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ELEKTOR ELECTRONICS JULY 1992
THERE IS STILL SPACE FOR FAMILY FIRMS

Information Technology—IT—has revolutionized our lives. The radical acceleration in the pace of change has had the same impact on the industry that sparked the revolution as it has in every other sector of the economy. Electronics is the most global of industries. And, with a fiercely competitive market driven by frenetic technical change, it is certainly one of the most free-wheeling. Yet, in this world-wide market dominated by giants of industry, there are still niches for smaller family firms that set their own agenda and business strategy.

The father and son team of Brian and Adam Chinery, based at Coventry, is a model example. The Chinerys are independent integrated circuit distributors (chip brokers) who buy and sell excess microchips on the world market. Their company, Dionics, has grown four times in four years, and its overseas business has increased to nearly 50 per cent of total turnover (£1 million plus). The company’s strategy is based on the old-fashioned values of a family firm: prompt response to enquiries and reliable service.

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The Modern Amateur Electronics Manual has over the past four years established itself as the most comprehensive electronics reference work ever produced. It has been purchased by many thousands of enthusiasts, students, training organizations and companies throughout the UK and around the world.

Following the closure of its UK publishing branch, WEKA Publishing have sold the UK title to Wimborne Publishing Ltd, who will continue to produce supplements and update the manual in line with previous policy. Wimborne Publishing Ltd, 6 Church Street, Wimborne BH21 1JH, England.

NEW PROGRAMMER FROM DATAMAN

Dataman has just released its fourth generation programmer: the Satly-Four or simply S4. The new unit encompasses the recent advances in surface mount design, intelligent microprocessor control and advanced power management to provide development engineers with the world’s most powerful handheld programmer/emulator.

Designed to programs EPROMs, Flash EPROMs and EEPROMs, S4 runs on high-power NiCd batteries. It can program up to 1000 PROMs on a single one-hour charge. Lithium battery back-up means your valuable data is never lost, even if you allow the NiCd batteries to run down. It is priced at £495 plus VAT.

Dataman, Station Road, Maiden Newton, DT2 0AE, England. Phone (0300) 20719.

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ITT Instruments, 346 Edinburgh Avenue, Slough SL1 4TU, England; Telephone (0753) 511 799.

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B.K. Electronics, Units 1 & 5, Comet Way, Southend-on-Sea SS2 6TR, England; Telephone (0702) 527 572.

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Danish Audiophile Loudspeaker Industries, or Dali, is a well-established, innovative Danish hi-fi manufacturer with a policy to be used for troubleshooting training.

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NEW from Maplin Electronics, one of Europe's largest electronic component distributors, is a compact, cost-competitive video düber and enhancer unit that will help to minimize signal degradation when dubbing and/or monitoring video recordings. The quality unit has inputs and outputs for two VCRs plus two monitor outputs.

**IBF LOUDSPEAKERS FROM B.K. ELECTRONICS**

As UK distributors for IBL, B.K. Electronics are taking off with flight-cased models. These include 12 in, 100 W r.m.s., 12 in, 200 W r.m.s., and 15 in, 200 W r.m.s. units. All models are fitted with wide-dispersion horns and include grilles, factory fitted to the die-cast aluminium speaker cabinets. These models are priced, respectively, at £159.00, £175.00, and £229.00 per pair incl. VAT.
points of this unit include the creation of a natural musical performance and a precise, realistic sound-stage. The 310 is a 20-litre, two-way bass reflex model incorporating a 6½” in (162.5 mm) cast chassis, polypropylene-coned woofer with a 32 mm long-throw voice coil and special low-loss surround for exceptional linearity and impulse performance at high and low levels. Treble is handled by a 1-in (25 mm) soft-dome drive with rear venting into an anti-compression chamber for a low resonance frequency.

Price at £399.99 incl. VAT, the Dali 310 is distributed in the UK by C.S.E., 1-3 Haywra Crescent, Harrogate HG1 5BG, Telephone (0423) 528 537.

VIDEO, CHROMA & DEFLECTION FOR MULTILINGUAL TV

AVAILABLE from Toshiba Electronics (UK) is an integrated video chroma deflection IC, Type TA8783N, for use in multi-standard colour televisions. housed in a single 64-pin shrink DIP, the device combines video, chroma and deflection processing with a teletext interface, all under the control of the industry-standard 12C bus.

To enable its use in multi-standard TV receivers, the TA8783’s chroma processing circuitry automatically identifies whether the incoming signal is PAL, Secam or NTSC, and distinguishes between 4.43 MHz and 3.58 MHz chroma sub-carriers. Colour and line controls are provided to NTSC standards, and RGB demodulation outputs are provided. The teletext interface features RGB and luminance inputs, and includes a half-tone function.

The IC runs from a 12 V supply and draws a current of not more than 130 mA.


NEW RANGE OF IMO AUDIBLE DEVICES

A new range of IMO buzzers, piezoe and transducers has been announced by the Electronic Components Division of IMO. The range has not been previously available in Britain and includes many devices with specifications not offered on the UK market until now.

IMO Precision Controls Ltd., 1000 North Circular Road, Staples Corner, London NW2 7JP, England; Telephone 081 452 6444.

EVENTS

IEE & IEEIE PROGRAMME

1 July - Periodic inspection and testing of electrical installations.
6 July - Portable appliance testing.
3-7 July - International Broadcasting Convention (Amsterdam)
10-12 July - History of Electrical Engineering.
12-15 July - Local telecommunications.
14 July - Initial verification of electrical installations.
26-31 July - Satellite communication systems.
27-31 July - Optical fibre communications.
11-13 Aug - Intelligent control.
19-21 Aug - Intelligent systems engineering.
24-28 Aug - European microwave conference.

Further information on these, and many other, events may be obtained from the IEE, Savoy Place, London WC2R OBL; from the IEEIE, Savoy Hill House, Savoy Hill, London WC2R OBL; from the IEE, Savoy Hill House, West Horsley, Surrey. KT24 6DJ; and from the IEE, Savoy Hill House, West Horsley, Surrey. KT24 6DJ.

FOUR ELECTRONICS SHOW

12-15 July - Local telecommunications.
24-28 Aug - European microwave conference.
6 July - Portable appliance testing.
3-7 July - International Broadcasting Convention (Amsterdam)

Further information from ElectroTech, Wix Hill House, West Horsley, Surrey. KT24 6DJ; Telephone (0483) 222 888.

TELECOMMUNICATIONS AND THE EUROPEAN BUSINESS MARKET

The Financial Times conference on Telecommunications and the European Business Market, the eighth in this annual series, will be held in London on 6 and 7 July 1992.

The conference will focus on the liberalization of the European telecommunications market. In the second half of this year the European Commission will have to decide whether to abolish the monopoly rights enjoyed in most countries over voice services and running networks. At the heart of the discussion is how to create a more dynamic telecommunications market, with lower prices and more services.

Details from The Financial Times Conference Centre, 126 Jermyn Street, London SW1Y 4UT; Telephone 011 925 2323.

COIL WINDING EXHIBITION

The Coil Winding International Exhibition will return to the Wembley Conference Centre, home to many of the industry’s past events, in September this year.

The exhibition, sponsored by the International Coil Winding Association and organized by the Evan Steadman Communications Group (now part of Reed Exhibition Companies), will be held from 29 September to 1 October in Wembley’s new Hall 3.

Further information from the Evan Steadman Communications Group, The Hub, Enmore Close, Saffron Walden, Essex CB10 1HL, Telephone (0799) 26699.

ANTENNAS AND PROPAGATION CALL FOR PAPERS

Papers are invited for the Eighth International Conference on Antennas and Propagation (ICAP’93), which will be held at the Edinburgh Conference Centre, Heriot-Watt University, Edinburgh, from 30 March till 2 April 1993.

The conference, which is being organized by the Institution of Electrical Engineers, is a leading international forum for the presentation and discussion of advances in antenna systems and electromagnetic wave propagation research.

Anyone wishing to offer a contribution should submit a synopsis of not more than one A4 page by 30 July, 1992, to: IEE Conference Services, IEE, Savoy Place, London WC2R OBL, England; Telephone 011 240 1871, Ext. 222.

SIXTH NORTH WALES RADIO & ELECTRONICS SHOW

The Sixth North Wales Radio & Electronics Show will be held on 31 October and 1 November, 1992, at the Aberconwy Conference Centre, Llandudno. The show opens at 10.00 a.m. on both days.

Further information from B. Mee, GW7SEXH, Anncoit, Hylas Lane, Rhuddlan, Clwyd, LL18 5AG; Telephone (0745) 591 704.
**OPTO CARD FOR UNIVERSAL PC I/O INTERFACE**

Whenever an interface is connected to a circuit with a supply voltage higher than 5 V, there is the risk that an error during experimenting, or a faulty component, will cause serious damage to the computer system. The opto card described here has been designed to afford complete electrical isolation between the computer and the (cruel) outside world, which is the only way to prevent system down time and expensive repairs caused by incompatible signal levels.

Design by J. Ruiters

In this article we present the second extension card for the Universal I/O Interface For IBM PCs, described in Ref. 1. While the relay card for this bus (Ref. 2) offers electrically isolated outputs, the present opto card is designed to process input signals in the safest possible way. By the way, the multi-purpose Z80 card described elsewhere in this issue may also be used as a controller for the universal bus.

**Eight optocouplers**

The circuit diagram of the opto card is given in Fig. 1. If you compare it with the circuit diagram of the relay card, you will find quite a few similarities. That is not surprising, because the functions of the two cards are closely related, one being a parallel output device (relay card), the other a parallel input device (opto card). The address decoding logic, for instance, is identical, consisting of a number of gates and a bidirectional buffer. With reference to Fig. 1, we are talking of IC9, IC10 and IC12. How the card is addressed, that is, how it complies with the rules of extension card addressing that apply in the universal bus system, will be reverted to below.

The opto card is actuated when bus signals A0, A1, RD and ENABLE go logic low. Consequently, bus buffer IC10 is enabled, and data is conveyed towards connector K1, i.e., towards the PC. At the same time, the OC input of data latch IC11 is pulled low, which enables the latch outputs. The data on the databus is clocked into the latches on the negative (falling) edge of the OC signal. This means that data applied to the optocoupler inputs is captured right at the start of a read cycle of the computer system, which ensures that data is stable on the bus during the actual read operation.

The eightfold optocoupler input circuit is all plain sailing. The only parameters to keep in mind are a couple of maximum specifications. To begin with, the input voltage is limited to ‘low voltage’ (in most countries, this is defined as 42 V a.c., or 60 V d.c.). This limitation is not caused by the optocouplers, but rather by the printed circuit board and a few other components. When designed to handle the 240 V (110 V) mains voltage at the input, the PCB would have become much larger to meet the relevant safety requirements. A further point to note is the specification of the series resistors with the optocouplers (R2, R4, R6, R10, R12, R14 and R16). The indicated resistors (1 kΩ, 0.25 W) may be used when the input voltage is between 2 V and 15 V d.c. Each optocoupler is protected against reverse voltages by a diode connected in anti-parallel. If voltages greater than 5 V d.c. are applied to the inputs, the
series resistors have to be increased accordingly, or resistors with a higher permissible dissipation must be used. The latter solution is not very elegant because of the larger size and the heat developed. The resistor values are calculated such that the LED current is a few milli-amperes at the given input voltage.

The construction of the opto card is entirely straightforward, and therefore not discussed further.

The bus system
As shown in Fig. 2, the address of any extension card connected to the universal bus system is determined by its position in the chain of extension cards. Unconventionally, DIP switches, jumpers and the like are not used. Apart from the beautifully simple and inexpensive hardware, the advantage of this system is mainly that you can not make address setting errors because there is nothing to set: the card address is determined by its physical position in the system. Note, however, that you must not confuse the bus-IN and bus-OUT connectors. Remember, the incoming A0 signal is inverted on every extension card, and swapped with A1 on the bus-OUT connector. This is done to enable any extension card to be selected when both A0 and A1 are logic low, although the actual address to be supplied by the PC to select a particular card is determined by the number of cards connected ahead of that card.

Those of you who have recalled from the earlier articles that the bus system has only four addresses may be surprised to see eight extension cards in Fig. 2. This is simple to explain. Any bus address can be read from, or written to. In other words, there are four 'read' addresses, and four 'write' addresses, which makes a total of eight. This difference is of no consequence as long as you do not wish to use more than four cards, which can then be chained via linking cables without problems. The difference between reading and writing is not in order until you use more than four cards. Let us assume that you wish to hook up four relay cards and four opto cards. This requires the positions with the same card number, e.g., 1 and 1', to be occupied by one relay card (write only) and one opto card (read only). You can not fit two

Fig. 1. Eight optocoupler inputs ensure that input signals can not cause damage to the computer.

![Diagram](image)

Fig. 2. The bus system can always accommodate four extension cards. However, it is also possible to fit up to four more cards, depending on their type (read or write function).
cards with 'write' functions, or two cards with 'read' functions, in a position with the same card number. If you still do so, the computer will forever be unable to 'see' the card with the accented (') number. Fortunately such an 'impossible' connection will not damage the hardware, because the first card addressed in the chain keeps the ENABLE signal for the rest of the chain logic high, whereby all other extension cards are disabled.

A different kettle of fish are extension cards with read and write functions (we have not published any of these, but you may have ideas ...). Such cards must always be fitted in one of the first four positions, but as far as possible towards the end of the chain. If a 'read/write' card is fitted in position '1', and three relay cards in positions '2', '3' and '4', it would appear that you can not fit an opto card up to position '2'. That will not work, however, since there is first card '1', but that position is blocked by card 1. The upshot is that first four positions must always be occupied by cards that can only be read from or written to. This leaves the next positions available for cards with the double 'read/write' function.

Summarizing the above:
- the address occupied by the card is determined by its physical position in the chain;
- at every address, a distinction is made between reading and writing;
- if an address is used for writing or reading only, the corresponding accented position may be occupied by a card with the complementary function only.

References:

COMPONENTS LIST

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<td>K2</td>
</tr>
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<td>Type 222 (Heddic)</td>
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Fig. 3. Track layouts (mirror images of the component side and the solder side) and component mounting plan of the double-sided PCB for the opto interface.
Synchronization circuit

The signals in a number of circuits of the processor must be matched to the input signal to prevent a wandering or distorted picture. The circuit responsible for this is the synchronization circuit, whose diagram is shown in Fig. 6.

The sync circuit is based on sync separator IC601. This circuit is fed with the VBS signal via T601 and provides three output signals: a sandcastle pulse at pin 7; a square wave at pin 3, whose first transition (leading edge) determines the onset of the horizontal sync pulse; and the vertical sync pulse at pin 10.

The signals at pin 3 and pin 7 are not yet suitable for further use. The sandcastle pulse contains as yet no information on the vertical blanking that is essential for colour demodulator IC101 (Fig. 4). Also, a signal that merely determines the onset of the horizontal sync pulse is not enough for further processing of the video signal. To make it into a true sync signal, a fly-back pulse is required. In fact, three pulses are needed: a horizontal and a vertical blanking pulse (12 µs and 1.9 ms respectively) and a horizontal sync pulse of 4.7 µs. These pulses are generated by monostables IC602, IC603a and IC603b, respectively.

The horizontal blanking pulse generated by IC603 (pin 6) is superimposed on the sandcastle pulse via R607. The modified sandcastle pulse has exactly the right shape to ensure correct functioning of IC101.

The reason for IC602 being fed directly by the sync separator, whereas both IC603a and IC603b are fed via potential dividers, is that IC603 operates from the 12 V supply line and the other two from the 5 V line.

Gate IC604 combines the horizontal and vertical blanking pulses into a composite sync signal. Note that the vertical blanking pulse is applied to the gate via XOR gate IC604a, because this signal is difficult to load owing to the impedance R617-R618.

Audio frequency circuits

To enable background music or a commentary to be added to the audio signal, the processor is equipped with a mixer and tone control.

The input of the audio circuits shown in Fig. 7 enables up to three signals to be mixed: the audio signal associated with the selected video source (master), a randomly selected audio signal (line) and a signal from a stereo microphone.

Selection of the master signal is facilitated by an input selector switch, consisting of analogue multiplexer IC404, and two electronic switches, IC405a and IC405b. The multiplexer selects four of the five possible audio signals, while the electronic switches take care of the fifth. Three audio signals are available from the SCART connectors in Fig. 3 and connector BU301 in Fig. 5. The other two arrive from the S-VHS connector and the BNC socket.
The line signal at sockets BU406 and BU407 is connected directly to the line control potentiometers on the mixer.

The level of the microphone signal is well below that of the line and master signals and is, therefore, amplified in IC401 (left-hand) and IC402 (right-hand).

The mixer consists of three linear stereo slide potentiometers, whose wipers are connected to the relevant left-hand or right-hand line via a summing resistor. The use of linear potentiometers may not seem right, but their wipers are loaded by the summing resistors in a manner that results in traditional logarithmic volume control.

Since not many video signals carry stereo sound, the audio circuit can be switched to mono by closing electronic switch IC405a.
which is controlled with switch TA401 and a bistable, IC403. The left-hand and right-hand signals from the mixer are then interconnected; their sum is seen by the remainder of the circuit as a mono signal.

The signal is then applied to two non-inverting opamps, IC410a and IC410b, and from these to control amplifier IC406. The control amplifier enables the volume, bass and treble, and the balance to be adjusted. The control range of the bass frequencies can be enlarged by using R436, C434 and C438 (marked with an asterisk). Without these components, the bass range is the traditional ±20 dB. The output of IC406 is available at output sockets BU408 and BU409.

The headphone amplifier enables the signal before and after the control amplifier to be listened to, depending on the position of electronic switches IC405a and IC405b. These switches are controlled by push button TA501 and the circuit based on IC501. The signals from the two sections of IC405 are applied to potentiometers R456 and R457, with which the volume and balance of the signal to the headphones can be adjusted. The signal to the headphones is amplified in dual opamp IC409.

**Power supply**

The circuit of the mains-operated power supply is shown in Fig. 8. The rectifier may look unfamiliar, but it is a straightforward bridge, whose negative connection is not directly to earth, but via three diodes. The drop across these diodes is used as an auxiliary negative supply (-2.5 V) for electronic switches IC405a and IC405b.

Fig. 8. Circuit diagram of the power supply.

Transistor T501 serves as the on/off control; it is operated by push button TA501 and the circuit based on IC503. That IC is fed by the potential across C501 to ensure that it can always be powered. That voltage is too high, however, and it is, therefore, lowered by zener diode ZD501. The output of IC503 is additionally protected against too high voltages by zener diode ZD502.

The circuit following switching transistor T501 is traditional: two three-terminal regulators ensure stable output voltages of 5 V and 12 V. The capacitors following the regulators are decoupling devices used at miscellaneous positions in the circuits.

The next and final instalment, dealing with the construction and calibration, will appear in our September issue (there is no August issue).
Many electronic instrumentation and data acquisition circuits must deal with low-level signals in the presence of strong interfering signals. If the signal level is small enough, even the noise produced by amplifiers and passive components can obscure the desired signal. In this article we will look at several strategies for solving problems with low signal level amplifier systems. These techniques include use of a low noise amplifier (LNA), filtering, circuit shielding, input leads shielding (including professional guard shielding techniques) and isolation of the circuit from the power mains.

By Joseph J. Carr

Noise, etc.

Noise can be defined as any unwanted signal, even though a somewhat narrower definition is sometimes sought in textbook treatments of the subject. But in the context of this article, noise can mean the internal ‘hiss-like’ noise generated in any amplifier, the atmospheric noise in radio receivers, 50 or 60 Hz hum picked up from the power mains, and interference from nearby sources of electromagnetic radiation (e.g., radio stations or other RF devices). Noise signals mix with, and either distorts or obscures the desired signals.

Several different forms of noise signal can be recognized: white noise, impulse noise and interference noise.

White noise supposedly contains all possible frequencies, so gets its name from analogy to white light, which contains all colours. Such noise is also called gaussian noise, although in reality it is neither ‘white’ nor ‘gaussian’ unless there are no bandwidth limitations placed on the system. True gaussian noise can be eliminated absolutely by low-pass filtering, because it by nature integrates to zero, given sufficient time. Bandwidth-limited noise, however, does not integrate to zero, but to a low value. The effect of low-pass filtering on pink noise is therefore not total reduction. An analogy to pseudo-gaussian or pink
noise is the 'hiss' heard between stations on an FM broadcast band receiver. Much of the noise in instrumentation systems is due to thermal sources, and has an RMS value of:

\[ U_n = \sqrt{4KTB\frac{R}{f}} \]  

Where:

- \( U_n \) is the noise signal in volts (V);
- \( K \) is Boltzmann's constant \( (1.38 \times 10^{-23} \text{ joules per Kelvin}) \);
- \( T \) is the temperature in Kelvin (K);
- \( B \) is the bandwidth in hertz (Hz);
- \( R \) is the circuit resistance in ohms (Ω).

Noise can be generated in a passive component such as a resistor by virtue of its resistance. According to Eq. (1), in a circuit with a 1,000 Hz bandwidth and a resistance of 100 kΩ, there is 0.6 microvolts (µV) of noise created by molecular motion due to temperature. Although this signal may appear to have a very low amplitude, keep in mind that many signals found in practical systems have the same order of magnitude. For example, in medical electronics, the electroencephalograph (EEG) machine records minute scalp potentials generated by the human brain's electrical activity, and may have components as low as 1 to 2 µV, with peak amplitudes in the 10 to 100 µV range. In that application, 0.6 µV represents a significant artifact especially when amplified 5,000 to 10,000 times, as is common practice in EEG machines.

Part of the solution to this type of problem is to keep circuit impedances in the early stages — i.e., those stages that most of the gain follows — very low so that the resistance term in Eq. (1) is reduced to a minimum practical value. Additionally, low-pass filtering, bandpass filtering or other methods might be employed to keep the bandwidth term low.

There are several sources of noise that are peculiar to solid-state amplifiers: shot noise, Johnson noise, and flicker noise. In some amplifiers these noise sources can add up to a significant amplitude. Although low-pass filtering offers relief, it is better to specify a low-noise amplifier for the earliest stages in the system.

Friis' equation uses the noise factors (i.e. ratio of input to output signal-to-noise ratio) to show us that low noise amplifiers in the input stages provide most of the noise relief for the entire system. It is for this reason that satellite communications or TV earth stations use Low Noise Amplifiers (LNA) as preamplifiers on the dish antenna. Similarly, analogue instrumentation and data acquisition amplifiers use a single LNA in the front-end, and then ordinary amplifiers throughout the rest of the circuit. The Friis equation for a cascade chain of amplifiers such as Fig. 1 is:

\[ N_{F_{\text{total}}} = N_{F_1} + \frac{N_{F_2}-1}{G_1} + \frac{N_{F_3}-1}{G_1 G_2} + \cdots + \frac{N_{F_n}-1}{G_1 G_2 \cdots G_{n-1}} \]  

Where:

- \( N_{F_{\text{total}}} \) is the noise factor of the entire cascade chain;
- \( N_{F_1}, N_{F_2}, \) etc. are the noise factors of the individual stages;
- \( G_1, G_2, \) etc. are the gains of the individual stages.

Thus, we can use a single, usually premium low noise amplifier device for the first stage, and regular amplifiers for all others. Low noise operational amplifiers are a good choice, but are sometimes rather expensive. A low cost alternative for many uses is the CA3130, CA3140 or CA3160 device in the 8-pin metal can package (not the mini-DIP!). Use a flexible heatsink of the type used for TO-5 metal transistor packages on the op-amp package, and operate the device from ±5 V dual polarity d.c. power supplies. This treatment (heatsinks and low power supply voltages) will mimic low-noise operation.

**Other noise problems**

Impulse noise is due to local electrical disturbances such as arcs, lightning bolts, electrical motors and so forth. Part of this same general type is general electromagnetic interference (EMI) problems. Such interference is usually caused by nearby radio transmitters, or other RF sources. It is not usually possible to force the transmitter off the air, even when it is an amateur operator, because they are licensed by the Government to be there ... while you are not.
Fig. 6. Differential instrumentation amplifier, based on operational amplifiers, using an active guard shield driver (amplifier A4).

From an engineering point of view, your equipment might be very expensive and quite good, and still be very poor from an EMI point of view. The purpose of any electronic equipment is twofold: a) it must respond to proper signals, and b) it must reject improper signals. It is point 'b') where most improperly designed equipment fails most significantly.

Shielding and filtering of signal lines is the key to EMI problems. Figure 2 shows a generic circuit with several of the possible correction types used. First, note that the entire instrument is built inside a shielded metal box, and the box is grounded. Points of entry and exit are passed through feedthrough 'EMI filter' capacitors. Feedthrough capacitors C1 through C3 have values of 50 pF to 2 nF (0.002 μF), depending on the circuit impedance and which capacitor is specified. For example, the signal line capacitors C1 and C2 will have smaller values, while power supply capacitor C3 should be larger than 1 nF (0.001 μF).

Each stage in Fig. 2 is isolated from other stages by a resistor, and has its own decoupling capacitor (C5 and C6). The main power bus is decoupled (C4a), and has a series radio frequency choke (L2) to prevent RF that gets past C5 from interfering with the operation of the circuit. The input leads are similarly filtered with L1 and C4.

The input resistance ($R_{in}$) of the amplifier and capacitor C4 also form a low-pass filter with a frequency response that rolls off at a -3 dB/octave rate from the -3 dB point defined by:

$$ F = \frac{1}{2 \pi R_{in} C_4} $$

Where:
- $F$ is the frequency in hertz (Hz)
- $C_4$ is in farads (F)
- $R_{in}$ is in ohms (Ω)

Not all of the techniques of Fig. 2 are needed, or even appropriate, in all circuits. Their inclusion was meant to show the possibilities, rather than form a recommendation for all applications. Select those that are appropriate, or practical, for your particular application.

### Suppressing local interfering signals

Local interfering signals are created by other electrical devices close to the circuit being operated, and by the 50/60 Hz electrical power mains in the building. Consider Fig. 3, where a low-level signal source is connected to an amplifier at the input of a larger circuit. The signal source might be a sensor such as a Wheatstone bridge strain gauge, an electro-optical detector. Alternatively, it may be a biopotential such as the EEG brain wave signal or electrocardiograph (EEG) heart signal.

The common factor shared by these signals is that they produce low level signals, and often must operate in a high interference environment.

A common solution to these problems is to use a differential amplifier at the input of the circuit. One of the properties of the differential amplifier is that its common mode rejection ratio (CMRR) tends to suppress interfering signals from the environment. It does this job because the inverting (−) and non-inverting (+) inputs offer equal gain, but are of opposite polarity. If identical signals are applied to the two inputs simultaneously, the net output voltage will be zero.

When a differential amplifier is used in a situation where it is connected to an external signal source through wires, those wires are subjected to strong local signals such as the 50/60 Hz a.c. fields from nearby power line wiring. Fortunately, in the case of the differential amplifier the field affects both signal equally, so the induced interfering signal is canceled out by the common mode rejection property of the amplifier.

### Guard shielding

Unfortunately, the cancellation of interfering signals by the input amplifier CMRR is not total. There may be, for example, imbalances in the circuit that tend to deteriorate the CMRR of the amplifier. These imbalances may be either internal or external to the amplifier circuit. Figure 4a shows a common sensor interface scenario similar to Fig. 3: a differential amplifier connected to shielded leads from the signal source, $U_{in}$. Shielded lead wires offer some protection from local fields, but there is a problem with the standard wisdom regarding shields: it is possible for shielded cables to manufacture a valid differential, but erroneous, signal voltage from a common mode signal.

Figure 4b shows an equivalent circuit that demonstrates how a shielded cable pair can create a differential signal from a common mode signal. The cable has capa-
citance between the-centre conductor and the shield conductor surrounding it. In addition, input connectors and the amplifier equipment internal wiring also exhibits capacitance. These capacitances are lumped together in the model of Fig. 4b as C1 and C2.

There are also resistances in the circuit. The signal source resistances $R_{S1}$ and $R_{S2}$ are generally low, but in some cases (e.g., EEG, ECG, pH electrodes, optoelectronic sensors, etc.) they may be quite high. In addition, there are also input impedances, both differential and unbalanced to ground (see Fig. 4a).

As long as the sum circuit resistances are equal, and the two capacitances are equal, there is no problem with circuit balance. But inequalities in any of these factors (which are commonplace) creates an unbalanced circuit in which common mode signal $U_{cm}$ can charge one capacitance more than the other. As a result, the difference between the capacitance voltages, $U_{C1}$ and $U_{C2}$, is seen as a valid differential signal by the amplifier.

A low-cost solution to the problem of shield-induced artifact signals is shown in Fig. 5a. In this circuit, a sample of the two input signals are fed back to the shield, which in this situation is not grounded. This type of shield is called a guard shield circuit. Either double shields (one on each input line) as shown in Fig. 5a or a common shield for the two inputs as in Fig. 5b, can be used.

An example of guard shielding for the standard three op-amp instrumentation amplifier, a very common differential front-end for electronic instrument circuits, is shown in Fig. 6. The instrumentation amplifier consists of $A_1$, $A_2$, and $A_3$, with associated resisters. If $R_2=R_3$, $R_4=R_5$ and $R_6=R_7$, the voltage gain of the circuit is given by:

$$\frac{A_v}{R_1} = \frac{10k\Omega}{R_1} \left( \frac{R_6}{10k\Omega} \right)$$

(All resistance in kilo-ohms)

In Fig. 6, the gain can be set by selecting values for $R_1$ and $R_6$, which implies also a value for $R_7$ (which is equal to $R_6$). Variable gain control is provided by making $R_1$ variable. Keep $R_6$ away from zero ohms, however, or the gain will get very high very quickly.

In the circuit of Fig. 6, a single shield covers both input signal lines, but it is possible to use separate shields. In this circuit a sample of the two input signals is taken from the junction of resistors $R_3$ and $R_5$, and fed to the input of a unity gain buffer-driver 'guard amplifier' (A4). The output of A4 is used to drive the guard shield.

Perhaps the most common approach to guard shielding is the arrangement shown in Fig. 7. Here we see two shields used: the input cabling is double-shielded insulated wire. The guard amplifier drives the inner shield, which serves as the guard shield for the system. The outer shield is grounded at the input end in the normal manner, and serves as an electromagnetic interference suppression shield.

**Power line noise**

Another potential source of interference is noise and EMI signals arriving on the a.c. power mains. I can recall digital instrumentation and computers in a medical school building that acted in a schizophrenic manner until it was identified that the a.c. power mains were the source of the problem.

A humorous event while this problem existed came about when the medical (M.D.) and medical sciences (Ph.D. and D.Sc) students took the standard multiple choice national examination in human physiology. They used a 'mark-sense' answer sheet on which they use a pencil to darken the letter corresponding to the printed candidate answer they believe is correct. These papers were then taken to an optical scanner that inputs the answers to a computer. While the scanning was going on one year, some ac power line switching equipment started operating, sending high voltage transients over the mains. The result was that the entire freshman class of medical and sciences students flunked the national exam!

Where sensitive scientific instruments are used, one might want to consider designing the ac electrical power mains system to be either isolated from the building system, or having a separate system that keeps a separated neutral and ground conductor all the way back to the service entrance of the building.

Figures 8 and 9 show methods for dealing with severe power mains noise. In Fig. 8 we see an L-C power line filter wired in the North American standard manner. These filters are shielded low-pass filters, and are mounted inside of equipment as close as possible to the point where a.c. enters the cabinet. Some filters are available molded into the a.c. chassis connector. Exterior to the filter is a Metal Oxide Varistor (MOV) device used to suppress a.c. line transients above the normal peak a.c. voltage (some high voltage transients can reach 2000 V for 30 μs).

The transformer in Fig. 9 performs two functions. First, it isolates the equipment electrical system from the mains electrical system. Second, it frequency limits the system to prevent high frequency transients and pulses from passing into the equipment. It is my opinion, shared by many other engineers, that no computerized or other digital equipment — and many types of analogue equipment — should be operated in a noisy environment without one of these transformers. If the equipment is life-support, or life-saving, as it often is in medical applications, then it is probably engineering malpractice to design a piece of equipment without the transformer.

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**Fig. 8. Shielded LC EMI filter for the a.c. power mains (wiring shown common in North America).**

**Fig. 9. Line isolation transformer used with digital instruments, analogue instruments and computers to eliminate high voltage transients, mains voltage fluctuations and other problems. This transformer is manufactured by Topaz in the USA.**
A.F. DIGITAL-TO-ANALOGUE CONVERTER

PART 1

Design by T. Giesberts

Twenty-bit converters, ×8 oversampling and high-quality analogue stages are used to translate the digital output of CDs, DAT recorders or radio broadcasts into the desired analogue information as accurately as possible. The converter is a stand-alone unit that is eminently suitable for use with the CD player published earlier this year.

During the past decade, digital techniques have assumed an increasingly important role in audio engineering. Hailed, somewhat exaggeratedly, as 'perfect' and 'ideal', quite a few improvements have been found necessary since those early years in appliances using these techniques. No doubt, others will be found indispensable over the next few years. There is also a growing band of audio pundits who foresee the end of both the CD and DAT before the year 2000.

Be that as it may, at present, the CD player is the second most important unit in the audio chain (world-wide, the compact cassette is still way ahead as the most popular music medium). It is well known that the output sections of that unit play a vital role in producing a near-perfect reproduction of the original sound. Unfortunately, a CD player cannot be upgraded as easily as a record player in the past. Then, a better quality tone arm or stylus could be added without any trouble. In a CD player, the only improvements possible are in the digital or analogue sections and they are not so easily implemented. There are, of course, two other possibilities: buy a new CD player or an add-on digital-to-analogue converter—DAC. Neither of these is a simple solution, although in the case of the add-on unit, it should be borne in mind that it can be used with a number of different appliances.

Furthermore, there are not that many commercially available stand-alone DACs on the market, probably owing to their high price; at present this can vary from a few hundred to a few thousand pounds. A build-yourself design was until now not really feasible owing to the non-availability of certain parts and components. The design presented costs about one third of a commercially available unit with near-identical specifications: £250-£400.

The design

The digital-input selection in the block schematic in Fig. 1 accepts four digital signal sources, which may be connected by fibre-optic or coaxial cable. The tape select stage enables any one of the four signals to be applied to the digital tape output.

![Fig. 1. Block diagram of the digital-to-analogue converter.](image-url)
The selected input signal is applied to a special Yamaha IC Type YM3623. The circuit of this chip is shown in Fig. 2. Its PLL (phase-locked loop) produces a clock from the input signal for the subsequent stages. The range of the PLL is wide enough to enable the processing of all current sampling frequencies. In the absence of an input signal, the IC’s crystal oscillator generates a stable frequency for the digital filter and the actual DAC stages.

The most important task of this chip is, however, the analysis of, and error detection (Philips-Sony format) in, the audio data of the incoming signal before this is applied to the digital filter.

Basically, the data provided by the IC could be processed directly by the DAC, but that would create problems in the analogue section, because the sampling frequency must be sufficiently suppressed there without introducing amplitude and phase errors in the audio range. That introduction is combated by oversampling, which involves the computation by a digital filter of intermediate steps that cause the sampling frequency to be shifted upwards artificially. The more intermediate steps, the higher the sampling frequency, and the easier the design of the analogue filters.

The digital filter, a Burr-Brown Type DF1700 dual channel type, provides x8 oversampling, which means that each sample is converted into eight discrete levels. These levels make possible smaller steps than the various original levels. The x8 oversampling converts the input data frequency of 44.1 kHz (with CD reproduction) to 352.8 kHz. This means that a third-order filter can be used in the analogue section, which reduces filter phase non-linearities.

The interposition of additional steps and the computation of the intermediate values increase the resolving power at the output of the digital filter compared with that of the 16-bit input. If a converter with an accuracy of more than 16 bits is used, the increased resolving power is retained in the conversion, so that in the present design, the converter processes 20 bits. This is, by the way, the maximum resolution the DF1700 can provide at its output. The pass-band ripple of the DF1700 is <0.00005 dB.

Although one-bit converters are currently in fashion, they do not really give satisfactory results for top-of-the-range equipment. In the present design, a 20-bit monolithic IC Type PCM63P from Burr-Brown was chosen. Burr-Brown supplies many manufacturers of CD players and currently this chip is their top multibit converter.

The PCM63P—see Fig. 4—contains two 19-bit converters: one each for the positive and negative halves of the signal. This design has the advantage that it prevents bipolar zero distortion (traditional DACs usually switch the most significant bit around the bipolar zero, which may give rise to glitches and non-linear distortion). To ensure exact synchronization of the two converters, they use the same R-2R ladder network and the same reference voltage. Because of the 20-bit conversion, the harmonic distortion is lower:

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**Fig. 2.** Block schematic of Yamaha’s digital audio interface receiver Type YM3623.

**Fig. 3.** Block schematic of Burr-Brown’s Type DF1700 dual-channel digital filter.

**Fig. 4.** Block schematic of Burr-Brown’s 20-bit monolithic audio DAC.
The analogue output section consists of a very fast current-to-voltage converter following the DAC and a third-order low-pass filter in which no opamps are used. Its design is known as a Generalized Impedance Converter-GIC. In this, an opamp configuration to ground acts as a second-order frequency-dependent passive element. This has the advantage that the audio signal does not have to pass through an opamp.

The analogue signal is fed to the output via a buffer stage. Relays short-circuit the output when a high level of noise is present or when the de-emphasis circuit is switched in, if the input signal makes that necessary.

The design of the power supply ensures complete separation of the digital and analogue sections, and it, therefore, used two transformers. No fewer than ten regulator ICs ensure optimum supply quality.

### The circuit

The digital signal from the CD player, DAT recorder, or DCC player enters the circuit in Fig. 5 via R1. Resistor R1 ensures correct termination of the coaxial input cable to obviate possible reflections. The bi-phase signal is enhanced by two inverters, IC1a and IC1b. With the aid of R2 and R3, the former is arranged as an analogue amplifier that raises the 500 mV signal six-fold. Inverter IC1b produces a pure TTL signal with improved transitions (edges). The signal is then applied to the Digital Audio Interface Receiver-DIR-IC2.

When the digital input selector (to be published shortly) is used, the input signal is applied from that selector to A—resistors R1, R2, and R3, capacitor C1 and IC1 can then be omitted.

The YM3623B requires only few external components. Crystal X1 provides a stable output frequency in the absence of input data; the internal PLL is then switched off. Capacitor C4 and resistor R6 form the integrator network for the VCO-voltage-controlled oscillator—of the internal PLL. The values of these components have been chosen to ensure that the 32–48 kHz frequency range is scanned with the minimum of phase jitter. Network R5-C3-D6 resets the clock switch-over circuit in the IC at power-on. The supply line is decoupled by R9-C7-C8.

Note that the crystal frequency is usually chosen to give an oscillator frequency of exactly 44.1 kHz (after scaling). In practice, it transpired that that created a lot of interference in the PLL, since both operate at about the same frequency. Therefore, a crystal frequency that is not a multiple of the sampling frequency was chosen. This does not detract from the operation of the circuit, because the oscillator is in any case used only as an emergency frequency source for the internal logic of the IC and subsequent circuits in the absence of input data. The crystal frequency may be 16–20 MHz.

The YM3623B provides, apart from the audio data, also additional information contained in the bi-phase signal. For instance,
Fig. 5. Circuit diagram of the audio-frequency digital-to-analogue converter.

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outputs S1 and S2 carry the current sampling frequency. That information is made visible with the aid of IC2a on one of three LEDs: 32 kHz; 44.1 kHz; 48 kHz.

The copy bit contained in the information is not made visible, because it is not used by any manufacturer.

The ERR(or) output indicates when an error has been detected in the data input. The signal available at this pin is lengthened sufficiently by pulse stretcher R3-C3-D7 to switch on T2, switching so that D5 lights. The ERR signal is also used to energize relays Re2 and Re4 via T3, whereupon the analogue outputs are switched off. Transistors T2 and T3 are MOSFET types to keep the switching currents on the printed-circuit board low.

The DEF output signals the presence of pre-emphasis on the input, whereupon the de-emphasis network and T1 are switched on via Re1 and Re3. As soon as T1 conducts, D4 lights.

Subsequently, the audio signal is fed from pin 17 of IC2 to digital filter IC. This IC needs clocks to function correctly: one is derived from the bi-phase signal and is available at pin 8 of IC, while the other is the timing signal for writing serial data and is available at pin 12 of IC. It also needs a multiplex signal that indicates whether the current data are for the left-hand or right-hand channel; this signal is fed from pin 15 of IC2 to pin 28 of IC.

The supply lines to the filter are decoupled by R14-C10 to C11.

After the filter has translated each data word into eight new values, the DACs can be driven by these data (DOR=Data Out Right; DOL=Data Out Left). Again, some control signals are needed: bit clock (BCKO) and word clock (WCKO).

The RC networks inserted into the lines from the filter to the DACs filter out any RF interference and noise signals.

As an aside, assuming that a CD signal is input, the system clock from IC0 to IC13 is 16.9344 MHz—the L/R clock is, of course, 44.1 kHz. The clock (BCO) for writing serial data is 2.8224 MHz. The BCKO clock and WCKO clock between the digital filter and the DACs are 8.4672 MHz and 352.8 kHz respectively. It is clear that these are all RF signals, and it is imperative to keep them—and their harmonics—away from the analogue section.

Presets P1 and P2 (P3 and P4) enable the setting of the MSB—most significant bit—of each 19-bit converter in IC5 and IC13. Precision test equipment is required for this, however; if that is not available, the presets can be omitted. The ICs are available in three versions indicated by (a) no letter after the type number (least expensive); (b) a J after the type number; (c) a K after the type number (most expensive).

Since the divorce of the analogue and digital sections comes about in the DACs, attention must be paid to the power supply. As already stated, the supplies for the two sections are completely separate. All supply connections are decoupled independently. Moreover, on the relevant printed-circuit board, the earth connections for the two
sections have been kept separate.

The output of the converters is applied to a current-to-voltage converter, IC6 and IC14, for the left-hand and right-hand channel respectively. These are needed, because virtually all converters provide an analogue output current rather than a voltage. Undistorted conversion of the voltage waveform (analogue signal with superimposed steps on the scanning frequency) requires an IC with a very high slew rate. That requirement is met by Analog Devices' Type AD844, an opamp with a bandwidth of 60 MHz and a slew rate of no less than 2000 V/µs.

The output filter is a third-order pseudo passive design, that is, there are no active components in the signal path. It consists of R27-C36 followed by the GIC configuration of IC1 and IC6, which is followed in turn by passive section R27-C36.

The emphasis network is entirely passive and consists of R23-R24-C31. It is switched into circuit by relay R21, as soon as this is energized, whereupon R23 is short-circuited (left-hand channel).

Finally, buffers IC9 (left-hand) and IC1; (right-hand) provide a high load for the low-pass filter on the one hand, and sufficient current for driving a (pre)amplifier at the output of the converter.

The impedance of the analogue outputs is 50 Ω (R21 and R28). Resistors R31 and R50 also make it possible for the mute relays, R5 and R6, to short-circuit the outputs to earth without detriment. Energizing of these relays is delayed by the circuit based on T6. This gives the entire circuit time to settle in after power-on before any relays are energized. In the case of error indications, the mute relays are de-actuated off via T5.

There is a wide choice of opamps for the low-pass filter and buffer stages. Ten different models were tried in the prototypes, but their performance was more or less uniform. The parts list (next instalment) gives an NE5534A, which was the least expensive of the ten. Note also, that there are commercial CD players costing over £1000 that also use the NE5534A.

It will have been observed that the entire analogue section is DC coupled and also that there is no output capacitor. To ensure that the direct voltage at the output is always exactly zero, a servo control, IC10 and IC16 respectively, has been added between the output of the buffer opamp and the non-inverting input of the current-to-voltage converter opamp. Although the off-set current of the DACs is virtually zero, this arrangement ensures that very-low-frequency signals, as well as the off-sets of the various opamps, are nulled.

The servo control is an integrator that monitors the output voltage of the buffer, and the basis of which it adjusts the direct voltage of the D/A converter so that the output voltage remains zero. The control has absolutely no effect on signals in the analogue circuits above 10 Hz. To ensure purity of these signals in the servo control, which is after all a sort of high-pass filter with low cut-off frequency, all capacitors are MKP (metal-plated polypropylene) types.

The power supply—see Fig. 6—is clearly in two parts. That for the digital section delivers +5 V (B1, IC1, and IC3). In that for the analogue section, rather more components are used, because it was thought important to give the left-hand and right-hand channel its own regulation. Bridge rectifier B2 is followed, therefore, by two positive and two negative regulators. A 'star' earthing point obviates any problems with earth currents.

There is a separate 12 V supply, provided by D1 and D2, for the mute relays. The value of capacitor C35 has been kept small to ensure that the mute relays are de-energized the instant the power is switched off.

Each of the supplies uses its transformer. It is, of course, possible to use one with split secondaries, but that will have to be wound to order. In the prototypes, toroidal transformers were used, which keep the stray field to an absolute minimum.

The second and final part of this article, describing the construction of the converter, will be published in our September issue (there is no August issue).
THE history of the modern capacitor goes back to antiquity when the Greeks made a study of electrification of amber by friction. After that, very little more about the subject of electrostatics emerged until the middle of the eighteenth century when, in 1746, the Dutch physicist Pieter van Musschenbroek (1692–1761) discovered, by accident, the principle of the capacitor in the form of the celebrated Leyden jar. Having received a powerful electric shock from his experiments, arising from an attempt to electrify water in a bottle, van Musschenbroek confessed that he would not take such another shock ‘for the kingdom of France’. Many modern experimenters have, no doubt, uttered words to the same effect when carelessly handling circuits containing charged capacitors.

We have seen in the previous parts of this article that inductive and conductive effects depend upon the magnetic field; here we shall see that capacitance and conductive effects depend upon the electric field. The existence of an electric field depends itself on the presence of electric charges. When the charges are removed, the field vanishes. The nature of the field and the direction taken by the electric flux is a function of the magnitude of the charges and their distribution.

The fundamental electric charge, e, resides in the electron, which carries a charge of \(-1.6\times10^{-19}\) coulomb (C). Electrons may be added to a body so as to give that body an excess of electrons; the body then exhibits a negative electrostatic charge. In the same way, electrons may be removed from a body, giving it a deficit of negative charges; the body is then positively charged. In normal circumstances, bodies have neither an excess nor a deficit of electrons and are uncharged, neutral or at ‘earth’ potential. So the general mass of the earth is a permanently neutral body and materials may carry charges in the form of an excess or a deficit of electrons with respect to earth. There should be no confusion here with the ‘excess’ or ‘deficit’ of charge carriers in semiconductor materials, where the crystal remains electrically neutral.

As the Greeks established, without being aware of what they were handling, the removal from, or the addition to, a body of electrons may be accomplished by purely mechanical means, as when a piece of amber is rubbed with silk, or more practically for our purposes today, by applying a potential difference between conducting materials.

A field of electric force, like a magnetic field, may be represented in magnitude and direction by drawing lines of force in the region surrounding the charged body or between charged bodies. The line densities being an indication of the field strength. The relationship between these convenient though imaginary lines and the field strength E (a vector) is that the tangent to a line at any point gives the direction of E at that point and the number of lines per unit cross-sectional area (that is, the flux density, D, in C m\(^{-2}\)) is proportional to the magnitude of E. A number of typical and idealized fields for isolated and adjacent charges is shown in Fig. 1. Unlike the magnetic field, electric lines of force do not form closed loops, but are taken to emanate from positive charges and terminate on equal and opposite negative charge. When, therefore, a charged body is brought into proximity of another uncharged body, an induced charge of opposite sign appears on the near surface of the body, resulting from the line termination points. Hence, a net attraction is set up between the bodies because unbalanced forces act on the induced surface charges as illustrated in Fig. 2: a phenomenon known to the Greeks when their charged pieces of amber attracted wisps of straw.

It was believed at one time that a metallic body could not be given a charge by frictional means as could bodies of non-conducting or insulating materials; however, provided that the metal is supported or held by an insulating substance, such conductors can be charged just as well as anything else. In metals, only the negative charge is free to move; the positive charge is as immobile as it is in glass or any other insulator.

The electric field

To get a proper understanding of the phenomenon of capacitance, we need a proper appreciation of the field concept. In both magnetic and electric systems, energy is stored in the field, not in the component parts producing the field. Before Faraday’s time, the force acting between charged particles was thought of as a direct and instantaneous interaction between the particles, and Faraday himself always thought of the field in terms of lines of force. This concept still provides us with a convenient way of visualizing field patterns as we have seen, but it is really necessary to think in terms of charge acting on a field or of a field acting on a charge, and not as charge acting upon charge as the action-at-a-distance concept would have us suppose. Look at Fig. 1c for a moment; suppose particle A carrying the positive charge suddenly moved to the left; how soon after this would the charge on particle B learn that A has moved and that the force of attraction it has so far experienced decreases? If action-at-a-distance were true, the information would be transmitted instantaneously to particle B, but this does not accord with commonplace experience. Moving charges in the aerial system of a radio transmitter, for example, establish an electromagnetic field and so influence electrons in a distant receiving aerial system, but only after a finite time, determined by the distance travelled and the speed of light.

When the flux density, D, changes, the electric force E changes proportionally and there is a constant relationship for a field established in air, the ratio D/E is designated \(\varepsilon_0\). This is the permittivity of free space or the free space constant. Its value is found ex.

**Fig. 1.** Some typical electrostatic field configurations using the concept of lines of force.
and tend to align it with the electric field.

This constant, that is, the capacity of the conductor, is called the capacitance. The m.k.s. unit of capacitance is the coulomb/volt, more commonly called the farad (F), named in honour of Michael Faraday. Thus, one farad = one coulomb/volt; that is, 1 coulomb of charge raises the potential of unit capacitance by one volt. The farad is a very large unit, and in practice the microfarad, μF (10⁻⁶ F), the nanofarad, nF (10⁻⁹ F), and the picofarad, pF (10⁻¹² F) are used, although memory back-up values of several farads are now commonplace.

How might the capacity of an inductor be increased? What follows is true for conductors of any shape, but to make the explanation easier, it is assumed that a conductor in the form of an isolated metal plate, A, carries a positive charge, q+, coulomb, giving it a positive potential. In Fig. 3a, this potential is represented by the line U+. The line marked U represents a neutral plane or earth line. Suppose a second metal plate, B, carrying a negative charge, q-, coulomb, also isolated, has a corresponding potential, U-, represented by the line U-.

The potential difference between the plates is clearly U=U+-U-. This is proportional to the charge on either plate.

Let the two plates now approach each other; the charge on the plates will be unaffected, but what happens to their potentials? If a positive charge is brought near to an isolated conductor, the potential of that conductor will be lowered, since a negative charge will be induced on it: in the same way, the proximity of a negative charge will raise the potential. Thus, the potential of the positively charged plate A, will be lowered by the nearby presence of the negatively charged plate B from U+ to some lower value U+. Similarly, the potential of plate B will be raised from U- to some higher value -U-. These new potentials are shown in the figure: the changes are indicated by the vertical arrows.

The same effect can be produced if a single plate, carrying, say, a positive charge q+, is approached by an isolated neutral plate as shown in Fig. 3b. Lines of force from the charge on A will terminate on the inner surface of plate B and an induced negative charge will be established there. The inner plate surfaces now carry equal but opposite charges. The induced charge on B produces its own field and, in the same way as described above, the potential of A consequently falls. Hence, the 'capacity' of A has been effectively increased by the presence of plate B. We conclude that the potential difference between two conductors that carry constant, equal charges of opposite sign is reduced as the conductors are brought closer together.

Devices that operate on these principles are capacitors; perhaps their old name of 'condenser' was not so inappropriate when it is considered that the electric field is concentrated by such means and so made capable of storing additional electrostatic energy.

### The parallel plate capacitor

The parallel plate is the most basic of all capacitor designs: all other varieties are simply adaptations of it. Suppose two parallel plates are connected into a circuit as shown in Fig. 4. Starting with plates A and B uncharged, let a voltage be applied by the closing of switch S. A positive charge +q then appears on the left-hand plate and a negative charge -q on the right-hand one, so that finally the potential difference between the plates is equal to the applied voltage, U. The plates are then charged.

For these charges to be established, there must have been a movement of electrons around the circuit; that is, a current must have flowed in the direction A to B for the time during which the equalization of the voltages was attained. This displacement current can be detected on an ammeter wired in series with the circuit. If the battery is switched off, the charge on the plates persists, as does the potential U across the plates. Thus, the capacitor stores electrical energy. It should be noticed particularly that q coulombs is the quantity of charge on either plate; it must not be taken as the net charge on the capacitor, which is zero.

If the plates are now connected together by a piece of wire, the capacitor will discharge: a momentary displacement current follows from plate B to plate A to restore the neutral condition of the plates and reduce the terminal voltage to zero. No current passes through the capacitor; there is simply a movement of electrons away from the positive plate and towards the negative plate; and these return in the opposite direction when the device is discharged.

### Effect of a dielectric

When a slab of insulating material is placed in the space between the plates of a capacitor, it is found that the capacitance is increased. This comes about because the molecules of the dielectric, as the insulator is called, have what are known as electric dipole moments which may be permanent in some material and tend to align themselves with an applied electric field, as illustrated in Fig. 5. The dipoles have random orientations in the absence of an external field, but experience a torque tending to align them with the field when this is applied. Complete alignment does not occur, because of ther-
nal agitation of the molecules, but the alignment increases as the field is increased or the temperature falls. Molecules that do not have permanent dipole moments will, nevertheless, acquire them when subjected to an electric force. The field tends to separate the negative and positive charges on the molecule, so creating an induced dipole that will tend to alignment only when the field is present.

When a slab of dielectric is introduced between the plates of a capacitor, assumed to carry a fixed charge \( q \), the effect of dipole alignment is to separate the centre of positive charge of the slab slightly from the centre of negative charge. The dielectric, although remaining electrically neutral, becomes polarized. The overall effect is the appearance of positive charges on one surface of the slab and of negative charges on the other. These charges must be equal in magnitude; within the slab itself, there is no displacement of electrons and no transfer of charge over large distances. As Fig. 6 shows, the induced surface charges will always be established in such a way that the electric field they themselves set up, \( E_1 \), will oppose the applied field. The resultant field in the dielectric, \( E_2 \), is then the (vector) sum of \( E \) (the applied field) and \( E_1 \), and this is always smaller than \( E \). We conclude that, owing to the presence of a dielectric in a field, induced surface charges tend to weaken the original field. This weakening shows itself as a fall in the potential between the plates; hence, the capacitance, for a constant \( q \), must be increased.

Faraday first investigated the effect of a dielectric on the capacitance of a conductor. In an experiment similar to that shown in Fig. 7, two capacitors were charged to the same potential, \( U \), by the battery. Unlike the argument made earlier, in this system \( U \) does not change, so the charge \( q \) on the capacitor with the dielectric must increase above that on the air-spaced capacitor, which Faraday found was so. Thus, since \( q \) is greater for the same potential and \( C=q/U \), the capacitance must increase through the introduction of a dielectric. If we give the capacitance with dielectric the symbol \( C_d \) and that without a dielectric \( C_0 \), the ratio \( C_d/C_0 \) is called the dielectric constant or, more generally, the relative permittivity, \( \varepsilon_r \). The dielectric constant is unity for air spacing (strictly, for a vacuum), which, as we have noted, has a free space permittivity \( \varepsilon_0 \). The absolute permittivity when a dielectric is used is \( \varepsilon_0 \varepsilon_r \). Hence, the capacitance of any capacitor can be expressed as \( C=C_0 \varepsilon_r/\varepsilon_0 \), where \( L \) depends on the form of construction and has the dimensions of length. For a parallel plate capacitor, \( L \) is \( \frac{d}{2} \), where \( d \) is the plate separation.

**Energy and losses**

When a capacitor is charged to potential \( U \), energy is stored in the electric field established between the plates. When the capacitor is discharged, the field energy is returned to the external circuit in some way, often as a spark that generates heat and light. For a perfectly efficient capacitor, there would be no energy loss and all the charge put into the system would be returned by it. In a dielectric, work is done in turning the molecular dipoles, and if the capacitor is charged and discharged periodically, this process causes heat to be generated in the dielectric which represents energy loss; this loss can be represented as a small resistance in series (or a large resistance in parallel) with a loss-free component. It is usual to express such a loss in terms of the angle by which the lead of the current on the voltage falls short of 90°.

The energy stored in a capacitor is potential energy, which is dependent on the potential difference set up across the plates; in an inductor, it is comparable with kinetic energy associated with the movement of electrons in a wire. It is easily shown that the energy stored in a capacitance \( C \) is given by \( CU^2/2 \) joules.

**Time constant**

Since current is the rate of change of charge, \( i=\frac{dq}{dt} \), the charge on a capacitor cannot change instantaneously, because that would require an infinite current and hence an infinite rate of change of voltage. This does not accord with experience. In effect, since the capacitance is constant, the voltage across a capacitor must momentarily remain the same before and after any abrupt change in the

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Fig. 6. (a) The applied field \( E \) separates the centre of +ve charge from that of -ve charge. (b) The surface charges set up a field \( E_1 \) which opposes the applied field \( E \). The resultant field \( E_2=E-E_1 \) is thus weaker than \( E \).

Fig. 7. The principle of Faraday's experiment to show that a dielectric increases the capacitance.

Fig. 8. The meaning of time constant for the charging and discharging of a capacitor through a resistor.
The voltage across C cannot rise immediately to the level of the applied voltage, $U$, when switch $S$ gets closed. In fact, it remains at zero. So, the capacitor behaves as an instantaneous short-circuit and the applied voltage appears across $R$. This situation cannot remain, however. $C$ begins to charge with a consequent rise in its terminal voltage and the potential across $R$ falls correspondingly. The charging current will, therefore, be reduced from its initial value of $i = \frac{U}{R}$ to $i = \frac{(U-U_c)}{R}$. That is, as the voltage across $C$ rises, the current and the rate of charging falls. Since $dq = idt = C\frac{du}{dt}$, at the onset of charging, the initial current, $i = \frac{U}{R}$; hence, $UIR = C\frac{du}{dt}$, and the initial rate of rise of voltage is $\frac{U}{t}R$ V s$^{-1}$. If this rate could continue unopposed, the voltage on $C$ would reach its final value $U$ in a time given by $t = \frac{U}{IR} = CR$ seconds as shown in Fig. 8. One important point arises here: if the charging current is kept constant, the charging curve will be linear. This fact is used in circuits where a voltage with a linear sawtooth waveform is required.

With the existing situation, of course, the voltage across $C$ continually rises and opposes the supply voltage, thus making the charging cycle non-linear. At any particular instant, the capacitor still has to be charged $(U-U_c)$ volts, and if this in turn were to continue at a constant rate, the time for the completion of the charge would be $U - \frac{1}{2} U_c = \frac{1}{2}U_c = CR$ seconds as before. Hence, at any point on the charging curve, the time remaining to complete the charge at a linear rate is $CR$ seconds. In theory, then, the capacitor can never be charged completely.

Figure 8 shows the actual form of the charging curve which is exponential in form, just as the rise of current in an inductor was seen to be in the first part of this article.

The product $CR$ has the dimensions of time, and it is not too difficult to show that in a time equal to $CR$ seconds, the charge will have reached a level given by $U_c = 0.63U$.

The curve also shows the discharge condition, assuming that the charged capacitor discharges through the same resistance. Again, if the initial rate of fall in the capacitor voltage were to continue at a constant rate, the cycle would be completed in $CR$ seconds, and this would be true for any point on the curve. Notice that the level actually reached in $CR$ seconds is $0.37V$.

The product $CR$ is known as the time constant of the circuit and is a very important and fundamental aspect of capacitance principles. Note that the time constant is normally given the symbol $\tau$, thus $\tau = CR$.

To summarize: the electronic age could not exist without capacitors and inductors. Both of these passive components are used in conjunction with other each other and with other devices to produce tuned circuits, oscillatory systems, smoothing circuits, the transmission of signals and to provide time delays, to name but a few. Perhaps these brief studies will afford the 'passives' a little more respect than they usually get.

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**ACOUSTIC CRYSTAL TESTER**

A crystal cannot be tested acoustically, unless, that is, its output is scaled down to the audio frequency range by a circuit as shown in the diagram.

A divider that is particularly suitable for this purpose is the Type 4060 CMOS IC. This circuit contains not only a 14-stage binary scaler, but also a complete oscillator.

The crystal to be tested is connected across the input terminals and $S_2$ set as indicated in the table. The crystal frequency is scaled down in $IC_1$ and, depending on the setting of $S_2$, one of the outputs of the 4060 drives transistor $T_1$ via $R_3$. The transistor, in turn, drives a small loudspeaker. $L_3$. The power delivered to the speaker is limited by $R_5$ to prevent damaged eardrums.

<table>
<thead>
<tr>
<th>Position of $S_2$</th>
<th>Crystal Frequency</th>
<th>Oscillator Frequency</th>
<th>Time Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>$&lt;1\text{ MHz}$</td>
<td>$32,768\text{ kHz}$</td>
<td>$32,768\text{ kHz}$</td>
<td>$256\text{ Hz}$</td>
</tr>
<tr>
<td>$1-10\text{ MHz}$</td>
<td>$1\text{ MHz}$</td>
<td>$1\text{ MHz}$</td>
<td>$244\text{ Hz}$</td>
</tr>
<tr>
<td>$&gt;10\text{ MHz}$</td>
<td>$27,145\text{ kHz}$</td>
<td>$9,048\text{ kHz}$</td>
<td>$1,051\text{ Hz}$</td>
</tr>
</tbody>
</table>

It is, of course, not possible to use one scale factor for all sorts of crystal, and that is why switch $S_2$ enables selection of one of three different factors. For crystals $<1\text{ MHz}$, the scale factor is 128; for crystals in the range $1-10\text{ MHz}$, the scale factor is 4096; and for crystals $>10\text{ MHz}$, the scale factor is 8192.

Also, crystals operating above $10\text{ MHz}$ oscillate readily at somewhat higher voltages than low-frequency ones. That is why $S_2a$ and $D_2$ lower the supply voltage to $4.7\text{ V}$ when crystals below $10\text{ MHz}$ are tested.

If a mains-operated power supply is preferred over a $9\text{ V}$ battery, a $12\text{ V}/50\text{ mA}$ one is recommended. In that case, $D_2$ must be a $6.8\text{ V}$ type.

[K.H. Lorenz - 924042]
Many people intending to go on a camping or caravanning holiday will appreciate this 600 VA converter that provides a standard 240 V 50 Hz a.c. supply from a 12 V vehicle battery.

A perennial problem in the design of an inverter is the waveform of the output voltage. A sine wave is, of course, the ideal, but unfortunately the losses in the inverter are then unacceptably high. It is far better to design a rectangular waveform output for which the losses are much smaller. That in turn allows a compact design that can deliver a fairly high power. This has, however, the serious disadvantage that virtually all appliances that are to be powered by the inverter require a more or less sinusoidal supply.

In the present design, a compromise between these opposites has been reached: the output voltage has a trapezoidal waveform. This is near enough to a sine wave to enable standard domestic appliances to operate from the inverter and, moreover, it does not appreciably add to the cost or size of the unit. Repeatability and reliability were two important aspects of the design. Efficiency and control action from no-load to full-load conditions in the prototypes are excellent.

Design

The block diagram of the inverter is shown in Fig. 1. A power stage, connected to the battery via a polarity protection circuit, converts the battery voltage into a low-level alternating voltage, which is applied to the secondary winding of a mains transformer. The turns ratio of the transformer is such that across the primary winding an alternating voltage at a level of 240 V is generated.

A controller stage provides the necessary correcting voltage to the power stage. The controller receives information from a temperature monitor, an input-current limiting stage, and a quartz oscillator (clock).

The oscillator provides a stable 100 Hz signal to which the frequency of the output voltage is locked.

An optoisolator supplies part of the output voltage to a voltage regulator. If the optoisolator should fail, a small piezo buzzer warns that the voltage regulator is inoperative. The output of the regulator is applied to the controller.

The circuit

In the circuit diagram in Fig. 2, four modules, each containing four power transistors, make up the push-pull power stage of the inverter. The power transistors in each of the modules are connected in parallel via R5-R8 and R51-R55. The low-value resistors in the emitter circuits provide current feedback, which ensures near-equal currents through the transistors. Modules 1 and 2 form one branch of the push-pull power stage, and modules 3 and 4 the other. The four secondary windings of Tr2 are connected to the collector circuits of the power transistors and to the positive terminal of the battery via relay contact R25. The power transistors alternately connect and disconnect the secondary windings to and from earth.

The control voltage for the power stages is derived from crystal oscillator IC1. The
I2 VDC TO 240 VAC INVERTER

The clock signals are applied to the bases of Type BD679 darlington transistors, T3 and T4, via AND gates IC3a-IC3d and IC3e-IC3h respectively. These transistors provide sufficient base current for drivers T3 and T5, each of which forms a darlington configuration with the power transistors in modules 1–2 and 3–4 respectively.

In this kind of configuration, it is essential that the drive signals for the two push-pull branches do not overlap. If that were to happen, all power transistors would conduct simultaneously, albeit for a very short time. This would, however, cause an unnec-
cessary drain on the battery, and also be a potential risk to the power transistors. This situation is obviated by shortening the clock pulses to about 0.1 ms by network R2-C3.

The control and protection functions are provided by opamps IC3 and IC5. The 100 Hz output signal of IC2a is applied to the inverting input (pin 2) of integrator IC4a. The output of this opamp is a triangular waveform (pin 1).

The triangular waveform is compared by IC3b with the signal at its non-inverting input (pin 5). That signal consists of three components that are OR-linked via D6 and D7.

The output (pin 7) of the comparator is a rectangular waveform, whose duty factor depends on the level of the voltage across R29-C7.

The clock provided by bistable IC2b is combined with the output of the comparator by IC3b and IC3d. If the width of the pulses provided by IC4b is small, the push-pull output is switched on only relatively briefly, which results in a low inverter output. The greater the width of the pulses, the higher the inverter output will be.

The components of the signal at pin 5 of IC4b are provided by temperature monitor IC5, voltage regulator IC4, and current limiter IC4a.

The inverting input of IC5 is fed with the voltage across an NTC (negative temperature coefficient) resistor R49, which is mounted on one of the heat sinks for the power transistors. The non-inverting input is at a fixed potential provided by R30-R31. When, owing to a rising temperature, the value of R49 becomes low, the output (pin 6) of IC5 toggles to near-earth potential, which is indicated by the lighting of D3. Feedback resistor R32 provides a hysteresis of about 10 °C, so that IC5 switches off at about 60 °C and switches on again at around 50 °C.

Because of its low output level, IC5 pulls the non-inverting input of IC4b to ground via D2. This results in a lowering of the duty factor of the output signal of IC4b, and this in turn reduces the mean current through the power transistors and, therefore, the output voltage of the inverter.

Voltage regulator IC4c compares the reference voltage at pin 10, which is held steady by R30-D4, with the potential at its inverting input (pin 9). That potential is derived from the inverter output by optoisolator T1-L1-L2. The light bulbs operate from a 9 V supply provided by T1.

The base of phototransistor T1 is at a temperature-dependent potential derived from the 8 V supply via P2-R45-R46. When it receives a large light flux, the potential across R45 rises above the reference voltage at pin 10 of IC4c, the output of that opamp drops and D12 lights. At the same time, the voltage at pin 6 of IC4b drops, resulting in a lower duty factor and, consequently, a reduced output voltage. When the inverter output has dropped to a value that causes pin 9 of IC4c to become more negative than pin 10, the inverter output rises again. The regulation is set with P2 to obtain a stable 240 V output.

Piezo buzzer monitors the two light bulbs. In normal operation, there is no drop across it, but when one of the bulbs burns out, it is connected to the 9 V secondary of T1 via R50 or R51 and sounds an alarm.

The control board is connected to the battery via a separate earth line. The current in that line is so small that for all practical purposes the earth potential of the board is the same as that of the battery. The output stages and T2 are connected to the (earthed) battery terminal via R2. This is, however, not a

Fig. 3. The printed control circuit board.
## PARTS LIST

### Control circuit

<table>
<thead>
<tr>
<th>Capacitors:</th>
<th>Resistor:</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1, C2 = 47 pF</td>
<td>R1 = 10 MΩ</td>
</tr>
<tr>
<td>C3 = 1.5 nF</td>
<td>R2, R28, R24, R33, R38, R45 = 100 kΩ</td>
</tr>
<tr>
<td>C4, C13 = 470 nF</td>
<td>R3, R14, R27 = 1 kΩ</td>
</tr>
<tr>
<td>C5 = 1 µF</td>
<td>R4, R13, R50, R51 = 82 Ω</td>
</tr>
<tr>
<td>C6 = 220 nF</td>
<td>R25 = 27 kΩ</td>
</tr>
<tr>
<td>C7 = 22 µF, 25 V, upright</td>
<td>R26 = 1 MΩ</td>
</tr>
<tr>
<td>C8 = 10 µF, 25 V, upright</td>
<td>R28 = 39 kΩ</td>
</tr>
<tr>
<td>C9 = 330 nF</td>
<td>R29 = 4.7 kΩ</td>
</tr>
<tr>
<td>C10 = 6.8 µF, 25 V, upright</td>
<td>R30, R34 = 10 kΩ</td>
</tr>
<tr>
<td>C11 = 100 µF, 25 V, upright</td>
<td>R31 = 6.8 kΩ</td>
</tr>
<tr>
<td>C12 = 100 nF, 630 V</td>
<td>R32 = 12 kΩ</td>
</tr>
<tr>
<td>C14, C16, C17, C18 = 100 nF</td>
<td>R35, R43, R49 = 560 Ω</td>
</tr>
<tr>
<td>C15 = 470 µF, 16 V, upright</td>
<td>R36 = 1.8 kΩ</td>
</tr>
<tr>
<td>Semiconductors:</td>
<td>R40, R42, R44 = 220 Ω</td>
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<tr>
<td>D1, D2 = BYW81P1-200 (SGS Thomson)</td>
<td>R41 = 3.3 kΩ</td>
</tr>
<tr>
<td>D3, D5, D7, D14 = 1N4148</td>
<td>R46 = 10 kΩ (at 25°C) NTC</td>
</tr>
<tr>
<td>D4 = 4.7 V, 400 mW, zener</td>
<td>R47 = 270 Ω, 1 W</td>
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<tr>
<td>D8-D10 = 1N4007</td>
<td>R48 = see text</td>
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<tr>
<td>D11 = 5 mm LED, red</td>
<td>R49 = 47 kΩ NTC</td>
</tr>
<tr>
<td>D12 = 5 mm LED, green</td>
<td>P1 = 1 kΩ multiturn preset</td>
</tr>
<tr>
<td>D13 = 5 mm LED, yellow</td>
<td>P2 = 100 kΩ multiturn preset</td>
</tr>
<tr>
<td>T1 = BPY62/2</td>
<td>IC1 = 4060</td>
</tr>
<tr>
<td>T2, T4 = BD679</td>
<td>IC2 = 4013</td>
</tr>
<tr>
<td>T3, T5 = TIP3055</td>
<td>IC3 = 4081</td>
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<tr>
<td>IC4 = LM324</td>
<td>IC5 = TLC271</td>
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<tr>
<td>IC6 = 7808</td>
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### Power circuit (for one module)

<table>
<thead>
<tr>
<th>Capacitors:</th>
<th>Resistors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>C4, C13 = 470 nF</td>
<td>R5-R8 = 0.1 Ω, 3 W</td>
</tr>
<tr>
<td>C15 = 470 µF, 16 V, upright</td>
<td>Resistors:</td>
</tr>
<tr>
<td>Semiconductors:</td>
<td>R52-R55 = made from 0.433 Ω/m resistance wire (see text)</td>
</tr>
<tr>
<td>D1, D2 = telephone bulb, 12 V, 100 mA</td>
<td>R5-R8 = 0.1 Ω, 3 W</td>
</tr>
<tr>
<td>K1, K2, K3, K4, K5, K7, K8, K9 = 2-way spring-loaded PCB terminal board, 5 mm grid</td>
<td>Resistors:</td>
</tr>
<tr>
<td>K3 = 3-way spring-loaded PCB terminal board, 5 mm grid</td>
<td>R52-R55 = made from 0.433 Ω/m resistance wire (see text)</td>
</tr>
<tr>
<td>K6 = 2-way spring-loaded PCB terminal board, 7.5 mm grid</td>
<td>Semic conductors:</td>
</tr>
<tr>
<td>S1 = on/off switch, 2 A</td>
<td>T6-T9 = 2N3771</td>
</tr>
<tr>
<td>X1 = crystal, 3.2768 MHz</td>
<td>Miscellaneous:</td>
</tr>
<tr>
<td>Re1 = 12 V relay, contact rating 8 A</td>
<td>K10 = 3-way flat-cable connector for PCB mounting</td>
</tr>
<tr>
<td>Re2 = 12 V relay, contact rating 70 A</td>
<td>Heat sink Type SK85/75/SA†</td>
</tr>
<tr>
<td>Bz1 = 5 V d.c. buzzer</td>
<td>Heat sink Type WP40/30/SA†</td>
</tr>
<tr>
<td>Tr1 = secondary 9 V, 1.5 A</td>
<td>PCB Type 920038-2</td>
</tr>
<tr>
<td>Tr2 = primary 9.3 V, 50–60 Hz, secondary 240 V*</td>
<td>*Available from Amplimo BV, Vassenbrinkweg 1; 7491 DA Deielen, The Netherlands; Phone 05407 62024; Fax 05407 63132</td>
</tr>
<tr>
<td>2x heat sink Type 129/37.5 SA′</td>
<td>+ Available from Dau Ltd, 70–75, Barnham Road, Barnham PO22 5ES; Phone (0234) 55303</td>
</tr>
<tr>
<td>Enclosure 165x440x350 mm</td>
<td>† Available from Monacor, P022 OES; Phone (0234) 55303</td>
</tr>
<tr>
<td>PCB Type 920039-1</td>
<td></td>
</tr>
</tbody>
</table>

### Miscellaneous:

- Resistance wire (see text)
- Heat sink Type SK85/75/SA†
- Heat sink Type WP40/30/SA†
- PCB Type 920038-2
- PCB Type 920039-1

---

**Fig. 4.** The printed power circuit board.

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resistor, but represents the resistance of the connecting cables. The (load-dependent) drop across this resistance ensures that the potential at the inverting input (pin 13) of IC4d is positive with respect to earth. The voltage level at the non-inverting input (pin 12) is preset with R1. When the drop across R1 exceeds the level at pin 12, IC4d changes state, whereupon its output (pin 14) pulls pin 5 of IC4b to ground via D7. That results in a lowering of the duty factor and, consequently, a drop in the inverter output. This situation is indicated by the lighting of D13.

Polarity protection is provided by relays Re1 and Re2 and diodes D8 and D9. The inverter can be switched on by S1 only if the battery is connected with correct polarity.

Finally, IC5 and C13-C15 provide traditional regulation of the power supply for the control board.

**Construction**

The inverter is built on five PCBs: one for the control, monitor and regulation stages—see Fig. 3, and four for the power stages: one for each module—see Fig. 4.

Populating the control board is straightforward. Driver transistors T3 and T5 must be fitted on a suitable heat sink as shown in the parts list, but IC6 does not need one. Sockets should be used for all ICs.

The optoisolator is constructed from a length of pipe of suitable, but not too large, diameter, into which the phototransistor and the two 12 V (telephone) light bulbs are fit-

---

**Fig. 5. Wiring diagram of the inverter.**
12 VDC TO 240 VAC INVERTER

Fig. 6. A self-adhesive foil for the front panel is available through our Readers' services.

ted at either side. When the connecting wires of these components have been taken outside, the pipe must be sealed light-tight.

Transformer Tr2 is a special type that may have to be made to order, although it is available from certain retailers. It is not cheap: of the order of £75-80.

Each of the four power boards is designed for use with an angle-profiled heat sink, for instance, that shown in the parts list, that must be sawn to the required length of 16 cm. The necessary holes are best drilled in it by using the a photocopy of the board as a template. The power transistors are mounted on it with insulating washers, non-metallic screws, nuts and washers, and a good helping of heat conducting paste, before the board and heat sink sink are fastened together.

Power resistors Rs2-Rs5 consists of a 50 mm (2 in) length of resistance wire, whose ends must be cleaned with wire wool and bent at right angles. Solder them on to the board so that they are a few millimetres above the surface.

Boards and power stages are best interconnected with the use of bullet plugs and sockets as used in vehicle wiring.

Wiring to the power transistors should be flexible and have a cross-sectional area of not less than 2.5 mm². The sixteen wires to the emitters and those to the collectors should preferable have the same length.

The inverter should be connected to the battery with starter cable of cross-sectional area not less than 10 mm², but preferably 16 mm².

Testing

The control board is best tested before it is assembled with the remainder of the inverter with the use of a 12 V battery or laboratory power supply. Connect the +ve line to K7B and the –ve line to K8. With an oscilloscope or voltmeter, check the voltages at the output of drivers T3 and T5 (K3 and K4 respectively). After power-on, the duty factor of this waveform should change slowly to 1:1.

Next, apply a variable direct voltage across the secondary of Tr1. When the level of that voltage is raised to about 6-8 V, the effect of the voltage regulation on the duty factor should be quite clear.

For the remainder of the tests, it is better not to use a power supply, but a car battery. Also, take care with using a multimeter: its 20 A d.c. scale may be overloaded at even low loading of the inverter.

Connect a 100 W light bulb to the output socket of the inverter. A few seconds after power-on, this should attain maximum brightness. Adjust P2 until it has the same brightness as when it is connected to the mains. It is, of course, also possible to carry out this adjustment with the use of a moving-iron voltmeter. Note that moving-coil and digital voltmeters are not suitable. Recheck the output voltage after the inverter has been on for about 10 minutes. If it has risen by more than 3-5 V, replace Rs6 by a 5 k (25 °C) type.

The current limiting must be set after the inverter has been loaded with six 100 W light bulbs in parallel. Adjust P1 such that current limiting just begins. If the inverter is to be used at high loads regularly, P2 should be re-adjusted accordingly. The inverter can deliver up to 800-1000 W for short periods, provided the battery is capable of this. If the inverter is to be used in high ambient temperatures, a cooling fan should be added.

Under no-load conditions and operating from a well-charged 12-V battery, the inverter draws a current of about 1 A; with a 100 W load, the drain is around 10 A; with a 300 W load, the current is some 30 A; and at full load, the drain is around 80 A—a current that most car batteries cannot deliver for long periods. Apart from the fact that the battery gets hot, its efficiency drops sharply. For instance, an 80 Ah battery delivering 40 A has an effective capacity of only 60% of its rated capacity.

It is, of course, possible to operate from a 24-V power source (battery or solar cells, for instance), when all currents mentioned will be halved. The changes necessary to do this are: (a) the primary of Tr3 should be rated at 21 V; (b) the power transistors should be Type 2N3772; and (c) the relays should be 24-V types. Furthermore, it would then be prudent to fit Ic on a small heat sink.

Efficiency

A variety of tests, carried out on the author's prototype operating from a 12-V battery, showed that the efficiency varied from 68% to nearly 75% when the battery voltage and load ranged from 11.3 V and 970 W to 12.5 V and 29 W.
This instalment of the course is devoted to arithmetic operations, which are used in nearly every program, however small. In addition, some programming techniques will be discussed to show how simple calculations can be performed on the basis of the arithmetic instructions. Finally, two example programs are given that take the theory into practice: a capacitance meter and a noise generator.

**Addition**

The 8051 family of microcontrollers has the following instruction to add two 8-bit values (bytes):

```
ADD A, BYTE-OPERAND
```

The result of the above instruction is left in the accumulator. The carry bit (CY, sometimes also referred to as C) is set if there is a carry out from bit 7, and cleared otherwise. When adding unsigned integers in the range 0 to 255, a set carry flag indicates that an overflow occurred.

The auxiliary carry (AC) flag, used for BCD number adding, is set if there is a carry out from bit 3, and cleared otherwise. The AC flag is used by the DA A instruction discussed further on.

The overflow (OV) flag is set if there is a carry out of bit 6, but not out of bit 7, or a carry out of bit 7, but not bit 6; otherwise OV is cleared. This flag allows an overflow to be detected when adding two signed integers in the range -128 to +127.

The meaning of the flags depends on whether the bytes involved are unsigned integers, signed integers, or BCD numbers. The difference between these three requires a short discussion, given below.

**Number notation**

1. **Unsigned numbers**

These are all numbers that can be written with the aid of a sequence of weighted binary values \(2^n\), where \(n\) is 1, 2, 4, 8, 16, etc.: \[128 \times \text{bit}^7 + 64 \times \text{bit}^6 + \ldots + 4 \times \text{bit}^2 + 2 \times \text{bit}^1 + 1 \times \text{bit}^0\] In this way, a byte (eight bit positions) can be used to represent all unsigned values between 0 and 255.

2. **BCD numbers**

The BCD (binary coded decimal) number notation is based on indicating the left and right 4-bit groups (nibbles) contained in a byte as binary coded decimal numbers. The advantage of this notation is that it is simple to output. However, it also has a disadvantage: binary addition and subtraction can not be used just like that on BCD numbers, since additional corrections (decimal adjustments) are required. The DA A instruction is capable of performing these corrections, for which it uses the AC flag.

3. **Signed numbers**

The world is, unfortunately, not completely positive. Many computer applications require the use of negative values, which forces us to think of ways to add the minus sign to a number. In computer number notation, this is usually done as follows for the range \(-128\) to \(+127\): if the number to be represented, \(x\), is positive or nought, it is simply written as a byte without a sign. If \(x\) is negative (and has an absolute value smaller than or equal to 128), it is represented as \((256+x)\), i.e., the value lies between 128 and 255. This means that bit 7 is set to indicate that the number is negative. To be able to output a negative number, the value has to be formed first, so that the new (then positive) number can be output with a minus sign in front of it. Negative numbers are also treated separately with multiplications and divisions.

Table 1 shows some examples of bit patterns that represent different numbers.

**4. Large numbers**

It often happens that the eight-bit positions of a byte are not sufficient to represent all numerical values needed to achieve a certain accuracy. 16-bit values, for instance, are represented by two bytes; 32-bit values by four bytes, and so on. The bytes that form a 16-bit or 32-bit number are kept together as a group stored in successive memory locations. In this course, the lowest order byte is always stored at the lowest address. The bytes stored in the next higher locations are either signed, unsigned or BCD numbers. When two bytes are used to represent numbers, the following 16-bit ranges are available:

<table>
<thead>
<tr>
<th>Table 1</th>
</tr>
</thead>
<tbody>
<tr>
<td>Binary</td>
</tr>
<tr>
<td>00000000</td>
</tr>
<tr>
<td>00000001</td>
</tr>
<tr>
<td>00000010</td>
</tr>
<tr>
<td>00000100</td>
</tr>
<tr>
<td>00100001</td>
</tr>
<tr>
<td>11111111</td>
</tr>
<tr>
<td>10000000</td>
</tr>
<tr>
<td>01111111</td>
</tr>
</tbody>
</table>
Table 2 shows a few examples.

### Adding with carry

If 16-bit or 32-bit numbers are to be added, it is required that a carry resulting from a byte position be taken into account when the next byte is added. This is achieved with the instruction

\[
\text{ADDC A, BYTE, OPERAND} \quad \text{add operand+CY to A}
\]

The result of this instruction is left in the accumulator. The following program example adds a number NMBR2 to a 16-bit number NMBR1, which is contained in RAM:

\[
\begin{align*}
\text{MOV} & \quad \text{NMBR1} \\
\text{ADD} & \quad \text{NMBR2} \\
\text{MOV} & \quad \text{NMBR1} \\
\text{ADDC} & \quad \text{NMBR2} \\
\text{MOV} & \quad \text{NMBR1}
\end{align*}
\]

BCD correction

To obtain a BCD number as the result of adding two BCD numbers, the instruction

\[
\text{DA A}
\]

is used. DA A stands for decimal adjust. After the DA A instruction, the CY flag again signals a carry, which means that the BCD value is 100.

### BCD correction

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\[
\text{DA A}
\]

is used. DA A stands for decimal adjust. After the DA A instruction, the CY flag again signals a carry, which means that the BCD value is 100.

### Multiplication and division

The 8051 features an auxiliary accumulator, referred to as register B, and located at SFR address 0FOH. It is used for multiplication and division of unsigned 8-bit numbers. The relevant instructions are

- MUL AB : multiply A and B
- DIV AB : divide A by B

where A is contents of the accumulator contents, and B that of register B.

The result of the MUL instruction is that the low-order byte of the 16-bit product is left in the accumulator, and the high-order byte in register B. If the product is greater than 255, the overflow (OV) flag is set; otherwise it is cleared.

The result of the DIV instruction is that the accumulator receives the integer part of the quotient, while the integer part of the remainder is stored in register B. Both the carry and the overflow flags are cleared. Only if B contained 00, the overflow flag is set to indicate a division by zero.

The power of the MUL and DIV instructions offered by MCS-51 devices is limited by the fact that they do not allow direct 16x16-bit multiplication, nor division of a 16-bit number by an 8-bit number. Later, enhanced, versions of the 8051, like Siemens’ 80537, have much more powerful multiplication and division instructions.

To compensate the ‘lack of arithmetic power’ of the 8051, the course monitor program, EMON51, offers a number of subroutines that may be used to perform 16x16-bit multiplications with 32-bit results. Simply study the relevant sections of EMON51.LST to see how this is done. The monitor is also capable of doing 32-bit/16-bit divisions.

### Capacitance measurement using the V24 port

The practical use of the above arithmetic instructions will be illustrated by an example. We have in mind a capacitance meter that transmits the value of an unknown capacitor to the terminal (display) via the V24 serial interface on the 80C32 SBC. Interestingly, the measurement prin-
The principle adopted allows resistance and time measurement also.

The hardware (shown in Fig. 22) is monostable ICs. The operation of the capacitance meter is classic and simple: first, monostable ICs is triggered by the program. Next, the program measures the monostable period, i.e., the time that elapses between the triggering instant and the instant the Q output of ICs toggles. The measured time is converted into an equivalent capacitance value, and subsequently sent to the display. What sort of program is required to realize such an instrument?

1. Time measurement

The timer of the program is measured by a 16-bit number. MT time is reset to 0 (label 1: lines 20 and 21). Next, the monostable trigger pulse (positive edge) is output via port line P1.2. The loop that starts at label 1pp lengthens the start pulse a little, to allow some recovery time for the monostable before a new trigger action.

After sending the trigger pulse, the program enters the time measurement loop, labeled MLP. The program checks if the monostable is still triggered by monitoring bit 3 of RAM address 00001. If so, the variable MT time is increased by one. If not, the measurement is finished, and the program jumps to the label END MEAS. Lines 29 to 34 show how MT time is increased, and thus, more generally, how 16-bit and 32-bit variables are treated when it comes to using them in calculations. Starting with the lowest-order byte, the calculations are performed in a step-by-step manner, taking carry-overs into account at all times. After increasing MTIME, the program returns to the start of the loop.

2. Overflow detection

It may happen that the measured capacitance is increased. In such a case, the program sends an overflow report and calculates the result to PROD32.

The flow diagram of the capacitance meter program is shown in Fig. 23.
Theory and practice

Unfortunately, the results of the above calculations must be taken with a pinch of salt, mainly because of tolerances on the 47-kΩ resistor and on the 74HC123. The latter tolerance is particularly troublesome as it is fairly large and temperature-dependent. The program therefore includes variables to calibrate the capacitance meter, not by adjusting the resistor, but by a more clever approach. First, \( p_1 \) and \( q_1 \) are turned into constants set to 1.000. Next, the program is started, so that the value of \( \text{MTIME} \) is output rather than the capacitor value, although a close-tolerance 1.5-µF capacitor is connected. The resulting value of \( \text{MTIME} \), say, 1,540, is noted, and assigned to \( q_1 \) in the final program. The other variable, \( p_1 \), remains at 1,000, so that \( \text{MTIME} = 1,540 \) when a 1-µF capacitor is connected. Multiplied by \( p_1 \) (1,000), and divided by \( q_1 \) (1.540), this gives the correct readout: 1,000 nF. Since the monostable period is virtually proportional to the connected capacitance, the meter can be calibrated in this way without problems.

The capacitance measurement program, simple as it may be, already goes to show that converting measurement values into a meaningful indication requires a thorough

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understanding of various programming steps involving arithmetic instruction sequences. Those of you who wish to practise with these may do so by writing small programs that make use of the arithmetic routines built into the system monitor program, EMONS1.

### Shift and rotate

The following instructions are available to shift (or rotate) a bit pattern in the accumulator to the left or to the right:

- **RL** A: rotate accu left
- **RRC** A: rotate accu left through carry flag
- **RLC** A: rotate accu left
- **RR** A: rotate accu right
- **RRR** A: rotate accu right through carry flag

The operation of the rotate instructions is illustrated in Fig. 25. By clearing the C flag before an RRC or an RLC, a 0 is shifted into the extreme left hand or extreme right-hand accu bit position, respectively. The rotate instructions can be used in shift register simulations (see below), or in arithmetic operations (see, for example, the realization of the DIV routine in EMONS1).

### A noise generator

Reference 3 describes a noise generator of which the central part is formed by a shift register with feedback as shown in Fig. 26. As a programming example, we will implement this circuit on the 80C32 SBC and its extension board. The result is a programmable noise generator. The noise will be output via the loudspeaker, which is driven via port line P1.1.

The function of the program we are about to discuss is based on the noise generator hardware shown in the top drawing in Fig. 26. The shift register is stored at four consecutive locations in the internal RAM: the extreme right-hand bit is contained in the byte at the lowest address. This storage arrangement corresponds to the structure of a 32-bit number. This number is shifted to the left, where the XOR gate determines which bit is fed back into the sequence, at the far right end. The highest-order bit (bit 31) drives the loudspeaker.

The program listing of the noise generator is shown in Fig. 27. It is a straightforward implementation of Fig. 26, only the realization of the XOR gate (lines 15 to 20) is based on a 'trick.' Essentially, the function of the XOR gate is translated into software by combining bits in the sequence into a new bit that is added to the rightmost end of the 32-bit word. This creates feedback in the shift register, so that the output is a pseudo-random bit sequence with a long repeat time. To be precise: one loop iteration takes 23 μs, and the shift register goes through about 2 million states before a bit pattern is repeated. The pseudo-random signal thus has a sample rate of about 43 kHz, and the first 'duplicate' occurs after about 13 hours, so pretty random it is!

When used as a noise generator for audio measurements, the system requires a low-pass filter to shape the output spectrum. However, even without such a filter, the loudspeaker will produce quite a bit of real noise (a practically 'white' spectrum is produced with components up to about 20 kHz).

### Assignments

Based on what has been discussed so far, consider the following assignments. Add ranges to the capacitance meter that allow it to measure smaller as well as larger capacitors. You may also try your hand at implementing an autorangeing function. Another interesting subject could be the realization of a 'tolerance window' that tells the user of the capacitance meter instantly whether or not the value of a ceramic capacitor is within the required range. Producing a program to realize this will enable you to get a thorough grip on the possibilities of the arithmetic instructions.

Finally, how about turning the noise generator into a rhythm generator?

### What's in store

This month's instalment nearly completes the discussion of the 8051 instruction set, which has been elucidated, where necessary, with examples. This allows us to concentrate on more hardware-oriented aspects of the 8051 in the following instalments. The subjects to be discussed will be timers, LCD connection, serial interface and D-A/A-D conversion.

**Reference:**

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**Fig. 27.** Listing of the noise generator assembler program.
**DIMMER FOR NEON TUBES**

A neon tube cannot be dimmed as easily as an incandescent lamp because the tube can start only at a voltage much higher than the mains, after which it will remain lit at the mains voltage. The level of both the starting voltage and the working voltage depends on the temperature of the tube.

Normally, the high starting voltage is obtained by interrupting the current through a choke. This is usually done by the starter, which also ensures that a fairly large current flows through the filaments of the tube. This heats the ends of the tube, which makes starting easier.

These tasks of the starter are taken over by the circuit shown in the diagram, which also enables the tube to be dimmed.

During the zero voltage crossings of the applied mains voltage, the triac will instantaneously switch off. At those instants, capacitor C₃ will be charged rapidly, which results in the instantaneous voltage, whose phase has shifted relative to that of the current, being applied across the tube. Capacitor C₃ and the choke form a resonant circuit that raises the sudden voltage across the tube to a very high value, whereupon the tube starts.

The larger the angle of the mains voltage during which the triac conducts, the larger the current through the tube filaments, which results in a lower starting voltage. At the same time, since a larger part of the current flows through the triac, that through the tube will be reduced, so that the tube will light more faintly.

When the tube is first switched on, the dimmer control, P₁, should be set for maximum brightness of the tube to facilitate starting.

The triac used should have a high du/dt value, otherwise the steep voltage transitions occurring across the tube, and thus across the triac, during the zero voltage crossings would cause the triac to remain on.

(SGS Thomson - 924041)

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**POWER-ON DELAY FOR ATARI ST**

When the Atari ST is provided with an external hard-disk drive, it has to be reset after about 15 seconds, as otherwise the drive is not enabled. The delay circuit presented here obviates that inconvenience by ensuring that the computer is not switched on until a (presettable) period of time has elapsed. The circuit may also be used with the combination of MS-DOS computer and HP DeskJet printer, since the latter can be switched on only after the MS-DOS machine is powered.

If that type of printer is used with an Atari ST, the printer must be switched on before the computer is. No doubt, there are other situations where this delay may prove useful.

Operation of the circuit is fairly simple. After the mains has been switched on, capacitor C₃ is charged via R₅ and P₁. When, after a period of time set with P₁, the potential across C₃ has reached 12 V, zener diode D₁ conducts and switches on thyristor T₁. That device ensures that bridge B₁ provides an a.c. connection between the mains input and the mains output. At the instant the thyristor is switched on, owing to the anode-cathode current, its gate voltage rises slightly. This results in C₃ being charged a little more and thus serves as gate-current buffer to ensure that the thyristor remains on during the mains zero voltage crossings. Therefore, once the thyristor has been switched on, it stays on.

The delay is best built into a small man-made fibre box with integral mains connectors.

The circuit is best built on a small piece of veroboard or other prototyping board, bearing in mind that it will carry the full mains voltage. This means that tracks carrying the mains must be separated by at least 3 mm and preferably 5 mm. This probably entails the removal of tracks between the mains-carrying ones.

[R. Lucassen - 924017]
MOST VHF/UHF amplifiers for use with a symmetrical antenna (such as an open or closed dipole) have a balun at the input. A balun (short for 'balanced to unbalanced') is an inductive device that converts the symmetrical (balanced) RF signal into an asymmetrical (unbalanced) signal that can be applied to the base of a transistor. Unfortunately, baluns have an inherent loss of 2–3 dB, while most input transistors (typically a BFR91 or similar) have noise figures not better than about 2 dB. This explains the rather poor overall noise figure of 4–5 dB of this type of input stage.

A much lower noise figure is achieved by the circuit shown here, which does not incorporate a balun, and uses the BFG65 low-noise transistor as the amplifying device. The combination of a telescopic rod and a low-noise wideband RF amplifier is referred to as an 'active antenna'. The design shown here has two options: (1) if used with an existing antenna, it ensures a much better S/N (signal-to-noise) ratio in the receiver, or (2) the same S/N ratio can be achieved using a much simpler antenna.

The antenna proper is an open dipole with a total length of 1.6 m, which works as a 0.5 lambda dipole from 60 MHz to about 187 MHz, or as a multi-lambda V-dipole up to about 900 MHz. Unusually, the balanced-to-unbalanced conversion is done at the output of the amplifier, with the aid of a length of coax cable that functions as a kind of choke. The construction of the amplifier on the PCB shown, and its electrical behaviour, enables the antenna to 'see' a balanced load.

The two telescopic rods are connected directly to the solder pads marked 'ANT.1' and 'ANT.2'. Usually, telescopic rods have a kind of 'knee' construction at the base that enables them to be rotated as well as bent up and down. With some dexterity, this mechanism may be retained for use with the present amplifier.

The output inductor, L1, consists of 10 turns of 2.5 mm dia. 60 Ω coax cable wound on a 10 cm long ferrite rod with a diameter of 10 mm. The phantom supply for the amplifier is not contained on the printed circuit board, but is simple to install at the input of the receiver, since it consists of a resistor and a capacitor only. If the receiver is located further than about 2 m away from the active antenna, it may be connected to K1 via a length of ordinary coax cable like RG58.

Provided the antenna is used in an area with relatively low local field strengths, you may lower the noise figure of the BFG65 by increasing R3 a little. This should not be done, however, if there are strong signals around, in which case the result is an increased risk of cross-modulation. With R3=560 Ω, the current consumption is about 20 mA. The gain of the dipole signal amounts to about 12 dB, while a noise figure of about 1 dB is achieved. This will ensure results comparable to those of a much larger antenna (yagi), provided signal reflections and multipath reception are not a problem.

(J. Barendrecht - 924102)
INFRA-RED HEADPHONE TRANSMITTER

The proposed transmitter provides an optical (infra-red), that is, wireless, connection to a headphone. The receiver is described in the next article.

Three infra-red (IR) LEDs are provided with a quiescent current by $T_1$. The level of that current is set with $P_1$. When an audio signal is applied to the gate of $T_1$, the current through the LEDs is modulated. Consequently, the light emitted by the diodes is also (amplitude) modulated.

To prevent overdriving of the gate causing too high a current through the LEDs, a current limiter, consisting of $T_2$ and $R_3$, holds the current below 100 mA.

The maximum dissipation of a BS170 is 830 mW at an ambient temperature of 25 °C, while the maximum drain current is 500 mA. Therefore, even when the FET is overdriven, those limits are not exceeded.

The prototype transmitter drew a current of about 60 mA at a supply voltage of 9 V. It is, therefore, advisable to use a mains adaptor, because that current is just a little too high for a PP3 battery. Keep the earths of the mains adaptor and the audio signal separated as shown in the diagram to prevent feedback of the LED current to the input.

The gate-source voltage of a BS170 may be up to 15 V. If you use a signal source that delivers a higher level, it is advisable to incorporate a simple protection circuit (for instance, a 10-V zener diode in parallel, or a resistor in series, with the input).

The optical connection is fairly directional, but this can be improved by placing the LEDs at varying angles. Also, the distance of operation can be extended appreciably by fitting reflectors behind the LEDs.

The optimum input level for an operating distance of some metres (4-8 ft) is 100-200 mV.

(Amrit Bir Tiwana - 924069)

INFRA-RED HEADPHONE RECEIVER

This receiver is meant to complement the transmitter described in the previous article. Its design is based on just one FET. This has the advantage that construction is simplicity itself, and the disadvantage that for a sufficiently low output impedance the value of drain resistor $R_3$ has to be fairly low. That results with correct operation of $T_1$ to a fairly large (certainly for a battery) current. The value of $R_3$ was chosen, as a compromise, at 560 Ω, which makes driving 600 Ω headphones possible. The load seen by $T_1$ is then 300 Ω.

Both usable types of receive diode shown in the diagram have a daylight filter and are tuned to the wavelength of the LEDs in the transmitter (950 nm at 25 °C). At a couple of metres (5-8 ft) distance, a (no-load) output voltage of 200-300 mV is obtained, which is quite sufficient for most headphones. The circuit then draws a current of 9-10 mA.

The setting of $P_1$ is fairly critical, but its control range may be reduced by adding a small resistor at either side of the preset. The preset should be adjusted for minimum distortion. This is best done by applying a 1 kHz audio signal at a level of 150 mV to the transmitter and adjusting both circuits for minimum (audible) distortion. This should be done without electric light, because the transmitter does not modulate the audio signal on to a carrier, so that light bulbs, and particularly neon tubes (which emit an appreciable amount of IR light at 950 nm, modulated with 100 Hz) can cause quite a hum. Even normal ambient light causes a deterioration of the signal-to-noise ratio. However, with a little ambient light and a distance between transmitter and receiver of 3-4 m (10-13 ft), the distortion was 1-2%, which is not bad for such a sparse design.

(Amrit Bir Tiwana - 924070)
METAL DETECTOR

This detector will help you find fairly large objects that consist of materials with a relatively high permeability. Also, it indicates whether the magnetic object inside the detection coil has good or bad conductive properties. Examples of materials that couple good magnetic properties to a fairly high electrical isolation are ferrites pressed from metal oxides. The detector is not suitable for 'coin digging', for which it is not sensitive enough. The more fanciful stuff like bombs and treasures left by pirates, is, however, reliably located.

The metal detector is powered symmetrically by two 9-V batteries, each of which is loaded with about 15 mA. The detection coil, L₁, forms part of a sine wave oscillator built around transistor T₁. Normally, the central frequency of the VCO (voltage-controlled oscillator) in the PLL (phase-locked loop) contained in IC₁ equals the oscillator frequency of T₁. That

**PARTS LIST**

**Resistors:**
- R₁ = 6kΩ
- R₂, R₃ = 4kΩ
- R₄ = 680Ω
- R₅, R₆ = 5kΩ
- P₁ = 10kΩ, multturn preset
- P₂ = 470Ω linear potentiometer

**Capacitors:**
- C₁, C₂ = 100μF 16V radial
- C₃ = 68nF
- C₄ = 15nF
- C₅, C₈ = 10nF
- C₆, C₇ = 1nF

**Inductor:**
- L₁ = details in text

**Semiconductors:**
- T₁ = BC547B
- T₂, T₃ = BC557B
- IC₁ = NE565N

**Miscellaneous:**
- S₁ = miniature double-pole change-over switch
- B₁₁, B₁₂ = 9-V battery with connecting clip
- M₁ = centre-zero ±50μA moving coil meter

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ELEKTOR ELECTRONICS JULY 1992
changes when a metal object (ferromagnetic or non-ferromagnetic) enters the field induced by L1. When that happens, the sine-wave oscillator is detuned, and the voltage difference between pins 6 and 7 of IC1 indicates the difference between the sine oscillator frequency and the VCO frequency. This difference causes moving-coil meter M1 to deflect. The needle deflection itself is a measure of the frequency change, while the direction of the needle depends on the type of material detected by the coil.

Next, adjust P1 such that the edges of the rectangular signal at pin 4 coincide with the positive peaks of the sine wave at pin 2. Next, null the meter by turning potentiometer P2. Since the null adjustment will drift a little as the battery voltage drops, it will be necessary to redo the balance adjustment every now and then during use.

(K. Kraus - 924038)

CB-TO-SW DOWN CONVERTER

This converter enables long-distance (DX) reception of AM or SSB stations in the 27-MHz citizens' band (CB) on a short-wave or medium-wave radio (note, though, that AM or SSB modulation in the 27-MHz CB band is no longer allowed in a number of countries).

The converter consists of a prestage, T1, and a mixer/oscillator, IC1. The antenna signal is coupled inductively to the gate of dual-gate MOSFET T1 via tuned circuit L1-C1, which acts as a 27 MHz input filter. The operating point of the MOSFET is determined by resistors R1-R2 connected to the gate-2 terminal. The amplified signal is fed to the mixer amplifier via a coupling capacitor and a second tuned circuit, L3-C2. The oscillator on board the S042P IC (manufacturer: Siemens) works with a quartz crystal X1. The frequency selected such that the difference frequency produced by the mixer (also on board the S042P) falls within the tuning range of the radio connected to the output of the converter. For instance, if an inexpensive 26,800 MHz crystal is used, the frequency of the received 27-MHz station, \( f_{\text{in}} \), is

\[ f_{\text{in}} = 26,800 \pm f_{\text{dia}} \]

where \( f_{\text{dia}} \) is the frequency read on the radio’s tuning scale (in this case, the received stations will appear in the medium-wave band). Other crystal frequencies may be used, e.g., 10 MHz, to move the CB band into the SW range (17 MHz).

Construction of the converter is fairly easy on the printed circuit board shown here. Parts shown with a dashed outline are fitted at the solder side of the board. The converter is shielded all around to prevent spurious radiation. The antenna and the IF output are best connected via coax sockets (SO295 or BNC). The two inductors in the converter are simply adjusted for best reception.

The converter is powered either by a regulated 9-V adaptor, or by the radio is is connected to, if this is capable of furnishing 9 V at a few tens of mA.

(924001 — Dr. U. Kunz)

For drawings of the PCB, see p. 99

PTER PARTS LIST

Resistors:

- R1;R2 = 100kΩ
- R3 = 56kΩ
- R4;R5;R6 = 330kΩ

Capacitors:

- C1,C2 = 56pF ceramic
- C3 = 2nF ceramic
- C4 = 10pF ceramic
- C5 = 1nF ceramic
- C6 = 47µF ceramic
- C7 = 22µF ceramic
- C8;C9 = 10pF ceramic
- C10 = 27pF ceramic
- C11 = 220pF ceramic
- C12 = 1nF ceramic

Semiconductors:

- T1 = BF982
- IC1 = S042P

Inductors:

- L1;L3 = 113CN2K50989189ADZ (Toko)
- L2 = 1mH choke

Miscellaneous:

- K;K2 = BNC socket
- X1 = see text

ELEKTOR ELECTRONICS JULY 1992
LOW-DROP REGULATOR

NOWADAYS the only reasons for not using a voltage regulator in a power supply are: I have not got one: I need an 'odd' voltage: I want to keep the current drain very low.

The regulator shown in the diagram is suitable for currents of 5-10 mA. The two transistors draw only a tiny current. The drop across the regulator depends on the load current and lies between 0.5 V and 1.4 V. The output voltage may be preset between 1.8 V and 8 V.

On power-up, there is no voltage at the source of T1, so that the FET conducts. Current amplifier T2 then draws base current and is switched on. This arrangement means that the reference (gate) voltage may be taken from a high-resistance potentiometer. The quiescent current depends on the level of the preset output voltage: at 5 V, it is a mere 1 µA.

When T2 is switched on, the output voltage will rise to its preset level. The base potential of T2, and thus the source potential of T1, remains about 0.6 V higher than the output voltage, so that it rises in step with the output voltage. The gate of T1 is, however, connected to the wiper of P1, whose voltage rises more slowly than the output voltage, because the preset is a potential divider. Consequently, the gate of T1 becomes more and more negative with respect to its source. An equilibrium is soon reached, whereby the FET reduces the base current of T2 to a degree that ensures stability of the output voltage.

In normal circumstances, the output voltage vs load current ratio is of the order of 9 mV/µA.

TELEPHONE MONITOR

MODERN telephones can easily be connected in parallel, enabling a household to have one in the bedroom, kitchen and hall or study. It is, however, not always easy to see (or hear) at one position whether the phone at one of the other positions is in use or has taken up the call.

The d.c. level on most telephone lines drops from 48 V to about 8 V when one appliance is in use and to around 5 V when two telephones are in use. Also, its polarity changes over when a telephone is in use.

Since the polarity changes over, the circuit is of necessity a bridge type as shown in Fig. 1. To keep the current drain small and relatively constant at varying voltage levels, a current source is needed for the high-efficiency indicator LEDs. At 50 V across a and b, the circuit is inactive: when that potential drops to about 8 V, the current source is on; and D7 is short-circuited. As soon as a second receiver is lifted, the line voltage drops to some 5 V, resulting in T3 being switched off, so that D7 lights.

The current through the LEDs is necessarily small and should in any circumstances not rise above 5 mA. With two telephones that means that R5 must be 270-330 Ω, and for three telephones, 390-470 Ω. As already stated, this means that high-efficiency LEDs are essential.

(P. Holmes -- 924030)
**CROWBAR PROTECTION**

Quite a few electronic components, particularly active ones, cannot withstand too high voltages. Preventing costly circuits dying a premature death because the supply voltage has risen too high, an overvoltage protector is no luxury. Such a protector must, of course, act swiftly, otherwise the deed has been done before it has a chance to act. Therefore, (slow-acting) relays are not suitable for this purpose.

The circuit shown here, a so-called crowbar, contains several fast-acting components. It is intended to be connected between the mains and the appliance to be protected.

The circuit depends, as it were, on brute force: when the supply voltage rises too high, a thyristor shorts the output. This means that the too high voltage is immediately removed from the connected appliance and also that fuse F1 blows. Brute force, indeed!

The voltage at which the crowbar comes into action is set between 5 V and 25 V with P1. This is done as follows:

1. Set P1 to its maximum value.
2. Replace the fuse temporarily by a wire bridge and connect the crowbar to a variable power supply. Set the current limit of that supply to 1 A and the output voltage to the value at which the crowbar is desired to act.
3. Turn P1 back slowly until the thyristor comes on, that is, when the current limiter of the power supply comes into action.

The crowbar is now set. Replace the wire bridge by the fuse (max. rating 5 A). In quiescent operation, the crowbar circuit draws a current of only about 1 mA.

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**CGA-TO-SCART ADAPTOR**

The adaptor has the great advantages of not needing a separate power supply and requiring so few external components that these can easily be accommodated in the SCART connector.

The signals on the R, G, and B pins of the CGA card are converted from TTL level to SCART level with the aid of transistors R4, R5 and R6 and the input impedance (75 Ω) of the SCART inputs. The output impedance of the so created voltage dividers is not exactly 75 Ω as specified for SCART video inputs, but in practice this does not appear to make much difference.

Resistors R1-R3 between pin 6 of the CGA card and the R, G, B pins of the SCART connector ensure that the brightness is brought down to 50% when pin 6 is active.

A composite synchronization signal and a blanking signal for the TV are derived from the horizontal and vertical sync pulses of the CGA card with the aid of R7-R11 and T1.

The 5-V supply for T1 is obtained from the CGA card via pin 7. Since that pin is not normally used, it must be linked to the 5-V line by a short length of wire.

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ELEKTOR ELECTRONICS JULY 1992
THE reliability of readings of battery-operated measuring instruments depends, of course, on the state of the battery. Even simple instruments are, therefore, provided with a facility (normally a push button) to check the battery voltage. If, however, the state of the battery must be monitored constantly, that type of check is not very reliable. And then there are instruments that have no battery check facility whatsoever.

The indicator shown in Fig. 1 solves all these problems. With jump links JP1 and JP2 in the positions shown, D2 serves primarily as on/off indicator. This low-current LED lights constantly as long as the battery voltage is sufficiently high and the optional alarm input (a) is not actuated. When the battery voltage drops below a value set by P1, oscillator IC1b is switched on, whereupon D2 begins to flash at a frequency of about 0.5 Hz. A second oscillator, IC1a, is activated when a high voltage level is applied to terminal a. The LED will then flash at a frequency of 10 Hz. This additional indication is often useful to alert the user to an operational error.

When JP1 and JP2 are in their other position, there is no on/off indication; the quiescent current then drops from about 4 mA to a few µA. If this condition is selected, the connections of D2 must be changed over (as shown in dashed lines).

The circuit uses the NAND gates of a Type 74HC132 IC, which is suitable for supply voltages, \( U_1 \), of 2–6 V. Since the change-over point of the indicator depends on the supply voltage, connected to \( U_1 \), this must be regulated. It is, of course, possible, even preferred, to connect \( U_1 \) to the regulated supply of the relevant measuring instrument. Here, an independent supply of 5 V is assumed. Since the regulator, IC2, is a low-drop type, the battery voltage may drop to about 5.5 V before the stability of the output voltage (at \( U_1 \)) becomes questionable.

The design may be adapted for use with higher supply voltages: if IC1a is replaced by a Type 4093, the circuit becomes suitable for an input voltage, \( U_1 \), of up to 18 V. The series resistor of D2 must, of course, also be adapted: for instance, D5 should be 3.9 kΩ when \( U_1 = 15 \) V.

The battery voltage is applied to the input of IC1a via potential divider R1-P1. Diode D1 protects the gate against too high a voltage from a fresh battery.

Although the function of C6 may not be immediately evident, it is a vital one: the capacitor ensures that the upper switching threshold (about 2.5 V) of IC1a is exceeded briefly on power-on. Oscillator IC1b is then reset. After about one second, the potential at junction R2-R3 is a measure of the state of the battery. If that potential lies below about 1.5 V (the lower switching threshold), the oscillator is actuated. If C6 were not in circuit, the oscillator would not be reset at low battery voltages, resulting in an erroneous low-battery indication.

Preset P1 (maximum resistance) is adjusted when a variable power supply is connected to \( U_{+} \) and set to an output level at which the indicator is to become operational, say, 6 V. Turn P1 slowly until D2 begins to flash.

The alarm input, a, may be adapted for use with negative battery voltages by the circuit shown in Fig. 2.
In fact, the two circuits together monitor a symmetrical battery supply: the positive line via $U_5^+$ and the negative line via the circuit in Fig. 2. When one of the batteries becomes low, $D_2$ begins to flash: at a frequency of 0.5 Hz in case of the + battery, and at about 10 Hz in case of the – battery. When both batteries become low simultaneously, $D_2$ flashes with interruptions.

Fig. 1 raises the negative battery voltage to above earth level. Note that, because of the loading of its output, $R_6$ in Fig. 1 must be removed. Connect the ++ line to $U_1$ in Fig. 1, the + line to the negative terminal of the symmetrical battery supply, and 0 to 0 in Fig. 1.

Table 1

<table>
<thead>
<tr>
<th>Input voltage $U_{\text{in}}$</th>
<th>Output voltage $U_{\text{out}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>70</td>
<td>180</td>
</tr>
<tr>
<td>85</td>
<td>198</td>
</tr>
<tr>
<td>100</td>
<td>260</td>
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<tr>
<td>125</td>
<td>325</td>
</tr>
<tr>
<td>160</td>
<td>212</td>
</tr>
<tr>
<td>175</td>
<td>233</td>
</tr>
<tr>
<td>200</td>
<td>270</td>
</tr>
</tbody>
</table>

[Provided all connections are correct, the total current drain rises by only about 10 μA over that of $\text{Fig. 1 (assuming two 9-V batteries in series).}$]

Preset $P_1$ is adjusted in a manner similar to that described above for $P_1$ in Fig. 1. When the upper switching threshold of $IC_1$ is exceeded, the gate oscillates and $D_2$ begins to flash. The power-on reset (terminal α becomes log 0) is provided by capacitor $C_2$. That capacitor is discharged via $D_1$ and the supply lines.

Diode $D_2$ protects $IC_1$ against negative input potentials, while capacitor $C_1$ prevents the circuit being triggered by possible brief noise peaks on the battery voltage.

[J. Ruiters - 924043]

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**VOLTAGE CONVERTER**

The limiter is capable of converting a 70–260 V r.m.s. alternating voltage into a 180–350 V direct voltage. For that, a full-wave rectifier, contained in an MC34161, is used as a voltage doubler at low input voltages and as a standard rectifier at high input voltages. In this way, an input voltage variation of x4 is reflected in an output voltage change of not greater than x2.

The MC34161 has an integral voltage reference source that provides a voltage of 2.54 V at pin 1. The level of a signal applied to pin 2 is compared internally with a potential of 1.27 V.

Voltage divider $R_2$-$R_3$ ensures that the internal comparator changes state when the input voltage rises above 135 V, pin 5 then becomes logic high. The potential at pin 2 is then lower than 1.27 V. The triac is switched off, and this removes the centre link between the output capacitors, $C_2$ and $C_3$, so that voltage doubling cannot take place. Diodes $D_1$-$D_4$ then operate as a normal bridge rectifier.

When the input voltage is lower than 135 V, the potential at pin 2 remains higher than 1.27 V; the internal transistor at pin 6 is then off. Because of $R_4$, pin 3 then carries the reference voltage of 2.54 V. That in turn results in the internal transistor at pin 5 being switched on, so that triac $Tr_1$ switches from the blocking state to the conducting state. Diodes $D_2$ and $D_3$, and capacitors $C_2$ and $C_3$, then function as voltage doubler.

Zener diode $D_5$, in conjunction with $R_1$ and $C_4$, ensures that the IC is supplied from a stable 12 V source. The time taken by the circuit to switch from standard rectifier to voltage doubler is determined by $R_4$-$C_1$.

The working voltage of capacitors $C_2$ and $C_3$ must be ≥250 V.

**WARNING:** the circuit carries high voltages and must therefore be treated with the greatest care.

[Motorola Application - 924075]
Although newer types of solar cell have come to the fore recently, amorphous cells are bound to be with us for some time to come, mainly because of their low cost. Most solar cells based on amorphous technology have a relatively high internal resistance, which results in large differences between the loaded and 'open-circuit' output voltages. Where a rechargeable battery is used as an energy storage device, a voltage regulator circuit is in order that charges the battery when the cell output voltage is high, and forms a minimum load on the battery when the cell output voltage is low.

For relatively small solar power systems, the parallel (or 'shunt') regulator is a viable alternative. Apart from a single (Schottky) diode, nothing is inserted between the solar cell and the battery. Since the supply voltage is furnished by the solar panel, the regulator works always, even if the battery is fully discharged or not connected. This ensures the best possible protection against overvoltage of all circuits powered by the solar cell (or array of cells).

The heart of the shunt regulator presented here is formed by a Type TL431LP precision voltage regulator from National Semiconductor. When the solar cell output voltage rises above the level set with preset P1, a current starts to flow through R3 - R2 - D1. When the current has risen to about 5 mA, transistor T1 starts to conduct. The transistor used here, a BD642, may be replaced by almost any other, similar, power darlington, for example, the TIP147. The collector of the BD642 is conveniently connected to ground, which means that the device can be bolted direct on to a heat sink.

Since the regulator is capable of shunting quite high currents, it has separate sense inputs, 'A' and 'B', to monitor the battery voltage. The solar cell is connected to terminals '+' and '−'. Resistor R4 limits the current through D2 in the event of the high-power series diode, D3, breaking down. Not shown in the circuit diagram, but required in the interest of safety, are fuses in series with the battery and the load(s). Also, do not forget to connect a surge arrester in parallel with the solar cell.

The Schottky diode used here, an SB530 from Conrad, is capable of passing up to 5 A. For panel output currents up to 3 A, it may be replaced by the more familiar 1N5401 (which, unfortunately, has a slightly higher forward drop voltage). Talking of ratings: the heat sink on to which the power transistor is bolted must...
If a stable 5-V direct voltage is to be derived from an already low supply voltage, the 4805 from SGS Thomson is probably the most suitable regulator IC available at present. The popular 7805 does not work so well with supplies lower than about 8 V. The 4805, on the other hand, needs an input voltage that is only about 0.5 V higher than its output voltage. Its data sheets state that its output voltage remains stable so long as the input voltage does not drop below 5.7 V. That voltage, by the way, is the worst-case voltage when the output current is 400 mA. In practice, therefore, the regulator normally works fine with input voltage as low as 5.4 V.

What to do if the output current is more than 400 mA? Well, in that case, and for output currents up to 1 A, the LM2940T appears to be just about the best available. This IC from National Semiconductor is available in three variations: 5 V, 8 V, and 10 V. The 5-V version, in which we are principally interested here, is type-coded LM2940T-5. These ICs are pin-compatible with the 7805 and 4805, which makes it possible for an existing low-drop supply based on one of these devices to be upgraded fairly simply.

For completeness sake, the circuit diagram shows a simple 5-V design. The only really important parameter here is the capacitance of decoupling capacitor C4. According to the relevant data sheets, its value should be not less than 22 μF to ensure correct stability.

If the input voltage does not go above about 7.5 V, a heat sink is not necessary. If the regulator is required to operate at a higher output voltage and the full output current, a suitable heat sink is imperative. A heat sink rated at 6.5 °C W⁻¹ makes the low-drop circuit suitable for input voltages up to 15 V at full output current or up to 25 V at 500 mA. The worst-case input voltage is 5.8 V.

The LM2940T, like the 7805, is short-circuit-proof, but it does draw a rather high quiescent current as shown in the characteristic curves.

Obviously, to cope with the increased power dissipation, the size of the heat sink must be increased accordingly.

The component values shown result in a battery voltage adjustment range of 13.4–17.6 V. This is on the high side for most (lead-acid and gel-type) batteries, which do not fare very well at charge voltages greater than 13.8 V. The circuit can be modified for an output voltage range of about 11–16 V by letting: \( R_5 = 220 \, k\Omega \); \( R_6 = 47 \, k\Omega \); \( R_\text{i} = 25 \, k\Omega \).

(K. Schönhoff - 924010)
WATER PUMP CONTROL FOR
SOLAR POWER SYSTEM

In most small solar power systems using a boiler it is required that the water circulation pump is not switched on until the temperature of the collector (the solar panel) exceeds that of the water in the vessel. Here, a two-sensor monitor is presented that enables this condition to be met. One sensor is fitted on the collector, the other on the water vessel. The control shown here has two adjustments: one for the temperature difference at which the pump starts to operate, and one for the temperature difference at which it is switched off. Although these settings are independent, the switch-off level must be lower than the switch-on level. Calibration in degrees Celsius is simple because the gradient of the voltage at the wiper of the potentiometers (or presets) that set the on/off temperatures is exactly 0.1 °C⁻¹.

The two temperature sensors Type LM334 are adjusted to supply a temperature gradient of 1µA °C⁻¹. Unequal sensor temperatures therefore produce a current flow at their junction. The voltage across R₁ is directly proportional to the temperature difference measured. This enables the switching thresholds of the on/off control to be set with the aid of two presets: the 'on' preset (P₁) is adjusted to, say, 3 °C, and the 'off' preset (P₂) to 1 °C. The range of the two presets is about 5 °C.

The sensors used here supply a current rather than a voltage. This eliminates thermocouple effects caused by temperature changes on the connecting cables between the sensors and the circuit. If voltage-type sensors such as PTCs or NTCs were used, the circuit would have become more complex because of the required compensation. The AD590 may be used instead of the LM334. Note, however, that the AD590 does not require an adjustment preset or resistor.

Relay Rₑ₂ switches the pump on and off. A second relay, Rₑ₁, comes on after Rₑ₂. It is optional, and may be used to switch the pump briefly to a higher speed, which is required in some solar heating systems to increase the initial water flow, or to fill the system.

The circuit is calibrated by setting equal sensor currents at equal sensor temperatures. The sensor current equals

\[ (273 + T_a)] \mu A, \]

where \( T_a \) is the ambient temperature in degrees Celsius. Thus, at a room temperature of 20 °C, presets P₁ and P₂ are adjusted until the current flow through each sensor is 293 µA. A few microamps more or less will not make much difference, as long as the sensor currents are equal.

It is best to first adjust one sensor only. Start by connecting a microammeter between 'A' and ground, and
adjust $P_1$. Next, adjust the other preset until the voltage across $R_1$ is nought. It will be clear that these initial adjustments require that the two LM334 are at the same temperature.

Current consumption of the on/off control is about 11 mA plus about 35 mA for each relay.

The dimensions of the printed circuit board are geared to the size of the box mentioned in the parts list. The potentiometers are fitted with the spindles at the track side of the board.

(K. Walraven - 924007)
**POWER SUPPLY TESTER**

This little circuit enables you to measure the so-called dynamic response of a d.c. power supply. A power MOSFET, T1, is used to switch the supply load on and off at a user selectable rate. The response of the supply to these fast load variations is displayed on an oscilloscope.

The switching rate is selected with the aid of a rotary switch, S1, which also serves as the on/off switch. The available switching frequencies are: 10 Hz, 100 Hz, 1 kHz and 10 kHz.

The well-known 555 timer IC is used to supply the switching signal. Diodes D3 and D4 cause the astable multivibrator to supply an output signal with a duty factor of about 0.5.

The switching transistor, T1, is protected against too high currents by a fast 10-A fuse inserted in the drain line. The tester may be powered by any regulated d.c. supply with an output voltage between 6 V and 15 V. However, this must not be the supply under test! Given the low current consumption of the tester (40 mA max.), a 9-V battery is an excellent power source.

The tester is extremely simple to use. First, select the load resistance of the supply you wish to test: say, 12 Ω/15 W for a 12-V, 1-A PSU. This resistor is connected between output R' of the tester, and the '+' output of the PSU. The '0' output of the tester goes to the '-' (or '0') terminal of the PSU. Next, connect the scope input to the PSU outputs, and the trigger input to K3 of the tester. Switch on the scope, the PSU and the tester. The scope will now display the dynamic regulation characteristic of the PSU at the given output current (1 A) and the selected switching rate (initially, 10 Hz).

Construction of the tester is straightforward on the small printed circuit board shown here. The power MOSFET is bolted on to a small PCB-mount heat sink, and will not run very hot even when the maximum permissible drain current (about 10 A) is approached.

(J. Ruiters - 924015)

### PARTS LIST

**Resistors:**
- R1, R2, R3 = 820Ω
- R4 = 470Ω
- R5 = 1kΩ

**Capacitors:**
- C1 = 100μF 16V radial
- C2, C3, C7 = 100nF
- C4 = 10μF 16V radial
- C5 = 10μF 16V radial
- C6 = 1μF 16V radial

**Semiconductors:**
- D1 = LED, red, 5mm
- D2 = 1N4007
- D3, D4 = 1N4148
- T1 = BUZ10
- IC1 = NE555

**Miscellaneous:**
- K1, K2 = 2-way PCB terminal block; pitch 5mm.
- K3 = panel mount BNC socket.
- S1 = 2-pole 6-way PCB mount rotary switch.
- F1 = 10A fast fuse plus PCB mount holder and cap.
- Heat sink 5K/W, e.g. SK129/38.1mm.
IT is an unfortunate and well-known fact that most PCs of the 'IBM and compatible' type make a lot of noise, which is both undesirable (as regards noise in the working environment) and unnecessary (as regards the actual power consumption, which is often quite low). Many fans in PC power supplies are over-rated, noisy, and run at a constant, high, speed. Fortunately, the authors found that such fans can do their protective job just as well at far lower speeds.

The speed controller presented here consists of (1) a temperature monitor based on a LED as the sensor device, and (2) an idle speed regulator. The combination of the temperature monitor and the idle speed regulator results in a linear relation between temperature and fan speed. In other words: the fan will never run faster than strictly necessary. By virtue of the two separate regulator circuits, there is no interaction between idle speed and temperature regulation, as with many other (less sophisticated) fan controllers. The result is low fan speed (low noise) at low temperatures, as well as good cooling and a safe start-up at all times.

When the temperature inside the PC rises, the voltage drop across the red LED, D₂, decreases by approximately -2 mV K⁻¹. This results in a higher output voltage of opamp IC₁A. Preset P₁ is used to set the start level, while P₂ determines the slope of the regulator characteristic. The adjustment range of the idle-to-full speed regulation is about 2 °C to 30 °C.

The idle speed is set to the desired value by adjusting P₃. Capacitor C₂ ensures a 4-second full-speed start up period, while diode D₃ restarts the fan after a short interruption on the mains.

The outputs of the temperature monitor and the idle speed regulator are 'joined' by two diodes, D₄ and D₅, at the base of T₁. The regulator is stabilized by feeding a small portion of the fan voltage back to the inputs of the opamps.

The circuit is built on a small printed circuit board that can be fitted into the PC's power supply. The LED may have to be moved off the board, and connected with wires, to enable it to be fitted in a position where temperature changes are best noticed.

(L. Svenkerud and A. Kristiansen - 924009)
HALOGEN LAMP PROTECTOR

HALOGEN lamps, particularly high-wattage ones, tend to draw very high currents when they are cold, because they then have a very low resistance: of the order of 0.1 Ω or lower. If such a light is operated from a 24 V battery, a switch-on peak current of over 200 A may flow. This is highly detrimental to the life span of the light, which after only a few switch-ons may give up the ghost.

That costly situation may be prevented by gradually building up the power supplied to the lamp. Since operation is from a d.c. source, the only practical way of so doing is by pulse-width modulation. With that technique, the voltage to the lamp is switched from zero with increasing pulse width, while the current is smoothed by a coil, so that its average level increases gradually. Switching is carried out with two MOSFETs Type BUZ11. This type is characterized by an extremely low channel resistance, which is typically 0.03 Ω for a gate-source voltage of 15 V. Moreover, the can handle continuous currents of up to 30 A and pulsed ones of up to 120 A. By connecting the two in parallel, the current is split two ways: not exactly 50/50, of course, but near enough to ensure that on switch-on, the devices remain within their limits.

The control circuit for the MOSFETs, T2 and T3, provides nothing new. A regulator, IC1, ensures that the supply to the circuit cannot rise too high: it is set for about 18.5 V. The battery voltage cannot be used directly, because, among others, the gate-source voltage of the MOSFETs must not be higher than 20 V.

A sort of triangular waveform is generated by Schmitt trigger IC2a; R3 and R4 ensure that the operating point of the opamp is half the supply voltage. Because of the feedback via R6-C5, the output waveform is rectangular. However, the voltage across C5 is an exponential waveform, that is more or less triangular. That voltage is compared by IC2b with the terminal potential of C6, which, after switch-on, rises gradually.

As long as the voltage across C6 is lower than that at the non-inverting input of IC2b, the output of that opamp will be high. As soon as the inverting input reaches a potential higher than that at the +ve input, the output changes state, which is accelerated by the positive feedback via R10. The higher the terminal voltage of C6, the longer the output of IC2b will remain high. Eventually, it will reach a value higher than the maximum voltage across C5; pin 7 of IC2b is then permanently high. The lamp is then no longer switched on and off, but remains on.

The Type CA3240 dual opamp in the IC2 position has the advantage that its output can become almost zero, but the disadvantage that it cannot sink relatively large currents. Because of that, T1 ensures that the gate capacitances of T2 and T3, together 2-10 nF, are discharged rapidly. Those gate capacitances are charged again (that is, T2 and T3 are driven into conduction) by IC2b via D2. An additional driver is not needed, because the CA3240 can source enough current at high levels. When the FETs are on, their gate potential is about 16 V.

When the lamp is switched off, C6 is discharged immediately, so that the circuit is ready at once to switch it on again. To ensure that on switch-off the induced potential across L1 does not rise above the maximum level of the drain-source voltage (50 V), the coil is shunted by D3. This needs to be a fast type (25 ns or better) that can handle currents of up to 30 A.

Resistors R11 and R12 provide the lamp with some voltage before the MOSFETs switch on.

The speed at which the circuit switches the lamp on is preset with P1; normally, it will be sufficient to set this to the centre of its travel.

It is advisable to mount the FETs on a heat sink, although their dissipation is of the order of 1.6 W only.

Capacitors C1 and C2 must be able to handle high-frequency pulse currents of up to 30 A.

Inductor L1 must ensure that the lamp current does not exceed a predetermined value: the larger the in-
ductance, the lower the maximum level of the current. However, the physical dimensions of the coil must be acceptable. In the prototype, the maximum lamp current was set at 30 A. At a switching frequency of 7 kHz, an inductance of 30 \( \mu \text{H} \) is sufficient. Moreover, to avoid saturation problems, the coil is an air-cored one.

It is made by winding 45 turns of 1.5 mm (1/16 in) dia. enamelled copper wire in three layers on a 24 mm (15/16 in) dia. round former. During the winding, apply some glue from time to time to the turns.

The current drawn by the circuit is primarily that through the lamp: with a 250 W lamp and a 24 V battery, the current is some 10 A.

[J.J. Paauwe - 924060]

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**EXTRA BRAKE LIGHT**

There are many cars on the road that could do with an extra brake light, particularly one high up that can be seen by the third, fourth or fifth car behind it. That proposed here consists of a running-light bar of LEDs that starts again and again when all LEDs light. In some countries, this may not be allowed—check with your local highway or police department, but it can be used, anyway, in model cars.

Power is taken from terminals of the present brake lights. Because of bridge rectifier \( D_1-D_4 \), the polarity of that voltage is immaterial. The voltage is kept steady at 5 V by IC4.

An oscillator, based on IC1, provides clock pulses to shift registers IC2 and IC3 as long as the brake pedal is pressed. The p.p.r (pulse repetition rate) is set with \( P_1 \).

LEDs are connected to the outputs of the shift registers via series resistors. Because input pins 1 and 2 of IC2 are connected to the positive supply line, each clock pulse generates a logic high at successive output pins, so that more and more LEDs will light. Since the last output of IC2 is linked to input pins 1 and 2 of IC3, the outputs of that register will also be supplied successively with logic 1s when all LEDs connected to IC2 light. When, finally, the last output of IC3 goes high, \( T_1 \) is switched on, which results in a reset of the shift registers and all LEDs going out. If the brake pedal is still pressed, the LEDs will light one by one again.

Network \( R_{19}-C_9 \) ensures that the shift registers are reset and thus the LEDs go out the instant \( T_1 \) is switched on.

[Soumya Mitra - 924059]
TWO ladder networks and a buffer each form a volume control with a range of 63 dB. Network R3-R17 provides fine control in steps of 1 dB, whereas network R20-R34 provides coarse control in steps of 8 dB. The desired attenuation is set with the aid of multiplexers IC1 and IC3, each of which is controlled via three digital inputs. The design is such that the binary code on the six-bit wide overall control input accords with the set attenuation.

Resistor R1 ensures that C1 can discharge, even if K1 is open-circuit. This resistor and network R3-R17 form an input impedance of 46.3 kΩ. The resistor also determines the maximum permissible input voltage. That voltage depends, in the first instance, on the supply voltage to IC1 and IC3 (±12 V). Resistor R1 plus R5-R17 attenuate the input signal x2.4 (7.6 dB). This means that the maximum input level must not exceed 20 Vpeak, that is, 14 V r.m.s. That means also that IC2a must not amplify to prevent too high an input to the second ladder network and buffer.

The amplification of the two opamps is determined by R18-R19 and R35-R36 respectively. As stated before, that of IC1a should be unity, in which case R18=0 Ω and R19 is omitted. If the amplification of IC2b is also unity, the overall control range is -7.6 dB to -70.6 dB. To obtain a control range of 0-63 dB (when the binary code on the control inputs accords with the actual attenuation), IC2b should provide an amplification of x2.4.

The current drawn by the circuit is determined primarily by the dual opamp and amounts to about 10 mA.

The overall distortion is <0.003% over the range 20 Hz to 20 kHz and an input signal of 1 V.

The control has one, small, drawback: when the volume is set, weak clicks occur (which are typical of all normal CMOS switches). That makes it less suitable for super-de-luxe applications, although many listeners will not even notice the clicks. And, in any case, the volume is not varied constantly.

(P.C. Hogenkamp - 924063)
AUDIBLE FLUID LEVEL INDICATOR

A 600 Hz signal at a level of 2.4 Vpp, generated by the oscillator on board an LM1830 (National Semiconductor) is applied to a probe via C2. The probe is immersed in the liquid whose level is to be monitored. Because of C2, there is no direct voltage at the probe, so that there are no electrolysis problems.

As long as the probe makes no contact with the (conductive) liquid, the signal level at the input of the detector is equal to the level of the oscillator signal. When the liquid touches the probe, the detector input is connected to ground (or nearly so). This causes the level at pin 10 to drop. When it becomes more than 0.6 V lower than the oscillator signal, the detector switches on the internal output transistor in the rhythm of the oscillator frequency, since that is not suppressed by the detector.

The consequent signal at pin 12 is used to drive a simple output stage, T1, which drives a small loudspeaker, LS1.

The supply for the circuit is best taken from a 9-V PP3 battery. In quiescent operation, the current drain is 3 mA; when the alarm sounds, the current rises to about 80 mA.

SUPER STARTER FOR CARS

The super starter makes it possible to start cars with ageing batteries and obsolescent (coil-based) ignition systems, particularly during cold or damp weather. During starting, the voltage of an ageing (and possibly cold) battery will be insufficient for the coil to generate a tension high enough to create a strong spark across the spark plugs. The circuit presented ensures that the coil is powered by a battery of NiCd cells; even in these arduous conditions, such cells will last up to ten minutes. After the engine has started and the dynamo voltage has risen, the coil is powered by the car battery again.

The circuit uses the D+ terminal of the charging current indicator lamp (Lai) to check whether the engine is running, since that terminal is connected directly to the dynamo. As long as the engine does not run, and the dynamo, therefore, does not generate a voltage, relay Re1 is energized via the ignition key (connection) and the low-resistance dynamo (connection to earth). The NiCd battery then provides plenty of power to the coil, irrespective of the state of the car battery. Once the engine has started, a voltage will be generated by the dynamo. There is then no potential difference across the relay coil and its contact changes over, whereupon the coil is supplied by the car battery (or the dynamo).

Diode D1 prevents two possible troubles. First, it prevents the relay interrupting the current to the coil when its contact changes over, which would result in a spark at a moment that the engine does not need one. Therefore, the coil is powered via the diode during the change-over period. This reduces the voltage across the coil by about 2 V, but that does not matter. Secondly, it ensures that the car can be started when the car battery is fully charged and the NiCd cells are flat, or have been removed for charging.

The relay must have a contact rated at not less than 8 A. Car relays with change-over contacts are not easily obtainable as a spare part, but can be often be found in car scrap yards (particularly in Citroen CX models).

The Type FR606 diode may be replaced by a Type BYW29-100; both can handle currents up to 6 A and their reverse voltage is high enough to withstand the inductive peaks generated by the coil.

The circuit is best built in behind the dashboard, although it is advisable to place the NiCd cells in a removable holder to enable them to be charged externally. It is, of course, possible to charge them from the car battery via a suitable resistor. The D+ connection at the charging indicator lamp is that which is at earth potential when the engine is not running, but the ignition is switched on. At the ignition switch is a cable that goes to the coil; that cable must be connected to the output, 15, of the circuit. The Ω terminal of the circuit must be connected to the freed contact at the ignition switch. The -ve line of the circuit must be connected to the car chassis.

ELEKTOR ELECTRONICS JULY 1992

[914110]

[L. Lemmens - 914110]

[J Vaessen - 924003]
**SIMPLE SIGNAL GENERATOR**

This signal generator provides a 440 Hz sine wave output at two levels. The power supply may lie between 1.5 V and 16 V, so that even a single 1.5 V battery can do.

Opamp IC1a operates as a rectangular-wave generator: the values of R4 and C1 determine the frequency at which the output of the device toggles. Preset P1 enables the output to be set as a square wave (it may be adjusted by listening for minimum distortion).

Network R5-R6-C2 reduces the output of IC1a by 3 dB (50%), after which the signal is superimposed on half the supply voltage (derived from the average level of the rectangular signal by C2). That voltage is needed for setting the d.c. operating point of IC1b.

Opamp IC1b forms a third-order Chebyshev filter with a cutoff frequency of 400 Hz. This filter removes the majority of harmonics from the rectangular signal, so that the output is a reasonably clean sine wave.

The level of the output signal is selected with S1 from potential divider R9-R10-R11, depending on the requirements of the circuit on test. With a power supply of 16 V, the output level is 1.5 V r.m.s. or 30 mV r.m.s.; with a 1.5 V supply, the output levels are 150 mV and 3 mV. The output frequency is somewhat dependent on the supply voltage and varies from about 440 Hz at 16 V to around 370 Hz at 1.5 V.

The circuit draws a current of 300 µA at 16 V and 80 µA at 1.5 V.

(C. Sanjay - 924026)

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**STARTER FOR MODEL AIRCRAFT**

MODEL aircraft tend to be realistic reproductions of the real thing. That means, among others, that starting the petrol engine must be done by hand or with an external, electric starter. Manual starting may be realized in two ways: with the circuit in Fig. 1 or with that in Fig. 2.

The circuit in Fig. 1 places a 3-V relay between the starter and the motor. When the starter switch is closed, nothing happens. But if then the propeller is turned by hand, the motor acts as a generator and, once it is turned fast enough, the generated voltage is high enough to energize the relay, whereupon the motor starts. An advantage of this is that it is a pure mechanical operation, which is readily incorporated. A disadvantage is that once the relay is energized, it is operated by a 6-V supply: a slight waste of energy.

The circuit in Fig. 2 is rather more economical, since it uses a 6-V relay. Again, the motor is used as a generator operated by turning the propeller. The generated voltage is applied to the base of T1 via R2. When the base is supplied with sufficient current, the transistor is switched on, the relay is energized and the motor starts. The starting point can be preset with P1.

(G. Bartelt - 924008)
FUSE MONITOR

When an appliance ceases to operate, there may be various causes for this, one of which is the blowing of the mains fuse. The monitor proposed here contains an LED that lights when the fuse is blown. It is suitable for use with fuses rated from milliamperes to amperes.

As long as the fuse is intact, the full mains voltage exists across C2-R3-D3. Capacitor C2 and resistor R2 serve to limit the base current of T1. Diode D3 prevents C2 from being charged, which would cause the base current to quickly drop to zero.

Capacitor C1 and resistor R2 limit the current through D1, while D2 ensures that the voltage across the LED does not exceed 2.7 V. At the same time, D2 prevents C1 from being charged.

As long as the mains voltage exists at junction F1-C2, transistor T1 conducts and short-circuits D1 and D2. When F1 blows, T1 is switched off, whereupon current flows through D1 and D2: the LED then lights.

Resistor R1 must conform to relevant safety regulations. Furthermore, capacitors that still carry the mains voltage after the appliance is switched off, must become discharged (via R1) within a stipulated time.

When the monitor is in use, remember at all times that certain of its parts are at potentially lethal, mains voltage.

PC COOLING FAN CONTROL

Most PCs are provided with a cooling fan to ensure that the internal temperature does not rise unduly. Unfortunately, in many PCs the fan noise soon becomes an irritant. Since for a large part of the time the fan cools the PC more than is necessary, it seems sensible to make the speed of the fan dependent on the ambient temperature. That is the purpose of the circuit shown.

The circuit, designed with discrete component, is intended for the control of 12-V fans that do not draw a current exceeding 200 mA.

To ensure that the fan operates satisfactorily in all circumstances, the supply to it must not drop below its starting voltage. That voltage is equal to the 12-V supply less the 'zener' voltage of T3-R6-R7. With values shown in the diagram, the supply to the fan will be at least 7 V.

If the fan does not start at 25 °C, replace temperature sensor temporarily by a 1.8 kΩ resistor and lower the value of R7. If the fan runs too fast, raise the value of R7.

Transistors T1 and T2 compare the fixed potential at junction R3-R4 with the temperature-dependent one at junction R1-R2. It may be found convenient initially to place 25 kΩ potentiometer meter in the R2 position, adjust this till the fan runs correctly, measure the resistance and then replace it by a fixed resistor of that value.

Place the temperature sensor in the warm air flow of the fan. When the computer is switched on, the speed of the fan, owing to C1, will be fairly high, but will soon drop to a minimum. With a thermometer, measure the temperature of the outflowing air close to the sensor. When the temperature has reached a value of about 35 °C, the control circuit should come into action, indicated by an increase in the speed of the fan or its supply voltage. If that does not happen, change the value of R3, or adjust the potentiometer in its place. When the temperature rises, the speed of the fan will increase. The maximum speed will be only slightly lower than that without the control circuit. This is thanks to the fact that T3 can be driven into hard conduction, so that the drop across it is only some tenths of a volt.

ELEKTOR ELECTRONICS JULY 1992
TELEPHONE GONG

Those of you who like the telephone bell louder may find this circuit of interest. It uses the telephone bell signal to actuate an oscillator, which in turn drives a relay that operates a standard door-gong.

The oscillator is an RC type based on a 4093 Schmitt trigger. Both the charging and discharge times of C4 are set with presets. The oscillator is followed by a kind of buffer that energizes the relay via T1.

The telephone bell signal is applied to terminals a and b. Capacitors C1 and C2 in the two lines isolate the circuit from the telephone network, at least as far as d.c. is concerned. To prevent the circuit responding to speech signals, level thresholds are provided by zener diodes D1 and D2. Network R1-D2 limits the 120-150 Vpp bell signal to about 12 V. That signal is rectified by D4 and smoothed by C3, after which it is used to switch IC1a. Capacitor C3 is discharged rapidly via R3 when the bell signal ceases. Resistor R2 prevents too high a drive voltage when there is speech on the input lines.

The relay contact is connected across the terminals via which the gong is operated. The circuit may be fed by the same transformer used for the gong. When the relay is energized, the total current drawn is only 35 mA.

The setting of presets P1 and P2 depends on individual taste how long the on and off periods of the gong are desired to be.

(T. Giesberts - 924050)

BOUNCE-FREE CHANGE-OVER SWITCH

Any push-button switch can be used as a bounce-free, change-over switch with the aid of two D-type bistables contained in a 74HCT74 and some external components.

In the circuit diagram, IC1b provides the change-over function. The Q output (pin 8) of this bistable is interconnected with its D input (pin 12), which results in the logic levels at the Q and D outputs alternately changing state when a leading transition (edge) appears at its clock input (pin 11).

Circuit IC1a serves as pulse generator and debouncing element. The push-button switch S1 is connected between its reset input (pin 1) and earth. Normally, because of R3, there is a high level at pin 1. When the push-button is pressed, IC1a is reset.

The clock input (pin 3) is also connected to the switch via R1-C1. When the switch is operated, C1 discharges rapidly via D1; when the switch is released, it takes a little while before C1 is recharged to a logic high level.

When S1 is open, pin 9 (IC1b) is low, while pin 8 and pin 5 (IC1a) are high. When the switch is closed, IC1a is reset immediately, resulting in pin 5 and pin 3 going low. When S1 is released, the reset is removed, but it takes a little while before C1 is charged to a logic high level. Only when that level is reached, and a leading transition appears at pin 3, does pin 5 go high again. This results in IC1b being clocked, whereupon its Q outputs change state.

(A. Rietjens - 924013)
FREQUENCY PROBE

The frequency probe enables you to 'listen in' to the speed of your computer. It is, however, also suitable for use with other digital circuits, because it makes high frequencies audible, so that signals can be conveniently monitored.

A 12-bit counter serves as 'frequency detector'. The signal measured in a computer or digital circuit is divided by 1024 and is output at pin 14. It is then used to control a transistor, T1, which in turn drives piezo buzzer B2.1.

The scale factor has been chosen to convert MHz into kHz, so that the clock frequency of, say, an XT computer will be heard as a shrill 8 kHz tone. If higher frequencies need to be monitored, scale factors of 2048 and 4096 can be obtained by connecting R2 to pin 15 or pin 1 respectively. If an HCT circuit is used, the measuring limit is some tens of MHz. For frequencies <4 MHz, a standard Type 4040 may be used: that has the advantage that the supply voltage need not be exactly 5 V.

The supply connections and the probe are best made from flexible wire terminated into crocodile clips.

SCANNER FOR PREAMPLIFIER

The scanner is an extension for the 'All-solid-state preamplifier' published in the December 1989/January 1990 issues of this magazine. As its name implies, it scans all inputs of the preamplifier to ascertain where there is an audio signal present. That input remains selected. After there has been no signal for some time, scanning is resumed.

The scanning action is provided by rectangular-wave generator IC2c. The output of this oscillator is applied to one of the input selector keys via buffer/inverter IC2d and diode D3. The diode prevents the key being disabled when the oscillator is off. The oscillator is switched on and off by IC2b, which in turn is controlled by IC1b. That opamp is configured as a comparator whose voltage threshold is preset with P1.

The inputs of the scanner are linked to the audio inputs on the volume control board of the preamplifier. The relevant signal is amplified x40 by IC1c and IC1d, after which summing amplifier IC1a combines the left-hand and right-hand signals. As soon as there is music or speech on the input lines, C1 will partly discharge rapidly. When the voltage across the capacitor drops below the level set by P1, IC1b disables the oscillator via IC2b, and the input selected at that instant remains actuated. As long as there is a signal coming in, part of the charge on C1 will ebb away via IC1b. If there is no signal for some time (presettable with P2 between 3 s and 25 s), the capacitor will be charged almost completely via P23, R8 and D1. Once the terminal voltage of C1 rises above the comparator threshold, the oscillator is enabled again and the scanning action resumes.

A total scan of all inputs is completed in 3 s (determined by R1-C2). The input sensitivity can be set between 10 mV and 4 V with P1. The current drawn by the scanner is not greater than 10 mA. 

(L. Soete - 924053)
STEREO PROTECTOR AGAINST D.C.

If a d.c. coupled output amplifier breaks down during operation, the loudspeakers, particularly the bass units, are at risk. The bass particularly so because it is not decoupled for d.c. by the capacitors in the crossover network. If, for instance, the output transistor has given up the ghost, the bass units will get the full d.c. supply voltage at their terminals.

A suitable circuit to protect the loudspeakers in such an eventuality, and at the same time to obviate the annoying 'plops' on switch-on is shown in the diagram. Interestingly, it operates from an unregulated, non-symmetrical power supply. Normally, it may be powered directly from the power supply of the output amplifier.

The a.c. component of the signals in the output stage is bypassed by R1 and the two anti-series connected capacitors, C2 and C3. The signal at the junction R1-R2 is, therefore, the d.c. component of the loudspeaker signal. From there it is applied to potential divider R2-R3 and then to window comparator IC1a and IC1b. Since the supply voltage is fixed at 10 V by R13-D7, the window height is fixed at 2 V by R5. In other words, \( u_2 = 6 \text{ V} \) and \( u_3 = 4 \text{ V} \). In the absence of d.c. at the output of the power amplifier, \( u_1 = 5 \text{ V} \). In this situation, the outputs of 'OR gates' D1 and D2 is logic high.

When the d.c. component at the output of the power amplifier is greater than ±2 V, \( u_1 \) is greater or smaller than either \( u_2 \) or \( u_3 \). The output of one of the opamps will then be logic low.

When the power amplifier is switched on in step with the present circuit and \( u_1 \) lies within the window, \( C_4 \) is charged via \( R_8 \). After about 1.5 s, 'Schmitt trigger' \( IC_{1d} \) changes state and its output becomes logic high. The relay is then energized and connects the loudspeaker to the power amplifier: no 'plop'.

If a defect occurs, or if the direct voltage at the output of the power amplifier rises, \( C_4 \) is discharged via \( R_7 \) within 50 ms. The output of \( IC_{1d} \) then goes low, the relay is deenergized and the loudspeaker is disconnected from the output amplifier.

Resistor \( R_{13} \) and the operating voltage of the relay must be suitable for the supply voltage. If that voltage is 20-40 V, a good value and rating for \( R_{13} \) is 4.7 kΩ, 1 W, while for 12-20 V, 1 kΩ, 1/4 W is right. If the supply voltage is, say, 36 V, the operating voltage of the relay should be 24 V. The difference of 12 V should be dropped across a suitable resistor.

If there is likely to be a requirement for switching off the circuit, \( S_1 \) should be incorporated. When that switch is closed, the relay is energized.

For a circuit suitable for a stereo power amplifier, only components \( R_1-R_3, C_2, C_3, D_1, D_2, D_5, D_6, IC_{1a} \) and \( IC_{1b} \) need to be duplicated. The additional circuit is connected in parallel with \( S_1 \). Note that the relay should then have two working contacts or two relays with their contacts in series should be used.

[T. Schaerer - 924051]
SIMPLE POWER SUPPLY CONCEPT

The best known alternatives to a 'quick and dirty' power supply are the three-pin fixed voltage regulator and the zener-plus-transistor combination. While these basic circuits will suit a good many applications, they do have their limitations, which can be frustrating at times. For example, most types of fixed voltage regulators are limited to an output current of about 1 A only. Where more power is required, a 'current by-pass' transistor is often added. However, while this boosts the maximum output current, the regulation of the supply suffers. Fixed voltage regulators with higher output currents (say, 5 A) are no alternative because they are notoriously expensive.

The second alternative, the zener-plus-transistor circuit, has limited use also because of its relatively poor ripple rejection and insufficient stability at output load variations.

The PSU presented here suffers none of the disadvantages mentioned above, and is simple to memorize as a multi-purpose concept. It is the perfect low-cost supply for a host of applications. At first glance, the circuit looks very much like the familiar zener-transistor combination. However, an essential difference is that feedback is implemented, which results in a 100-Hz ripple suppression of up to 55 dB—far more than can be achieved with the simple zener-transistor stabilizer.

The voltage reference used here is $D_1$, a TL431C from Texas Instruments. The internal structure of the TL431C is shown in the diagram. Here, $D_1$ supplies a base current to $T_1$ that results in 2.5 V across resistor $R_3$. This allows you to calculate the supply output voltage, $U_0$, from

$$U_0 = 2.5 \times (P_1 + R_2)/R_3\text{ volts}.$$  

The indicated component values result in an output voltage of 12 V. For other output voltages, simply adapt the output voltage divider, making sure that the current through $P_1$, $R_2$ and $R_3$ is at least 1 mA. This is required to ensure that the current flowing into the reference input of the TL431 is negligible (approx. 2 $\mu$A). The power transistor is a Darlington with a guaranteed current gain of 1000 or greater at an emitter current of 5 A. This means that only 5 mA of base current is required. Although this is not much, it has to be taken into account when $R_2$ needs to be given a different value. Also, $D_1$ requires a minimum cathode-anode current of 0.5 mA, which results in a total, minimum current of 5.5 mA through $R_1$. This design information, together with the lowest possible input voltage, $U_{in}$ (measured across $C_6$), and the base-emitter drop of $T_1$ (approx. 2 V), results in a theoretical value of the current limiting resistor:

$$R_1 = (U_{in} - U_{be} - U_0)/I_b\ \Omega.$$  

Because the current gain of the Darlington may be up to two or three times the guaranteed value mentioned above, it is often possible to give $R_1$ a higher value than calculated. Since a higher resistor value results in lower dissipation of $R_1$ and $D_1$, some experimenting is certainly worth while.

The PCB designed for the supply accommodates the complete rectifier section, that is, a bridge rectifier, a buffer capacitor and a fuse. The buffer capacitor, $C_1$, and the on-board heat sink for $T_1$ are large enough for output currents up to 2 A.

As already mentioned, this PSU is a concept. Those of you who do not need the rectifier section may omit it, and connect a d.c. voltage of 16 V to $K_1$. Note, however, that this requires wire links to be fitted in the positions indicated with dashed lines near the bridge rectifier.

If you require more output cur-
rent (say, up to 5 A), simply move the power transistor off the board, and fit it on a larger heat sink (see parts list). Also, increase the buffer capacitor to 10000 µF. Since such a capacitor (or array of capacitors) will not fit on the board, it is connected as an external part via heavy-duty wires and two spade terminals (marked ‘+’ and ‘-’ on the component overlay). A continuous output current of 5 A also requires the bridge rectifier to be cooled a little. This is best achieved by leaving it on the PCB, and clamping it on to a side panel of the metal enclosure used to house the supply.

(J. Ruiters - 924024)

**PARTS LIST**

<table>
<thead>
<tr>
<th>Resistors:</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>R1 = 470Ω 0.33W (see text)*</td>
<td></td>
</tr>
<tr>
<td>R2 = 6kΩ</td>
<td></td>
</tr>
<tr>
<td>R3 = 2kΩ</td>
<td></td>
</tr>
<tr>
<td>R4 = 1kΩ</td>
<td></td>
</tr>
<tr>
<td>P1 = 2kΩ5 preset H</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Capacitors:</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>C1 = 470µF 40V</td>
<td></td>
</tr>
<tr>
<td>C2 = 10µF 35V tantalum</td>
<td></td>
</tr>
<tr>
<td>C3: C5 = 100nF</td>
<td></td>
</tr>
<tr>
<td>C4 = 100pF 40V</td>
<td></td>
</tr>
<tr>
<td>C6 = 10µF 40V</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Semiconductors:</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>D1 = TL431C</td>
<td></td>
</tr>
<tr>
<td>D2 = LED, green, 3mm</td>
<td></td>
</tr>
<tr>
<td>T1 = MJ3001</td>
<td></td>
</tr>
<tr>
<td>B1 = B805000/3300</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Miscellaneous:</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>K1:K2 = 2-way PCB terminal block, pitch 5mm.</td>
<td></td>
</tr>
<tr>
<td>F1 = 2.5A fast fuse (6.3A)* and PCB mount holder.</td>
<td></td>
</tr>
<tr>
<td>Heat sink: SK201 (6KW) or SK71/75mm* (1.25KW).</td>
<td></td>
</tr>
<tr>
<td>Two ‘fast-on’ spade terminals for PCB mounting *.</td>
<td></td>
</tr>
<tr>
<td>Printed circuit board 924024.</td>
<td></td>
</tr>
</tbody>
</table>

* for 5 A version only.

**EXPERIMENTAL FAST NICD CHARGER**

The perpetual difficulty in designing fast NiCd chargers is determining when the battery is charged, that is, when to stop charging. The charger presented here is based on the latest developments as reported by several manufacturers. It is not at all certain whether, and under what conditions, the circuit will consistently give satisfactory results. The battery is charged with a current (in mA) that is ten times its nominal capacity (in mAh). That means that, for instance, an HP7 (AA; RG) type battery is charged at 5 A, a current 100 times larger than used in standard charging. The charging is controlled by Type 555 timer IC, here connected as an astable. If the IC’s output is high, charging takes place. There is, however, a fixed period of time (=R6C3) during which there is no charging. As soon as charging stops, C1 is connected across the battery by electronic switch IC3a. Its terminal voltage is then compared by IC1a with the maximum battery voltage set by P1. The output of the comparator is integrated by R3 and C2 and then used to determine the period of the astable. If the maximum battery voltage has not been reached, charging takes place for about 90% of that time. If the maximum battery voltage has been reached, charging takes place for 1% of the time (trickle charging). Do not leave the battery connected to the charger unnecessarily: when the LED lights, the battery is fully charged. During charging, owing to a variety of resistances, primarily in the supply leads and connections, the battery e.m.f. is an unreliable yardstick for determining the state of charge of the battery. Therefore, the e.m.f. is taken immediately after a burst charge, because then the voltage can be measured exactly. The important question is, of course, to what e.m.f. P1...
should be set: in other words, what is the correct battery e.m.f.? Opinions vary, but with our prototype we had good results with a value of 1.42 V at room temperature (21 °C).

The circuit draws a current of only 10–15 mA, which may be obtained with a 7805 regulator.

The proposed charger is intended for charging one 1.5 V NiCd battery in 8–10 minutes. The charging current for a 500 mAh battery is about 5 A, which need not be regulated, since it will be limited by R1.

The value of R1 is given by Ohm’s law. If, for instance, the charging current is drawn from an 8-V source, and assuming that the drop across the battery and T1 is 2 V, the voltage across the resistor is 6 V. Its value should thus be 6/5=1.2 Ω. Bear in mind that the power dissipated in it is 6x5=30 W; you will, therefore, have to connect a number of resistors in parallel.

If you want to charge a number of batteries in series, raise the level set by P1 accordingly (at 1.42 V per battery). Note, however, that you should use only batteries that have already been sorted for equal capacity by the manufacturer. Also, the supply voltage to the charger circuit must always be 2 V higher than the level set with P1.

Finally, after every five fast charges, give the battery a ‘normal’ (1/10 mAh over 14 hours) charge.

(K. Walraven - 924085)

**CONTINUITY TESTER**

A continuity tester is very useful for checking printed-circuit boards. It indicates a sound connection by a squeak from a buzzer: there is, therefore, no need to continually watch a meter.

The voltage across test terminals TP1 and TP2 is only 80 mV. That is not sufficient to test diodes, but it obviates the risk of damage to electronic components.

The design consists of a comparator, IC1, and an astable consisting of T1 and T2, whose frequency is around 1250 Hz for a supply voltage of 3 V. The astable is actuated the moment the output of IC1 (pin 6) goes logic high. This happens when the resistance between TP1 and TP2 is low, so that the voltage at pin 2 of the comparator is lower than that at pin 3, which depends on the setting of P1.

The circuit is powered by a 3-V battery (two 1.5 V cells in series), but may be maximum 12 V. At such a high supply voltage, it may be that the tone of the buzzer is too high; in that case, the values of C1 and C2 should be increased.

If a Type TLC271C is used instead of a TLC251 for IC1, the supply voltage must be not lower than 3 V. At 3 V, the circuit draws a current of about 1.4 mA.

(C. Sanjay - 924002)
NOISE GENERATOR

Noise generators are used for measuring the self-noise of amplifiers and receivers and for some acoustic measurements. The noise of traditional low-frequency noise generators is based on the stochastic properties of an ion current resulting from a gas discharge. A simple noise generator can, however, be designed without a special gas discharge tube: the reverse-biased base-emitter junction of a bipolar transistor a compact and inexpensive alternative.

In the circuit diagram, the noise voltage is taken from the emitter of $T_1$. The base-emitter junction of this p-n-p transistor begins to behave like a break-down diode at a reverse bias of about 9 V, but it really starts to generate noise at 10 V.

To ensure that the circuit works satisfactorily with a battery supply of 9 V, a step-up generator, based on IC$_1$, is used. This stage provides a rectangular output voltage at a frequency of around 2750 Hz.

The diode pump, consisting of C$_1$, C$_3$, D$_1$, and D$_2$, doubles the battery voltage, which results in a stable direct voltage of 10 V across D$_3$.

Low-pass filter R$_{17}$-C$_6$ prevents frequency components of the rectangular-waveform generator appearing in the noise spectrum.

Each of the opamps in IC$_2$ raises the noise voltage in the frequency range from 01 Hz to 300 kHz tenfold. The amplitude of the output voltage can be preset with P$_1$. The noise signal can be tested with the a.c. buzzer by closing S$_2$.

With a fresh 9-V battery, the circuit draws a current of 5-6 mA.

240 VAC-TO-110 VAC CONVERTER

From time to time one comes across appliances in second-hand goods stores that were designed for operation from a 110 V a.c. (50/60 Hz) supply. If such an appliance is a pure resistive load, such as a radiant fire, a soldering iron, or a melting furnace, the circuit shown may be found useful. Strictly speaking, it is a dimmer set so that the output voltage has an r.m.s. value of 110 V. It is, of course, possible to set it to a different output voltage if so desired.

To obtain an r.m.s. voltage of 110 V across the load, the phase angle at which the triac is switched on must be about 110°. There is no guarantee that this will be met exactly by the present design: owing to tolerances of the various parts, the phase angle may be quite different so that the r.m.s. voltage will be higher or lower than 110 V. It is, therefore, essential, to check the actual voltage across the load. Bear in mind that the circuit carries mains voltage and is thus potentially lethal. Checking the phase angle with an oscilloscope cannot be carried out safely without special precautions. The safest and most accurate way of measuring the voltage across the load is with the use of a true-r.m.s. voltmeter (which shows the r.m.s. value also of non-sinusoidal voltages). If the voltage across the load is not correct, the value of R$_2$ must be altered.

If you have no true-r.m.s. meter to hand, checking may be done in a slightly more primitive way. Use an incandescent 5 W, 240 V bulb as the load and place a thermometer close to it. Switch on the converter, wait till the thermometer gives a stable reading and note that reading (if the thermometer goes off its scale, place it a little further away from the bulb). Do not change the distance between the bulb and the thermometer, and connect a second 5 W, 240 V bulb in series with the first. Once the first bulb has cooled down sufficiently, connect the two in series across the 240 V mains supply (when the bulbs

ELEKTOR ELECTRONICS JULY 1992
will each drop 110 V). If the thermometer after a while has the same reading as before, you can be pretty certain that the converter provides an r.m.s. voltage of 110 V.

The converter may be remotely switched on and off via a direct connection between a suitable 240 V switch to the REMOTE terminals. The wire link should then, of course, be removed. It may also be switched on and off by a voltage of 3–32 V as shown. This optoisolator circuit has the great advantage of isolating the circuit from the mains.

If a triac Type TIC226 is used, the converter can handle currents of up to 2 A. If the triac is mounted on a heat sink, the current may go up to 4 A.

[J. Vanden Berghe - 924035]

VIDEO ENHANCER

The enhancer amplifies the high frequencies of a video signal, resulting in a sharper picture. It may be inserted between, say, the video recorder output and the SCART input of a television receiver.

The simple design is based on only three transistors. The first, T1, is a buffer. Resistor R1 ensures that the input impedance is of the order of 75 \(\Omega\). The signal is then applied to amplifier T2, whose gain is determined by the setting of P2.

The frequency characteristic of the signal at the base of T2 is shaped by P1, R6, and C6, and is, therefore, to a certain extent under the control of the user (by P1).

Buffer T3 provides sufficient current for correctly driving most 75 \(\Omega\) loads.

Preset P2 must be set to give an output voltage of 1 Vpp (terminated output; for an open-circuit output, the level should be 2 Vpp).

The enhancer draws a current of about 50 mA. Note that the 12 V supply should be regulated.

[J. Bodewes - 924079]
VOLTAGE CONVERTER II

Although a 9.5–24 V direct voltage can be brought down to 5 V easily by a standard regulator, the converter described here has the advantage that, since it is a switch-mode type, it hardly dissipates any heat. Its maximum, steady output current at 5 V is 250 mA, although it can cope with peaks of up to 750 mA.

The converter is based on a Motorola Type MC34161 circuit, to which a power stage Ti has been added. Inductor L1, C5 and D1 remove any ripple from the output.

The internal comparator at pin 2 of IC1 is connected to the output of the converter via potential divider R4–R5 to monitor the output voltage. The second comparator (pin 3) is used in the oscillator circuit and connected to pin 6 direct and to pin 5 via R3–C3.

When the supply is switched on, the output of the converter, and thus the voltage at pin 2, is 0: the oscillator operates normally. Transistor T1 charges C5 via L1. When T1 is off, L1 ensures a supply of energy to C5 via D1. As soon as the terminal voltage of C5 has risen to a sufficiently high level, the internal comparator at pin 2 changes state. The oscillator is then switched off via pin 6, so that T1 is also off. After C5 has discharged to an extent that its terminal voltage drops below the preset level, the oscillator is re-enabled, and C5 is charged again via T1.

The values of L1 and C5 determine the switching frequency: with values as shown, an input voltage of 12 V and a load current of 250 mA, the frequency is 18 kHz. At higher inductions and input voltages, the frequency drops.

It is essential that all earth connections are taken to the negative terminal of C5 as shown in the diagram.

The inductor is a standard triac choke, to which a number of turns have to be added. The inductance of the choke is L µH and its number of turns is n, the number of turns, n' required for the present inductor is given by n'=(470/L).

(Motorola Application - 924077)

CURRENT LIMITING FOR LM317 REGULATOR

Although the well-known Type LM317 voltage regulator is already short-circuit proof, there are cases where a limit of the heavy short-circuit current can be desirable. As the diagram shows, such a current limiting facility can be provided in a simple manner. Use is made of the fact that the output voltage, U0, is dependent on the feedback to the control input. As long as the current limiting does not operate, resistors R2 and R3, as well as T1, may be ignored. The output voltage is then:

\[ U_0 = 1.25(1+P_1/R_1)+U_{adj} P_1 \] (volts).

Since the maximum level of \( I_{adj} \) is 0.1 mA, \( P_1 \) can set \( U_0 \) to 1.25–27 V.

When the current through the regulator causes a drop of about 600 mV across R3, T1 will come on. This will cause a drop in the level at the control input of the regulator, and thus in the output voltage. With a value of R3 as shown, the current limiting will come into operation at a current of 0.6/4.7=120 mA.

(National Semiconductor Application - 924100)
FRONT-TO-REAR WIPER COUPLING

On some cars, it is convenient to couple the rear window wiper to the windscreen wiper. However, since the rear window does not get nearly as wet as the windscreen, particularly when the car is moving, the rear wiper should operate only once for every umpteen wipes of the windscreen. Note that your car may be one of the fortunate ones of which the rear window, when the car is moving in the wet, does not get wet at all because of the car's design.

The coupling shown in Fig. 2 ensures that the rear wiper operates once for every four or 16 wipes of the windscreen, depending on the setting of switch S1 (as shown, once every four).

The clock for the circuit is taken from the return (terminal 53-e – green/black wire on most cars) of the windscreen wiper motor—see Fig. 1. This signal, which is a square-wave, is applied to IC1 via K1. Its level is lowered to not more than 5 V by potential divider R1-R2 to prevent any damage to IC1. Any noise from the car's electrical system is bypassed by C1.

The Q2 output (pin 1) of IC1 goes high every fourth clock input, and the Q4 output (pin 3) once every eighth clock input.

The trailing edge of the signal at S1 is transformed into a trigger pulse by R3-C4. This pulse, whose length is determined by R7-P1-C6, is applied to monostable IC2. The length of the pulse should be set to about one second to give the rear wiper time to get going.

The monostable drives transistor T1, which in turn controls relay Re1. This relay is a motorcar type that can handle the large switch-on current of a wiper motor; it may be rated at 25 A or 35 A.

The supply for the circuit is provided by IC3, which brings down the car voltage of 12 V to 5 V. This IC also prevents large peaks on the battery voltage from reaching the circuit.

[I. Fietz – 924023]
THREE-PHASE SIMULATOR

MOST domestic consumers (in the UK) are provided with a single-phase supply, unless exceptionally heavy loading is foreseen. It may, however, occur that a low-voltage three-phase supply is required for experimental purposes and in such cases, the simulator can prove useful.

The source signal for the phases, R, S and T is generated with a standard Wien bridge. The sine-wave generator is formed by IC1a. Preset P1 enables the frequency to be set accurately to 50 Hz: the output level (pin 1) is set with P2 to 1 V (peak).

Circuit IC2a provides a constant load impedance for IC1a, which is important for the stability of the generated frequency. It also raises the signal level to 5.6 V (peak). The peak value of the phases is set to 0–12 V with P3. Series capacitor C9 prevents the offset voltage of IC1a and IC2a adding direct voltage to the outputs of IC2b and IC2d.

The R phase results from inverting the signal at the wiper of P3, that is, shifting it by 180°. Owing to low-pass filter R12–C11, the T output lags the signal at the wiper by 60°, while R9–C10 provide a 60° lead at the S output. There is, therefore, a 120° phase difference between any pair of phases.

Presets P4 and P5 need to be set only once and that in such a manner that the peak values of the three phases are identical.

Low-current LEDs D5–D7 light only if there is an alternating voltage present at the associated output. An accidentally short-circuited phase is therefore, detected immediately.

The opamps are short-circuit proof and can provide a current of about 10 mA. When a symmetrical ±15 V supply is used, the quiescent current is ±20 mA.

SIDAC NEON TUBE STARTER

THE sidac from Motorola is best compared with a triac of which the gate connection is missing. It switches on whenever the voltage across it exceeds a certain level. The polarity of that potential is immaterial. Unlike a triac, the sidac works equally well with direct and alternating voltages. Furthermore, when the sidac is on, it resembles a short-circuit and remains in that state until the level of the current drops below a certain value (the holding current), whereupon it switches off.

A series network of a sidac and a load connected to the mains results in a kind of dimmer whose, non-variable, phase angle depends on the starting voltage of the sidac. Sidacs are available for starting voltages between 104 V and 280 V.

A neon tube does not switch on as easily as an incandescent lamp because the tube can start only at a voltage much higher than the mains, after which it will remain lit at the mains voltage. The level of both the starting voltage and the working voltage depends on the temperature of the tube.

Normally, the high starting voltage is obtained by interrupting the current through a choke. This is usually done by the starter, which also ensures that a fairly large current flows through the filaments of the tube. This heats the ends of the tube, which makes starting easier.

These tasks of the starter are taken over by two 135 V sidacs (or a single 270 V one). The starting voltage is
thus 270 V, which is below the peak value of the mains (about 340 V), but higher than the working voltage of a 20–40 W neon tube.

As long as the tube has not started, almost the whole of the mains voltage is dropped across the starter. Assume for a moment that the polarity of the mains causes D_1 to be forward biased. When the instantaneous value of the mains voltage reaches the level of the starting voltage of the sidacs, these will short-circuit the starter; whereupon a fairly heavy current will flow through the filaments and the coil. This gives rise to a magnetic field around L_1. When the polarity of the mains voltage reverses, the positive current through L_1 will decrease gradually. When the level of the current approaches zero, the sidacs switch off. Whereupon the instantaneous negative mains voltage is applied across the tube immediately, because of C_1 being charged rapidly. This capacitor and the starter form a series resonant circuit that magnifies the sudden drop across the tube to way above the level of the mains voltage.

During the next positive period of the mains voltage, the sidacs switch on again, and the sequence repeats itself until after a few cycles the tube has warmed up enough to remain lit. The drop across the filaments does not exceed the starting voltage of the sidacs, so the electronic starter is switched off.

Capacitor C_1 not only suppresses any r.f. interference generated by the tube, but also makes the load on the mains supply less inductive (so-called cosφ improvement).

The capacitor and diodes can probably be fitted into the man-made fibre enclosure of the original starter.

(Motorola Application - 924106)

TOUCH ‘ON’, AUTO ‘OFF’ CONTROLLER FOR BATTERY-OPERATED EQUIPMENT

This handy circuit is intended for battery-operated appliances. It functions as a touch-operated supply on switch and delayed supply off switch. Figure 1 shows the design for applications that require only a few milliamperes. Figure 2 is identical but for the added FET at the output, which enables up to 300 mA to be switched.

The active electronics is formed by six Schmitt triggers contained in a Type 40106. The touch key consists of two small conducting plates that can be interconnected by our skin resistance. When the key is not touched, R_1 causes a high level at the input of IC_{1a}. That gate is followed by a diode, D_1, which ensures that C_1 can be charged only if the output of IC_{1a} is high. When the key is touched, C_1 is charged rapidly. The capacitor discharges slowly via R_2; the state of its charge is monitored by N_2–N_3–R_3.

The shunting of gates IC_{1b} and IC_{1c} by resistor R_3 greatly increases the hysteresis at the input of IC_{1b}. That means that the output of IC_{1c} goes high only when C_1 is almost fully charged, and changes state again only when C_1 is nearly discharged.

Gates IC_{1d–f} serve as output buffers.

The appliance connected to the output terminals will be provided with power as soon as the touch key is operated, because C_1 will then be charged rapidly. The discharge time of this capacitor (and thus the time that the appliance is powered) is fairly long and depends on the value of R_3 and the leakage resistance, and may, therefore, be extended by giving R_3 a higher value. Note that it is not advisable to increase the value of C_1 by much, because that increases the charging time and, worse, simply touching the key may not be sufficient any more. When C_1 is nearly discharged, the supply to the connected appliance is switched off.

(R. Evans - 924031)
Although most petrol-engined cars and lorries have a rev counter as standard, that is by no means the case in diesel-engined motor vehicles. The reason for this is that, since a diesel engine has no contact breaker, it is not so easy to derive pulses to drive a rev counter. There are, none the less, several possibilities of adding a rev counter if so desired.

Firstly, it would be possible to take the pulses from terminal W of the alternator. Unfortunately, that machine does not run at the same speed as the engine, so that some arithmetic unit would have to be added. Moreover, and more seriously, terminal W of modern alternators is no longer accessible externally.

A second way might be to attach a small magnet to each of the cranks on the crankshaft and so induce magnetic pulses in a fixed coil. The problem here is to attach these magnets securely.

A third method is an optical one proposed in this article. In this, the cranks are divided into sectors that are painted alternately white and black. A home-made light barrier is then used to evaluate the speed with which the sectors are rotating. If the cranks are divided into four sectors, pulses are generated that are suitable for driving commercially available rev counters, irrespective of whether the diesel engine has four, six or eight cylinders.

The circuit in Fig. 1 may, therefore, be considered as an adaptor for the rev counter.

A small 12-V bulb lights the crankshaft, whereupon the light reflected by the white sectors falls on to phototransistor T1. This transistor is connected in a darlington configuration with T2. The type of phototransistor is not important; it is thus not necessary to use the BP103 shown in the diagram. The output of T2 is applied via C2 to IC1, where it is chopped and amplified by about x50 to give a rectangular output signal of about 10 Vpp. That signal is perfect for driving a rev counter.

The circuit is best built on a small piece of prototyping (vero) board and then fitted in a tube, whose front is closed watertight by a circular piece of perspex. It may be necessary to separate the lamp and the phototransistor by a dark screen.

The circuit is connected to the rev counter by a three-core cable. The cable connections (+12 V, earth and pulse signal) to the circuit must be waterproof.

If the lamp is found to be too bright, it may be connected in series with a small resistor.

The circuit can be tested by measuring the voltage at the collector of T2, which should be 1-5 V when a white sector is being illuminated.

The maximum temperature at which the sensor can be used is 130 °C. The sensors are calibrated to an accuracy of ±0.25 °C during production.

In principle, it would be feasible to apply the rectangular signal to a moving coil meter. This would indicate a value that is directly proportional to the average voltage level of the rectangular signal and thus with the duty factor and the temperature.

The actual temperature data are stored in the duty factor (that is, the ratio of the pulse width to the pulse spacing). There is a linear relation between the temperature, T and the duty factor:

duty factor=0.32+0.0047xT.

Thus, at a temperature of -45 °C, the lowest at which the sensor can be used, the duty factor is 0.109.

The maximum temperature at which the sensor can be used is 130 °C. The sensors are calibrated to an accuracy of ±0.25 °C during production.

In principle, it would be feasible to apply the rectangular signal to a moving coil meter. This would indicate a value that is directly proportional to the average voltage level of the rectangular signal and thus with the duty factor and the temperature.
In practice, it is, however, much more sensible to connect the sensor to a digital input port of a peripheral interface or a microcontroller. Sampling the rectangular signal enables the computer system to carry out temperature measurements with a minimum of external components as shown in the diagram. A suitable PC measurement card was published early last year (see Ref. 1). Connector K₁ is a 26-way box header that is linked to connector K₆ of the measurement card via a short length of flatcable. The 5 V supply is taken directly from the computer. On the prototype, L₁⁻C₁⁻C₂ proved essential to prevent jitter of the signal, which caused the first digit after the decimal point to move to and fro.

The program for controlling shown here is also available, with other basic routines for the measurement card through our Readers' services (ESS 1753)

(J. Ruiters - 924110)

Reference
FUZZY LOGIC: AN INTRODUCTION

Fuzzy logic is a kind of statistical reasoning, whose foundations can be said to have been laid in the 18th century by the British philosopher Thomas Bayes. With this technique, large amounts of data can be condensed into a much smaller set of variable rules than with rigid logic. The result is an expert system that can process information faster, and provide a more flexible, more human-like response than conventional logic.

The great German polymath, Gottfried Leibniz (1646–1716), dreamed about devising a way whereby a couple of philosophers could discuss and settle any human argument once and for all by pure logic. But he and many other thinkers after him discovered that there are many problems that cannot be solved by just logic. This realization gave rise to another way of attempting to solve problems: the use of statistics. In statistical reasoning, probabilities express the idea of ‘perhaps’. One method of statistical reasoning, whose foundations can be said to have been laid in the 18th century by the British philosopher Thomas Bayes, is called fuzzy logic. In fuzzy logic, there is not just ‘true’ and ‘false’, 1s and 0s, ‘black’ and ‘white’, but also all the various grades of grey in between. Fuzzy logic can condense large amounts of data into a much smaller set of variable rules than rigid logic. The result is an expert system that can process information faster, and provide a more flexible, more human-like response than conventional logic.

For example, a washing machine controlled by fuzzy logic will wash very dirty clothes very hard: not-so-dirty clothes get a milder wash, and so on. Fuzzy logic is already being used in many domestic appliances, cameras and passenger trains.

Traditional control technology is based on a mathematical model that describes the control process. Although this is perfectly satisfactory for simple processes, it gets more difficult as the process becomes more complex. In such cases, the solution is derived from a simplified model or from a set of values that was determined empirically.

An example from everyday life would be when you are driving along in your car and you want to turn left (or right): without conscious calculation you determine the moment when you have to start turning the steering wheel. Without precise information of the width of the roads, the position of your car, the way the front wheels of your car react to the turning of the steering wheel, the wheelbase of your car, and so on, you will normally act so that you do not get on the wrong side of the road or on the pavement. In the same easy manner, you can steer your car, or that of your neighbours, through completely different bends. What you are doing is reacting in a ‘fuzzy-logical’ way to the effect of an action. You turn the steering wheel a little, your eyes register the effect of this and your brain corrects, if necessary and without complex analyses and calculations, the action. If we were to have this, to us simple operation carried out by a digital control system, we would have to design a surprisingly complex system that would, moreover, require a fairly large computer power. However, for a control system based on fuzzy logic, the rules would be based on human practical experience. For instance, at home:

- If the room temperature is much too low, turn up the thermostat to maximum;
- If the room temperature drops slowly, turn up the thermostat a little.

But how do we define too low, slowly, and a little? Fortunately, fuzzy logic can cope with these terms, as we will see later on.

Collecting data

An important principle of fuzzy logic is set theory (in mathematics, a set is a collection of elements chosen for membership of the set because it possesses some required property). This may be illustrated by, say, our desire to go out and buy fragrant red roses. We may go to a market and find a stall that sells flowers. We make our wishes known to the stall-holder, who subconsciously may reason: ‘if the flower is a rose, and if it is red, and if it is fragrant, then the customer will buy a bunch’. In other words, if the flower is an element of all three collections (rose, red, fragrant), it is the desired one. This is illus-

Fig. 1. Venn diagram of choices in a flower shop.
Fig. 2. Various basic operations in Set Theory.
FUZZY LOGIC: AN INTRODUCTION

In the logic rule we use to arrive at a final conclusion, we make use of a number of basic operations that are illustrated in Fig. 2, another Venn diagram. Figure 2(a) shows the simplest situation that can occur: from a set B, a new set A is formed, such that all elements of A are also elements of B. Such a set A is called a subset of B. Mathematically, this is expressed as \( B \supseteq A \), read as 'A is a subset of B'. A more general situation arises when two sets A and B are involved, each of which possesses elements that are not common to the other, so that neither \( A \supseteq B \) nor \( B \supseteq A \) is true. The set of elements C that is common to the two sets is called the intersection of sets A and B and is written \( C = A \cap B \); this is illustrated in Fig. 2(b). Another important set related to sets A and B is the set C containing all the elements belonging to both A and B. This is called the union of sets A and B and is written \( C = A \cup B \), read as 'A cup B'; it is illustrated in Fig. 2(c). Finally, in connection with sets A and B, there is the complement of B relative to A, which is written as \( A \setminus B \) and read as 'A minus B'; this is illustrated in Fig. 2(d).

All this is still clearly defined, but in the earlier instance of the red roses, we could ask: 'What is red; where does pink begin?'. In general, the colours red and pink will be recognized as such by most people, but in between there is a range of hues that is not clearly red or pink. That sort of problem is solved by the use of fuzzy sets, in which a clearly red flower is entirely common to the red set and not at all to the pink set. A flower with a colour in between red and pink is common to both sets, for instance, 70% red and 30% pink. This is called the degree of association, \( \mu \). For example, the degree of association of an element \( x \) to a set \( A \) is written as \( \mu_A(x) \). The degree of association is shown by the curves in Fig. 3.

The type of characteristic shown in Fig. 3 is an important aid in the application of fuzzy logic in control engineering, because it enables measured values to be arranged in sets. The measured values (input signals) are entered on the x-axis, while the degree of association curves for a number of sets are plotted on the y-axis—see Fig. 4. In the design of control systems, it is usual to take an odd number of sets and to place the central one in a position where it coincides with the desired value: here, 50.

Another instance of allocating likely values of ambient temperatures to fuzzy sets is shown in Fig. 5. The boundaries between areas are not always clearly defined; in fact, in this way it may be determined how 'fuzzy' the boundary between two sets is. It is customary, but not obligatory, to allow the boundaries to overlap to such an extent that the combined border areas have a degree of association of 100%, that is \( \mu = 1 \).

It will have been noticed in these examples that the curves are trapezoidal. This is the most customary shape, since it allows straightforward arithmetic. Other shapes are possible, as long as they are convex, that is,
their edges should not have transitions as shown in Fig. 6. It is, however, possible to omit the horizontal top of the trapezium so that the curves attain a triangular shape. This is done, for instance, in the case of a set that represents the desired value of a control system to obtain a very accurate setting. It is always done when the subdivision of the output signals is fuzzy—see Fig. 7, which shows the positions of a boiler valve in a heating system. It may look strange that the trapezoid characteristic for 'closed' extends to -25% and that for 'open' to +125%, but that is how these sets are weighted to the same scale as the other three positions. Once the output signals have been brought back to concrete values, the valve can be set exactly between 0% and 100%, no more, no less.

Examples of how not to subdivide input and output signals are shown in Fig. 8. The curves in Fig. 8 do not overlap, which means that there is no defined It for a number of values. A well-known example of this is the inputs of TTL-gates—see Fig. 8b. In these, a certain range of values belongs to the set 'low' and another range to the set 'high'. Values between these ranges will lead to unpredictable behaviour. This is, by the way, a special fuzzy set: a so-called crisp-set. In Fig. 8c, the edges of the various curves spill over into various other sets: this will lead to instability.

Logic combining of fuzzy sets

We have seen how input and output signals can be divided into fuzzy sets. To use these to make a practical control system, certain rules are required to indicate the logic connections between input and output sets. These rules, which describe and determine the behaviour of the system, can be arrived at through practical experience of the system or by trial and error.

As an example of how to go about setting the rules, we will use a system that has a switch-on behaviour as shown in Fig. 9b. This is quite a common behaviour: for a little while after switch-on, the measured value will swing around the wanted value. It is the task of the control system to bring the measured value to the wanted value quickly and to keep it there in spite of possible interference. To design the system, we set out the rules, which describe and determine the behaviour of the system, can be arrived at through practical experience of the system or by trial and error.

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Fig. 8. Examples of how not to sub-divide fuzzy input and output signals.

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Fig. 9. Example of how to compute a control system.
Fuzzy logic arithmetic

The function \( \mu_\alpha(x) \) enables us to calculate to what degree an element \( x \) is common to set \( A \). If we use a microcontroller and an analogue-to-digital (A-D) converter, the calculation becomes very simple, because the converter provides concrete values. For each of these values, the associated \( \mu \) can be found in relevant tables.

Evaluating the logic rules is normally simplicity itself. There are three basic operations: AND, OR and NOT. In an AND-operation, the smallest \( \mu \)-value of the relevant sets is allocated to the output set. If, for instance, two inputs, \( x \) and \( y \), have values of \( \mu(x) = 0.8 \) and \( \mu(y) = 0.3 \), it follows from the logic rule 'A AND B gives C' \( (A \land B = C) \) that the smallest \( \mu \) that is, 0.3, must be allocated to C. This is done by adding an element of value 0.3 to set C. This operation is called minimum-operator \( (\text{MIN}(A,B)) \) and is one of the special instructions in fuzzy logic.

The OR-operation allocates the largest \( \mu \)-value to the output set. This is called maximum-operator \( (\text{MAX}(A,B)) \). With the values from the previous example, the logic rule 'A OR B gives C' \( (A \lor B = C) \) gives an element with a \( \mu \)-value of 0.8.

The NOT operation is just as simple: deduct the \( \mu \)-value from 1: \( \text{NOT} A = 1 - \mu_\alpha(x) \).

Compensating operations yield results that lie somewhere between AND and OR; they add a sort of 'but' to the logic rule. For instance, you want to buy a house. It must be sound, in a good position, and not too expensive, but, if it is very nice and well situated, it may cost a little more. In pure logic terms, such a consideration is difficult to realize, but compensating operators make it possible. The most important of these is the gamma-operator. If the value of \( y \) lies between 0 and 1, this operator can be set between AND and OR. The simplified form of the gamma-operator (for three sets) is

\[
\mu(x) = \max\left[ \min\left( \mu_\alpha(x), \mu_\beta(x) \right) \right] = \max\left[ \min\left( \mu_\alpha(x), \mu_\beta(x) \right) \right],
\]

where \( \mu = \mu_{\text{result}} \) and \( 0 \leq \mu \leq 1 \).

Fuzzy becomes distinct

After working through the logic rules, we have a number of indications (21 in Table 1) of what has to happen next. All the rules in Table 1 refer to a signal that controls the relevant process. In complex processes, more control signals may be used. For the calculation of concrete values for these signals, a method is needed that somehow combines the results relevant to the signals. There are two usual methods: min-max-median and product-sum-median. The former is simple, but not suitable for the example in Table 1, because several logic rules apply to the same output set. In that case, the product-sum-median method must be used.

Assume that on evaluation of the logic rules the following results have been obtained: 0.7 median; 0.2 and 0.3 high; and 0.2 very high. How the product-sum-median method works is shown in Fig. 10. For each element, we multiply the height of the triangle indicating the \( \mu \) of a set with the value of the elements and use the results to draw four new (smaller) triangles. We then add the areas of the triangles together and determine the median position of the resulting figure: the value on the x-axis underneath that position is the value we seek.

All this may sound pretty complicated, but the arithmetic is not too difficult. If we assume that the functions of \( \mu \) for the control signals are isosceles triangles, the calculation becomes:

\[
S_c = \frac{\sum_{x=1}^{q} a(x) \times S_a \times A}{\sum_{x=1}^{q} a(x) \times A}
\]

where:
- \( S_c \) is the value of the control signal;
- \( q \) is the number of logic rules whose value is relevant to the magnitude of the control signal;
- \( a(x) \) is the result of logic rule number \( x \);
- \( S_a \) is the value of the control signal immediately underneath the apex of the triangle (set) to which logic rule \( x \) refers;
- \( A \) is the area of the relevant triangle (set) \( (A = \frac{1}{2} \times \text{base} \times \text{height}) \).

This formula is worked out in a computer in seconds: \( S_a \) and \( A \) are fixed data for all sets, which, therefore, can be stored in memory and need not be computed. If all triangles have identical areas, as in the example, the calculation becomes even simpler, because \( A \) can then be ignored both in the numerator and the denominator.

Reference

Kluwer Academic Publishers
RS-232 QUICK TESTER

No PC interface has attracted more attention in the electronics press, and caused hotter debates, than the RS-232 interface. It is unfortunate but true that both hardware and software appear to be open to different interpretations when it comes to connecting RS-232 devices like printers, plotters, mice and modems. The tester described here is a handy tool to help you locate possible hardware errors, when hooking up a new RS-232 peripheral is not going as smoothly as you would have hoped.

Design by A. Rietjens

THE RS-232 interface is an old faithful in the telecommunications industry, its protocol and hardware being geared to conveying data over long distances. In its most rudimentary form, the interface could be reduced to only two lines: ground and data, which are used alternately by the transmitter and the receiver. First, data is sent from location A to location B. Next, the control software changes the direction, and location B starts to transmit to location A. This type of half-duplex communication is nearly extinct these days, when a three-wire connection appears to be the minimum. Provided software is used to deal with the handshaking function, such a three-wire link could be used, theoretically, to implement full-duplex communication. In most cases, the handshaking software is realized on the basis of the Xon/Xoff protocol. In this system, the receiver sends the Xoff code as soon as it has received the maximum amount of data it is able to handle. Transmission of the Xoff code causes the communication to be interrupted until the receiver transmits the Xon code. Communication is resumed on receipt of this code at the transmitter side of the link.

The main advantage of the Xon/Xoff protocol is that only three wires are required to set up a bidirectional data link. Also, the simple electrical connection is a boon because it prevents hardware problems. However, true full-duplex communication is not possible using the Xon/Xoff protocol.

Faster!
The fastest and most versatile version of the RS-232 interface is its complete implementation on the basis of hardware handshaking. In addition to the two data lines, the 'full' version has a number of handshaking lines (usually five) to control the data exchange between connected devices. Unfortunately, in particular the connection between transmitter and receiver has given rise to much confusion. This is mainly caused by poor understanding of the terms DTE (data terminal equipment) and DCE (data communication equipment) used in handbooks and system documentation.

A DTE is, for instance, a PC or a terminal, while a modem or a line printer is usually a DCE. By virtue of the RS-232 protocol, it is possible to interconnect not only a DTE and a DCE (the original aim), but also two DTE devices, without the need of inserting a modem. A DTE-DTE connection, however, requires some cross-links to be made in the connecting cable. Figure 1 shows the most frequently used connection options.

The RS-232 connector

Most of you will recognize the 25-way RS-232 sub-D connector found on virtually all PCs these days. The 9-way (AT-style) version, which omits some of the less important handshaking lines, is also used increasingly. The lines discussed below are available on both the 25-way and the 9-way sub-D connector — see the overview in Table 1.

As already mentioned, eight pins on the RS-232 connector are essential for a reliable data link. Pin 2 carries the TxD (transmit data) signal. This line is used by a DTE device to send data to a DCE device. Pin 5 is used for the complementary function, RxD (receive data), which carries data from the DCE device to the DTE device. Pin 4 carries the RTS (request to send) signal, which is sent by the DTE to the DCE to indicate that it is about to send a dataword. The DCE responds via pin 5, the CTS (clear to send) line. In addition, there is the DSR line (pin 6). Via this line, the DCE informs the DTE that it is 'on line' and ready for use. The DCD (data carrier detect) pin carries a signal that is used by the DCE to set up a stable data connection with another DCE. The last line is DTR (data terminal ready), via which the DTE tells the DCE that it is on, and ready to use the DCE to set up a data link. Figure 1 shows how a DTE is connected to a DCE: a pretty straightforward connection.

In practice, a communication sequence via an RS-232 link would look something like this. Initially, a PC (DTE) and a modem (DCE) are switched on. The PC actuates the DTR line, and the modem actuates the DSR line shortly afterwards. Next, the modem
Fig. 1. Familiar, but never easy to remember, these RS-232 connection options!

sets up the link to another modem, and subsequently actuates the DCD line. This is detected by the PC, which decides that communication can be started, and actuates the RTS line in order to do so. If the modem is capable of processing the transmit command, it actuates the CTS line, whereupon the communication, i.e., data exchange, can commence. One wonders what could possibly go wrong if everything appears to be as simple as that. Well, a lot! Let us look at some possible problems.

First, there is quite some confusion about the signal levels on the RS-232 interface. According to the RS-232 standard, the control signal levels may lie between +3 V and +27 V for a logic '1', and between -3 V and -27 V for a logic '0'. This applies to control lines DTR, DSR, DCD, RTS and CTS.

The opposite applies to the data lines, RxD and TxD, which use active-low levels, i.e., a logic '0' corresponds to a level between +3 V and +27 V, while a logic '1' corresponds to a level of -3 V to -27 V. Evidently, manufacturers of equipment that uses only +5 V and ground for an RS-232 interface violate the standard by not meeting the required voltage swing (and polarity).

A second source of trouble is the connection of two DTE devices. Although there should be no problems with such a connection, cross links are required in practice on all lines except DCD.

A third problem arises when the RS-232 interface is not complete at the DTE or DCE side. In the worst case, this requires some hard thinking before a link can be set up successfully. The best known 'trick' to fool the handshaking circuits at either side of the link is the zero-modem, which simulates signals normally supplied by the 'other party' (see Fig. 1). This is also referred to as 'local echo'.

RS-232 quick tester

We all hope that there will be no problems when serial equipment is properly connected. After all, a good quality cable has been used, connectors have been secured, and matching data transfer parameters (number of data bits, stop bits, and parity) have been set at both sides after wading one's way through hefty manuals, and some cajoling with DIP switches. Alas, problems are still likely to occur. Time to get out your RS-232 tester!
As shown in Fig. 2, the circuit consists merely of a number of connections, a set of switches, and six LEDs. The tester is inserted between the DTE output and the cable to the DCE, or another DTE. It should be noted that switches S2 and S3, and S4 and S5, form pairs, and must always be operated simultaneously. Two separate switches are used rather than a fourfold slide switch because the latter proved difficult to find at a reasonable price.

LED D1 indicates the level on the TxD line, D2 that on the RxD line, D3 that on the RTS line, D4 that on the CTS line, Ds that on the DSR line, and, finally, D0 that on the DTR line for the DCE. The LEDs are bicolour (red/green) types, which enables high and low signal levels to be easily seen. Red means a positive line voltage; green a negative line voltage.

Figure 4 shows the connections that are established when the switches are operated. If all switches are set to position ‘A’, all cross-links required to connect two DTEs via a standard RS-232 cable are made. So, if you want to set up an RS-232 link between two PCs, insert the tester between one of these and the cable. The LEDs will indicate how the communication is getting on.

Normally, a DTE is connected to a DCE via a standard RS-232 cable. If the switches are set to position ‘B’, the tester may simply be inserted in the link. Here, too, the LEDs show what is happening (or not).

Switches S2, S3, S4 and S5 also have a position ‘C’, which serves to set up a null-modem connection. A null-modem causes local feedback at the peripheral and the terminal. This disallows hardware control over the data exchange, and may be required either when the Xon/Xoff protocol is used, or a peripheral device is connected to a PC, or any other computer system, that does not have hardware handshaking. Examples of equipment that lacks hardware handshaking on the serial port include the 80C32 single-board computer (Ref. 1), and many types of serial A-to-D converters.

Table 1. RS-232 connector pinning overview. Essential signals are in bold print.

<table>
<thead>
<tr>
<th>D-25 pin</th>
<th>D-9 pin</th>
<th>Signal</th>
<th>Function</th>
<th>DTE</th>
<th>DCE</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3</td>
<td>CG</td>
<td>chassis ground</td>
<td>out</td>
<td>in</td>
</tr>
<tr>
<td>2</td>
<td>2</td>
<td>TxD</td>
<td>transmitted data</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>RTS</td>
<td>request to send</td>
<td>out</td>
<td>in</td>
</tr>
<tr>
<td>4</td>
<td>7</td>
<td>CTS</td>
<td>clear to send</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>5</td>
<td>8</td>
<td>DSR</td>
<td>data set ready</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>6</td>
<td>6</td>
<td>SG</td>
<td>signal ground</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>7</td>
<td>5</td>
<td>DCD</td>
<td>data carrier detect</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>8</td>
<td>1</td>
<td>positive test voltage</td>
<td>negative test voltage</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>9</td>
<td>4</td>
<td>SDCD</td>
<td>secondary DCD</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>10</td>
<td></td>
<td>SCTS</td>
<td>secondary CTS</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>11</td>
<td></td>
<td>STxT</td>
<td>secondary TxT</td>
<td>out</td>
<td>in</td>
</tr>
<tr>
<td>12</td>
<td>9</td>
<td>TxC</td>
<td>transmit check (DCE)</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>13</td>
<td>8</td>
<td>SrxD</td>
<td>secondary RxD</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>14</td>
<td></td>
<td>RxC</td>
<td>receive clock</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>15</td>
<td></td>
<td>SRTS</td>
<td>secondary RTS</td>
<td>out</td>
<td>in</td>
</tr>
<tr>
<td>16</td>
<td>7</td>
<td>DTR</td>
<td>data terminal ready</td>
<td>out</td>
<td>in</td>
</tr>
<tr>
<td>17</td>
<td>6</td>
<td>SQ</td>
<td>signal quality detect</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>18</td>
<td>5</td>
<td>SEL</td>
<td>speed selector DTE</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>19</td>
<td>4</td>
<td>TCK</td>
<td>speed selector DCE</td>
<td>out</td>
<td>in</td>
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<tr>
<td>20</td>
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<td>SRTS</td>
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Fig. 3. Track layout (mirror image) and component mounting plan.

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<td>CTS</td>
<td>clear to send</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>5</td>
<td>8</td>
<td>DSR</td>
<td>data set ready</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>6</td>
<td>6</td>
<td>SG</td>
<td>signal ground</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>7</td>
<td>5</td>
<td>DCD</td>
<td>data carrier detect</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>8</td>
<td>1</td>
<td>positive test voltage</td>
<td>negative test voltage</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>9</td>
<td>4</td>
<td>SDCD</td>
<td>secondary DCD</td>
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<td>out</td>
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<td>secondary CTS</td>
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<td>secondary TxT</td>
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<td>TxC</td>
<td>transmit check (DCE)</td>
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<td>out</td>
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<tr>
<td>13</td>
<td>8</td>
<td>SrxD</td>
<td>secondary RxD</td>
<td>in</td>
<td>out</td>
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<tr>
<td>14</td>
<td></td>
<td>RxC</td>
<td>receive clock</td>
<td>in</td>
<td>out</td>
</tr>
<tr>
<td>15</td>
<td></td>
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<td>secondary RTS</td>
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<td>in</td>
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<td>21</td>
<td>2</td>
<td>SEL</td>
<td>speed selector DTE</td>
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<td>out</td>
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<td>speed selector DCE</td>
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<td>in</td>
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<td>0</td>
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<td>out</td>
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<td>TCK</td>
<td>speed selector DCE</td>
<td>out</td>
<td>in</td>
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</table>
Construction

The design of the printed circuit board for the RS-232 quick tester is shown in Fig. 3. Start the construction by fitting the six wire links on the board, followed by the resistor array (note the orientation!) and the five switches. As already mentioned, switches S2-S5 are formed by two double-pole, 3-way slide switches, because a 4-pole switch with 3 positions proved difficult to get. The equivalent switch pair will be fine as long as you remember always to operate them together. With some dexterity, the switches may also be coupled by a small piece of plastic glued in between the levers.

Continue the construction by fitting the two connectors K1 (male) and K2 (female), followed by the LEDs, D1-D6. Note that all LEDs should be fitted with the same orientation. The flat side of a bicolour LED should be at the side of the cathode of the diode symbol printed on the overlay. Take care to ensure the correct LED orientations, or you will not be able to tell the RS-232 signal level for sure. To make it sturdy and easy to handle, the tester may be fitted into a small plastic enclosure.

The circuit is now ready for use. Set all switches to position 'A' for a DTE-DTE link, or to position 'B' for a DTE-DCE link. If you need a null-modem link, set S4 and S5 to position 'C'. A null-modem may be called for in a DTE-DTE link as well as in a DTE-DCE link. Well, at this stage proper communication via the RS-232 ports can only be disrupted by software faults, in which case Murphy's laws apply.

Reference:

NEW PRODUCTS

LogicWorks logic simulator now available for DOS.

LogicWorks has been the number-1 package for teaching digital circuit design on the Macintosh since 1985, and is now available to the wider DOS audience. LogicWorks brings circuit ideas to life on the screen with fully interactive digital simulation. Probes, displays and switches can be placed right on the circuit diagram, and operated just like a real circuit breadboard. Device delays, clock rates and signal connections can be changed with a few mouse clicks, allowing circuit ideas to be tested out in seconds. More info from Capillano Computing Systems Ltd., Canada. Tel. (604) 522-6200, Fax (604) 522-3972.
The original experimental QTC transmitting loop antenna was built in late 1990, tested immediately, and details were published in the June 1991 issue of this magazine. The design raised a lot of interest and a number of fellow amateurs in both the UK and the USA advised the author that they were apartment dwellers (no outdoor antenna possible!) and were going to construct it. Others asked whether a two-band 80/40 metre (3.5–7.0 MHz) version was feasible. That request is answered in this article.

In response to the comments and questions of a number of readers, particularly in the UK and the USA, a Mark 2 QTC loop antenna was constructed to cover the 80 m and the 40 m bands. A similar octagonal configuration as for the original was used, but the tuning/coupling impedance matching circuits were redesigned for two-band operation after a number of experiments and tests. The result is even simpler than the original.

The original QTC loop was a 30 in. (75 cm) diameter octagonal spiral loop, of novel design, whose general circuit and configuration is shown again in Fig. 1. The loop was designed for indoor use, set up on a table alongside an 80-m low-power CW transmitter. It was intended as a possible answer for those amateur transmitting enthusiasts who, for one reason or another, cannot erect an outdoor antenna.

Theoretically, a small, indoor loop using a maximum of 10 W RF power cannot be expected to produce a signal comparable with the conventional 100 W transmitter/outdoor dipole set-up. However, in practice, it occasionally does; we think for the following reasons.

1. The QTC loop being directional, and comparatively narrow-band, reduces incoming interference (and outgoing, such as TVI).
2. Being 'on hand' alongside the operator, the QTC loop can be peak resonated for absolute maximum performance and direction on the spot frequency being used.
3. Some outdoor dipoles (and other types) are erected to textbook dimensions, with little or no regard being paid to the height and surrounding objects, resulting in an off-tune compromise, which, when fed with, say, 100 watts, produces results that are accepted without the realization that these could be greatly improved if the antennas were tuned to optimum frequency at optimum height. To which should be added the antenna compass alignment, and the type, length and condition of the coaxial feedline, and so on. The overall result is that the original 100 W signal is largely dissipated or attenuated in the system.

Reason 3. became apparent during various contacts with continental European stations when using the QTC Loop, and later the Mk. 2 QTC Loop, and discreet questions were asked about the height, length and surrounds of their antennas.

Nevertheless, it must be stressed that in most cases the QTC loop signal was somewhat down on the outdoor dipole/transmitter set-up. It was, however, quite adequate for 2-way communication, which must be considered satisfactory when an enthusiast is lumbered with a 'no outdoor antenna' situation, for whatever reason.

The circuit

The original QTC Loop was intended for operation on the 80-m (3.5 MHz) amateur band. The circuit and general layout are given in Fig. 1. A spiral 5 1/8 turns loop, L1, and inductance L2 were resonated to frequency by bandset and bandspread variable capacitors C1–C2. There was also an optional tuning meter.

The Mark Two QTC 80/40 Loop uses an identical spoke framework—see Fig. 2. The total number of turns of L1 is five for 80 metres and an optional tap for 40 metres. It is resonated by C1–C2, which is a good-quality receiving type two-gang variable capacitor (2x125 pf) with the built-in padders removed. Coupling to the transmitter (and receiver) is by C3 (300 pf ceramic variable) to 48 in (120 cm) of RG58 coaxial feedline. The original tuning meter—see Fig. 1—can be used with advantage.

Construction

The construction is shown in Fig. 3. It consists of an eight-spoke octagonal frame, a vertical member, and a heavy base to prevent it tipping over. On to this is fixed the sim...
ple resonating/loading unit, which is constructed on a standard 9x4" (23x10 cm) fibre-glass board, copper-clad on one side. Ideally, this should be replaced by a suitable box. However, as it was anticipated that the loop would be used as a test bed for future experiments, a board was used as this can be re-used, or quickly replaced, at little cost—see Fig. 3 and Fig. 4.

The loop frame consists of eight spokes made from four lengths of moulded hardwood, each 30 in. long, 5/8 in. wide and 1/4 in. thick (760x16x6 mm). A 2BA (5 mm dia) clearance hole is drilled in the centre of each spoke. The spokes are then glued and clamped together with a long 2BA (5 mm dia) bolt, nut and washers. The spokes must overlap in the order shown in Fig. 3 and evenly spaced apart. The assembly should then be allowed to dry out thoroughly, after which it is given a coat of polyurethane varnish. A six-way, 2 A polythene terminal block (cut from a standard 12-way block) is screwed to the end of each spoke to provide the necessary securing, spacing and insulation of the wire turns.

A 23x0.8x0.8 in (585x20x20 mm) vertical wooden support is glued, screwed and bracketed to a suitable 12x8x0.5 in (300x200x12 mm) wooden base, the whole being teak-wood stained. The loop frame is then glued and bolted to the top of the vertical support, as shown.

The loop winding, L1, uses PVC covered 7/0.2 mm wire with an outside diameter of 1.2 mm and rated at 1 kV, 1.5 A. After loosening the grub screws in the terminal blocks, thread the wire through, turn by turn, for five turns, starting at the outer hole of the bottom right-hand spoke, threading through all the blocks in an anti-clockwise direction, and terminating at the inner hole of the bottom left-hand spoke. Tighten the grub screws while going along just enough to hold the wire in place, and firmly when all the wire is threaded through and pulled tight. Leave long wire tails at both ends; these can be cut back and soldered later—see Fig. 3.

Next, secure the fibre-glass board to the wooden upright—see Fig. 3 and Fig. 4. Positioning the capacitors depends on their actual size and shape, but should be more or less as shown in Fig. 4. Capacitor C1-C2 must be fastened to the panel as shown and fitted with an extension shaft, with front support bracket and panel bush as shown in Fig. 4, with a wooden support platform, cut to size, to support the bracket.

Capacitor C5 is mounted at an angle and fitted with an insulated extension shaft as shown in Fig. 3 and Fig. 4. It must be completely insulated from the copper-clad fibre-glass panel. Also, it should be fitted such that the knob is well away from the loop winding. This capacitor requires setting in only one position for 80 m (3.5 MHz) and 40 m (7.0 MHz) and may, therefore, be considered preset.

A two-way polythene terminal block is screwed to the vertical support just inboard of the bottom inner end of L1—see Fig. 3. This is to give the 40 m (7.0 MHz) facility later.

---

Fig. 3. Construction of the Mark 2 QTC Loop.
The outer end of L1 must be cut back and soldered to the junction of C2 and C3 as shown in Fig. 3. The inner end of L1 is cut back and inserted into the bottom connection on the terminal block—see Fig. 2 and Fig. 3. Solder a flexible plug lead, cut to length, to the stator of C1. A 2 mm plug is a snug fit in the block terminal insert.

Selecting of the 40 m (7 MHz) tap is discussed in ‘Testing and operation’ later.

All other behind-the-panel wiring must be in 16 SWG tinned copper wire, with the copper cladding of the board used for the ‘earth’ connections—see Fig. 4.

Fit a large 3 in. dia (75 mm) control knob to C1–C2.

Sectae a 48 in (120 m) length of RG58 coaxial feedline with cleats to the wooden base.

Testing and operation

1. 80-metre (3.5 MHz) band

   Initial tests must be made with a receiver. Rotate the resonating control, C1–C7, to ensure that it covers the 3500-3800 kHz (3500-4000 kHz in the USA) band. Then rotate C3 for maximum signal in the centre of the band.

   Feed a small amount of RF to the loop and re-adjust C1–C3 to resonate at the exact transmit frequency. Capacitor C3 may also require a small re-adjustment for maximum loading (or lowest SWR if a SWR meter is available). Check the loop radiation with a nearby field strength meter (the tuning meter used on the original QTC Loop is ideal). It should now be possible to fully load the loop. Capacitor C3 should not require any further adjustment over the band.

   Check the directional properties by tuning the receiver to a signal and rotating the loop for maximum signal—this is also the position for maximum radiation to the desired station.

2. 40-metre (7 MHz) band

   The approximate position of the 40-m tap on L1 is shown in Fig. 2 and Fig. 3. The exact position must be selected carefully so that C1–C2 are enmeshed about 15% at 3800 kHz. The tap is made with a flexible lead.

   To avoid mutilating the PVC wire covering on L1 when seeking the correct ‘tap’ position, take a lead from the top insert on the two-way terminal block. Solder a thin sewing needle at the other end.

   With the receiver tuned to about the centre of the 40-m band, insert the 2 mm plug into the two-way terminal block at the top and push the needle through the PVC wire cover in the position shown in Fig. 2 and Fig. 3. Resonance should be obtained at 3800 kHz with C1–C3 enmeshed about 15% and C3 peaked for maximum signal. If necessary, move the ‘needle’ tap slightly to the left or right as required. The tapping point will vary between loop models, depending on the construction. Once the tap point has been found, the PVC can be removed, and the lead end soldered on in place of the needle. It will be found that any tiny needle holes will not show if the wire is squeezed between finger and thumb.

   The transmitter can now be resonated/loaded to the loop as in the 80-m band.

3. Other bands

   There seems to be no obvious reason why a further tap cannot be added for another band, such as the 20 m (14 MHz) band using a similar technique to the one described. A slow motion drive could be added between the knob and C1–C2. The loop would probably be less directional on the higher frequencies.

Safety

The QTC 80/40 Loop is designed for use with low transmitter power. The prototype has been operated with up to 10 watts CW. It has been tested up to 20 watts, but any higher powers would necessitate higher voltage variable capacitors and thicker wire for L1.

In the interest of domestic household safety, 10 watts should not be exceeded. There will be no prizes for the operator for setting fire to the curtains or giving the kids a nasty RF shock, or scaring the living daylights out of the cat.

Reference

The last printed circuit board to be discussed is the LED-based S-meter (signal strength) unit, which is fitted onto the front panel of the tuner enclosure.

The S-meter is a simple circuit (Fig. 16) based on the familiar LM3914 LED bar driver IC from National Semiconductor. This IC allows LEDs to be connected direct, i.e., without the usual external current limiting resistors. Here, the LM3914 is used in ‘bar’ mode, which allows rectangular face LEDs to be used to mimic a continuous horizontal scale. The driver is powered from the 5-V supply line in the tuner. The bar indication is a good alternative to a moving-coil meter because the input voltage range starts at 0 V (in spite of the single power supply), and the circuit is simple to adapt to the required full-scale voltage level. The full-scale indication is determined by a voltage reference source that outputs 1.25 V between pins 7 (REFOUT) and 8 (REFADJ). The resistor connected between these two pins has two functions. Firstly, it determines the current through the LEDs driven, according to

\[ I_{\text{LED}} = \frac{12.5 \, \text{V}}{R_i} \]

The indicated value of 1.2 kΩ results in a LED current of about 10 mA.

Secondly, the combination of \( R_1 \) and \( R_2 \) determines the full-scale input voltage level, \( U_{\text{max}} \):

\[ U_{\text{max}} = 1.25 \, \text{V} \times (R_1/R_2 + 1) = U_{\text{RH}} \]

The voltage at RH (pin 6) determines the end-of-scale value, while that at RLO determines the start of the scale. The resistor values used here result in a scale range of 0 to 4 V.

The supply voltage of the LED driver is not limited to 5 V only. In fact, you may use any supply voltage between 3 V and 12 V, as

**COMPONENTS LIST**

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<tr>
<th>Resistors:</th>
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<td>1kΩ</td>
<td>R1</td>
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<td>1</td>
<td>2kΩ</td>
<td>R2</td>
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<td>10μF 16V</td>
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<td>D10</td>
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<tr>
<td>1</td>
<td>LM3914</td>
<td>IC1</td>
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<tbody>
<tr>
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<td>Printed circuit board</td>
<td>920005-6</td>
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</table>
Fig. 18. Track layout (mirror image) and component mounting plan of the keyboard/display PCB.
COMPONENTS LIST

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<th>Capacitors:</th>
<th>Semiconductors:</th>
</tr>
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<td>1 220nF</td>
<td>33 1N4148</td>
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<tr>
<td>49 330Ω</td>
<td>1 22µF 16V</td>
<td>1 1N4001</td>
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<tr>
<td>5 1kΩ</td>
<td>1 100µF 16V</td>
<td>7 74HC4511</td>
</tr>
<tr>
<td>1 390Ω</td>
<td>1 22µF 25V</td>
<td>1 74HC138</td>
</tr>
<tr>
<td></td>
<td>2 100nF</td>
<td>7 DP352PK (red, cathode)</td>
</tr>
</tbody>
</table>

Miscellaneous:

- 5 Digit key with wide (17mm) cap
- Printed circuit board 920005-4
- Front panel foil 920005-F

COMPONENT SUPPLIER INFORMATION

- **FD12 tuner module:**
  Restek Electronic Products GmbH
  Industriegebiet Richard-Roosen-Strasse 15
  3500 Kassel-Waldau
  GERMANY.
  Telephone: +49 561 585325
  Fax: +49 561 581664.

- **E400 Müfler:**
  GERMANY.
  Telephone: +49 212 951250
  Fax: +49 212 7430.

- **214KCS-10115X Inductor:**
  Contact your national Toko distributor (equivalent types may be supplied).

Keyboard PCB

Figure 18 shows the printed circuit board for the keyboard/display unit of which the circuit description was given in last month's article. On this PCB is a 100-nF decoupling capacitor, C106, not indicated in the circuit diagram (Fig. 14). It is connected in parallel with D134.

The functions of the keys are apparent from the front panel foil layout given in Fig. 19. This self-adhesive foil is available

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RADIO AND TELEVISION

ready-made through our Readers Services, and fits perfectly on the front panel of the 19-inch rack enclosure used to house the receiver. It should be noted, however, that the order of the function keys as indicated in Fig. 19 should be corrected to (left to right):

NORMAL WIDE EFFECT STEREO MONO

Ensure this order by cutting out the rectangular, lettered, pieces from the foil, and sticking them on to the right keycaps.

The LEDs indicate the following functions (left to right):

WIDE EFFECT MUTE STEREO

FD12 tuner modifications
Those of you keen on taking the performance of the FM receiver to an even higher level may be interested to know how the noise figure of the FD12 tuner module can be lowered. Well, this may be achieved by replacing the BF900 dual-gate MOSFET in the RF input amplifier by a later, improved, type, the BF982.

To gain access to the BF900, remove both covers of the tuner module. Locate the transistor (Figs. 20 and 21), and remove it from the board, noting its orientation. The BF982 is pin compatible, and can take the place of the BF900 without any modification to the circuit. Make sure to treat the BF982 with care: like the BF900, it is a static sensitive device, which requires the soldering iron tip to be grounded. That completes the low-noise modification. Fit the covers again.

Conversely, it may happen that the tuner has to cope with very high signal levels, for instance, from a radio/TV cable network in which amplifiers are used to boost certain FM band signals. High signal levels may cause intermodulation, and require a separate, attenuated, receiver input, which has to be provided on the rear panel of the enclosure. Use any suitable RF input socket, and solder a 8242 resistor between the centrepin and ground. Next, connect a 1-kΩ resistor between the centre pin of the coax socket and the signal wire in the coax cable to the terminal marked ‘Ant.’ on the main tuner board.

Case and front panel
The photographs in last month’s instalment give a good impression of the internal construction of the FM tuner. The rear panel of the 19-inch case contains the two audio output sockets (RCA or ‘phono’ style), the isolated RF input socket, and the mains appliance socket. The latter may, of course, be substituted by a fixed mains cord with a feed-through grommet and a strain relief clamp at the inside of the enclosure. In any case, the metal enclosure must be connected to the protective earth terminal (E) on the mains socket.

The main tuner board, the synthesizer board and the power supply board are mounted on 10 to 15 mm high PCB pillars secured to the bottom plate of the enclosure.

The other three boards (mode control; S-meter; keyboard/display) are fitted at the in-
Table 1. Wiring overview

<table>
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<th>Synthesizer board</th>
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<tbody>
<tr>
<td>0 (K3)</td>
<td>+33V</td>
</tr>
<tr>
<td>+32V (K3)</td>
<td>+33V</td>
</tr>
<tr>
<td>0 (K2)</td>
<td>Synthesizer board</td>
</tr>
<tr>
<td>+5V (K2)</td>
<td>+ (near C414)</td>
</tr>
<tr>
<td>0 (K2)</td>
<td>Synthesizer board</td>
</tr>
<tr>
<td>+5V (K2)</td>
<td>+ (near C414)</td>
</tr>
</tbody>
</table>

Main tuner board

| +33V | Main tuner board +33V |
| +33V | Main tuner board +33V |
| UABST| Synthesizer board +33V |
| M1   | Display board +33V   |
| GND  | Mode control board   |
| +15V (+C67) | Mode control board +15V |
| BASIS B LED | Mode control board B |
| P-Stere | Mode control board D |
| STEREO-LD | Mode control board D |
| MODESELECT A | Mode control board D |
| MODESELECT B | Mode control board D |
| MONO  | Mode control board   |
| MUTE (at R49) | Mode control board   |

Synthesizer board

| UNLOCKED | Mode control board |

Keyboard/display board

| D0 | Synthesizer board D0 |
| D1 | Synthesizer board D1 |
| D2 | Synthesizer board D2 |
| D3 | Synthesizer board D3 |
| D4 | Synthesizer board D4 |
| D5 | Synthesizer board D5 |
| D6 | Synthesizer board D6 |
| IORW| Synthesizer board IORW |
| + (near D134) | Synthesizer board + (near C414) |
| 0 (near C106) | Synthesizer board + (near C414) |
| E (near S115) | Synthesizer board + (near C414) |
| F (near S115) | Synthesizer board + (near C414) |

Wiring

To keep synthesizer noise to a minimum, the ends of the signal wire in the coax cable must be kept as short as possible. This means that the signal wire must be screened over the maximum possible length between RF input socket (on the rear panel of the enclosure), and the 'Ant.' and 'GND' terminals on the main tuner board. The shielding braid of the coax cable must be connected at both ends: to the 'GND' terminal next to the 'Ant.' terminal, and to the shaft of the RF input socket, which is fitted isolated from the rear panel.

The audio output signals are taken from the ROUT and L-OUT terminals on the main tuner board to the output sockets on the rear panel. The cable screening is connected to the respective ground (GND) terminals on the main tuner board. The output sockets are also the best location to make the only connection between the circuit ground (GND carried via the screening of the audio cable) and earth (carried via the metal enclosure).

The connections of the components connected to the mains must be carried out with great attention paid to proper isolation, and using appropriate wire. All mains wiring must be secured to the enclosure, and kept as far as possible from signal wiring.

The connections between the printed circuit boards are listed in Table 1. Needless to say that a carefully done wiring job may prevent precious time spent on fault finding. Note that the tuning voltage (UABST) is carried via a single-core screened audio cable.

Provided the main tuner board has been adjusted as outlined earlier, the FM tuner is ready for use when the wiring is finished. Simply connect the antenna, an audio amplifier, and power up. Program your favourite stations, and ... happy listening!
A capacitor can be defined by the fact that
\[ i = C \frac{dV}{dt} \]
is the formula relating voltage, \( V \), current, \( i \), and time, \( t \), when the voltage across the capacitor, \( C \), changes.

Similarly, although rarely used now except in r.f. work,
\[ V = L \frac{di}{dt} \]
is the formula used to define an inductor, \( L \).

Oliver Heaviside used his transform to define the "fractional integral" in the following terms:

If \( F(t) \) has Laplace transform \( F(s) \), then
\[ \int_0^t F(t) dt \]
has Laplace transform \( f(s/s) \) and \( dF/dt \) has Laplace transform \( f(s)f(0) \).

Now, the half-integral
\[ \frac{1}{2} \int_0^t F(t) dt \]
has the Laplace transform \( f(s)/s \). When defined, the half-derivative is the inverse of the half-integral. Call this operator \( D_{1/2} \).

This can be done as the half-integral of the derivative:
- If \( y = \exp(at) \), then \( dy/dt = a \exp(at) \).
- If \( F(t) = \exp(at) \),
  then \( f(s) = a \delta(s-a)f(s) / \rho(s) = \pi(s-a) \) (where \( \pi \) is \( i/s \)),
  for which \( F(t) = \exp(at) \exp(\pi^2s) \) is the inverse Laplace transform. For the ability to calculate this, we are much indebted to Oliver Heaviside.

Thus, the admittance of the half-capacitor at angular frequency \( \omega \) is given by:
\[ Y_{1/2}(j\omega) = \frac{1}{C \pi(\omega^2)} \] and the impedance of the half-inductor is given by
\[ Z_{1/2}(j\omega) = \frac{1}{L \pi(\omega^2)} \].

Erf \( (x) \) is the integral from 0 to \( x \) of \( \exp(-x^2)x \).

The product of the two impedances is \( VZ \), and \( t \) is the time since turn-on. Note that after the circuits have settled down \( t \) is large, \( \pi(\omega^2) \) is very close to one. Thus, the formulas are much simplified and become:
\[ L(1+j)/2(\omega^2) + L(1-j)/2(\omega^2) = L(1+2j) \]
which is a real, frequency-dependent resistance; note, however, that \( LC = 1/\omega^2 \) compared with the usual formula \( LC = 1/\omega^2 \) for a tuned circuit at resonance.

The parallel connection produces exactly the same result: let \( z = L(\omega^2/2) \), then
\[ z (1+j)/z (1-j) = z = L^2(1/2) \].

Note that with normal tuned circuit, the parallel impedance at resonance is infinite, and the series impedance is zero.

Is \( Z \) maximum or minimum when \( LC\omega = 1 \)?
Let \( LC\omega = 1 \), then the series connection is
\[ L(1+j)/2(\omega^2) + L(1-j)/2(\omega^2) = L(1+2j)/2(\omega^2) \]
which, with magnitude squared, becomes
\[ \omega^2(1+k^2) \].

This is a minimum when \( k = 0 \), unlike the case for a usual tuned circuit.
Let \( z = L(\omega^2/2) \), then for the parallel case:
\[ (1-j)/2(\omega^2)(1+2j)/2k = (1+k)(1-j)/2k \].

The parallel impedance is then:
\[ 2k(1-k)(1+j) \].

Let \( k \) be positive, then the minimum impedance occurs when \( k = 0 \), and the maximum when \( k \) is very large.

Zero resistance

The condition \( LC = 1/C \), which can be brought about by the use of negative impedance converters—NICs—leads to zero resistance in both the parallel and series connection. Then,
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gives an impedance \( z = jL P(i/2) \).
For a half-capacitor in series with a 1-Ω resistor, a law like this is obeyed:
\[
V = v = CD \sqrt{t}(i),
\]
where \( V \) is the supply voltage. Therefore, the solution of half-differential equations like this is specifically interesting to see what kind of evolution of voltage across the half-capacitor occurs in practice.

Duration of the initial transient

So far in this analysis, we have adopted a simplified expression for the half-reactances, ignoring the error function \( \text{erf} \); this is justified, since the time-dependent feature vanishes very quickly at all but the very lowest frequencies, and this we will establish now.

As time increases, the function \( \text{erf}(t) \) rises to 1.0 and then stays near 1 indefinitely for increasing time. \( \text{erf} \) constantly increases to this level. When \( t = 0.75 \), \( \text{erf}(t) \) is close to 0.1414 = 0.7071, and thereafter is greater. Thus, we must consider:
\[
\text{erf}(i/2) > 0.7071.
\]
This occurs when \( \omega t > 0.55 \); since \( \omega = 2\pi \), we have:
\[
\omega t > 0.872.
\]
This implies that for frequencies above 10 Hz, this feature can be ignored after 0.2 s.

The effect of \( j \) within the square root has been ignored here, but analysis in terms of functions like Fresnel sine and Fresnel cosine shows that the same general behaviour is displayed, although the complex number introduces some small wobbles in the tendency to unity.

Since the use of \( \exp(j\omega) \) in circuit analysis implicitly applies only to the steady state when \( \omega \) = 0, the use of the simplified formulae is justified. The fact that, just after initial transients, \( dy/dx \) has other solutions is important, even for physics. Since the half-derivative may have a direct bearing on the thinking behind Dirac's equation of the electron. If time is never measured directly, how can we assert confidently that the first derivative, and not the half-derivative, is appropriate for measuring change?

Laws obeyed by half-reactance

Laws obeyed by half-reactance will apply in the steady state after the levels within the circuit have become quiescent. There are two kinds of element, say, \( cZ \) and \( fZ \), where the small letters denote 'half'.

1) \( v = e^{-cZ}(t0) \)
2) \( f = e^{-fZ}(t0) \)
3) \( cZ \times eZ \) is a capacitance;
4) \( fZ \times eZ \) is an inductance.

There are other relations that can be related to the original defining relations:
\[
i = CD \sqrt{t}(i),
\]
and
\[
V = D \sqrt{t}(i).
\]

Evidently, when a half-capacitance is in series with a 1-Ω resistance, there is a relation between the voltage on the resistance and that on the half-capacitance. Let \( V \) be the applied constant voltage across the series pair; then:
\[
i = V - v = CD \sqrt{t}(i).
\]
Half-differentiating this yields
\[
V(i) = -C \sqrt{t}(i),
\]
that is, after the circuit has settled down, \( i \) and \( v \) are related by the law of a negative capacitance. The same applies to the half-inductance in parallel with a 1-Ω resistance; this becomes a negative inductance. This does apply not only to a constant voltage, but also to a sinusoidally varying one.

But does following the law of a negative capacitance necessarily entail that we have a negative capacitance? Say, \( i = -C \sqrt{t}(i) \) and \( v = \sqrt{t}(i) \), then
\[
i = -C \sqrt{t}(i) \cos \sqrt{t}(i);
\]
let \( i = \sqrt{t}(i) \), then
\[
\sqrt{t} = v = -1/joC.
\]
which is the formula for the impedance of a negative capacitance; so, the answer is yes.

Practical use of half-reactance

Can these circuit elements be built? The essential requirement is the square-rooting of some impedance. Now, suppose we had a device to square any impedance presented to the input terminals, so that the square of the impedance appeared at the output terminals. We could then compare the output with, say, an inductance and feed back any error signal to the input of the squaring device appropriately transformed, and thus have the appearance of a half-inductance at the input.

Thus, fabrication is possible, because impedance squaring devices are known to exist. The case of a capacitor is not so easy, because some active devices don't work well with a true capacitive load; but certainly a half-capacitor in parallel with a very high impedance could be made. The effect of 1-Ω in series can then be checked empirically. (A circuit element that produces a negative capacitance or a negative inductance when combined with standard components cannot fail to have many uses if practically produced, but until this is checked empirically there is uncertainty about whether this can be done practically; a practical implementation may just oscillate.)

More significantly, a circuit that has an impedance magnitude that is not simply proportional or inversely proportional to frequency, but to the square root of frequency, is the reciproc of this, is a device that has many immediate uses in the design of different types of filter: new types of filter with a more gentle roll-off than usual.

In particular, the ends of the frequency band would have responses less dependent on component tolerances, so that the frequency response of, say, audio filters at high and low frequencies would be more predictable, and the spread in manufacture would be less.

Also, the half-capacitance and half-inductance, apart from theoretical interest, open the door to devices whose impedance magnitude depends on half-integral powers of the frequency domain.

We must restore here that these devices are defined in practice by the half-derivative operation (or half-integral) and not by the simple square-root form, which is approximate.

Quiescence

We have already determined how long the quiescent phase lasts at different frequencies, that is, before the approximations used in the laws of the last section become accurate. This is a consideration that relates to the use of the device itself, but, on a more general note, this is serious, because we never directly measure time (think about how time is determined). We cannot say whether changing parameters should be differentiated or half-differentiated to represent the rate of change. Consider that there is another way of considering the time variable, that is, change in parameter \( t \) is represented by \( D^2q(t) \), where the half-derivative is with respect to a variable \( f \) and:
\[
D^2q(t) = dq/dx = (df/dx)(d/df)
\]
by the chain rule. We may then consider that \( t \) and \( f \) are related by some non-trivial func-
The prime example of this is the 'American East Coast Blackout' after which it was found that only one man knew how to turn on all the stations and switches in the right order without causing fatal instabilities. That is a large-scale version of the same phenomenon.

It is a daunting thought that the Universe, which is some ten thousand million years old, may still be affected by a switch-on transient (of course, this is only on one view of cosmology).

More importantly, this analysis indicates that under a.c. conditions, not d.c. (which is what the erf alternative refers to), we cannot necessarily apply the standard analysis methods to the frequency dependence of the circuit at very, very low frequencies, since erf forms are equally appropriate. Especially when half-capacitances and half-inductances are built in, this feature must be checked carefully. In essence, 

\[ D(y) = y + \text{(terms decaying with time)} \]

\[ D(y) = y + 1/t^2(\pi t) \]

has a solution similar to 

\[ y = \exp(t/\text{erf}(t/|t|)) \]

our erf function.

That is, blithe claims that 'this circuit works down to zero frequency' must be checked very carefully, since reactance is always present.

References


### DIFFERENTIAL TEMPERATURE INDICATOR

The circuit in the diagram enables the monitoring of two temperatures, \( t_1 \) and \( t_2 \). The sensors are NTC resistors \( R_3 \) and \( R_4 \), which may be connected to the circuit via lengths of circuit wire that may be up a few metres long.

Diodes \( D_1 \) and \( D_2 \) indicate whether \( t_1 \) and \( t_2 \) are close to each other or not. 'Close' means that the difference between the two voltages from the sensors is smaller than the level set with \( P_2 \). If the temperatures are close, both LEDs light; if they are not, one LED will go out: \( D_1 \) if \( t_1 \) is higher than \( t_2 \), and \( D_2 \) if \( t_1 \) is lower than \( t_2 \). Apart from this optical indication, it is possible to obtain an acoustic one by connecting a d.c. buzzer to terminal \( S \).

It is also possible to connect a 7-12 V relay with a maximum energizing current of 400 mA to this terminal. Free-wheeling diode \( D_5 \) protects \( T_1 \) against a possible, destructive back-e.m.f.

The circuit draws a maximum supply current of 35 mA, largely on account of the LEDs.

A certain temperature off-set may be preset with \( P_1 \). Normally, this preset will be at the centre of its travel: when \( t_1 = t_2 \), the potential at the wiper will then be half the supply voltage.

The 'window' within which temperatures are monitored may be preset with \( P_2 \) to a value of 1-25 °C at a regulated supply voltage of 8 V and \( t_1 = 25 \^\circ C \).

[Amrit Bir Tiwana - 924082]
READERS’ CORNER

CORRECTIONS

Wide-band active telescopic antenna
(page 46)
In contrast to what is stated in the article
and in the Readers’ Services (page 102), a
printed-circuit board for the antenna is avail-
able at £2.75 plus 48p VAT (No. 924102).

CB-to-SW down converter
(page 49)
Drawings of the PCB for this unit are omit-
ted from the article: they are shown here.

LETTERS

Dear Editor: Would you please inform me if
Elektor Electronics are going to do a project
on the ‘Sine wave converter’ as was proposed
for March 1990?

L.J. Burns
New Zealand

Unfortunately, the designer could not ob-
tain the repeatability of his prototype that is
absolutely vital for our readers. Furthermore,
the designer has since moved away. Sorry!
[Editor]

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ing order if possible. Reasonable price paid.
Phone Sean on (0256) 57517.
EVEN since the long-wave broadcast station Kalundborg changed its frequency from 245 kHz to 243 kHz to comply with the CCIR's recommendations for a 9 kHz raster in the LW and MW bands, it has become possible to use the carrier in the locked frequency standard described in Ref. 1. Kalundborg is a 300 kW long-wave transmitter in Denmark, with a range of about 500 km (300 miles). This article is, therefore, of particular interest to our Scandinavian readers.

Briefly, what is proposed here and in the next item ('preamplifier for Kalundborg frequency reference') is to change the divider in the frequency reference such that a carrier input frequency of 243 kHz can be used instead of 77.5 kHz (the transmit frequency of DCF77 in Germany), for which the circuit was originally designed. The changes are outlined in the simplified block diagram, Fig. 2, which is best compared with Fig. 1 in Ref. 1. In practice, the original circuit of the reference is changed to the extent that a new circuit diagram is required—see Fig. 1. The new circuit is much simpler than the original, mainly because some circuit sections could be omitted, including the VLF preamplifier (T1-T4 in the original design), the 10 MHz 'locked only' output (IC7 and T12 in the original design) and the 'error' detector (N2, N3, N4 and the beeper in the original design).

In the 'Kalundborg' circuit shown here, the 10-MHz signal supplied by X1 and T3 is multiplied by 3 by parallel tuned circuit L4-C39. The 30-MHz signal is subsequently divided by 100 (IC5) and again by 100 (IC6) to obtain the 3 kHz reference for multiplier IC3. The 240 kHz signal used to heterodyne with the 243 kHz carrier is...
obtained by dividing the 30 MHz signal by 25 (IC5) and then by 5 (74LS90).

The antenna, formed by an inductor wound on a ferrite rod and resonated by capacitors C1, C2 and C3, is connected directly to the balanced inputs of the SO42P mixer (IC1). The error signal at the output of the multiplier (IC3) is filtered and converted into a tuning voltage, which is applied to a dual varicap, D5. The varicap is capable of detuning (to a small extent) the 10 MHz quartz oscillator, and so closes the phase-locked loop (PLL). Provided Kalundborg is received with adequate strength (rotate the ferrite rod), the LED at the output of IC8 lights, and a 'rock-steady' 10 MHz reference signal is available at the output of N2.

Figure 3 shows a detailed diagram of the ×3 multiplier. Inductor L₃ consists of 4.5 turns of 0.5 mm dia. enamelled copper wire on a former with a ferrite core. The former has an outside diameter of 6 mm, and L₃ is drawn out to a length of about 8 mm. The operation of the multiplier is easily checked with the aid of an oscilloscope and/or a frequency meter, the signal levels being quite high. The core in L₃ is adjusted for the highest 30 MHz level at pin 5 of N₄.

Reference:

PREAMPLIFIER FOR KALUNDborg
FREQUENCY REFERENCE

In all cases where reception of Kalundborg is marginal using the ferrite rod at the input of the SO42P, the two-stage preamplifier shown here can help you ensure longer 'lock' periods of the frequency standard.

The antenna inductor, L₁, consists of 150 turns of 0.3 mm dia. enamelled copper wire on a 12.3 mm dia plastic, thin cardboard or paper former, positioned centrally on a ferrite rod with a length of 238 mm and a diameter of 9.6 mm. The inductor is resonated at 243 kHz by a fixed capacitor, C₉, and a trimmer, C₁₀. To keep the load on the antenna inductor, as small as possible, the first stage of the preamplifier is formed by a FET Type BF245.

The amplified 243-kHz signal is fed to one of the inputs of the SO42P at the input of the frequency standard (see previous article). Using the preamplifier, the author (who lives in Denmark) measured a signal level of 1.4 Vpp on pin 6 of IC₂ in the frequency reference. The noise level normally measured after Kalundborg switches off at 0.30h CET is about 100 mVpp.

Reference:
(L. N. Jensen, OZ6LV - 924091)
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Omni-Pro II - The Next Generation

When you get a new product, what are your main concerns? Freedom from frustration is certainly one important consideration, for your time is valuable. You will want a product which is reliable and sophisticated, yet simple to use, with clearly written documentation. You will be looking for a high standard of technical support and regular upgrades for the product.

We at Dataman recognise how difficult it can be to choose between programmers which look and cost much the same. So, instead, why not concentrate your effort into choosing a reliable vendor. Dataman has been the leading vendor of low-cost programmers for as long as the market has existed. Any of our customers will tell you that Dataman has always supplied excellent well-supported products. That's why we're still here! We take technical support seriously, We give you your money back, if you're not satisfied.

These are important points to consider. But now let's take a look at some of the special benefits of owning Omni-Pro II.

What Benefits?

Well, for instance, the interface is not via the computer's parallel port, which is speed-limited, and probably connected to your printer. A dedicated plug-in half card performs fast data transfers. The software is a professional package in full colour that will run in only 400K of RAM. What's more it will run on any PC/AT or compatible - even the latest 486 machines. That's because Omni-Pro II has its own independent clock - some programmers rely on the computer for timing, and won't work with faster machines.

Ground pins are connected by relays - not by logic outputs. Some vendors won't approve programmers which don't ground pins in this way. The 40-pin Textool socket can be changed without even having to remove the cover. A complete range of PLCC adapters is available.

Truly Universal

Omni-Pro II has universal pin-drivers which will accommodate a very wide selection of parts. You can program BIPOLARS, PROMS, E/EEPROMS, PALs, GALs, PLAs, PROMs, E/E/EP LDS and MICRO-COMPILER. The latest FLASH EPROMS are supported too. The list has 1250 devices already and substantial numbers of new devices will be added FREE every quarter.

We provide optimised programming speeds, using algorithms like Quickpulse, Flashrite and TI Snap and have already gained parts approval from TI, NS, and ICT. We provide fast downloading of files in any standard format: Intel Hex, Motorola, Tek Hex, HP4000, ABSS, Binary. You can also send JEDEC files from all popular PLD compilers and JEDEC standard vector testing is supported: a full array of test condition codes can be generated.

Remember - you get a 30 day money-back guarantee, FREE quarterly software updates and FREE technical support - as much as you need. Phone now for a free Demo Disk and up-to-date Device-List.

Omni-Pro II comes with a FREE copy of NS's superb Open Programmable Architecture Language - OPAL Junior.

Gang-of-eight Programmer...£395

This production programmer from Dataman can handle all 25 and 27 pin EPROMS up to 512K bits. Programs eight copies from a master EPROM, or from an object-file. The G8 offers fast programming methods and three user-selectable programming voltages. G8 is clearly designed for the busy workshop being supplied, as standard, in a high quality steel case.

Strobe Eraser.........£175

UK customers please add VAT. Major credit cards accepted. UK delivery available next working day.

Software Development from £195

Dataman's Software Development Environment, SDE, comprises a two-window Editor, Macro Assembler, Linker, Librarian, Serial Comms and intelligent Make facility. The latest FLASH EPROMs are supported too. The list has 1250 devices already and substantial numbers of new devices will be added FREE every quarter.

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Simply the best!

THE VELLEMAN K4000.
GUTSY & GOOD LOOKING.
SOUNDS GREAT!
PRICE FOR PRICE, THE
BEST VALUE, BEST
SOUNDING, BEST LOOKING,
STATE-OF-THE-ART,
VALVE POWER AMPLIFIER
KIT THAT’S AVAILABLE.
VELLEMAN, SIMPLY THE BEST!

The Velleman name stands for quality, and the K4000 valve amplifier is supplied with everything you’ll need to build it, including a ‘Get-You-Working’ back-up service.

Delivering 95 watts in class A/B1, the K4000 is, without doubt, price for price, the best sounding, 'guttiest', most handsome valve power amplifier kit available anywhere.

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