHz). If these variations are unacceptable, use combinations of resistors to give values as shown in the table in the diagram.

The current drawn by the TL082 is about 4 mA.

Design by T. Giesberts
[954026]

3 OP AMP CURRENT SOURCE

A current source based on one op amp can not easily be made variable, nor is the current very stable when standard resistors are used. The present design may be made variable and the current is held within reasonably accurate limits.



In contrast to a single op amp source, the current design uses active feedback, which serves to make the potential drop across R_1 qual to the reference voltage. The output current is thus U_{ref}/R_1.

To keep the output voltage across R_1 constant, differential amplifier IC_{1b} monitors the potential at both ends of the resistor. This potential is applied to the inverting input of IC_{1a}, which compares it with U_{ref}. Depending on the result of this comparison, the potential at pin 1 will increase or decrease until an equilibrium is reached whereby $U_{R1} = U_{ref}$.

It is necessary that $R_2 = R_3 = R_4 = R_5$. Also, IC_{1c} prevents the current through R_1 being loaded by a possible measuring current, provided its input resistance is not too low. Capacitor C_1 is necessary for frequency compensation since the delay of the control loop is relatively long.

If these requirements are met, the design provides a good current source. If a variable output current is required, R_1 must be replaced by a combination of a fixed and a variable resistor.

Design by H. Bonekamp
[954013]

MAINS SLAVE

The principle of the circuit is that on the basis of the current drawn by a mains-operated master equipment another appliance, the slave, is switched on. Some units that provide this function make use of one or more diodes connected in series with the master power line. This inevitably means a loss of voltage to the master and (albeit slight) interference on the mains. The present circuit uses a current transformer, which causes a loss of voltage to the master of only a few millivolts.

The current through the primary of Tr_1 produces a proportional voltage across the scondary that is independent of the number of primary turns. Basically, therefore, one turn would suffice, but in practice two turns produce a mechanically more stable winding. The prototype reached saturation at about 0.5 A, after which the secondary voltage no longer increases. It is thus possible to detect the switching on of appliances with a standby power of roughly 100 W. In practice, the standby power is only a few watts, and the ratio of operating power to standby power will be large.

The secondary winding and C_2 and C_3 form a resonant circuit tuned to 55 Hz, which suppresses mains and r.f. interference, and increases the sensitivity. With a mains frequency of 50–60Hz, this is fine, but if it deviates largely from this, it may be necessary to adapt the number of (secondary) turns (fewer turns for a higher frequency). This may reduce the sensitivity, but that can be countered by the range of P_1.

Preset P_1 sets the level of current through the master at which the slave must be actuated. The values of R_6, R_7, and P_1 give a potential at junction R_6-P_1 of about 0.5 V above that at junctions R_4-R_5 and P_1-R_7. When that potential is exceeded, IC_1 outputs a 50 Hz pulsating voltage. This voltage is smoothed by D_6 and C_6 and then used to switch on T_1 via R_8, whereupon relay Re_1 is energized. The relay contacts then connect the slave to the mains.

Potential divider R_8-R_9 ensures that when the output of IC_1 is permanently low (it can not drop to zero), T_1 can not be switched on

Diode D_6 and C_6 provide instant switch-on, but delayed switch-off, of the slave

When smaller powers need to be detected, the number of primary turns may be increased. This means that the value of C_2 and C_3 must be altered as well to retain the resonant frequency. The inductance of the secondary winding is $L = A_L \cdot n^2$, where A_L is a factor determined by the the core of the transformer (in the prototype: 10 μH ±25%), and n is the number of turns. The total capacitance of the parallel connection C_2-C_3 = C is given by

$$C = 1/(2\pi f)^2 \cdot L \approx 8.4 \cdot 10^{-6} \cdot 1/L.$$

The 24-V supply for IC_1 and the potential dividers is derived directly from the mains voltage by capacitive divider C_1-D_5. The current through C_1 is about 30 mA, which is sufficient for the present circuit. When the mains voltage is appreciably different from 240 V, the value of C_1 must be recalculated. In that case, do not forget the voltage drop across R_1, D_1–D_4 and C_4, totalling about 32 V).

Resistor R_1 limits the current when the mains is switched on first. The total dissipation is about 1 W. The relay draws a current of about 28 mA.

It may be necessary with certain loads to place a snubber network (R = 220 Ω, 1 W; C = 100 nF, 630 V) across the contact of Re_1.

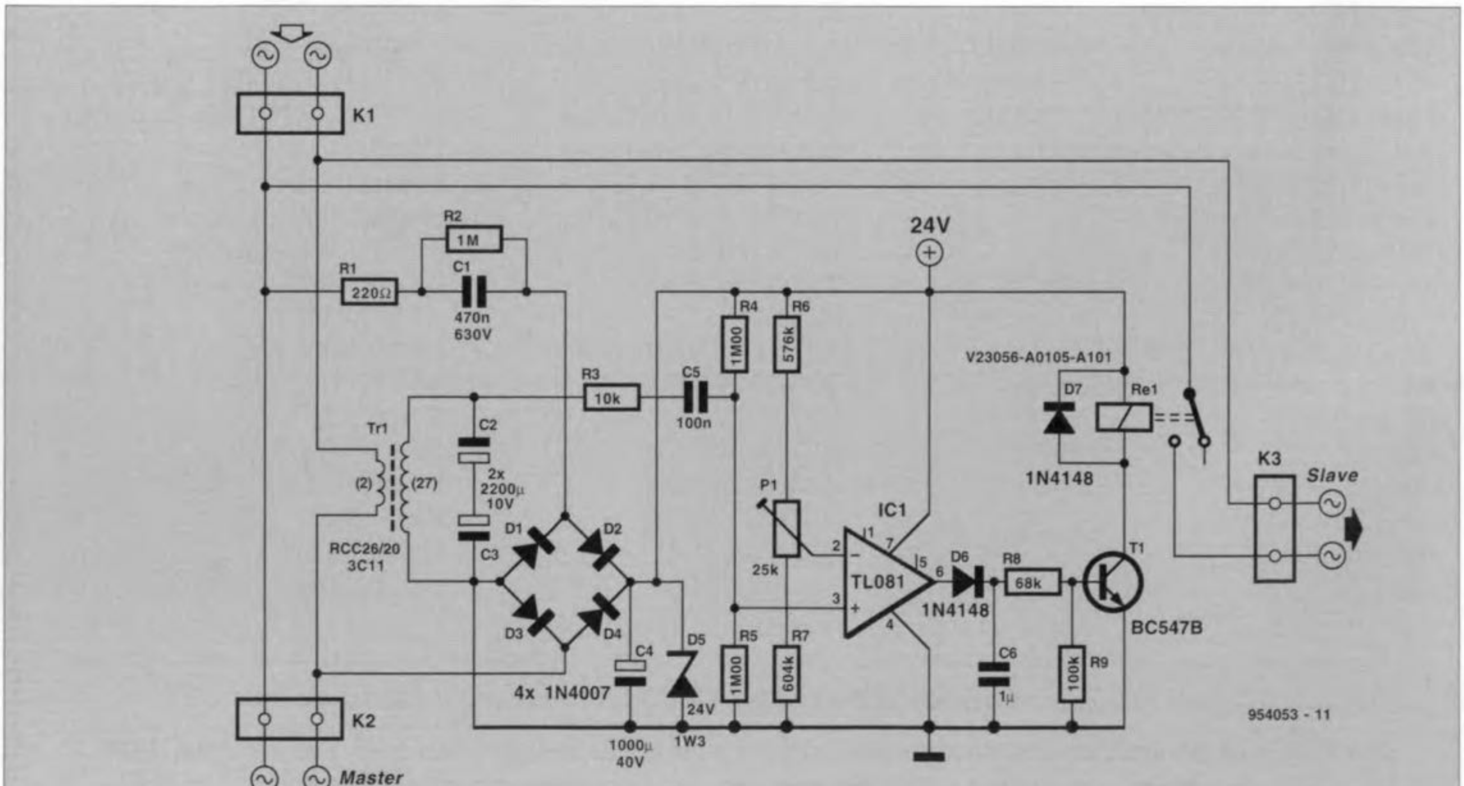
The secondary winding of Tr_1 is wound from 0.7 mm enamelled copper wire. The diameter of the wire for the primary winding is determined by the maximum permissible current to the master equipment.

If the circuit is followed faithfully, the potential difference between the primary and secondary side of Tr_1 is the sum of the forward voltages of the diodes, that is, about 27 V. In most other cases, the difference may be as high as the mains voltage.

GREAT CARE SHOULD BE TAKEN WHEN WORKING WITH THE CIRCUIT SINCE IT IS CONNECTED DIRECTLY TO THE MAINS.

Design by T. Giesberts

[954053]



LIQUID-LEVEL GAUGE

Since measuring liquids with a direct voltage quickly leads to catalysis effects and corrosion, the gauge uses a rectangular voltage. The gauge is intended primarily for measuring the level in a rainwater reservoir, but can, of course, also be used for measuring other kinds of liquid. In the prototype, eight sensors in the reservoir were used. Placing these sensors at well-thought-out locations will minimize any errors in the measurements.

Counter-oscillator IC_4 functions as the central control of the circuit. The oscillator frequency is determined by R_1-R_2-C_{20} and is here 76 kHz. The test voltage produced by IC_4 is applied to the sesors via potential divider R_3-R_4, multiplexer IC_5, capacitors C_{12}–C_{19} and K_1.

The sensors are scanned sequentially at a frequency of 76 kHz. When a sensor remains dry, the test voltage is unchanged. When it is under water, however, an additional load is placed on potential divider R_3-R_4, which results in the test voltage dropping to some extent. The test voltage is demodulated by R_5-C_1-D_1 and applied to comparator IC_2.

Circuit IC_4 not only selects the sensors, but also arranges, in conjunction with IC_3, the multiplexing of the eight indicator LEDs. For this latter purpose, pin 12 of IC_3 is 'misused' as enable input, whose level determines whether the relevant LED lights or not. It is basically possible to link pin 12 directly to the (inverted) output of IC_2, but this would result in a tiny cosmetic error. Owing to time constant R_5-C_1 the test voltage of a sensor is retained until just after the next switching action of the multiplexer and this causes the first LED to light very faintly although it should have remained off. Gates IC_{6a} and IC_{6d} obviate this happening.

Gates IC_{6b} and IC_{6c} create a clear 'empty' indication: D_{11} flashes if the reservoir is empty. In this situation, C_7 is charged via R_9, whereupon pin 4 of IC_{6b} goes high. A choice may be made with jumper JP_1 whether D_{11} flashes in rhythm with the output at pin 3 of IC_4 (about 4.7 Hz) or lights continuously. As soon as one of the sensors is under water, C_7 is discharged periodically via D_2, so that D_{11} can not light.

The gauge is intended to be built on the printed circuit shown, which is, however, not available ready made. Do not forget the wire bridges before populating the board.

The sensors may be made from nine pieces of bare wire of increasing length. The earth link must be as long as sensor 1 and should not be too far from the other sensors. In the prototype, a piece of flatcable was used, whose cores were cut to different lengths; about 3 mm of the insulation was removed from the ends. The earth link must, however, be completely bare.

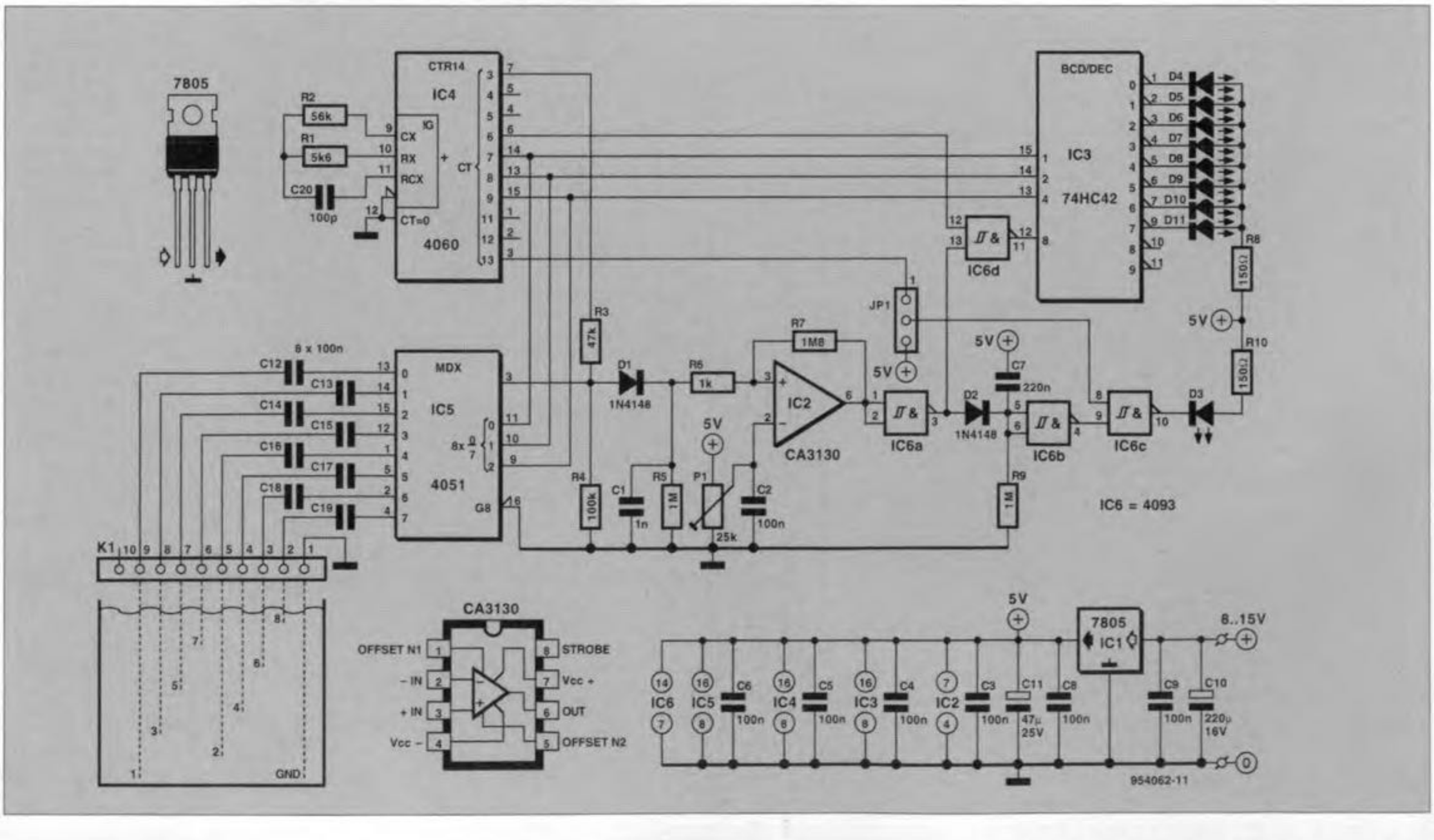
The gauge is calibrated by immersing the electrodes to a certain depth in, say, a glass of water and setting the reference level of IC_2 with P_1 such that the difference between dry and wet sensors is clearly indicated by the display. Once this is done, P_1 may, if desired, be replaced by two fixed resistors of appropriate value.

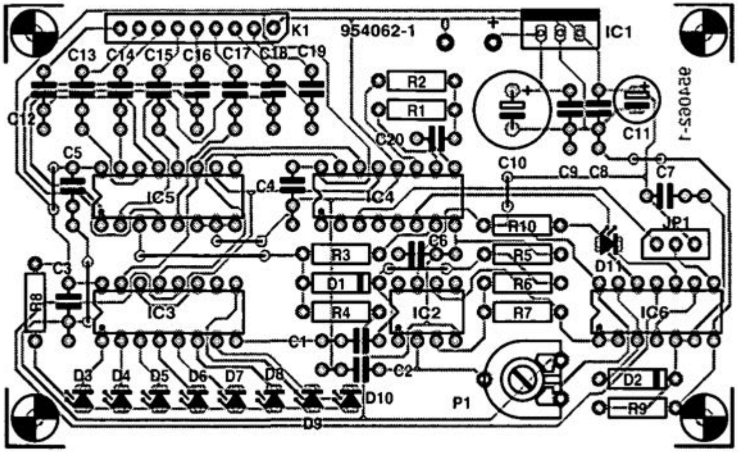
Power for the gauge may taken from a simple 8–15 V mains adaptor.

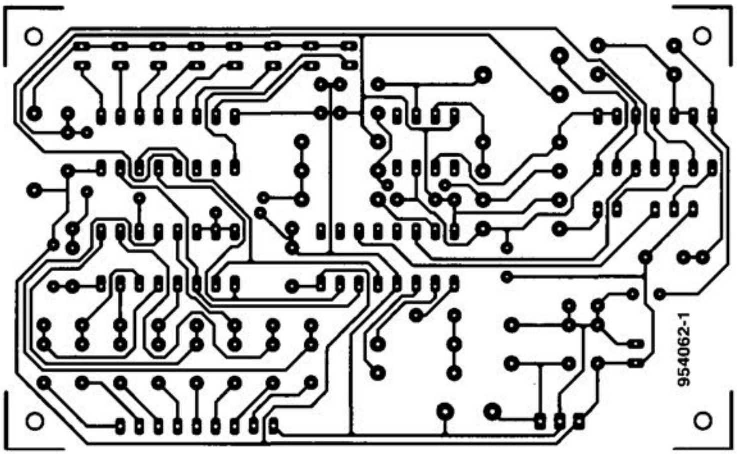
Parts list

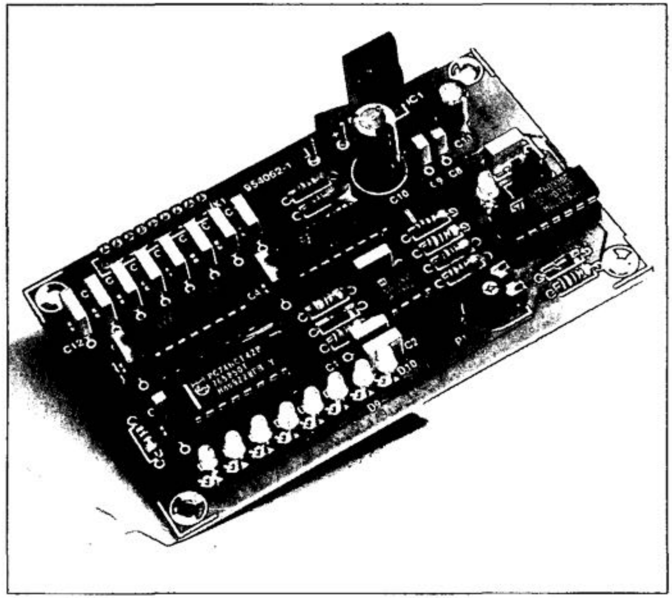
Resistors:
R_1 = 5.6 kΩ
R_2 = 56 kΩ
R_3 = 47 kΩ
R_4 = 100 kΩ
R_5, R_9 = 1 MΩ
R_6 = 1 kΩ
R_7 = 1.8 MΩ
R_8, R_{10} = 150 Ω
P_1 = 25 kΩ preset

Capacitors:
C_1 = 1 nF
C_2–C_6, C_8, C_9, C_{12}–C_{19} = 100 nF
C_7 = 220 nF
C_{10} = 220 µF, 16 V
C_{11} = 47 µF, 25 V
C_{20} = 100 pF









Semiconductors:
D_1, D_2 = 1N4148
D_3–D_{10} = LED, low current
D_{11} = LED, red

Integrated circuits:
IC_1 = 7805
IC_2 = CA3130
IC_3 = 74HCT42
IC_4 = 4060
IC_5 = 4051
IC_6 = 4093

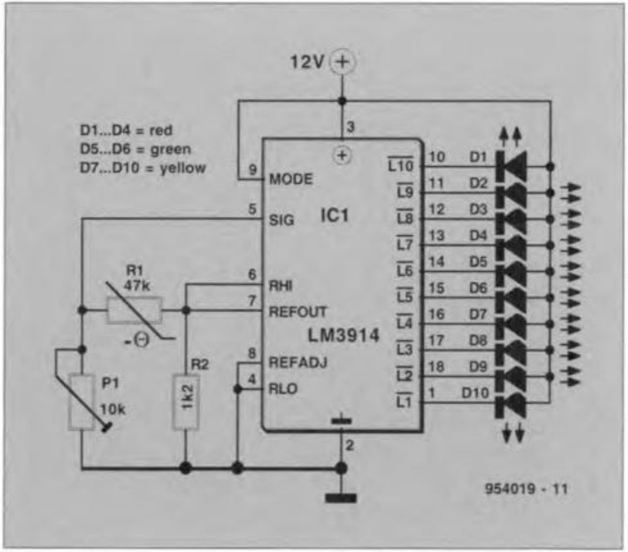
Miscellaneous:
JP_1 = jump lead
K_1 = SIL connector
PCB 954062 – not available ready made

Design by U. Werner
[954062]

MOTORCYCLE TEMPERATURE GAUGE

Many motorcycles have no temperature gauge – which means that their riders have to guess whether the engine is hot, cold or lukewarm (by putting their hand on or near the engine). The present circuit takes this guesswork away. It uses a linear voltmeter IC which drives a display of ten LEDs in a bar- or dot-mode.

The drive voltage (SIG) for the IC is derived from a potential divider consisting of a sensor, R_1 (a resistor with a negative temperature coefficient – NTC), and preset P_1. This combination is adjusted so that an operating temperature of 80 °C causes the first green LED to light (arranged with a bowl of water at this temperature).



Several types of NTC-resistor may be used. The most convenient is, of course, one with a screw fitting, since it needs to be fitted at or near the engine. Make sure that it can withstand the fairly high temperatures of the engine block: these may reach 120 °C.

The colours of the LEDs can, of course, be varied to individual taste; those specified are used on the prototype: four yellow ones for cold; two green ones for normal; and four red ones for hot. The LEDs are connected in the bar-mode by strapping pin 9 of the IC to pin 3, and in the dot-mode by leaving pin 9 open.

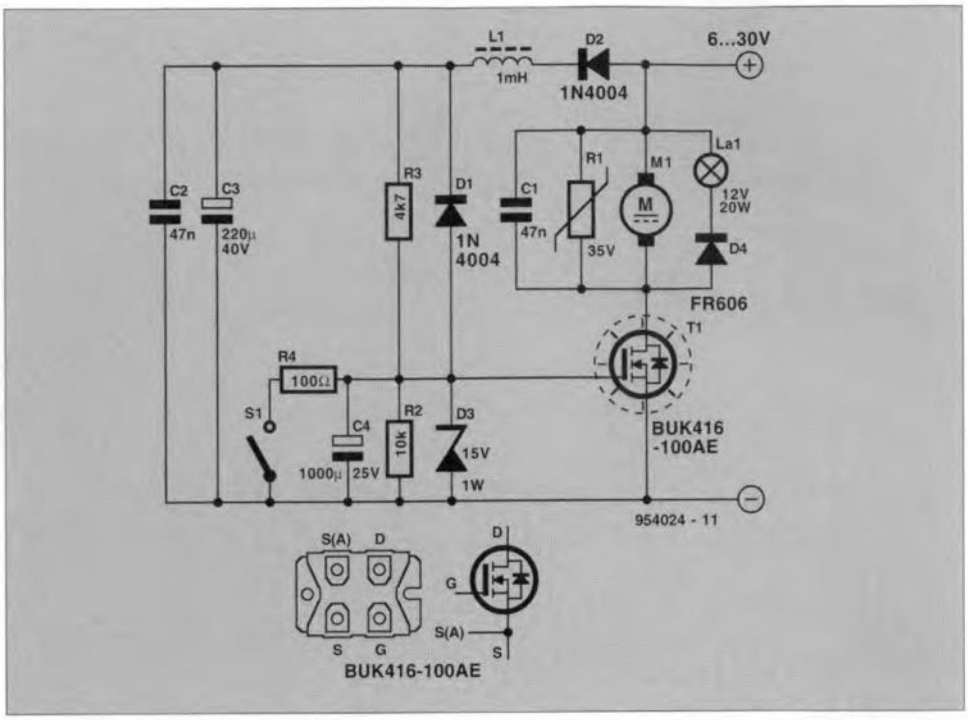
Bear in mind that, owing to the curved characteristic of R_1, the diodes do not give a linear display. Some experience, or calibrating, is necessary to know when the engine really gets too hot.

The current through the LEDs is ten times greater than that through R_2. At a reference voltage of 1.25 V, the current through the diodes will thus be 10 mA. In the dot-mode, the total current will remain below 20 mA, but in the bar-mode it may be up to 100 mA when all LEDs light.

Design by J. Bosman
[954019]

SOFT START SWITCH FOR DC MOTORS

The starting current of heavy-duty d.c. motors can be 3–8 times as high as the normal operating current, which puts a heavy load on the on/off switch. Also, when the motor is switched off, a breaking spark occurs which causes erosion and oxidization of the contacts. This obviously shortens the life of the switch appreciably. Moreover, the starting torque is much greater than the operating torque and this does neither the motor nor the load to which it may be coupled any good. The usual remedy for this deficiency is a series resistor in the motor leads which limits the starting current and is short-circuited shortly after the motor has started by a relay. It does not nullify the breaking spark, however.



A better solution is the circuit shown here, which uses a MOSFET as switch. Such a switch can be made to operate like a continuously variable resistor. In the present circuit, it allows the current to the motor to reach normal operating level within three seconds. The load on the on/off switch is a matter of a few milliamperes. The circuit can handle currents of up to 25 A, but, with suitable cooling, can be used with currents of 40–100 A.

Since the starting current is limited, the MOSFET should, of course, not be rated at the original starting current. This is a matter of cost, because the 100 A model used in the prototype, the BUK416-100A, is expensive. Therefore, for smaller loads, a much less expensive (and lower rated) MOSFET should be used.

In the quiescent state, S_1 is closed, so that the gate of the MOSFET is virtually at ground level (via R_4). When S_1 is opened, C_4 will be charged slowly via R_3 to about $^2/_3$ of the supply voltage. The MOSFET starts to conduct and is fully on when the gate potential has risen to 5 V. When the switch is closed again, C_4 is discharged rapidly via R_4.

When it is on, T_1 has an on-resistance, R_{DSon} of less than 13 mΩ, so that its dissipation is small. When the current through the motor is 20 A, the dissipation of T_1 is 4 W, so that a heat sink of 2.5 K W^{-1} will suffice. This rating must, of course, be increased when the motor current is larger. The BUK416-100AE need not be isolated from the heat sink. Its source and AS (auxiliary source) pins must be interlinked.

Such an expensive component needs to be protected, whence D_2 functions as polarity reversal protection. Diode D_4 (an FR606 or P600) short-circuits the current briefly when the motor is being switched off; during normal operation, it prevents inverse voltage peaks arising.

Zener diode D_3 makes it impossible for the gate potential of T_1 to be greater than 15 V.

Capacitor C_1 decouples the circuit for spurious r.f. signals.

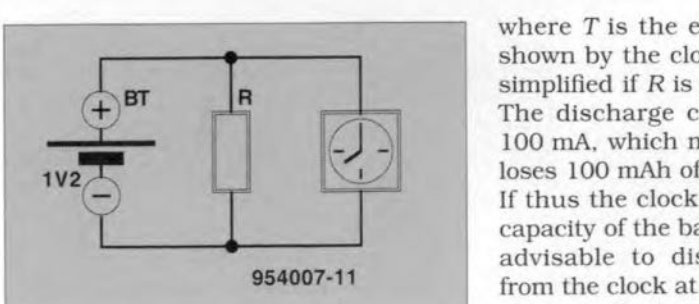
Voltage-dependent resistor R_1 eliminates any voltage peaks across the motor larger than 35 V.

Inductor L_1 and capacitors C_2 and C_3 ensure a smooth direct voltage when the motor is heavily loaded and brush sparks may create pulses on the supply lines.

Design by W. Zeiller
[954024]

NiCd-battery tester

A small quartz clock that is powered by a 1.5 V battery provides a simple means of determining the capacity of a NiCd battery. The circuit makes use of the property of such a clock to cease working when its supply voltage drops below about 1.05 V.



Set the clock to 00:00 and connect the fully charged NiCd battery to its battery terminals. Resistor *R* ensures a defined discharge current. Assuming that the nominal battery voltage, U_b, is 1.2 V, the capacity of the battery, *K*, (in mAh) is given by

$$K = T \cdot U_b / R,$$

where *T* is the elapsed time in hours shown by the clock. The calculation is simplified if *R* is given a value of 12 Ω. The discharge current is then about 100 mA, which means that the battery loses 100 mAh of its charge in an hour. If thus the clock stops after 5.5 h, the capacity of the battery is 550 mAh. It is advisable to disconnect the battery from the clock at that time to avoid discharging the battery beyond repair.

Design by J. Dietrich
[954007]

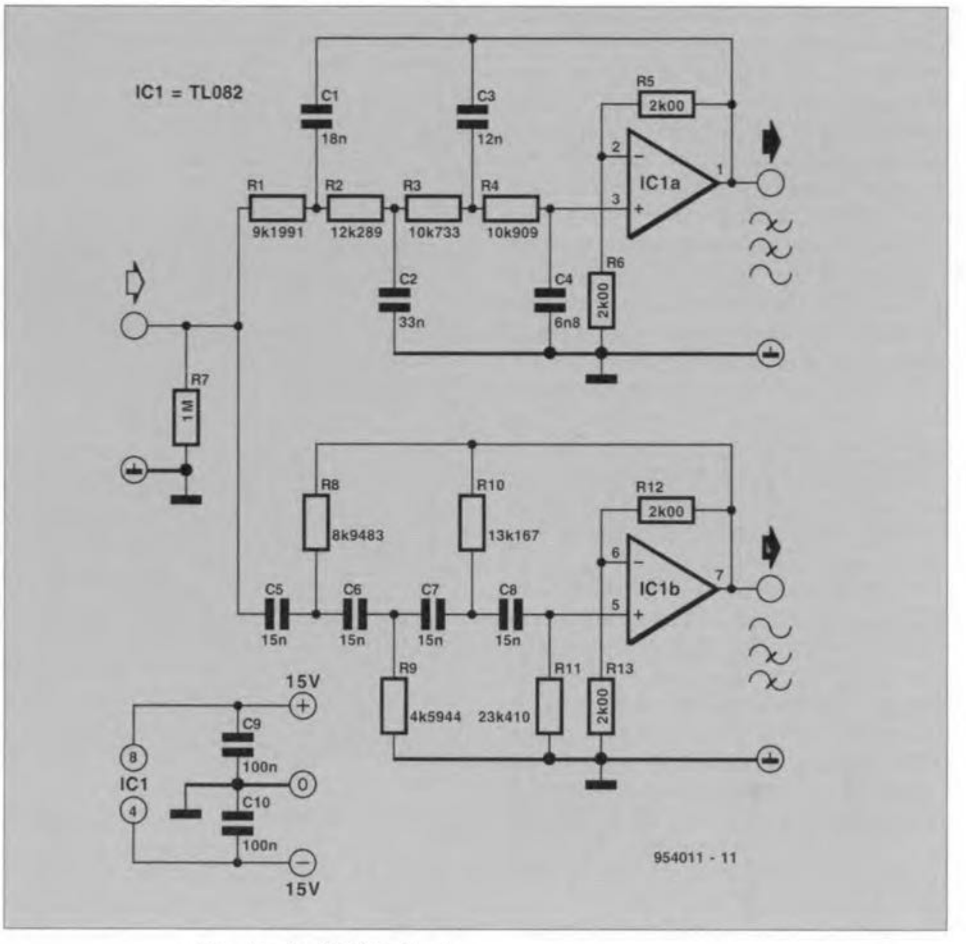
LINKWITZ-RILEY CROSS-OVER NETWORK

The Linkwitz-Riley crossover network is known for its homogenous performance and constant amplitude. The transfer is identical to that of two 2nd-order Butterworth filters in series. The Linkwitz-Riley filter is an example of an active crossover network for a two-way loudspeaker system whose crossover frequency is 1 kHz. It can be built around one quad op amp for a two-way or three-way speaker system and can be converted to provide alternative crossover frequencies by proportional adaptation of the resistors.

The design is based on E-12 series capacitors, which simplifies purchasing the components. Note that if different values are used, the entire filter may have to be recomputed, unless the values are ten times smaller or larger.

To keep the ratios C_1:C_2:C_3:C_4 and R_8:R_9:R_{10}:R_{11} as small as possible, the amplification of the buffer is ×2 (R_5/R_6 and R_{12}/R_{13}). The amplification at the crossover frequency is then 0.

When components with values as specified are used, the sum of the two filter sections will form a straight frequency characteristic. Exact resistance values can be obtained only by series/parallel combinations of resistors. If E-96 series resistors are used, that is not too difficult.

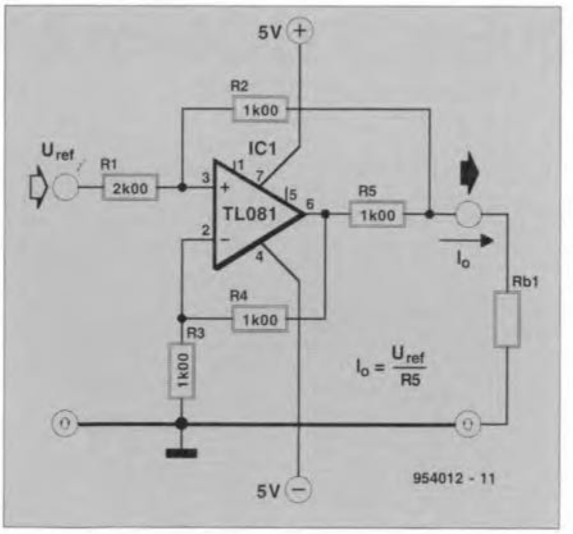


Design by T. Giesberts
[954011]

SINGLE OP AMP CURRENT SOURCE

Current souces are used in a variety of applications. They can be designed in a number of ways. One of the simplest is a common resistor, but this presupposes a constant load and a constant voltage supply. If those requirements are not met, the current will vary according to the circuit elements. A considerable improvement is achieved when an active element like a transistor is added to the resistor. Variations in the ambient temperature and in the threshold potential of the p-n junction then determine the accuracy of the source.

A better idea is to add an op amp to the circuit as shown in the diagram. This has the added advantage of being bipolar. In the diagram, resistors R_1, R_2 and R_5 render the non-inverting amplifier into a current source whose output is given by U_{ref}/R_5, provided that $R_2 = R_3 = R_4$ and $R_1 = R_2 + R_5$.

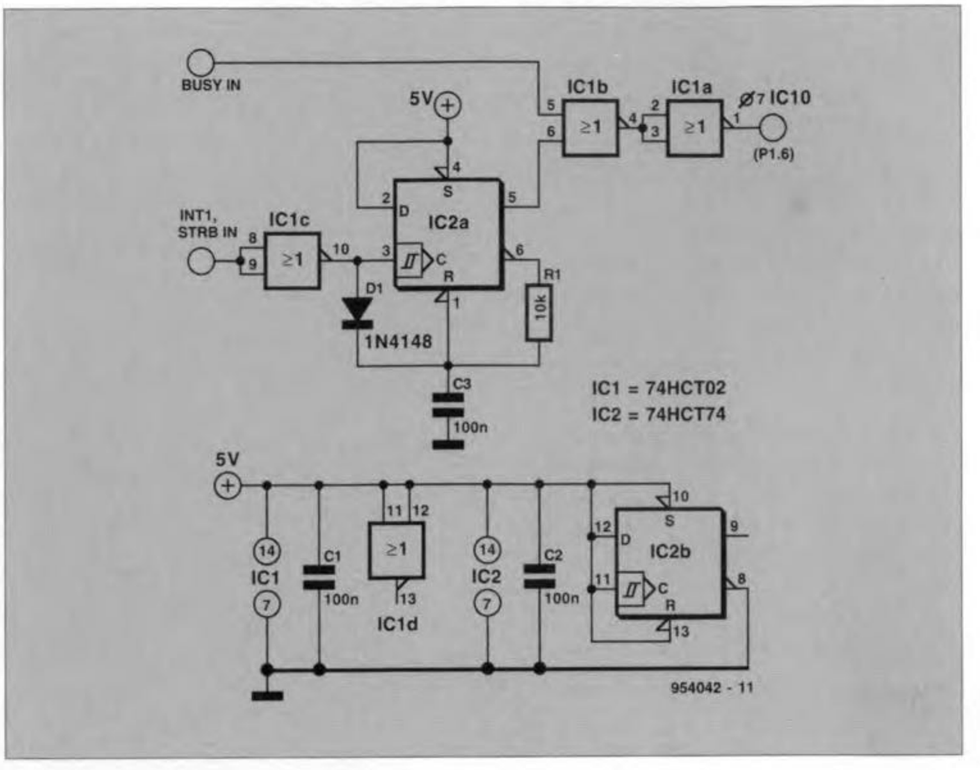


The operation of the source is fairly simple if it is assumed that the load is a short-circuit. In this case, a current of $U_{ref}/3 \times 10^3$ (3 kΩ) will flow through R_1 and R_2. The potential at the non-inverting input is then $U_{ref}/3$, which results in a voltage at pin 6 of $2U_{ref}/3$. The current through R_5 is $2U_{ref}/3R_5$. The current through the (short-circuited) load is $U_{ref}/3 \times 10^3 + 2U_{ref}/3R_5$. If R_5 = 1 kΩ, the current is U_{ref}/R_5.

Although the design is fairly well controlled, tolerances of the resistors may cause current variations. If these are unacceptable, a design with three op amps should be used (see p. 94).

Design by H. Bonekamp
[954012]

SPEED-UP FOR PRINTER BUFFER



Printer buffers serve to ensure that the data to be printed are taken from the computer as quickly as possible so that the computer can be used for other purposes. This is why a large buffer memory and fast communication between printer and computer are essential: the buffer prevents the computer having to wait for the much slower printer.

At the instant the first data are written into the buffer, the printer starts to print. This means that the buffer has to control two matters: the data traffic *from* the computer and that *to* the printer. If the printer is made to wait during the periods when data are being received, the buffer needs to control only one process at a time, which saves time.

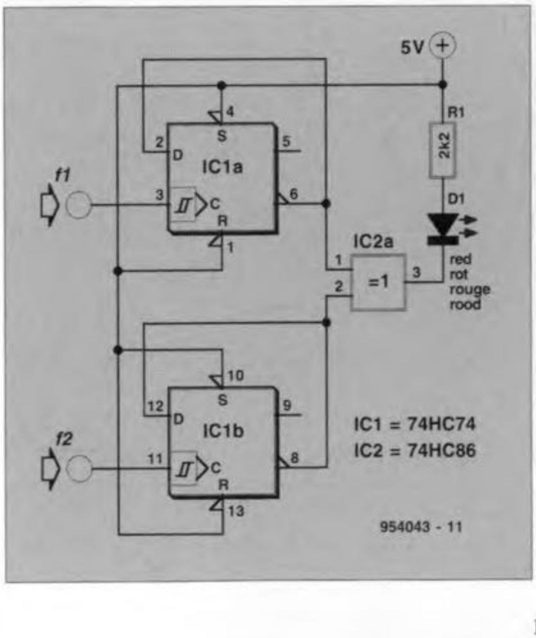
To arrange this, the strobe signal must be monitored by the computer. If there are strobe signals, the busy line from the printer must be actuated so that the buffer does not send data to the printer. When after about 1 ms no strobe signals have been received, the busy line is made low so that the printer is reenabled. This arrangement does not work so well when the buffer is nearly full, however. In that case, the present circuit may actually delay matters because the printer is reenabled only after a delay of about 1 ms. This is a good reason for making the RAM in the buffer as large as possible.

The strobe signal (STRB IN or INT1) ensures via NOR gate IC_1 that the bistable is set and C_3 is charged. When the bistable is set, the $\overline{Q}$ output is low and C_1 is discharged via R_1. If the strobe is not active for some time, C_1 will be discharged completely, whereupon after 1 ms the bistable is reset.

The output signal is combined with the busy signal, which means that in the existing buffer the link with pin 7 of IC_{10} (BUSY IN) must be broken. Connect IC_{1b} and IC_{1a} across this break so that the modified busy signal is applied to the microprocessor at the correct location. Connect pins 8 and 9 of IC_1 to the incoming strobe, STRB IN (external buffer) or INT (terminating card). In other words, the circuit may be used with an internal or an external printer buffer.

Design by A. Rietjens
[954042]

FREQUENCY COMPARATOR



Even with a good frequency meter, it is well-nigh impossible to set two frequencies to a fraction of a hertz. It can, however, be done with the aid of the circuit shown. It consists of two bistables (flip flops), an XOR gate which functions as a comparator, and an LED indicator.

The bistables derive from the two inputs signals with a duty factor of 0.5 and at half the input frequency. The output of the XOR gate is low when the input frequencies are equal. When they are not, this is indicated by the flashing of the LED.

When the frequencies are equal, the light intensity of the LED is a measure of the phase shift between them. The brighter the LED lights, the smaller the phase difference.

The maximum input frequency is determined by the peak operating rate of the bistables, and is here about 30 MHz.

The circuit draws a current of around 1.5 mA, which is mainly on account of the series resistor of the LED. The current is kept low by the use of a high-efficiency LED.

Design by H. Bonekamp
[954043]

CHARGER FOR ALKALINE/MANGANESE BATTERIES

Design by K. Walraven

In spite of several reports published in the past year or so, there is no possibility of recharging flat dry batteries of whatever type. This allegation has been finally put to rest after tests by various international technical and consumer organizations. What may have confused a number of people (no doubt driven by the scandalously high prices of dry batteries) is the emergence on the European market of secondary alkaline/manganese batteries which had, however, been known in the USA and Canada since the late 1980s* (see our May, 1995, issue). The charger presented here is intended for these secondary (i.e., rechargeable) batteries.

The construction of secondary alkaline/manganese batteries is different from that of the well-known primary (non-rechargeable) alkaline battery First tests in our design department showed that the secondary alkaline /manganese batteries could be properly charged like NiCd batteries. However, as far as their electrical properties are concerned, they have more in common with primary batteries than, e.g., with NiCd batteries. Their nominal e.m.f. is 1.5 V and the capacity of an HP11-sized type is 1000 mAh. However, their internal resistance is appreciably higher than that of NiCd batteries, so that they are not able to provide large currents. In fact, tests show that if a large current is drawn from them, their capacity drops within a few charge/discharge cycles (see our May, 1995, issue).

Charging these alkaline.manganese batteries is also quite different from that of NiCd batteries. According to the manufacturer's data sheet, charging should not be done at constant current, but at constant voltage (i.e., similar to lead-acid batteries). Furthermore, to prevent gas formation in the batteries, it is important not to use a constant, but a pulsating, current.

A further difference with other secondary batteries is that the alkalnine/manganese batteries must not be discharged too much, otherwise the checmical process can no longer be reversed, which means they can not be recharged properly. It is, therefore, recommended by the manufacturers that they are not discharged below an e.m.f. of 1.0 V (the absolute 'safe' lower limit is 0.9 V).

The charger

It is clear from the foregoing that secondary alkaline/manganese batteries

* See Crompton's authoritative *Battery Reference Book*, ISBN 0 408 00790 7 Butterworth Scientific Ltd, 1990

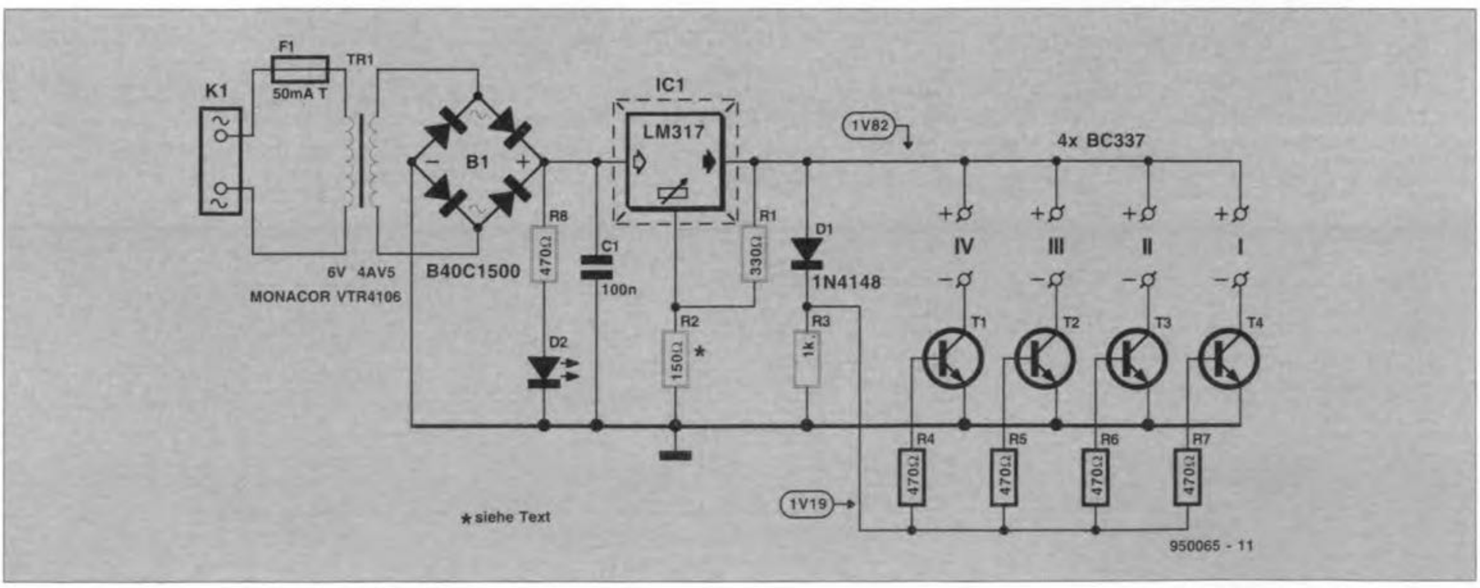


Fig. 1. Circuit diagram of the charger for alkaline/manganese batteries.

must be charged by a dedicated charger. Unlike NiCd batteries, they have no memory effect, so periodic completely discharging is not necessary; in fact, it may well be detrimental, but we have no practical experience of this as yet.

As already stated, these batteries must be charged at constant voltage with a pulsating current. This means that limiting the charging period, as with NiCd batteries, is not necessary. The upper limit of the charging voltage is 1.85 V. The maximum charging current for HP11 batteries is about 120 mA, which is very close to the well-known value of $^1/_{10th}$ of the capacity.

A first requirement of the charger is thus a constant-voltage source which limits the maximum charging potential. Notwithstanding what has been said, it appears useful to provide some sort of charging-current limiting. If the charging current is, moreover, made pulsating, the basic requirements are met.

A practical design, intended for charging up to four HP11 batteries in parallel, is shown in **Fig. 2**.

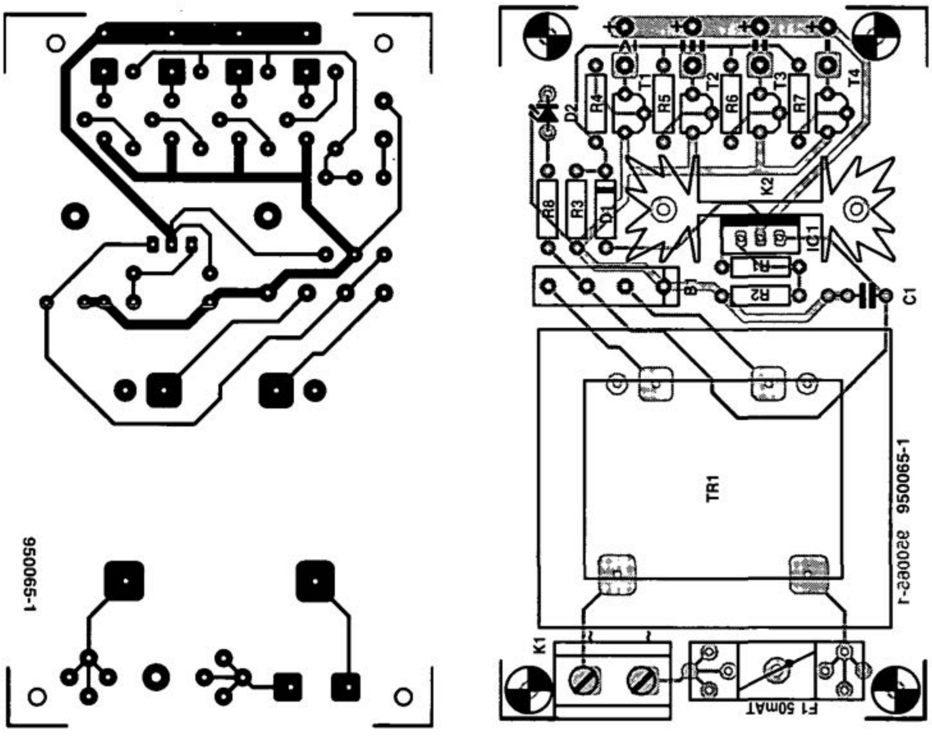


Fig. 2. Printed-circuit board for the charger (see p. 70).

The charging voltage is clamped at 1.85 V by integrated voltage regulator IC_1. The LM317 used is about the only readily available regulator that can provide such a low level (since its reference level is even lower at 1.25 V). The output voltage is determined by the ratio R_1:R_2.

The input to IC_1 is provided by a traditional mains transformer and bridge rectifier combination. A smoothing capacitor is not fitted so that the output voltage of IC_1 is pulsating.

Each of the (up to) four batteries to be charged is in series with a transistor, T_1–T_4, which functions as a current source, and also provide a degree of current limiting. However, in practice, this limit appears to be about several hundred mA, so it should, perhaps be seen as a protection rather than as a limiting. The transistors are Type BC337, because these can provide rather more current than the usual BC547, although their gain is about the same.

Thus, when the batteries are fairly flat, the initial charging current is limited to a few hundred mA. When the battery voltage rises, however, the current drops quite quickly to an appreciably lower level. When, after about 16–18 hours, the battery voltage has risen to 1.85 V, the current is almost negligible.

In this design, therefore, the charging current is not highly defined, because at all times it is determined by the charging voltage. However, for a large part of the charging period, the current drawn from the prototype charger was about 100 mA, which is an excellent level for these batteries.

Since there is no smoothing capacitor, the output voltage of IC_1, and thus the base potential of T_1–T_4, is pulsating. This action is magnified because the charging current has to overcome two thresholds: the diode junction of D_1 and the base-emitter junction o the transistors. This results in the transistors switching on quite abruptly, remaining on for about $^2/_3$ of the time and being off for around $^1/_3$ of the time. This pulse behaviour accords with the manufacturers' recommendations.

It is hoped that this makes it clear that the transformer and rectifier in no circumstances can be replaced by a mains adaptor. Such a device does not provide a pulsating current, but a smoothed direct current. Also, the transformer must not provide a voltage higher than 6 V, because that would make the intervals between the charging-current pulses too short.

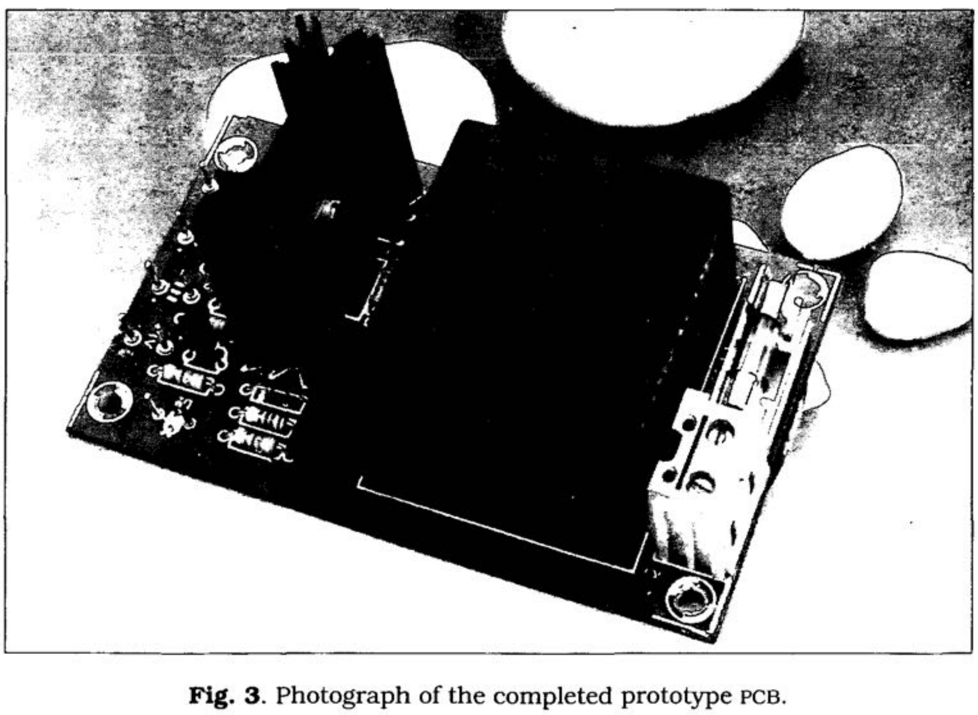


Fig. 3. Photograph of the completed prototype PCB.

Construction

The charger is intended to be constructed on the printed-circuit board shown in **Fig. 3**. As will be seen, this should not present any difficulties. It should, however, be noted that IC_1 should be provided with a heat sink. Since the metal base of the LM317 is internally connected with its input, the base is at the same potential. Since the heat sink is not connected to any other parts, insulating washers are not needed.

The finished board should be fitted in a plastic case, since several parts carry the mains voltage. Therefore, no parts of the circuit must be accessible from the outside during operation. Moreover, the mains input cable should be provided with a strain relief.

The battery holders should be mounted on top of the case. They are linked to the board with flexible, insulated circuit wire. It is advisable not to use a four-fold battery holder, because these normally have series connections which can not modified easily.

Finally, the indicator, D_2, must, of course, be clearly visible on the front panel.

The finished prototype board is shown in **Fig. 4**, while **Fig. 5** gives a view of the opened prototype charger.

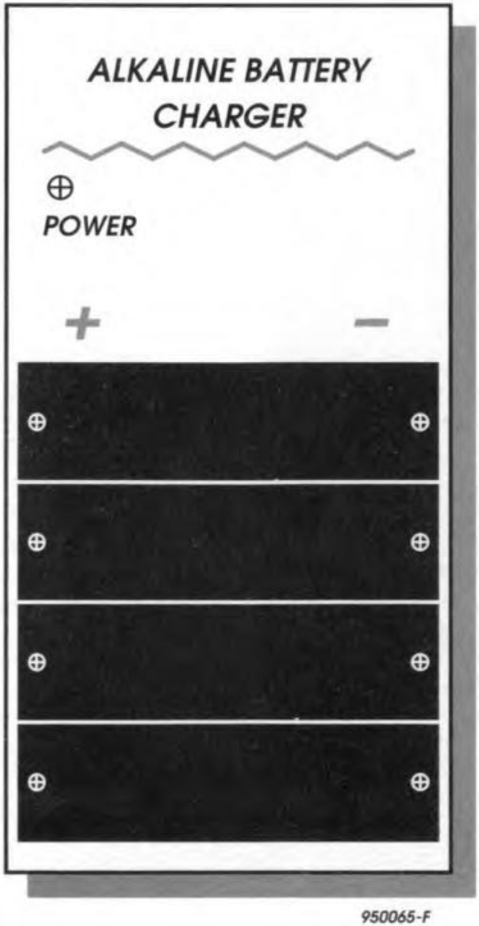


Fig. 4. Suggested front panel layout.

Setting up

Before the charger can be taken into use, the charging voltage must be set, for which a digital voltmeter is required.

Temporarily connect a 100 µF, 16 V, electrolytic capacitor in parallel with C_1. Measure the output voltage of IC_1, which should be 1.8–1.85 V.

If the voltage is too high, lower it by giving R_2 a lower value. This resistor need not be removed: simply connect another resistor, probably 470–1000 Ω, in parallel with it. If the measured voltage is too lowm a resistor should be connected in parallel with R_1.

Remove the 100 µF capacitor from across C_1.

The charger is now ready for use. When charging batteries, make sure that there is good contact between them and the holders, because bad contacts give rise to high resistances.

A charging time of 16–18 hours is recommended (in line with the advice of he manufacturers).

Parts list

Resistors:
R_1 = 330 Ω
R_2 = 150 Ω
R_3 = 1 kΩ
R_4–R_8 = 470 Ω

Capacitors:
C_1 = 100 nF

Semiconductors:
D_1 = 1N4148
D_2 = LED
T_1–T_4 = BC337

Integrated circuits:
IC_1 = LM317

Miscellaneous:
B_1 = rectifier Type B40C1500
F_1 = fuse, 50 mA, slow
1 off fuse holder
K_1 = terminal block, 2-way, pitch 7.5 mm
Tr_1 = mains transformer, secondary 6 V, 4.5 VA (e.g., Velleman 106005OM, available from Maplin)
Heat sink SK104, 37.5 mm (available from Dau components)
Case, 120×65×65 mm, e.g. OKW H120

PCB Order no. 950065 (see p. 70)
Front panel foil not available.

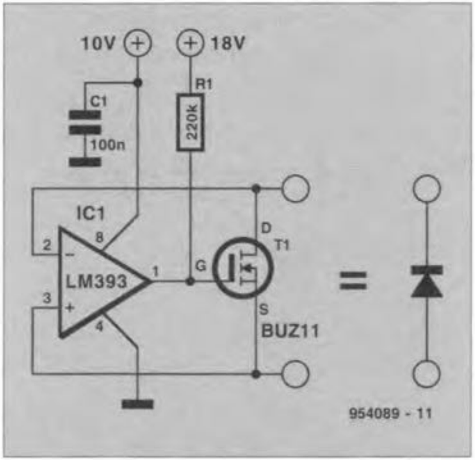
[950065]

DIODE SIMULATOR

In certain applications it may be desirable, in order to limit unnecessary power losses, to have the use of a diode with a forward voltage lower than 0.7 V. The present circuit fulfils that desire. At a forward current of 1 A, the total drop across the two (active) components is only 0.04 V – an appreciable reduction compared with a standard diode.

Normally, the current in an n-channel MOSFET flows from drain to source. In the present circuit, however, it flows in the opposite direction, which conforms to the polarity of the internal protection diode between drain and source. This prevents the whole circuit conducting in the reverse-voltage direction.

When the source is positive with respect to the drain, the comparator switches the MOSFET on. At a current of up to 1 A in the forward direction of



the FET, the voltage drop is then only 0.04 V.

When the polarity of the voltage is reversed, so that the drain becomes positive with respect to the source, the output of the comparator changes state (from high to low), resulting in the current through the FET being interrupted. In other words, the 'pseudo diode' conducts in only one direction.

The drain and source potentials of the BUZ11 must remain at least 1.5 V below the supply voltage to ensure that they fall within the input range of the LM393.

Resistor R_1 functions as pull-up resistance for the open-collector output of the comparator.

Design by A. Rietjens

[954089]

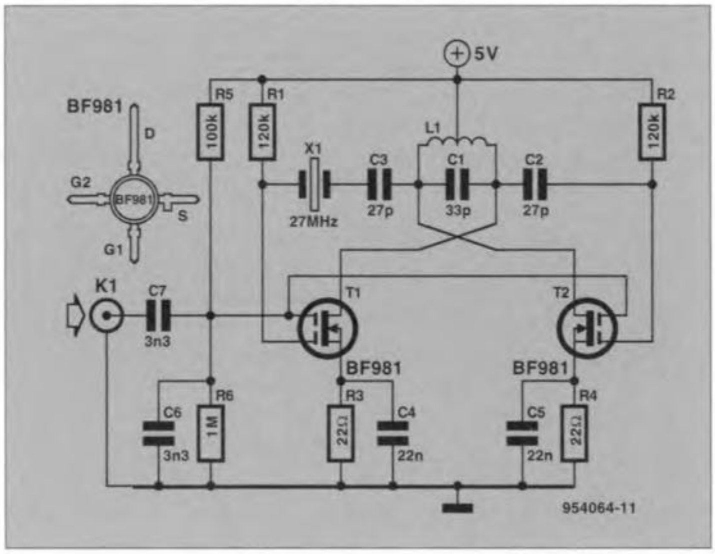
MODULATED PUSH-PULL CRYSTAL OSCILLATOR

Push-pull oscillators have an important advantage over conventional oscillators: they provide appreciably more power at the same supply voltage.

The present oscillator is based on a pair of gallium-arsenide, GAs, field-effect transistors, FETs. Controlled by a quartz crystal, it operates with capacitive feedback, which ensures high output power and high frequency stability.

The circuit may be used as a power oscillator or as a small transmitter in remote control applications. An a.f. signal may be applied to K_1 to provide amplitude modulation of the oscillator output.

The highly symmetric design allows further push-pull circuits to be controlled. Amplifiers, or an antenna (when the circuit is used as a small transmitter), can be coupled inductively to L_1 by 2–3 turns of enamelled copper wire.



At a supply voltage of 4.5 V, the circuit draws a current of 150 mA to provide a relatively high power output. When a higher supply voltage is used, the values of R_3 and R_4 must be increased accordingly to prevent the FETs being overloaded.

Design by L. Lemmens
[954064]

Kunz

FRONT/REAR CAR RADIO FADER

Although most car radio/cassette players produced in the past five years or so are fitted with a front/rear speaker volume control, there are still many about that have no such provision. A simple way of adding this facility is shown in diagram *A*, which is the principle on which many car radio faders operate. This arrangement gives a stepped front/rear speaker volume control with step ratios between 4:1 and 1:4. (Many modern car radios have a continuously variable control, however. Editor).

The arrangement has a few drawbacks in that power is dissipated (unnecessarily) in the resistors, which lowers the overall efficiency (down to 83% in positions 1 and 5, but as low as 67% in position 3), and the load impedance is lowered to 3 Ω in position 3 and to 2.67 Ω in positions 1 and 5.

A better arrangement is shown in diagram *B*. Since this is a series circuit, the load impedance can not drop to the dangerously low values met in *A*: it varies from 4 Ω in positions 1 and 5 to 8 Ω in position 3. Since the dissipation in the resistors is smaller, the efficiency is higher than in *A*: only in positions 2 and 4 does this drop to 88% – in the other positions it is 100%.

The volume control proper is similar to that in *A*, but in the extreme positions one speaker is on full, while the other is short-circuited.

Make sure that the switch can handle the power of the car radio.

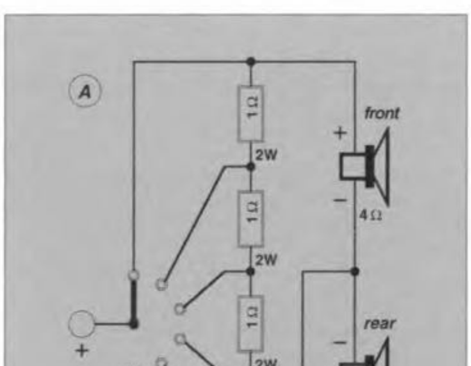
A:

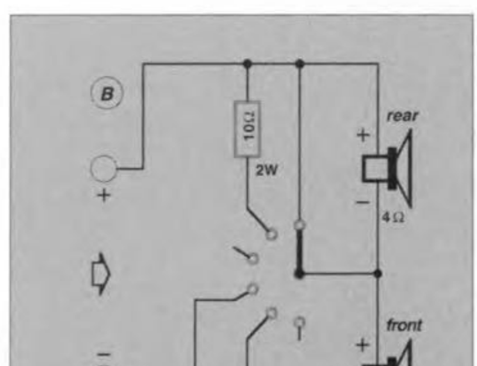
position	front/rear ratio
1	4:1
2	2:1
3	1:1
4	1:2
5	1:4

B:

position	front/rear ratio
1	∞
2	2:1
3	1:1
4	1:2
5	∞

Design by J. Seyler
[954085]





ZERO CROSSING DETECTOR

In many circuits that operate in synchrony with the mains, synchronization normally takes place during the zero crossing of the mains voltage. An auxiliary circuit as shown in the diagram is then required to detect the zero crossing. This pulls the output briefly to earth at the instant the mains voltage becomes positive after passing through the zero crossing.

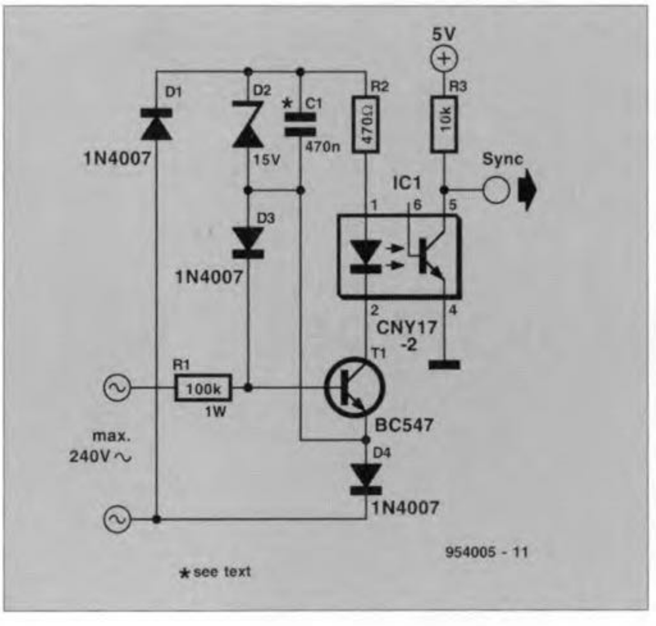
The present auxiliary circuit operates with the full mains voltage rather than with a transformed-down potential. This has the advantage that the zero crossing is detected much more accurately. Also, of course, there is no transformer required. This arrangement is particularly useful in computers in which there is no lowered mains voltage available.

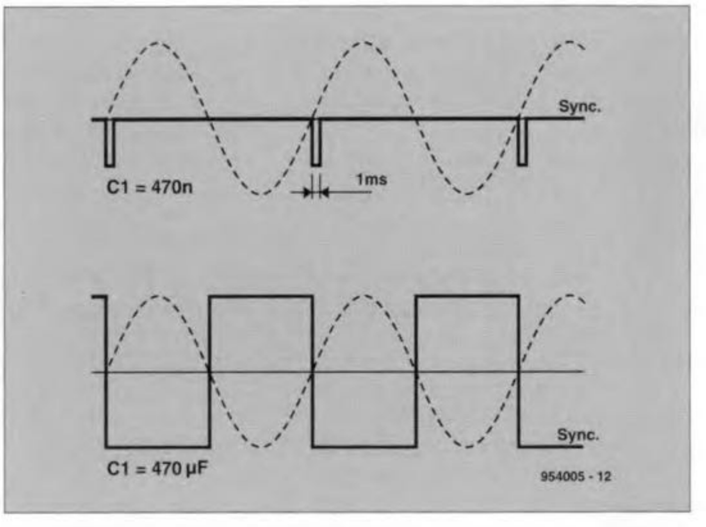
During the negative halves of the mains voltage, C_1 is charged via D_1, D_2, D_3 and R_1 to a potential of 15 V. This charge will leak away via the LED in the optoisolator when T_1 begins to conduct. This happens when the potential on the L(ive) line is about 1.4 V higher than that on the N(eutral) line (so that base current flows via R_1).

Since the collector current is fairly high, C_1 will be discharged quite rapidly. The diode will thus light only briefly, which means that T_1 conducts only for a short time. With the specified value of C_1, the sync output will be low for about 1 ms. This time can be varied by changing the value of C_1: when it is 470 µF, the pulse will last for about 10 ms.

Since the circuit is connected directly to the mains, it must be treated with respect and great care. Note that the optoisolator must be able to handle the mains voltage. The specified type is.

Design by J. Schuurmans
[954005]





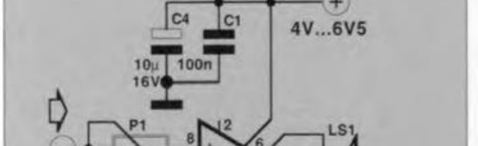
AUDIO POWER DOWN

The audio amplifier in a battery-operated scanner or walkie-talkie is often on standby for long periods. Of course, in that state it draws a much smaller current than during normal operation, but even so, this is a waste of energy.

A system in which the amplifier is switched by the squelch circuit is already a lot better as far as energy drain is concerned. Unfortunately, the switching clicks are all too audible.

However, there are special amplifier chips that have an integral power-down circuit, which comes into operation when there has been no signal at the input for more than 0.5 s. The current then drops from about 15 mA to 100 µA. Examples of such ICs are those in the TLC247x series from Texas Instruments (see diagram).

The ICs are invariably intended for speech, rather than music, applications. This is because an internal lowpass filter limits the frequency range to 5 kHz in the TLC24701/24721 or to 3.5 kHz in the TLC24711/24731.



The devices are also suitable for direct processing of pulse-width-modulated (PWM) signals. Such a signal is applied to pins 7 and 8. A potentiometer that varies the potential at pin 2 between 0 and $^2/_3V_{dd}$ functions as volume control. Note that the diagram shows a circuit that works with an analogue signal, whose volume is adjusted with P_1.

The circuit shown draws a current of not more than 250 mA. With a supply voltage of 5 V, the output stage can provide a peak power of 0.5 W.

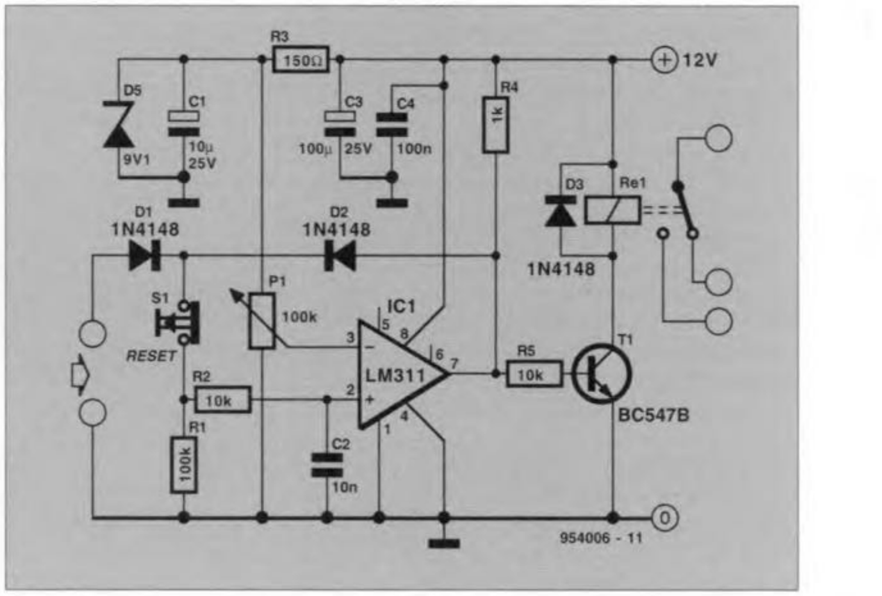
Design by L. Lemmens
[954003]

SWITCHED VOLTAGE COMPARATOR

A voltage comparator is a device that monitors whether an input potential is below or above a certain level. Depending on the manner in which the comparator is switched, its output is high or low respectively.

Since a comparator is able to liken two potentials, many sensor circuits use one, for example to show when the temperature is too high. The output of the comparator can then be used to switch on a cooling fan, for instance. When the temperature has dropped to below its upper limit, the circuit detects this and switches off the fan.

Apart from its application in sensor circuits, a comparator may also be used as an event detector. Often it is not just important to know whether a certain limit has been exceeded, but also when this happened. For such an application, the comparator must be cascaded with a bistable (flip-flop). This is set at the moment the comparator detects a limit being exceeded. This arrangement can also be made by adding a resistor and a diode to the comparator circuit to simulate the bistable operation.



The complete circuit is shown in the diagram. The input signal is applied to the input of the comparator via D_1. Depending on the setting of P_1, the potential at the non-inverting (+) input is then higher or lower than that at the inverting (–) input. If it is higher, the output will be nearly equal to the supply voltage. This potential is applied to the input via D_2, so that the comparator retains its state irrespective of the level of the input signal. This situation alters when the switch is pressed: the comparator then changes state (returns to its quiescent state).

As an example, a transistor is added to the circuit to operate a relay, but this addition is not necessary. The level at pin 7 of IC_1 can also serve to drive other circuits. Resistor R_4 is always required, however, because IC_1 has an open-collector output.

The circuit draws a current of only a few milliamperes when the relay is not energized.

Design by M.S. Nagaraj
[954006]

ZN436 REPLACEMENT

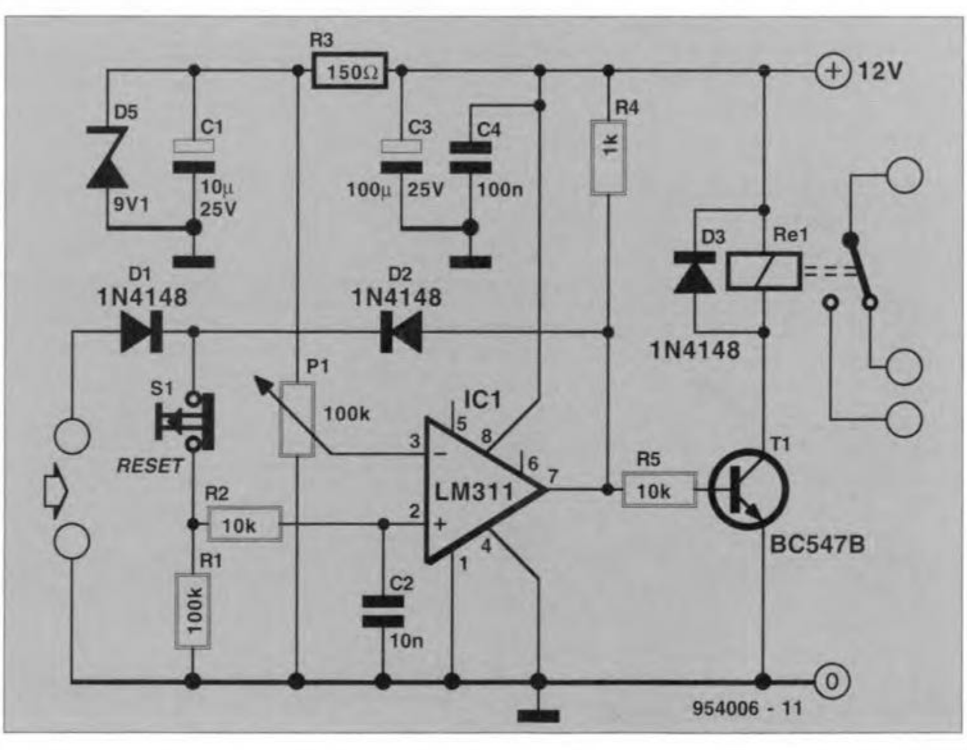
Tekening dubbel
Lees in 9509-75.

Since Ferranti have taken their well-known ZN436, a 6-bit digital-to-analogue converter, out of production, the ZN426 has 'replaced' it. This is, however, not a direct replacement, because, to start with, it is an 8-bit device. It is also not fully pin compatible as the diagram shows.

Note that the ZN426 has a 2.5 V internal reference source. Therefore, if the ZN436 to be replaced uses an external reference source, this can be omitted; pins 5 and 6 of the replacement ZN426 should then be strapped together. To convert the ZN426 to a 6-bit device, D_0 and D_1 are grounded.

If the device is mounted on a board and an external reference is used, pins 9 and 10 must be connected to earth. If the internal 2.5 V source is used, pin 5 must, of course, be disconnected from the external reference.

Design by A. Rietjens
[954010]



CORRECTIONS TO JULY/AUGUST ISSUE

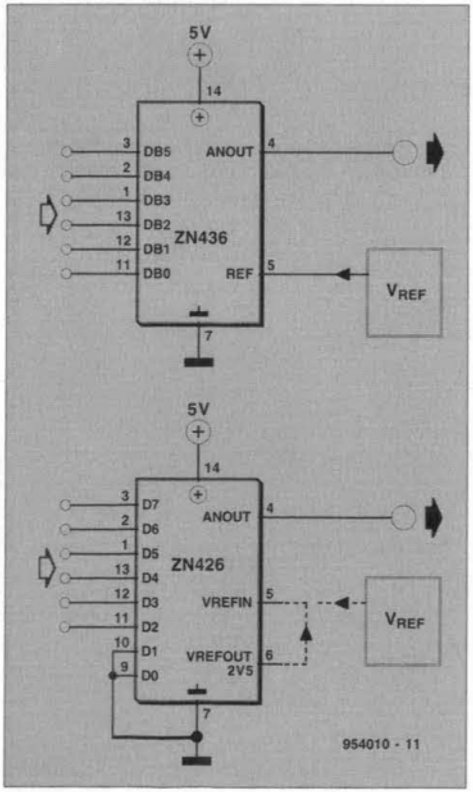
Owing to some last-minute alterations, a few small, but nevertheless annoying errors occurred in our July /August issue. The necessary corrections are detailed here.

The '**AF Power Indicator**' advised in the contents to be on p. 102 was omitted and is given here.

The page number of '**Speed-up for printer buffer**' was shown in the contents as 104, whereas this should have been 102.

The page number of '**Simple r.f. signal generator**' was given as 30: this should have been 10.

The circuit diagram included with '**ZN436 replacement**' on p. 109 is incorrect; the correct one is reproduced in column 1.



ZN436 replacement

AF POWER INDICATOR

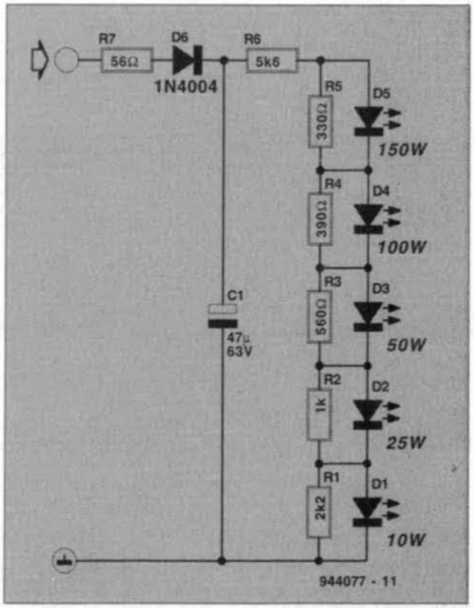
The indicator is intended primarily for heavy-duty PA (Public Address) loudspeakers. Connected simply to the loudspeaker input terminals, it displays the power applied to the speaker by a row of five LEDs: the larger the power, the more diodes will light. Since the power provided by the output amplifier also influences the total current through the diode array, the brightness with which the LEDs light also increases with increasing power. In this way, only five LEDs give a fairly good indication of the applied power.

The scale is geared to use with PA systems, where a few watts here and there do not matter. The stated values relate to peak voltages across 8 Ω. If the loudspeaker impedance is 4 Ω, the values are double the stated ones.

Since a clear display requires that the lighting times of the LEDs are not too short, a single-phase rectifier, R_7-D_6-C_1, is provided at the input. Because the capacitor is discharged each half period via R_6 and the remainder of the circuit, the afterglow lasts for some tenths of a second. Resistor R_7 limits the charging-current peaks to C_1 to about 1 A. The LEDs used are of the low-current type and light clearly at currents as low as 1 mA. Resistor R_6 ensures that at full drive the maximum permissible current through the LEDs is not exceeded (7 mA at an a.f. power of 150 W into 8 Ω).

If slightly longer afterglow times are required, the value of C_1 may be increased to some extent. However, this results in a power somewhere between peak and continuous values being measured rather than peak values. This is determined by the time constant of network R_7-C_1. Some experimenting with this may prove useful.

Design by T. Giesberts
[954077]



GAME PORT AS PC INPUT

Parallel inputting of data into a PC is not straightforward, unless an 8-bit slot is available into which an easily available card with two game ports can be inserted.

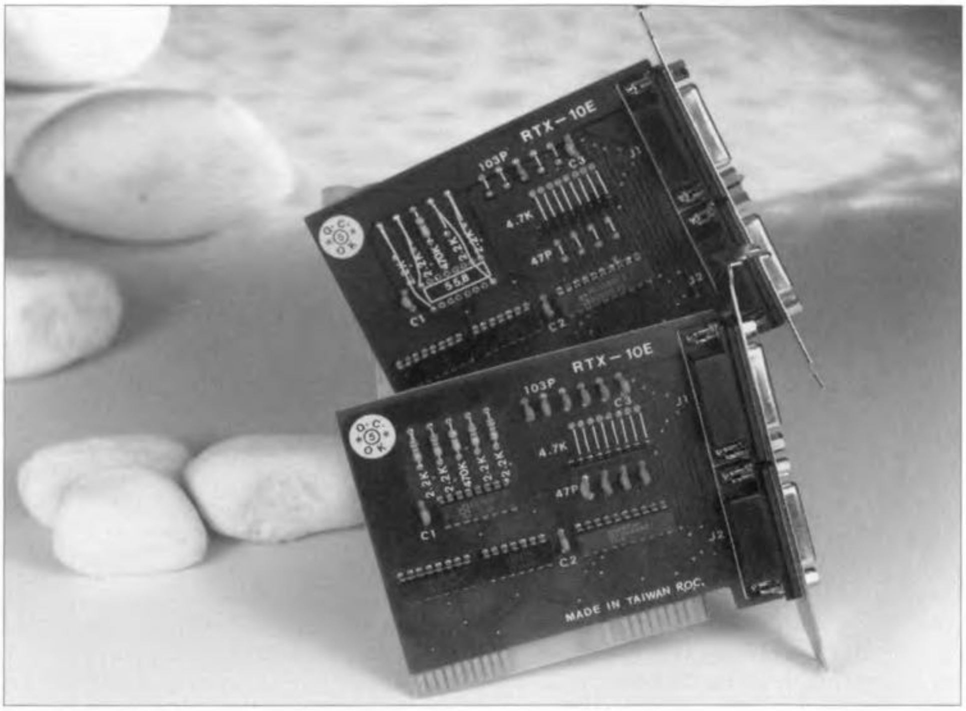
The card consists of an address decoder (2× 74LS138 and 74LS32 for decoding game port address 201_{HEX}) and a four-fold timer, NE558C, which converts the resistance values of the single joystick parameters into pulses of corresponding widths. Many game port cards only use an LS138 for the address decoding, so that addresses $200–207_{HEX}$ can be accessed. A 74LS244 functions as bus buffer and connects the outputs of the timers and the joystick to the PC bus.

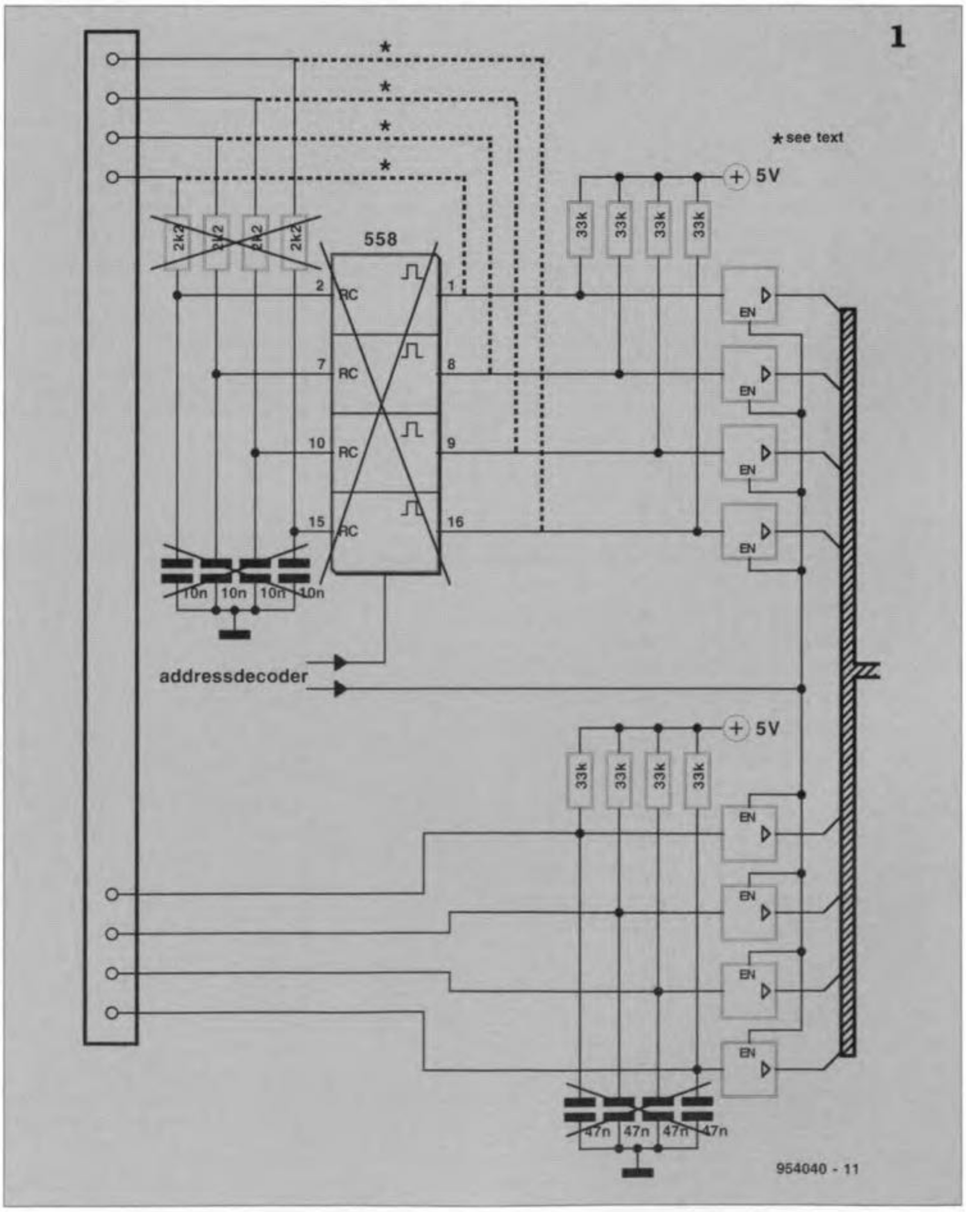
The four RC networks, four additional capacitors and the timers must be removed from the card as shown in Fig. 1. The timers are replaced by wire bridges between input and output. This gives allocations of the joystick terminals (A - top; B - bottom) to the set addresses as shown in the table. Since terminals A and B are connected in parallel, the corresponding allocations must not be occupied simultaneously to avoid short circuits.

The modification described causes the card to operate at address 201_{HEX}. If another joystick card is already used, it should be removed or deactuated, otherwise the game ports will be destroyed by the consequent bus conflict. A second joystick card can be used only after the standard address has been altered. To that end, the address decoder must be adapted as shown in Fig. 2. Break the link between output Q_0 of the LSB decoder and the buffer IC and make a new connection between Q_1 and the buffer IC. The card can then be accessed at address 203_{HEX} (or $208–20F_{HEX}$ if the card contains only a single address decoder IC).

```
10 DIM BIN(8)
20 FOR K=1 TO 8
30   BYTE = INP(&H201)
40   FOR J=0 TO 7
50     BIN(7-J) = 1
60     IF (BYTE AND (2^J)) = 0 THEN
BIN(7-J) = 0
70   NEXT J
80   FOR L=0 TO 7
90     PRINT USING "#";BIN(L);
100  NEXT L
110  PRINT " "
120 NEXT K
130 PRINT
140 GOTO 20
```

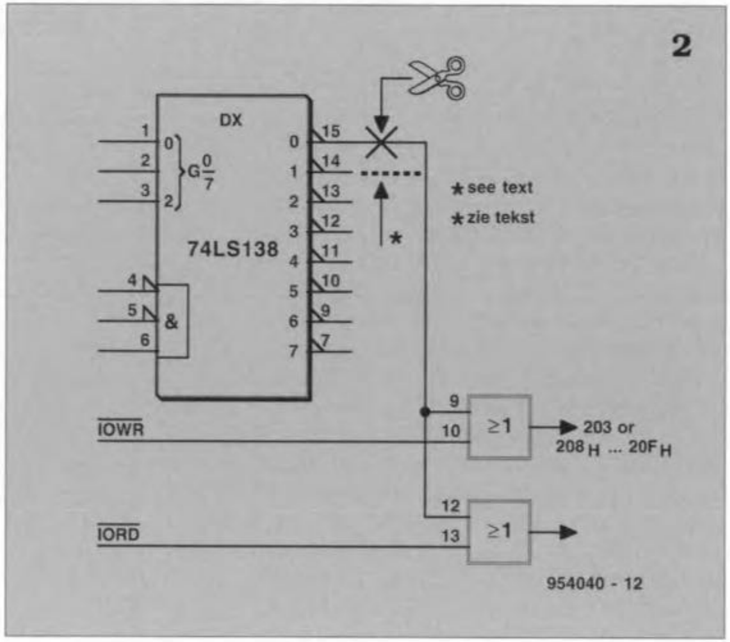
Sample program in BASIC





```c
#include <stdio.h>
#include <dos.h>
#include <limits.h>

#define ADDRESS 0x201

void bit_print(int);

void main(void)
{
  int value, i;

  while (!kbhit()) {
   i=0;
   do {
   value = inp (ADDRESS);    /* Read value at port */
   bit_print(value);       /* Print value bitwise */
   printf(" ");
   } while ((i++) !=7);
   printf("\n");
  }
  getch();
}

void bit_print(int a)
{
  int i;
  int n=CHAR_BIT;
  int mask = 1<<(n-1);

  for (i=1; i<= n; ++i) {
   putchar(((a & mask) == 0) ? '0' : '1');
   a <<=1;
  }
}
```

Sample program in C



Since all inputs of the card are TTL compatible, it may be used with most PCs. The output +5 V and ground may be used for experimental circuits.

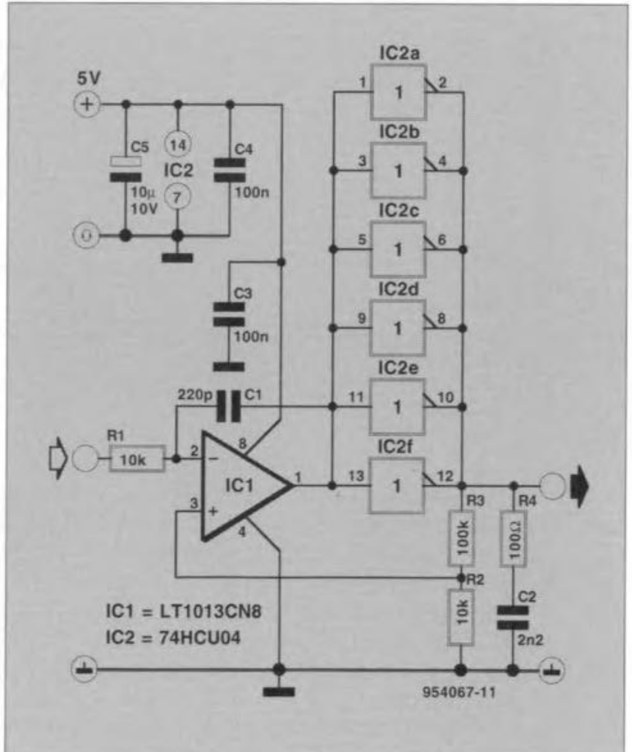
Complicated software is not required, since writing can be done in BASIC or C (INP(&H201) or Pascal ($201). The sample programs in BASIC and C represent the written values on the monitor.

Design by N. Koerber
[954040]

SIMPLE VOLTAGE AMPLIFIER

The resistive nature of the complementary outputs of a CMOS logic inverter may be used to create a voltage gain stage which allows an output swing of almost the supply voltage. In the circuit shown, CMOS gates are used to extend an amplifier's output swing to almost 5 V. The circuit is particularly useful in 5-V powered analogue systems where improvements in available output swing are needed to maximize the signal processing range and drive margin.

The paralleled inverters are placed within the LT1013's feedback loop, where they effectively drop the output resistance, aiding considerably in the swing capability. The inversion in the loop requires the feedback connection to go to the amplifier's positive input. Network R_4-C_2 serves as a lead/lag compensation. It eliminates oscillation in the inverter stage, which has a high gain-bandwidth when



operating in its linear region. Capacitor C_1 presents capacitive feedback to create loop compensation.

The output voltage swing of the circuit is 0 to 4.92 V with a load of 5 kΩ. A disadvantage of this method is the relatively high current drain of the inverters when the output voltage approaches the half-supply level. The application is intended mainly for d.c. circuits, because the distortion at full drive rises to unacceptable values above 100 Hz or so. Current drain of the circuit is ≤ 50 mA.

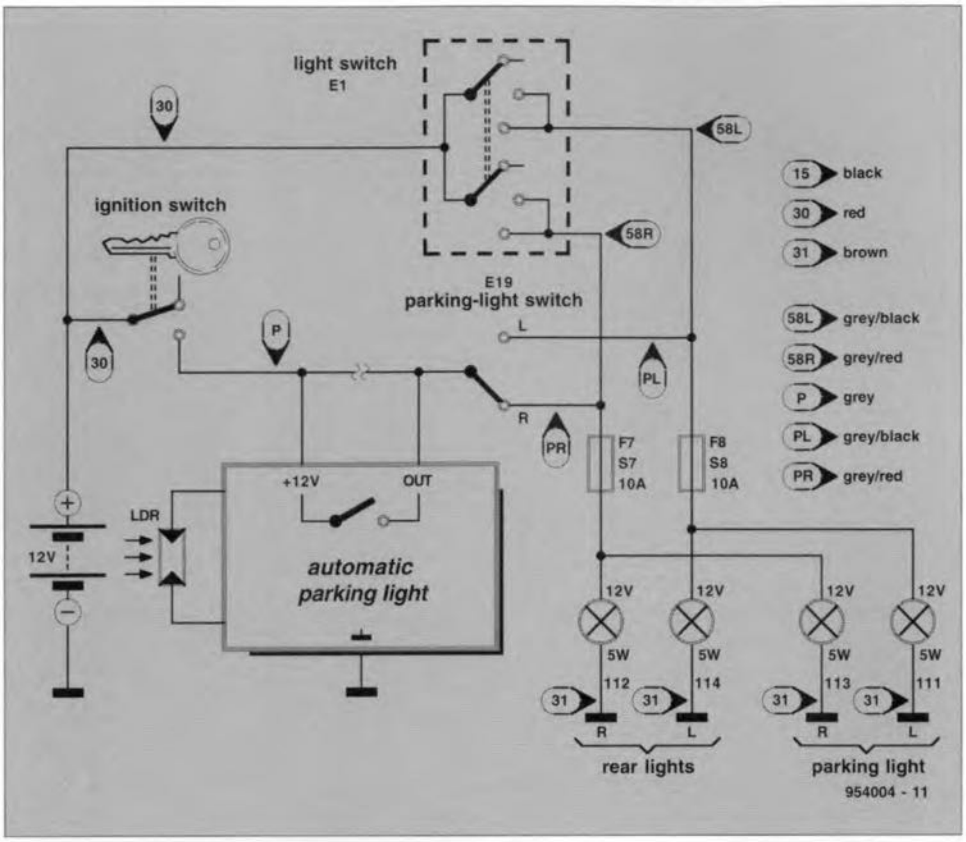
Linear Technology Application Note 18

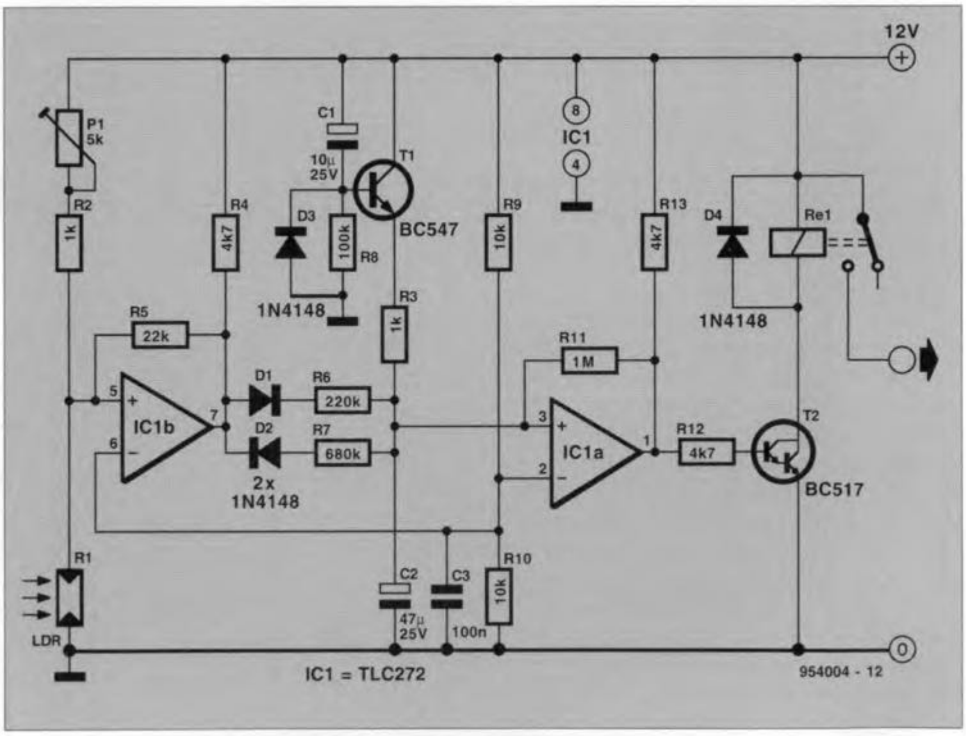
[954067]

AUTOMATIC PARKING LIGHT

In some countries, it is still obligatory for cars parked in the road at night to have their parking lights on. If the car is parked well before dark, there is an unnecessary drain on the battery. This may be obviated, however, by the present circuit, which ensures that the parking lights do not come on until it gets dark, and are switched off again when it gets light. It has integral delays to prevent the parking lights being switched off by the headlights of passing traffic. The circuit can be enabled only when the ignition is switched off and the parking lights are switched on.

How the circuit is built into the car electronics is shown in the top diagram. The figures near the various cables are standard in most European cars; the colour codes apply primarily to Volkswagen vehicles.





The circuit of the automatic parking light unit is shown in the lower diagram. The ambient-light sensor is a light-dependent resistor (LDR), R_1. The sensor and R_2 form a potential divider, whose junction is linked to the non-inverting input of comparator IC_{1b}. The inverting input of the IC is at half the battery voltage via R_9–R_{10}. Resistor R_3 provides a certain hysteresis, which prevents the comparator reacting to brief light pulses. Resistor R_4 is a pull-up resistor.

The circuit is made insensitive to brief darkening or lighting, as the case may be, by delay network R_6-R_7-C_2. Branch R_7-C_2 determines the delay with which the parking lights are switched on, while R_6-C_2 determines for how long the parking lights remain on when daylight breaks. With values as specified, these delays are 15 s and 10 s respectively.

Transistor T_1 provides an operating check by actuating the parking lights for 15 s after the ignition is switched off. To this end, C_2 is charged via T_1 and R_3, until the potential across C_1, which is charged via R_8, attains a level at which T_1 is cut off. If there is sufficient ambient light, C_2 is discharged slowly via R_7. If it is dark, however, C_2 can not be discharged, so that the potential across it stays high, and the parking lights remain on. Capacitor C_1 is discharged via D_3 when the supply voltage is switched off to ensure that the delay time is available when next the supply voltage is switched on again.

The delay networks are followed by a second comparator, IC_{1b}, whose inverting input is also at half the supply level via R_9–R_{10}. Hysteresis is provided by R_{11}. Resistor R_{13} is a pull-up resistor.

The comparator controls T_2, which can supply a current of up to 200 mA, which is sufficient to energize an inexpensive car-type relay, Re_1. The transistor is protected against inductive voltage peaks by D_4.

The entire circuit, except R_1, is built on a small piece of prototyping board. It needs only three leads to the car's electric system: +12 V, earth and output. These connections are best made with car-type in-cable connectors. The leads to the relay contact should be heavy-duty wire.

The LDR may be located in any suitable position on the car where it receives ambient light. If the connecting cable to it is longer than 10 m, screened cable should be used to prevent noise on the car's electric system causing spurious operation.

The values of R_2 and P_1 should be

matched to the type of LDR used. When it is light, they should provide a voltage at pin 5 of IC_{1b} which is appreciably below the potential at pin 6. When it is dark, they should provide a voltage at pin 5 that is appreciably higher than that at pin 6.

The time constants of the delay networks may be altered to individual requirements, but the value of C_2 should not be so high that the charging current through R_6 is no longer sufficient to charge the capacitor.

Design by G. Kleine

[954004]

MAINS VOLTAGE DETECTOR

It is normally easy to tell when the mains voltage fails because then the radio stops playing or you are in the dark. Joking apart, computers and other electronic equipment may require a special signal which provides a simple mains present/not present indication. Simple as it sounds, there is more to such a detector than you would expect. The main reason is electrical safety, and that is exactly the key feature of this little circuit, which offers total electrical insulation from the mains.

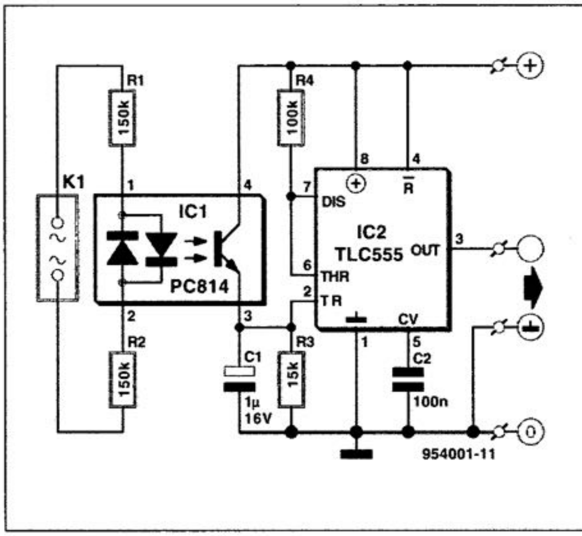
The mains voltage arrives at the input of optoisolator IC_1 via two relatively high-value resistors, R_1 and R_2. Inside IC_1 are two anti-parallel LEDs, which light alternately on each half-phase of the mains voltage. The phototransistor conducts during (nearly) the entire mains voltage cycle. The resulting voltage at its emitter is smoothed by capacitor C_1, and is turned into a TTL compatible digital switching signal by a TLC555 timer, which is configured as a Schmitt trigger. Capacitor C_1 is discharged via R_3.

The output of the circuit is logic 0 (low) when the mains voltage is present, and logic 1 (high) when the mains voltage fails at the input.

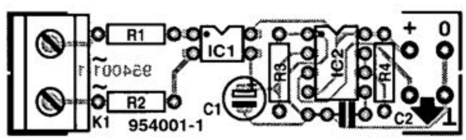
A simple BASIC program is shown to enable a PC, or indeed any computer capable of running BASIC, to monitor the output of the circuit. Obviously, the computer should not be on the same mains outlet as the detector (so get out your laptops and notebook PCs)! The output of the detector is connected to any data line of the Centronics port on LPT1:, and, of course, to the ground line of the interface.

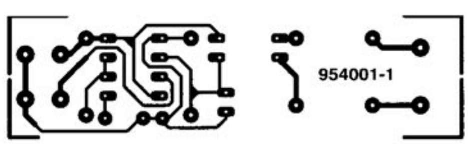
The printed circuit board shown here is unfortunately not available ready-made through our Readers Services, so you have to make it yourself, or have it made.

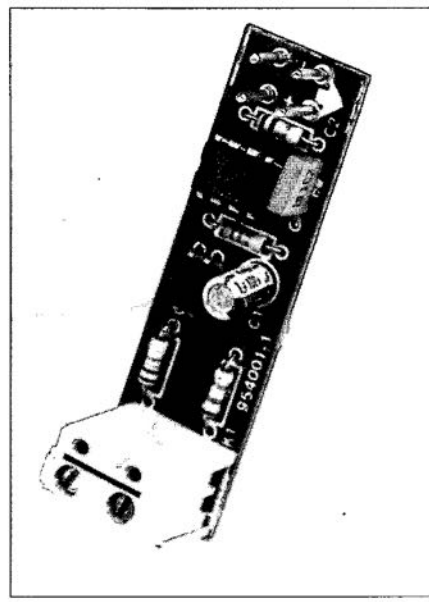
Warning The circuit should be built into an all-plastic case to eliminate any risk of the parts at mains potential being touched. It is recommended to use a mains adaptor case, which comes with a moulded mains plug. Never work on the circuit while it is connected to the mains.




```
CLS
key$ = ""
OUT &H378, 0
LOCATE 12, 29
COLOR 7
PRINT "state:"
WHILE key$ <> CHR$(27)
   x% = INP(&H378)
   LOCATE 12, 38
   IF x% > 0 THEN
     COLOR 0, 7
     PRINT " off"
   ELSE
     COLOR 0, 2
     PRINT " on"
   END IF
   key$ = INKEY$
WEND
```







Parts list

Resistors:
$R_1;R_2$ = 150kΩ
R_3 = 15kΩ
R_4 = 100kΩ

Capacitors:
C_1 = 1μF 16V radial
C_2 = 100nF

Integrated circuits:
IC_1 = PC814 (Sharp Semiconductors)
IC_2 = TLC555

Miscellaneous:
K_1 = 2-way PCB terminal block, pitch 7.5mm

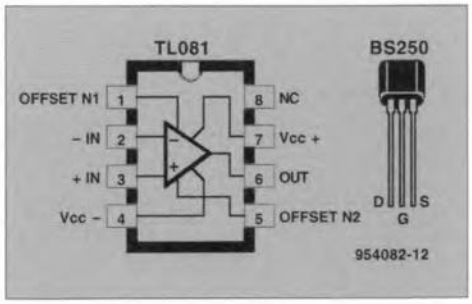
Design : O. Rennen
[954001]

CURRENT METER

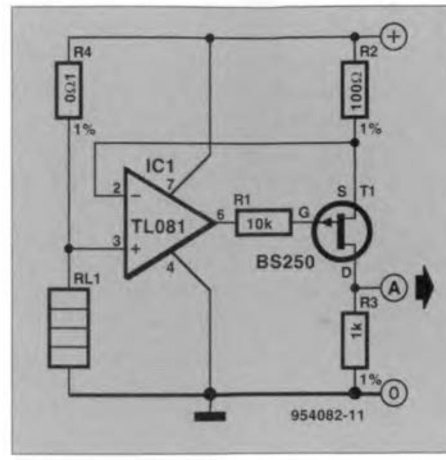
Measuring the current drawn by an apparatus is always a rather complicated matter. If a shunt resistor is placed across the positive supply rail, the measured voltage has, by definition, an offset equal to the supply voltage. It is, of course, much simpler to place a resistor in series with the negative supply line, but, owing to various circumstances, that is not always possible. A better way is using the circuit shown in the diagram.

The circuit removes the common-mode voltage and has variable gain so that the transfer ratio can be selected as desired. It draws a current of only a few milliamperes.

In the circuit, the measurement resistance is R_4. The current-dependent voltage drop across this resistor is applied to the non-inverting input of IC_1. The op amp will endeavour to bring its inverting input at the same potential by switching T_1 on or off. As a result, a current flows through R_2 that is proportional to the drop across R_4 and thus to the current through R_L. The current through R_2 causes a potential drop across R_3, which is proportional to the load current and is referred to ground. Thus, a voltmeter connected between *A* and ground gives an exact indication of the load current.



The sensitivity of the meter is determined as follows. A load current of 1 A causes a drop of 100 mV across R_4. The current through R_2 is then 1 mA, which causes a drop of 1 V across R_3. The transfer ratio is thus 1 V A^{-1}. This ratio can be varied by giving R_3 a different value. Resistor R_4 may also be adapted to particular measurements. If, for instance, the load current is small, it is advisable to increase its resistance to 1 Ω.



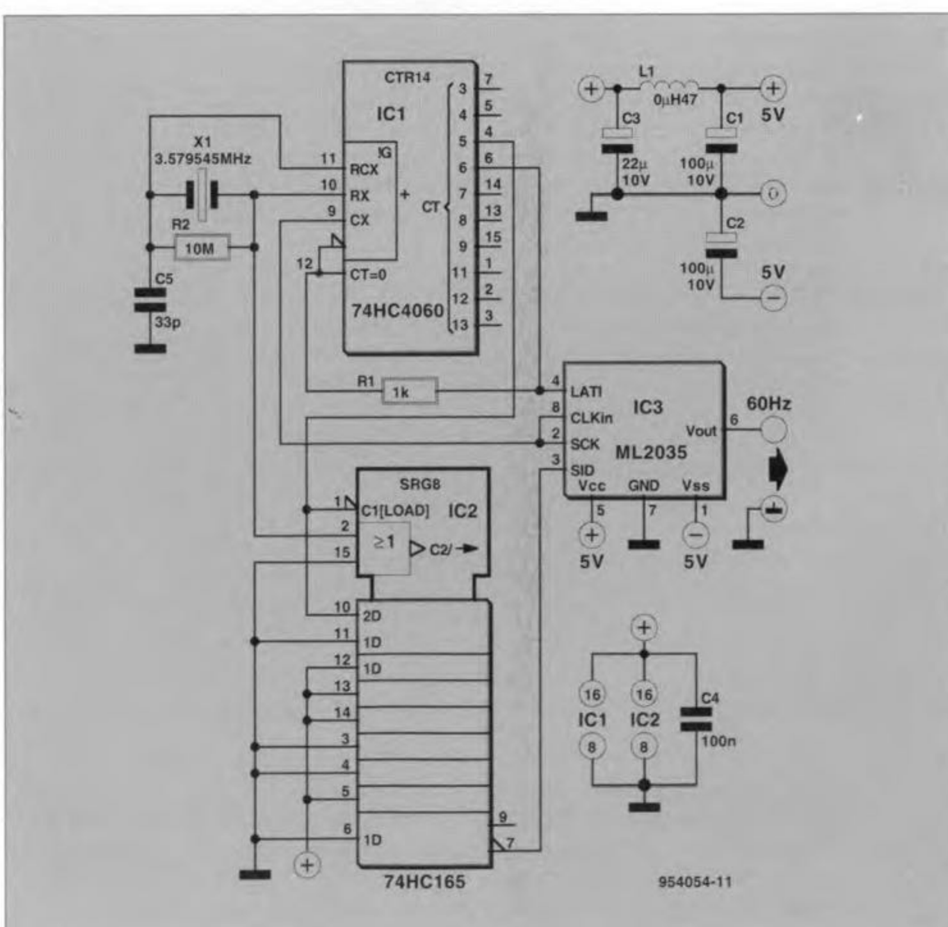
A National Semiconductor application
[954082]

PROGRAMMABLE SINE WAVE OSCILLATOR

Micro Linear's ML2035 monolithic programmable sine wave generator IC is used here without the usual microcontroller to supply the frequency determining code word. IC_1, a 74HC4060 oscillator/counter, supplies both the sine wave generator's clock and the gating pulse for shift register IC_2. When the Q5 output of IC_1 goes high, IC_2 starts to shift eight hard-wired bit levels into the SID (serial input data) input of the ML2035, followed by eight 1s. The shift-in operation is synchronized by the positive-going edge of the clock signal applied to the SCK pin. When all 16 bits are read, the frequency determining word is copied into the ML2035's internal data memory on the negative-going edge at the LATI input. The generator's output frequency, f_{out}, is computed from

$$f_{out} = \frac{f_{CLKIN} \times (D15 - D0)_{DEC}}{2^{23}}$$

The lower-order byte of the frequency determining word is hard-wired at the load inputs (A through H). The higher-order byte is always 255 (1111 1111). Because the $\overline{Q}$ output of the HC165 is used, inverted logic applies, and the frequency determining value should be written as:



NOT(1111 1111 HGFE DCBA)

The circuit can produce 50 Hz or 60 Hz sine wave signals from an inexpensive NTSC quartz crystal. The shift register load values for other popular crystal frequencies are shown in the table. The generator's output signal exhibits a maximum of 0.5% THD.

Current drain of the circuit is about 7 mA at +5 V, and 2 mA at –5 V.

A Micro Linear application
[954054]

Micro Linear, 2092 Concourse Drive,
San Jose, CA 95131, U.S.A.
Tel. (408)/433-5200.

$F_{CRYSTAL}$ (MHz)				74HC165 $\overline{\text{Code}}$		
	f_{OUT}	D_{10}	D_{HEX}	ABCD	EFGH	Error
4.00	50	105	69	1001	0110	0.14%
4.00	60	126	7E	1000	0001	0.14%
4.194304	50	100	64	1001	1011	0.0%
4.194304	60	120	78	1000	0111	0.00%
6.00	50	70	46	1011	1001	0.14%
6.00	60	84	54	1010	1011	0.14%
8.00	50	52	34	1100	1011	–0.82%
8.00	60	63	3F	1100	0000	0.14%

LOUDSPEAKER PROTECTION IN CARS

The output stages of many car radio/cassette players have no switch-on delay for the loudspeakers. This causes annoying clicks when the radio is switched on or off. Usually, the output stages are remotely controlled via switching voltage outputs of the radio. One of these outputs, V_s, is used by the present circuit to switch on the amplifier before the loudspeakers and to switch off the loudspeakers before the amplifier.

The realization of this is shown in the circuit diagram. When the radio is switched off, the inverter inputs are at ground potential via R_2. Capacitor C_2 is then discharged, the input of IC_{1a} goes low, T_1 is switched off, and relay Re_1, which switches control voltage V_s' (output G) is deenergized. The output of IC_{1c} is high, however, so that C_3 is charged and the output of IC_{1d} is low. Thus, T_2 is off and the loudspeakers are removed from the output stages by relay Re_2.

When the radio is switched on, the V_s input is 12 V, C_2 becomes charged and series-connected IC_{1a} and IC_{1b} change state. Transistor T_1 then conducts, whereupon Re_1 is energized, so that the output stages are switched on.

At the same time, IC_{1c} changes state, which causes D_4 to be reverse biased, so that C_3 is discharged slowly via R_6. When the potential at the input of IC_{1d} drops, the inverter changes state, so that T_2 conducts, whereupon Re_2 is energized and its contacts connect the loudspeakers to the output stages.

When the radio is switched off, the output of IC_{1c} goes high, causing C_3 to be charged, whereupon IC_{1d} changes state and T_2 is switched off. The loudspeaker connections are thus broken before C_2 has been discharged sufficiently via R_3 to cause IC_{1a} and IC_{1b} to change state, T_1 to be switched off, and Re_1 to be deenergized (which switches off the output stages).

The Schmitt trigger inputs of the inverters ensure faultless switching.

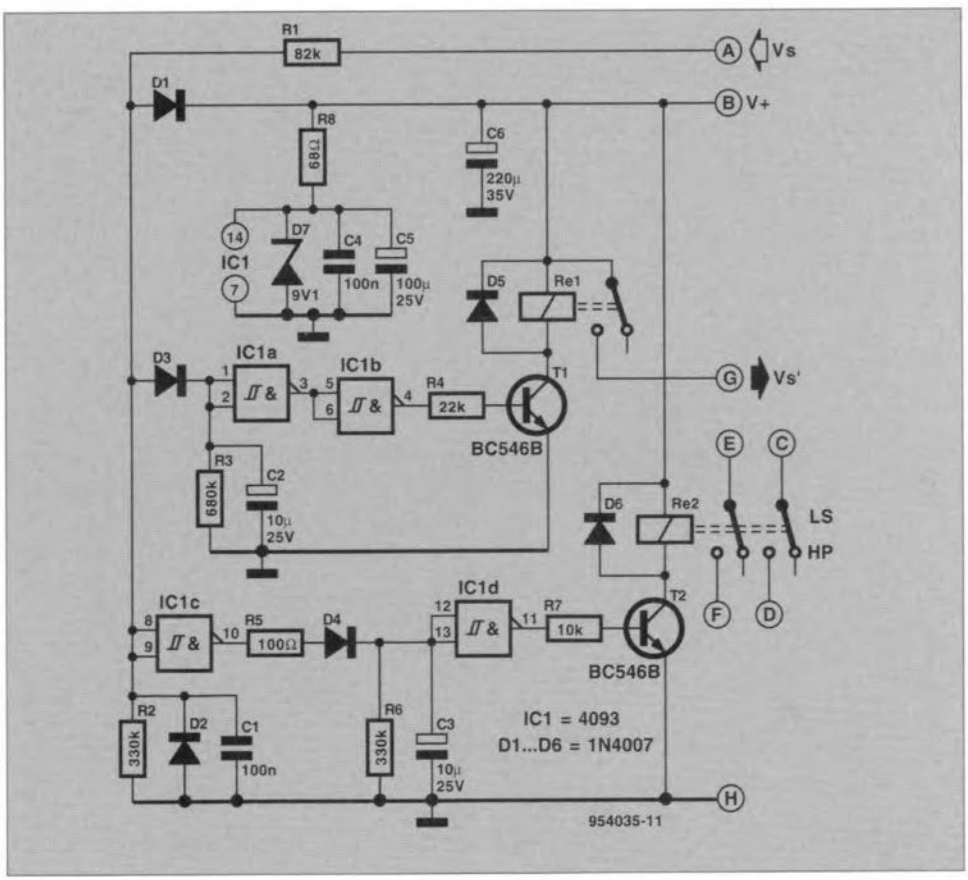
The supply voltage for the inverters is stabilized at 9 V by R_8 and D_7, and buffered by C_4, C_5. The control voltage lines are decoupled by C_1 and C_6. Diodes D_1 and D_2 short-circuit voltage peaks over 12 V and below –0.7 V respectively, thereby protecting the inputs of the inverters against excessive potentials.

Relay Re_1 may be a miniature type, since its contacts seldom handle currents higher than 250 mA. Relay Re_2, however, should be a heavy-duty type whose contacts can cope with the high currents flowing through the loudspeaker coils.

All connections are best made via car-type in-line connectors available from all good car accessory shops.

The protection circuit draws a quiescent current of only 4.5 mA.

Design by F. Hueber
[954035]

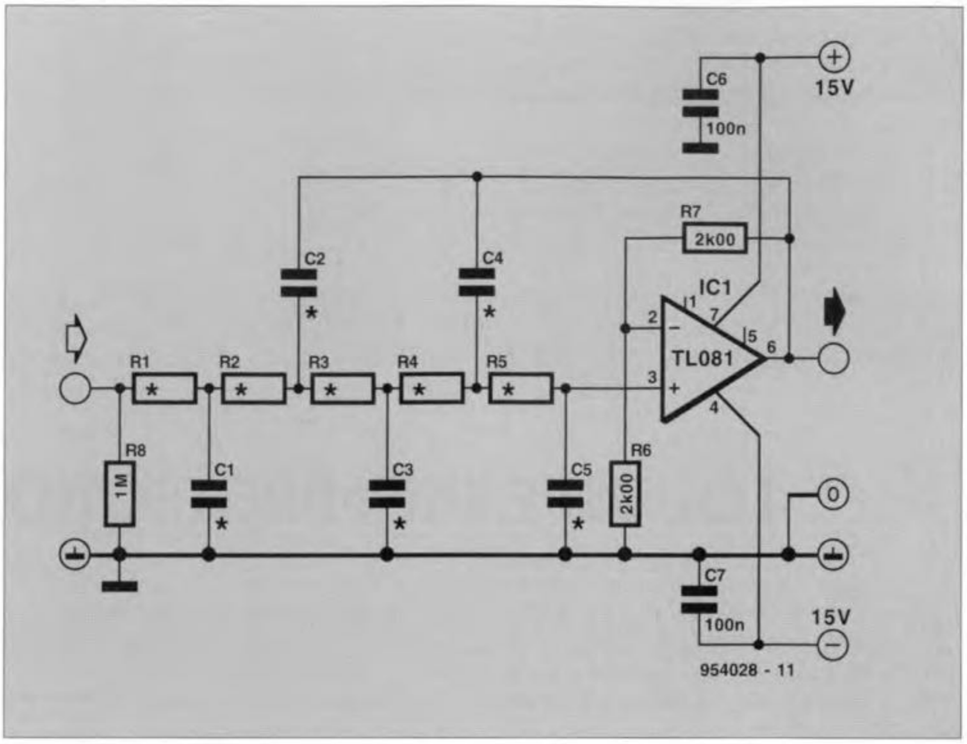


5TH-ORDER BUTTERWORTH FILTER

Butterworth filters have a characteristic which is virtually straight right up to the cut-off point and this makes them ideal for many applications. Unfortunately, the tolerance of the components is critical and should, therefore, be 1% or better.

The Type TL081 op amp used in the prototype may be replaced by any good op amp that is stable at an amplification of ×2 or more. The table gives values for an amplification of ×2 and a cut-off frequency of 1 kHz. Because of the amplification, the ratio of the capacitors is as small as feasible, while that of the resistors is as nearly equal as possible. The last line gives theoretical capacitor values for equal values of the resistors. These are required if the low-pass filter is to be converted to a high-pass section.

If the amplification presents a problem, it can be reduced by converting R_1 into a potential divider. Make sure in that case that the output resistance of the divider is equal to the value of R_1 in the table. Resistor R_8 provides bias for IC_1 when there is no d.c. coupling at the input.



The table gives resistor values for all combinations of capacitor from the E-12 series whereby their mutual ratio varies little if at all. Because of the rounding off of the capacitor values to E-12 values, some parameters of the filter may differ by a few per cent from the theoretical values.

The TL081 draws a current of about 2 mA.

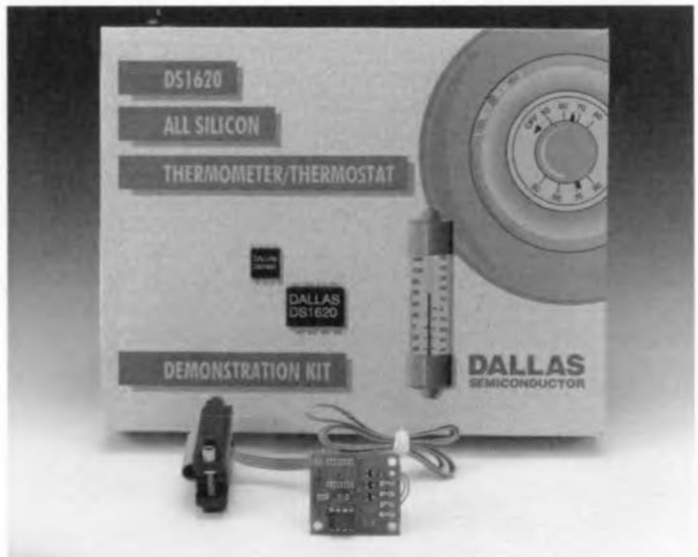
Design by T. Giesberts [954028]

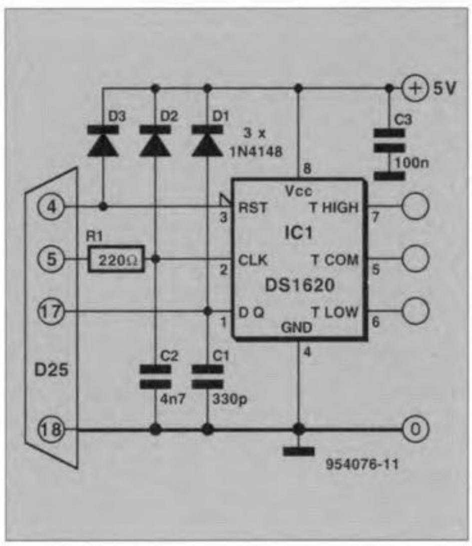
	C_1 (nF)	C_2 (nF)	C_3 (nF)	C_4 (nF)	C_5 (nF)	R_1 (kΩ)	R_2 (kΩ)	R_3 (kΩ)	R_4 (kΩ)	R_5 (kΩ)
1	10	5.6	8.2	3.9	1.5	33.216	36.632	32.664	29.103	32.864
2	12	6.8	10	4.7	1.8	27.462	30.613	26.568	24.129	27.447
3	15	8.2	12	5.6	2.2	21.710	25.189	22.187	20.526	22.550
4	18	19	15	6.8	2.7	17.261	21.891	17.364	17.049	18.416
5	22	12	18	8.2	3.3	14.059	18.145	14.531	14.231	15.053
6	27	15	22	10	3.9	11.931	14.055	11.926	11.481	12.799
7	33	18	27	12	4.7	9.340	12.147	9.654	9.621	10.713
8	39	22	33	15	5.6	8.158	9.783	7.988	7.509	8.970
9	47	27	39	18	6.8	7.147	7.629	6.755	6.265	7.305
10	56	33	47	22	8.2	6.198	6.137	5.611	5.080	6.011
11	68	39	56	27	10	5.115	5.099	4.821	4.119	4.917
12	82	47	68	33	12	4.243	4.227	4.001	3.341	4.104
13	35.076	19.362	25.933	11.803	4.9125	10.0	10.0	10.0	10.0	10.0

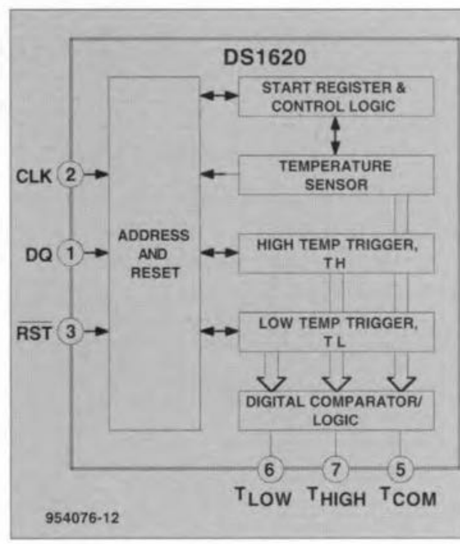
TEMPERATURE SENSOR

The circuit described is based on a Type DS1620 temperature sensor from Dallas Semiconductor. This device can measure temperature with a resolution of 0.5 °C over the range –55 °C to +100 °C. Moreover, since it contains two 8-bit non-volatile memories, it can give an alarm above and below programmed limits.

The IC has three switched outputs of which T_{high} is high when the temperature is above the upper limit; T_{low} is high when the temperature is below the lower limit; T_{com} is low with rising temperature, becomes high when the upper level is exceeded, and then becomes low again when the temperature has dropped below the lower limit.







The limits are programmed via a three-wire interface. The internal memories retain their data, so that the chip after having been programmed can be placed into a circuit without losing its programmed information. This makes it possible for a stand-alone unit to be designed. Loss of the supply voltage has no consequences.

Data are written on the leading edge of the clock. Use is made of an 8-bit serial format, LSB first. There is an internal configuration register that determines whether the chip operates in a stand-alone situation. In that case, a conversion can be started via a pulse at the clock input, or measuring may be continuous. During a measurement, which lasts only 1 second, the IC draws a current of 1 mA; during standby only 1 µA.

When the IC communicates with a PC or microprocessor, instructions such as 'read the temperature', 'read/write the temperature or status registers', 'start/stop conversion' are possible. More ICs may be accessed in turn by resets.

To measure the temperature, the instruction 'AAH' is sent to the IC. The master (PC or microprocessor) then emits nine additional clock pulses and, always at the leading edge, reads the data emerging from the IC starting with LSB. The temperature is given in 9-bit 2s complement format. Thus, 0 means 0 °, 32H means 50 steps of half a degree, that is, 25 °, and 1FFH means one negative step, i.e., –0.5 °.

The circuit as shown is contained in a demokit available from Dallas Semiconductor. The temperature, switching thresholds, and status of the three switched outputs are shown via the parallel port with the aid of a Visual Basic program. It shows clearly how simple the interface can be.

Power is provided to the IC via the three diodes.

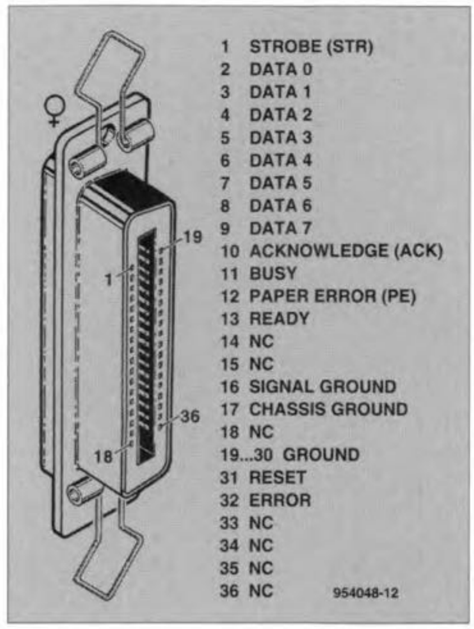
Design by K. Walraven
[954076]

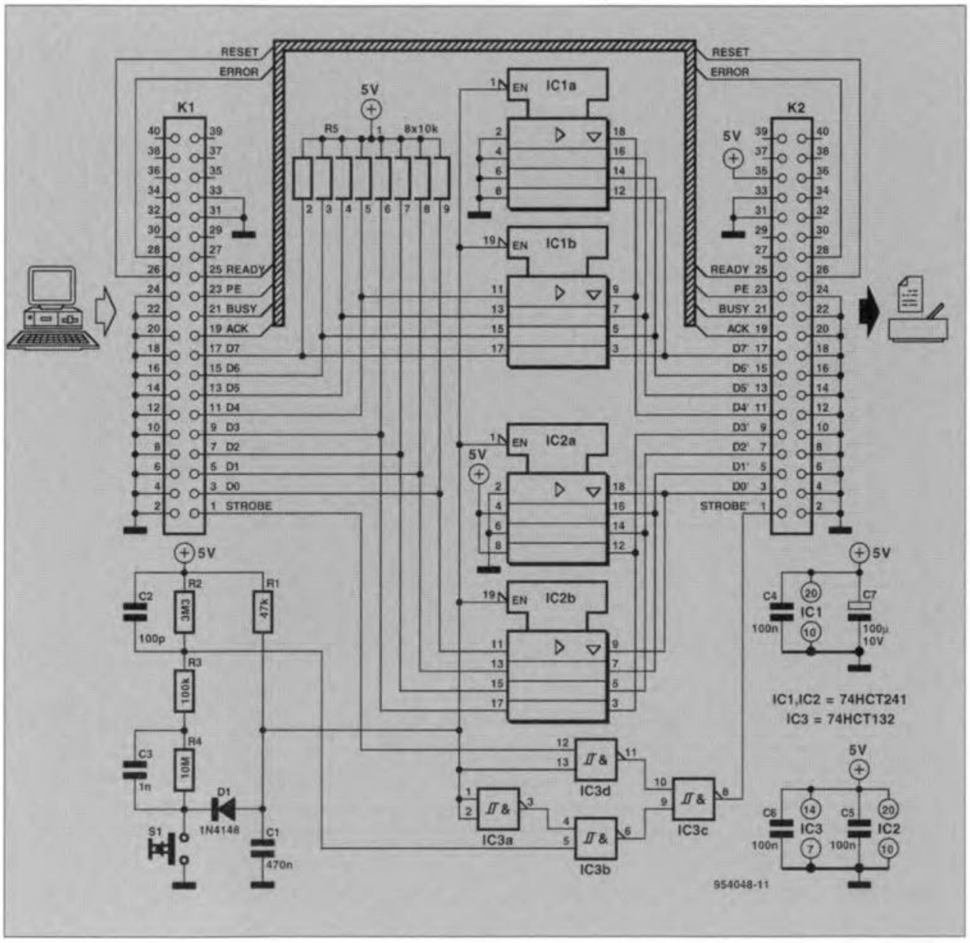
LINE FEED ADD-ON FOR HP DESKJET 520

A fine printer, really, that Deskjet 520 from Hewlett Packard, only it lacks a line feed button! This obvious deficiency is made good by the add-on circuit shown here. It is inserted in the Centronics link between the computer and the printer, and effectively supplies the code for a line feed (0A hex) plus the associated strobe pulse.

The line feed command is pre-wired as hex code 0A on pins 2, 4, 6 and 8 of IC_1 and IC_2. The two HCT241 bus transceivers switch between Centronics data and the line feed command. The Centronics data are distributed across the transceiver sections with 'high' enable. The line feed command is distributed across the transceiver sections with 'low' enable.

The strobe pulse required to copy the line feed command into the printer is generated by pressing S_1. The switch controls a simple timer based on *RC* networks and Schmitt trigger NAND gates, which also debounce the presskey. Gates IC_{3b} and IC_{3d} select between strobe pulses supplied by the computer, and the line feed strobe pulse. Obviously, the line feed button should not be pressed while the computer is still feeding data to the printer, since that would corrupt the file to be printed. The length of the line feed command is determined by the discharging of C_1, and is will be at least a few tens of milliseconds, depending on the time S_1 is held depressed. The strobe is actuated a little later (approx. 25 µs), and lasts about 200 µs.





The circuit has a current consumption in the µA region, so that it can be powered from the 5-V supply voltage which is present (in most cases) on pin 18 of the Centronics connector (check to make sure!). The input and output connectors of the circuit are 40-way boxheaders which accept 40-way IDC flatcable sockets. The pinout of the Centronics socket is given for reference. Note that only pins 1 and 36 are 'straight through' connections between the boxheaders and the Centronics sockets — the rest of the pins have no number correspondence.

Design by T. Giesberts [954048]

INTELLIGENT LOW-SIDE SWITCH

Although a digital circuit for driving a load such as a lamp or a d.c. motor is fairly simple, it has a few drawbacks. It does not make the layout and construction of a printed-circuit board simpler. Space needs to be reserved for at least one resistor, one transistor and one diode. If the circuit consists of a number of such stages, this can be a problem. It would be far easier to use an IC that has all these components on board. Several types of this kind of IC are available, such as Texas Instruments Type TPIC2404.

The TPIC2404 is a monolithic high-voltage, high-current, quadruple low-side switch especially designed for driving from low-level logic to peripheral loads such as relays, solenoids, motors, lamps, and other high-voltage, high-current loads.

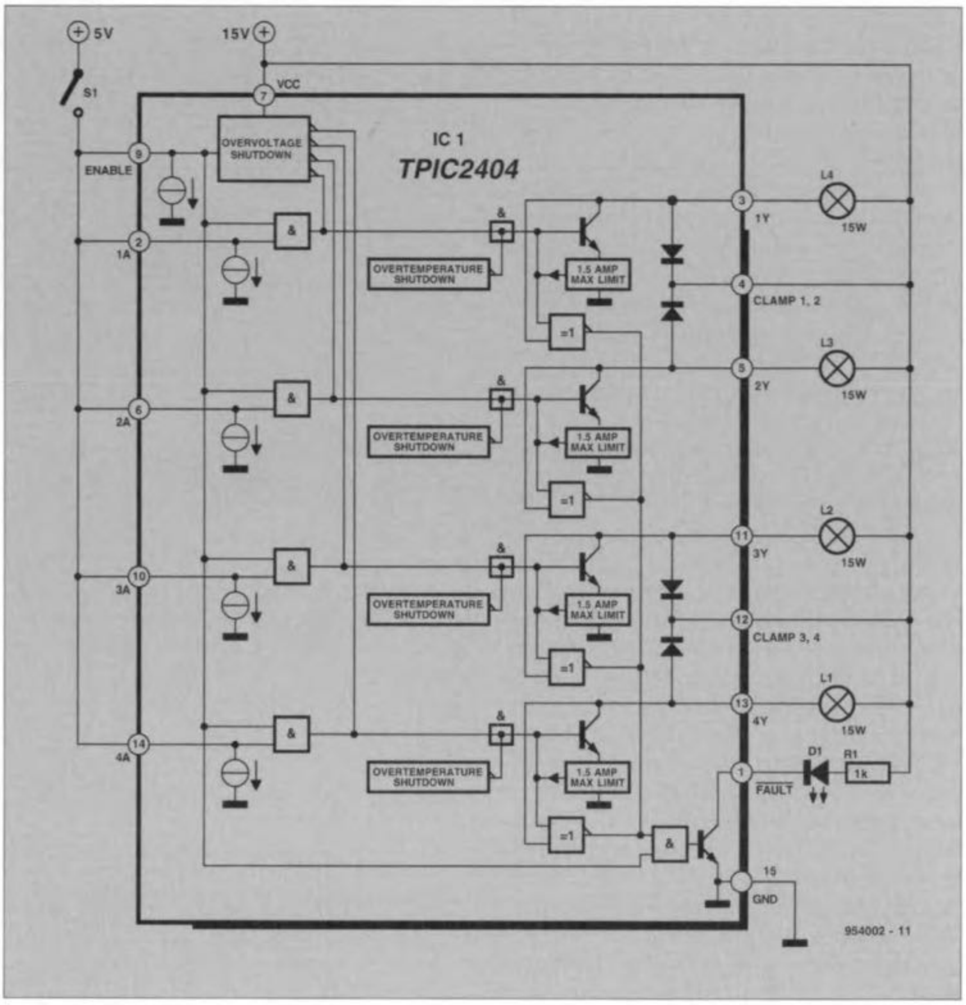
The device has four inverting open-collector outputs controlled by a common-enable input. When ENABLE is low, the outputs are disabled. An error sensing circuit monitors loads and device faults. When an error is sensed the FAULT output goes to a low state. In addition, the device has on-board V_{cc} overvoltage and thermal overload protection circuits, and the outputs are current-limit protected.

The output state is compared with the input value by four XNOR gates. If, for instance, an output is shorted to earth, this will be detected by one of the gates which causes D_1 to light. This diode also lights in case of overvoltage or a thermal overload.

TheTPIC2404 has five TTL or CMOS compatible inputs: four for control and for enable. When ENABLE is high, and switch S_1 (see diagram) is operated, lamps L_1–L_4 light simultaneously. The integral current limiting circuit restricts the current to 1.5 A.

Finally, the device is protected against inductive voltages at the outputs by four diodes in parallel with the load. This arrangement enables the device to switch inductive loads without any difficulty.

A Texas Instrument application [954002]



	Enable	A	Y	FAULT
Normal operation	H	H	L	H
Open load	H	L	H	H
Short to ground	L	X	H	H
Overvoltage shutdown	H	L	L	L
Thermal shutdown	H	X	H	L
Short to V_{cc}	H	H	H	L

FUNCTION TABLE

SMD-TO-DIL ADAPTORS

These adaptor boards are designed for readers who find it increasingly difficult to experiment with certain integrated circuits which are only available as SMDs (surface mount devices). Also, an increasing number of ICs is no longer manufactured inDIL shape, being replaced by SMDs (for example, the 74LS163A). Obviously, the problems are caused by the small size of these parts, which require dexterity, as well as by the fact that SMDs do not sit in sockets, and thus are not easily removed from a circuit for a quick test, or for faultfinding.

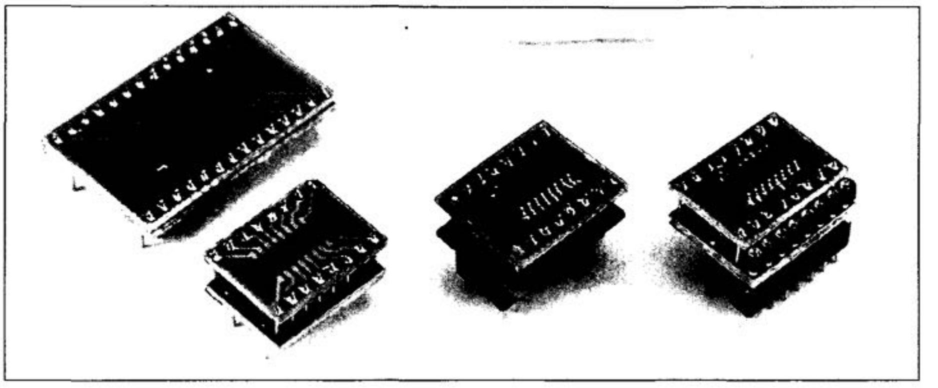
The PCB shown (which is not available ready made) enables you to build three adaptors. Each adaptor consists of two sub-boards. First, the SMD is carefully soldered at the copper side of the board with the SMD pin layout. Next, pins cut off from a turned-pin IC socket are pushed through the holes of the centre rows of the other sub-board, and soldered. Alternatively, short pieces of stiff wire can be used. Check that the the pins at the underside of the sub-board are properly aligned, and that they can be inserted into a regular DIL socket. Then secure the sub-board with the SMD on it on top of the other sub-board with short lengths of stiff wire which pass through the relevant holes at the outside of the sub-boards. In general, make the adaptor unit as compact as possible.

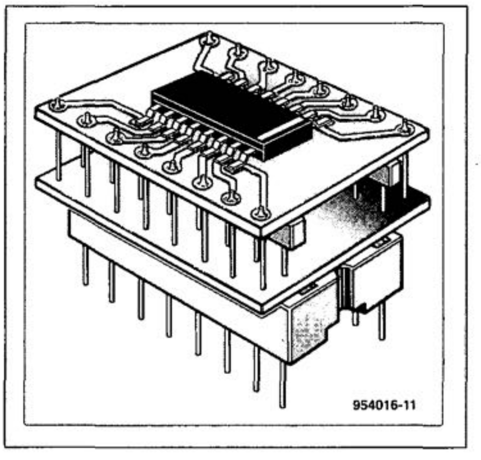
The 24-way adaptor is obviously also suitable for 18 and 20-pin ICs, provided there is enough space around the socket on the PCB for the adaptor to be inserted.

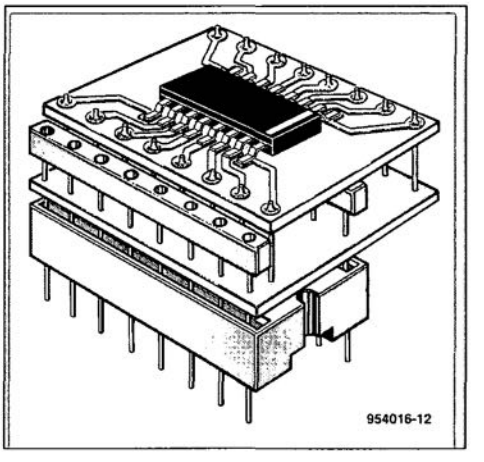
The 16-way adaptor with the IC offset from the centre is intended for applications where there is little space at one side of the socket into which the adaptor is inserted.

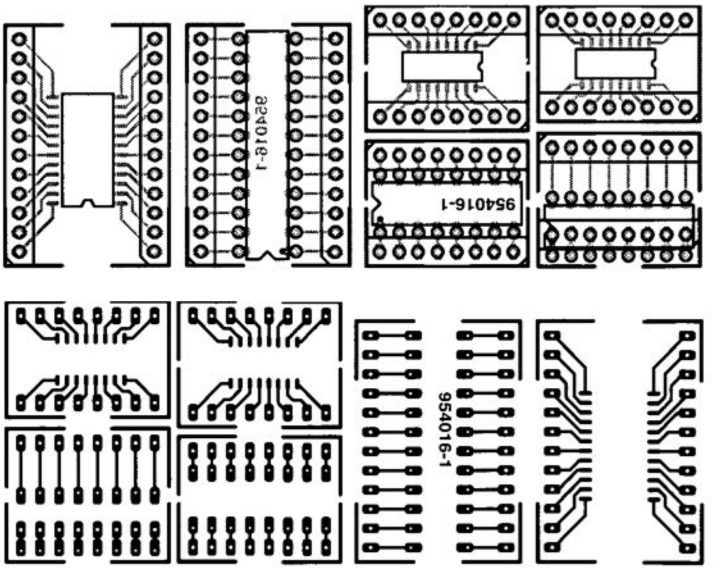
Finally, before inserting the adaptor, ascertain the orientation, and be sure the pin numbers of the SMD correspond one-to-one with those of the DIL version of the relevant IC!

Design by L. Lemmens
[954016]



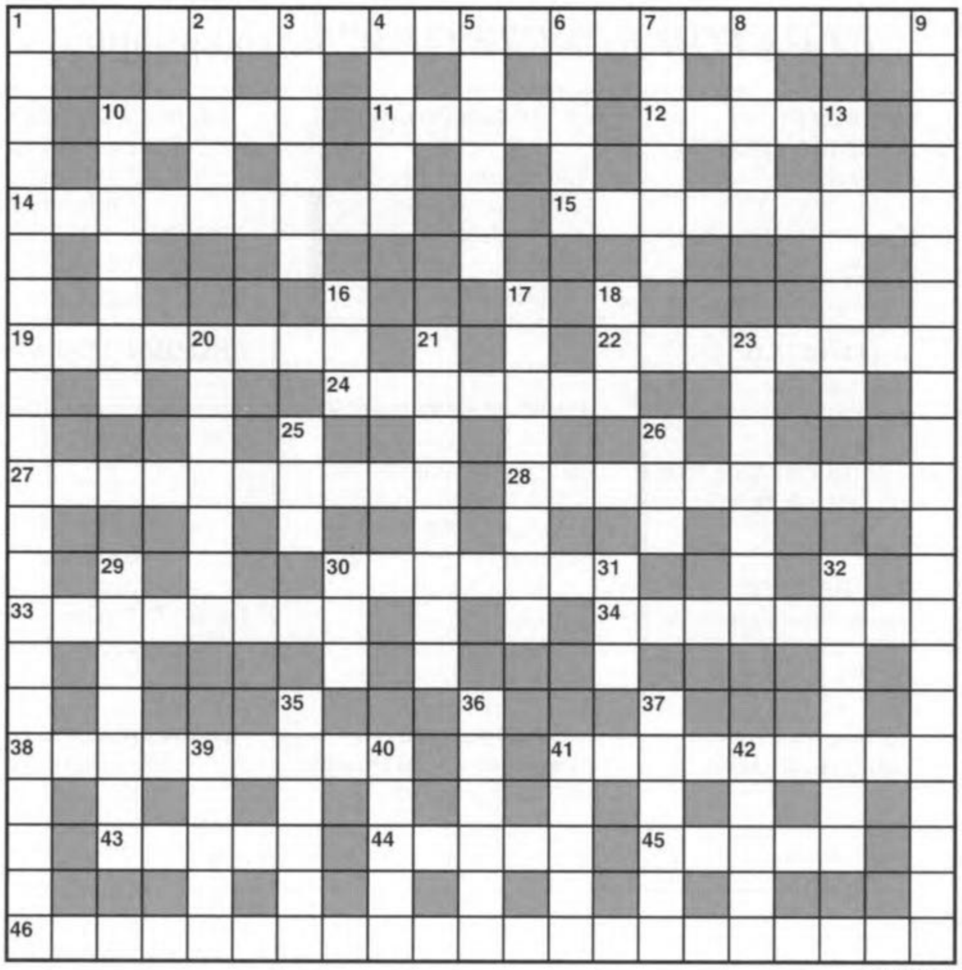






SUMMER CROSSWORD

By Klodz



950087-11

Across:

1. A proverb ruins picnic time - surely not Christian Science? But still is first word! (2, 9, 4, 6)

10, 34 Some dip around for a way to work (5, 8)

11 In which paragon advertises organ! (5)

12 Mountains where you find some sour Alsatian dogs? (5)

14 Gold sisters made to slow flow (9)

15 To rob Mig oil may cause mass confusion (9)

19 2000 pointless slaters getting wrong nets (8)

22 Cleaning lady's company, et al, produces fuel (8)

24 Interpol – EC – attempts to catch the beast (7)

27 A tenth some make out to be different (3, 3, 4)

28 Hill-dweller round with a twitch – a bit cold here! (10)

30 Deprecate? No thanks – about to come early (7)

33 Despair planet Earth – has one of many fliers (8)

34 See 10 across

38 Impress room-mate in Ed's place! (9)

41 Top bill, one on hill, keeps the charge! (9)

43 Lesotho losing the French, mixed sprout (5)

44 Preen? Clean out natural log division! (5)

45 Drive away the bird (5)

46 O, spectacle: Sid is an iron. But needs extra cares! (7, 14)

Down:

1 A tern in its lair on a port – strangely – comes and goes from these! (13, 8)

2 Industrial attempt to denude flower (5)

3 Throw gold to Heavenly twin (6)

4 Rampage South to contain attendants (5)

5 Gold to the North East, with rye, missing the point – plainly (6)

6 Do I raid spokes? (5)

7 Tom's opposable ones?! (6)

8 Orate strangely to Muse (5)

9 Strict Roman dips toilet in muddied conurbations (12, 9)

10 Dreams I haven't seen properly?! (7)

13 Colin is made of this element (7)

16 Initialize a signal processor? It didn't do Cleopatra any good! (3)

17 Cycle-man goes to repair a water-violet (8)

18 Cat-play? (3)

20 He has surface-mount integrated circuits held at every laboratory! (7)

21 ? (8)

23 Managed dog with nothing but bitterness (7)

25 One quiet tree (3)

26 Sheepish memory? (3)

29 Creeps around point to follow gyration (7)

30 Cat-scanner? (3)

31 Neon light has a long lifetime (3)

32 Right kitten mixes with wool-gatherer (7)

35 Nearly everyone chases drink at doorway (6)

36 Insist apes have these in ear (6)

37 Strange despair loses a spinner (6)

39 Gosh! I can play Japanese chess (5)

40 Inman goes to have a fruit! (5)

41 Rue CD player, but got better (5)

42 Riddle it soundly for perfume (5)