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ELEKTOR ELECTRONICS

THE INTERNATIONAL ELECTRONICS MAGAZINE

Electronics know-how: Batteries

BUMPER CHRISTMAS ISSUE

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RF immune power supply

PSD3xx chips

1000 Hz sine wave generator

The Digital Solution

Proximity detector

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ELEKTOR ELECTRONICS

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Apologies

We regret that owing to a reprographical error page 34 of the November issue was repeated on page 35. The missing page 35 is published on page 107 of this issue. Those readers who do not wish to buy the December issue can obtain page 35 on request from our editorial office in Dorchester.

We also regret that in certain issues incorrect glue was used to fix the subscription leaflet. Readers with a damaged contents page can obtain a new one on request from our editorial office in Dorchester.

Front cover

The photo shows the circuit of a 1-to-3-phase converter and some motors whose speed can be controlled with it. The unit, described on pp. 26–30, converts the single-phase mains supply to a 3-phase voltage which can control small three-phase motors rated at up to 725 W, irrespective of whether these are synchronous or induction types.

To all our readers: thank you for your continued support during the past year and may peace be with you in the coming year.

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1000-HZ SINE WAVE OSCILLATOR

Oscillators exist in a bewildering number of types and variants, each with its own, specific, features. A type known for its near-perfect sine wave output is the Wien bridge oscillator. This well-tried design is revamped here with a special trick to stabilize the output signal level.

Design by T. Giesberts

SPECIFICATIONS	
Frequency:	1 kHz
Max. output voltage:	8 V_{rms}
Distortion (THD+noise):	<0.0003%
Supply voltage:	±15 V
Current consumption:	approx. 10 mA

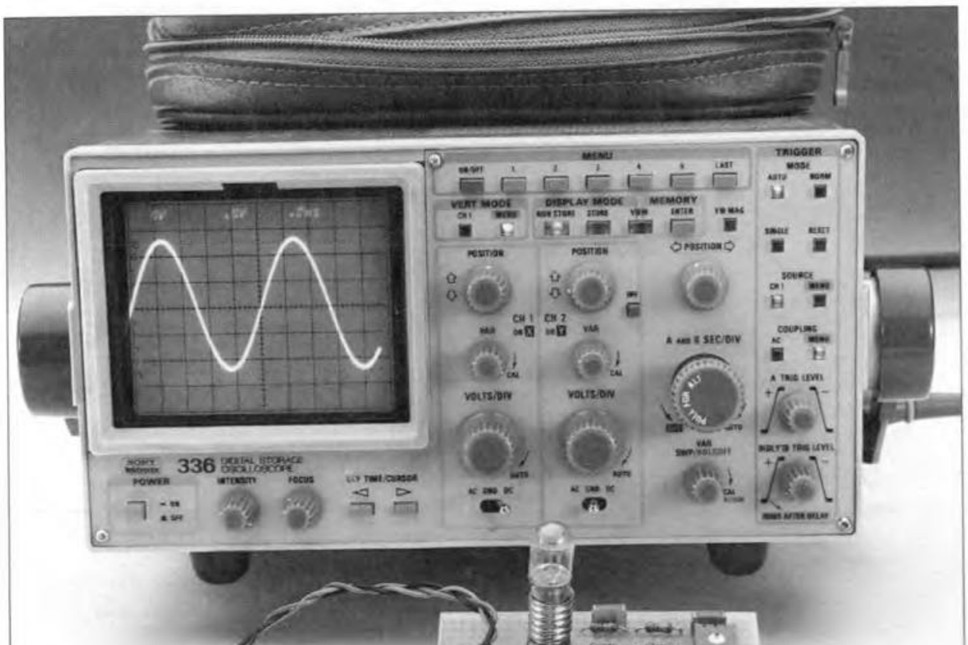
ONE of the areas of electronics which is known for its tremendous variety is oscillator design. You name an application and there is an oscillator design to fill the bill. There exist *RC*, *LC* and crystal oscillators, high and low frequency oscillators and high and low power types. The output signal is either a sine wave, a rectangular wave, or a sawtooth. Furthermore, many oscillators have output level and frequency controls. Either it exists, or it can be made.

If you want to do elementary tests on audio equipment (like faultfinding and repair), you probably require a simple signal source capable of supplying a 'clean' sine wave. The ultimate, of course, is to have an oscillator with adjustable frequency, although one with a fixed frequency, say, 1 kHz, is also very useful. The classic Wien bridge type oscillator is just the thing for this application. The design, although simple, is well-tried, and enables a good quality sine wave generator to be built.

The Wien bridge consists of an *RC* series network and an *RC* parallel network. The two capacitors have the same value, and the same goes for the two resistors. At the central frequency, the phase shift is nought because then the two networks introduce an equal phase shift with opposing signs. The attenuation is then three times. Consequently, the circuit can be made to oscillate by adding a simple amplifier with a gain of ×3, as illustrated in **Fig. 1**.

Filament control

Recapping the above, a sine wave oscillator based on the Wien bridge principle has few ingredients: two RC networks and an amplifier. Attractive as that may sound from a point of view of cost and component count, there are a few extra points to take into consideration if there are more than average demands on the shape of the sine wave produced by the circuit. To begin with, the amplifier used must be a near perfect type. Also, a smart control system is called for to make sure that sufficient gain is available to initiate and maintain the oscillation. However, the amplifier must not be overdriven,



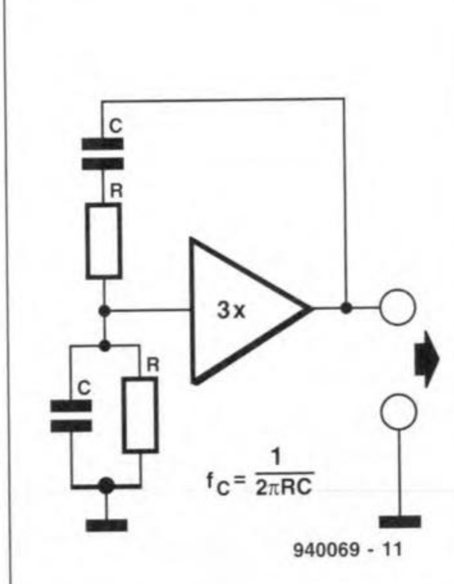


Fig. 1. A Wien bridge oscillator consists of an *RC* series network, an *RC* parallel network, and an amplifier to overcome the attenuation at the 'null' frequency.

because that, as is easy to understand, would cause 'damage beyond repair' to the beautiful waveshape you are after.

The first of the above points can be solved by using a good quality opamp in the amplifier. Since a wide range of excellent opamps is available these days, that need not be a problem. The second point, the control system, turns out to be unexpectedly simple as it can be realised by an ordinary lamp.

The circuit diagram of the oscillator is shown in **Fig. 2**. Arguably, the resemblance to Fig. 1 is striking. The Wien bridge proper consists of R_1, R_2, C_1 and C_2, and is arranged for a frequency of 1 kHz. Capacitors C_1 and C_2 are Siemens 'MKT' (metal theraphtelate) types which you will have to match for a tolerance of 1% or better to make sure that the circuit oscillates. If you want an exact output frequency of 1 kHz, these capacitors should have a (theoretical) value of 119.67 nF.

The amplifier is based on an opamp type OPA627. The required gain of about three times is defined by feedback network P_1-R_3-R_4-La_1. In this network, the lamp acts as a positive temperature coefficient (PTC) resistance. The PTC is provided by the lamp's filament, whose resistance is temperature dependent. When cold, the filament has a resistance of about 25 Ω, which doubles (roughly) when the normal operating temperature is reached. This characteristic is exploited here by incorporating the lamp in the feedback network. When the lamp is 'cold', the gain of IC_1 is determined mainly by the ratio R_3/R_4. As the output voltage rises, however, the current through the feedback network rises also, as does the resistance of the lamp. Consequently, the ratio of the network changes, causing the feedback to increase and the gain of IC_1 to decrease. As a result, the output voltage drops, and the current through the lamp drops also, causing the gain of IC_1 to rise again. After a short warming up period, the oscillator output level will settle at a stable value.

The electronic counterpart of the lamp in the lower branch of the feedback network is a preset, P_1, in the upper branch. This preset allows the bridge to be 'balanced' accurately. The operation of the output level control system is further stabilized by an additional current source, R_5, for the lamp. This resistor may have to be changed a little to match the type of lamp used. In the prototype of the oscillator, good results were achieved with a value of 3.3 kΩ. Should you find that the output voltage still rises a little after a few minutes, the value of R_5 must be decreased. Likewise, a slowly decreasing output voltage calls for a somewhat

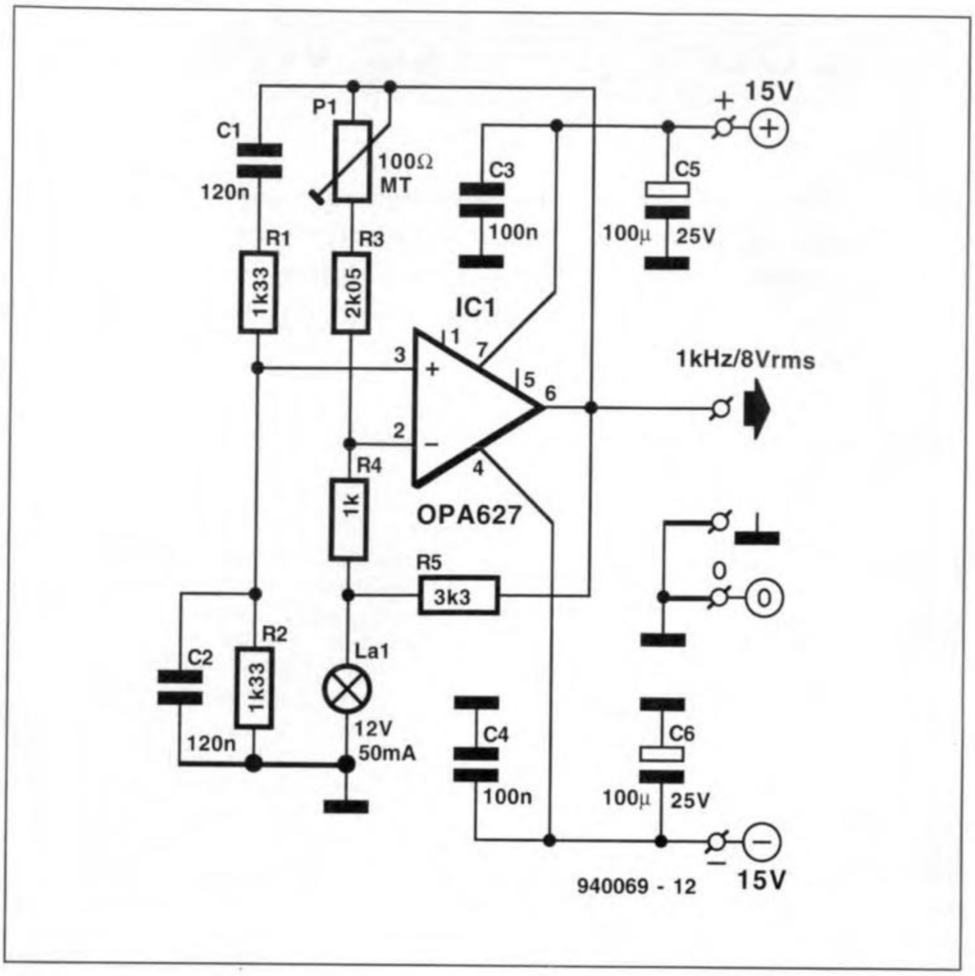


Fig. 2. Modern electronics with a touch of nostalgia: a small lamp is incorporated in the oscillator's feedback network. The PTC characteristic of the lamp filament is exploited to obtain an effective automatic gain control.

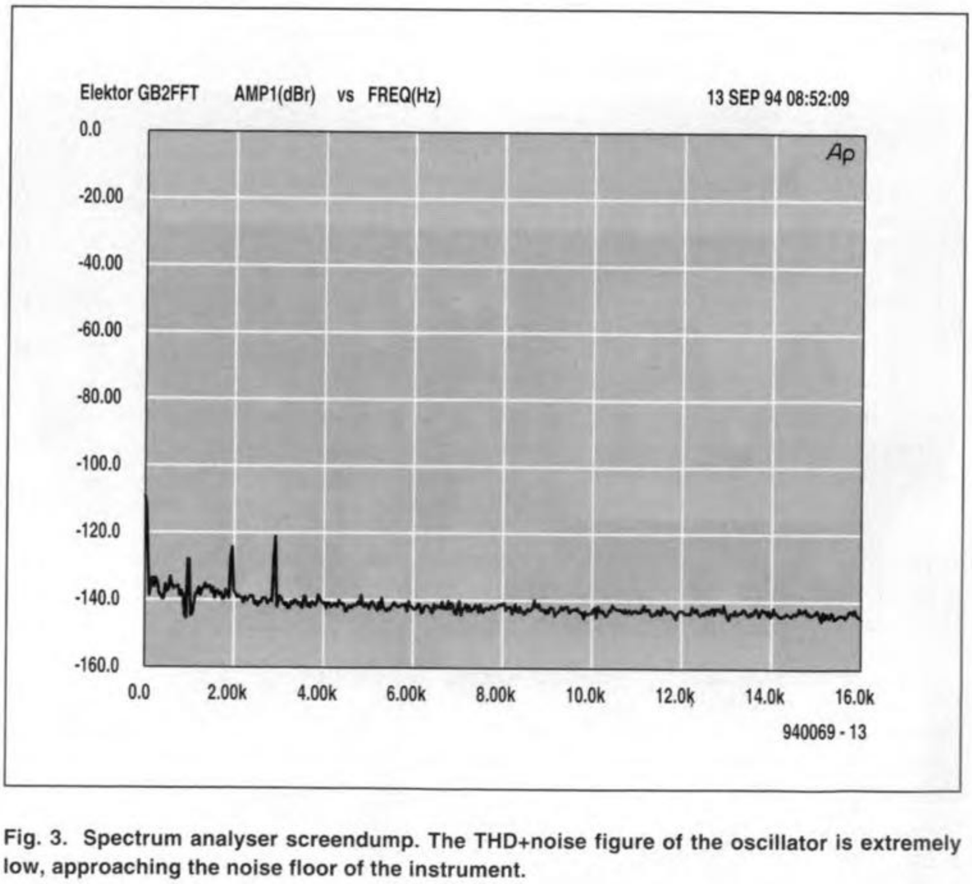


Fig. 3. Spectrum analyser screendump. The THD+noise figure of the oscillator is extremely low, approaching the noise floor of the instrument.

higher value of R_5. The optimum value can be found by trial and error only.

Good specifications

The use of a lamp as a PTC device demands that the resistors around IC_1 have relatively low values. Obviously, a reasonable current has to flow through the lamp to make it act as a PTC device, and that can only be achieved with low value resistors.

An advantage of low-value resistors is that their noise contribution is low. On the down side, an opamp has to be used which is capable of driving a load of 600 Ω (minimum) at a symmetrical supply voltage of ±15 V. That is not a problem for the OPA627 opamp used here, which has FET inputs, a low noise factor (5.6 nV/√Hz at 1 kHz), a low input offset (max. 100 µV) and a very low distortion. The latter specification is verified by the THD+noise characteristic shown in **Fig. 3**, produced by our laboratory spectrum analyser. The measurement was made at an oscillator output voltage of 8 V_{rms}, and a bandwidth of 22 Hz to 22 kHz. The actual THD consists almost entirely of the second and third harmonics, which together tot up to a remarkably low 0.00014%. The total THD+noise specification remains below 0.0003%, which is not bad for such a simple circuit.

Unfortunately, the OPA627 is a fairly costly device. If you can live with a slightly higher distortion, the OPA627 may be replaced by the (pin compatible) NE5534.

Construction and adjustment

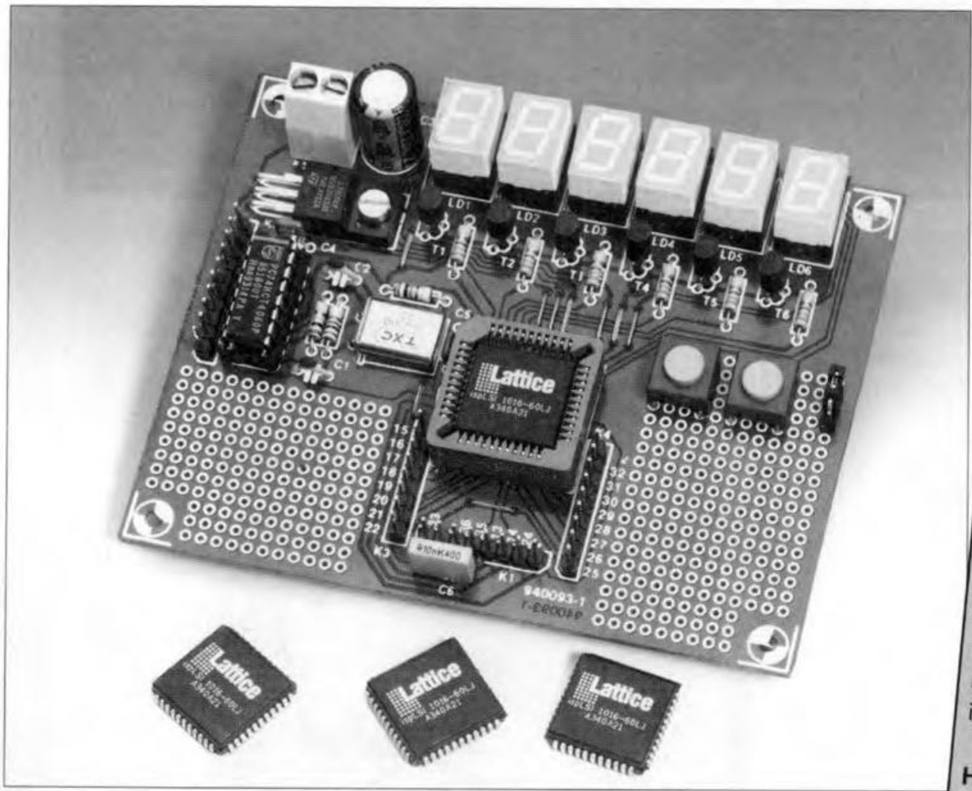
The circuit is so simple that a printed-circuit board is probably not necessary. The introductory photograph illustrates that the sine wave oscillator is fairly simple to build on a piece of veroboard. Since the current consumption of the circuit is a modest 10 mA or so, a small and simple symmetrical supply may be used. The smallest available mains transformer with a 2×15-V output, a bridge rectifier and two three-pin voltage regulators type 78L15 and 79L15 are perfect for the present application.

Take your time to adjust P_1. In practice, good results were obtained by adjusting the preset for the highest possible output level. That setting should also produce the lowest distortion, while it prevents the output voltage from rising to unexpected, and undesirably high, levels after the control system has settled.

(940069)

ispSTARTER KIT FROM LATTICE

Recent developments in electronics allow logic components to be programmed 'in-circuit'. A most interesting spin-off of this trend is that it is now possible, for the first time, to design your own logic parts as one-offs for small-scale applications.



ispStarter Kit Contents

Software

pD1016	pDS software (pLSI/ispLSI development system) for ispLSI1016
ispGAL	download software
ispGDS	compiler and download software
ispCODE	ANSI-C compiler source programming software

Datasheets

ispLSI1016
ispGAL22V10
ispGDS22/18/14

Samples

ispLSI1016-60LJ
ispGAL22v10B-15LJ
ispGDS14-7

Hardware

ispDownload cable

By our editorial staff.

Source: Lattice Semiconductor Corporation.

LIKE so many electronic components, those that fall into the 'programmable logic' category have advantages as well as disadvantages. The advantage is that users of special and 'made to measure' components can profit from the law of production volume. Because of the high production numbers these parts remain affordable even in small quantitities. On the down side, any type of programmable logic invariably seems to be tied up with special programming systems that essentially turn a 'blank' device into one tailored to the application. Apart from special hardware, programming also requires dedicated software.

The ispLSI series from Lattice Inc. heralds a new generation of logic components which do not have the traditional drawbacks of programable logic. The great thing about ispLSI is that it offers you all the advantages of programable logic without having to purchase costly programming equipment. In the case of ispLSI components, a simple cable and some hardware fitted into a 25-pin sub-D plug are used to link an MS-DOS PC to a special connector on the application board. All signals needed to program the ispLSI device are conveyed via this link. No programmer, no special voltages.

This article describes the hardware and software which are needed to get going with ispLSI (In-System Programmable Large Scale Integration) logic, and in particular the ispStarter Kit marketed by Lattice. A simple ispLSI programmer is described in a another article in this issue. That article also provides a general background to the exciting ispLSI technology.

Get started with ispLSI

Figure 1 shows the circuit diagram of a small development system based on an ispLSI1016, IC_1. The special connector which is needed to program the IC is hooked to header K_1. All signals on this connection are fed directly to the ispLSI device. The function of the signals on this 8-way connector is described in the 'Low-cost ispLSI programmer' article in last month's issue. For the time being, it is sufficient to know that the signals originate from a compact programmer interface which is connected to the Centronics port of a PC. With a few exceptions, all other processor pins are available for I/O functions. Pins 37-44 (I/O16 through I/O23) are connected directly to the segment terminals of six LED displays. The I/O lines on pins 3 through 8 (I/O24 through I/O29) are used to enable and disable each LED display via a switching transistor. The absence of current limiting resistors may strike you as odd, but bear in mind the duty factor of 1/6 (0.16) of the display drive signals. This can be done with impunity as long as four or more displays are used. If fewer displays are used, the multiplex ratio must still be 1/6. Alternatively, connect resistors in series with the display segments.

The non-used I/O lines of the

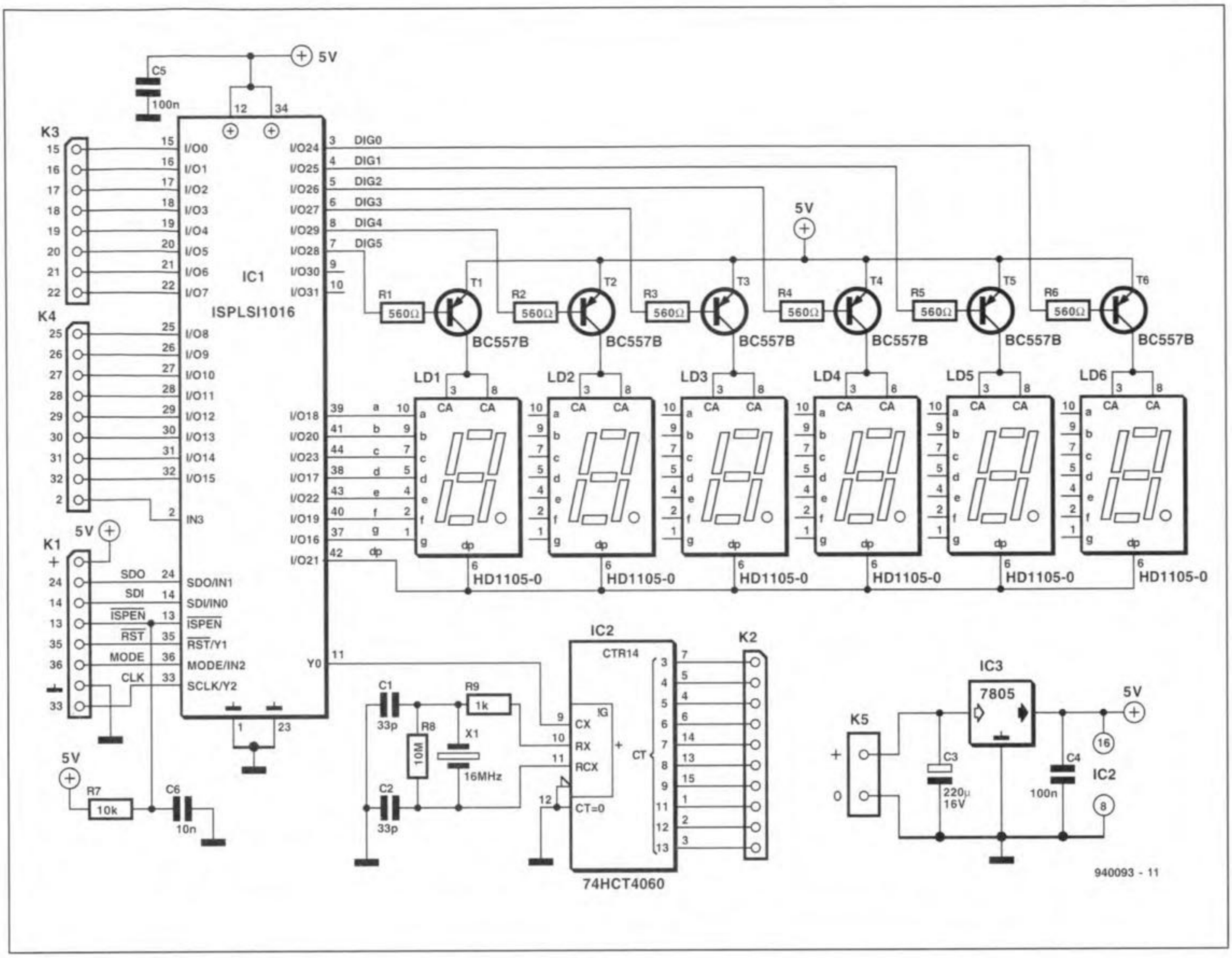


Fig. 1. Circuit diagram of the experimenting board on which ispLSI components can be programmed without additional hardware. Together with the ispStarter Kit supplied by Lattice, this board forms a powerful ispLSI development system.

ispLSI1016, I/O0 through I/O15, are available for experiments, and brought out to pins on connectors K_3 and K_4. Input IN3 is also 'free', and connected to K_4. The last part of the circuit is formed by IC_2. This CMOS 4060 generates the clock signal for the ispLSI1016. In the present circuit, the ispLSI1016 is supplied with a 16-MHz clock. Connector K_2 supplies the signals derived from this clock. These signals may be used to supply suitable clock signals to external logic circuits connected to the processor.

The 5-V power supply is conventional, based on a 7805 and two capacitors.

Construction

The circuit is best built on a prototyping board of which the artwork is shown in **Fig. 2**. This board has a large prototyping area which has plenty of space to build your own experimental extensions. The printed circuit board is available ready-made through the *Elektor Electronics* Readers Services.

The construction of the board is entirely straightforward. Fit a socket for the ispLSI device (IC_1), taking good care not to fit it the wrong way around. The same goes for the other components. Connectors K_1 through K_4 are single-row pin headers. The circuit is best powered by a mains adapter with d.c. output. The adapter is connected to PCB terminal block K_5. Do take care not to reverse the + and – wires on this connector. If you do, components will be damaged in the circuit.

Sample application: a clock

The diskette supplied with the printed circuit board for the ispLSI development board contains an example which demonstrates how a small clock can be realised with the system. Unfortunately this clock is a little fast. In spite of the many registers in the ispLSI1016 is not possible to derive a 1-second 'tick' from the 16-MHz device clock. This is, however, possible if an ispLSI1032 is used. Unfortunately, that device is not supported by the ispStarter Kit. However, the fact the 1016-based clock is a little fast should not distract from the usefulness of the example, which is first and foremost aimed at making you acquainted with ispLSI programming.

The time is displayed on LD_3-LD_6. A jumper at pin 15 allows you to select between 12-hour and 24-hour indication.

The clock is adjusted with two press-keys. It is advanced slowly by pulling pin 17 of the ispLSI1016 to ground. That is done with a press-key. Similarly, the 'fast' setting is accomplished with the aid of the press-key connected to pin 16. The pulses to be counted originate from pin 3 of IC_2, and are applied to pin 20 of IC_1.

The example programs on disk

clearly illustrate how to go about programming ispLSI devices. However, to actually be able to use the program, you also need software that forms part of the ispLSI Starter Kit supplied by Lattice.

The software

The ispStarter Kit supplied by Lattice contains development software for the ispLSI1016 (pDS software; pLSI Development System), software for the ispGAL22V10 (yes, a GAL in isp technology), an ispGDS (isp generic switch) Compiler with downloader, and an ANSI-C compiler with source codes for the download routines. The kit also contains the cable needed to connect the experimenting board to your PC. Plus there's datasheets and one sample each of the ispLSI1016, ispGAL22V10 and ispGDS22/18/14 devices. A really wonderful, and affordable, kit which allows you to get started straight away.

The development software can be used on any PC running Windows 3.0 or 3.1. The flow diagram in **Fig. 3** shows the different steps that make up the design process. To begin with, the desired circuit is divided into logic blocks (GLBs, Generic Logic Blocks) and I/O (input/output) cells. This step is necessary to check beforehand if the design can actually be fitted into, say, an ispLSI1016. The logic functions offered by this device are listed in **Table 1** in last month's introductory article on ispLSI. The standard blocks available are:

- Generic Logic Blocks (GLBs);
- Output Routeing Pools (ORPs);
- I/O cells;
- a clock signal distribution network;
- a connection matrix.

The basic functions of these blocks are discussed in the 'Low-cost ispLSI Programmer' article presented in last month's issue of *Elektor Electronics*.

Use the software to formulate the logic relations associated with the GLBs and the I/O cells. The graphics editor enables these functions to be

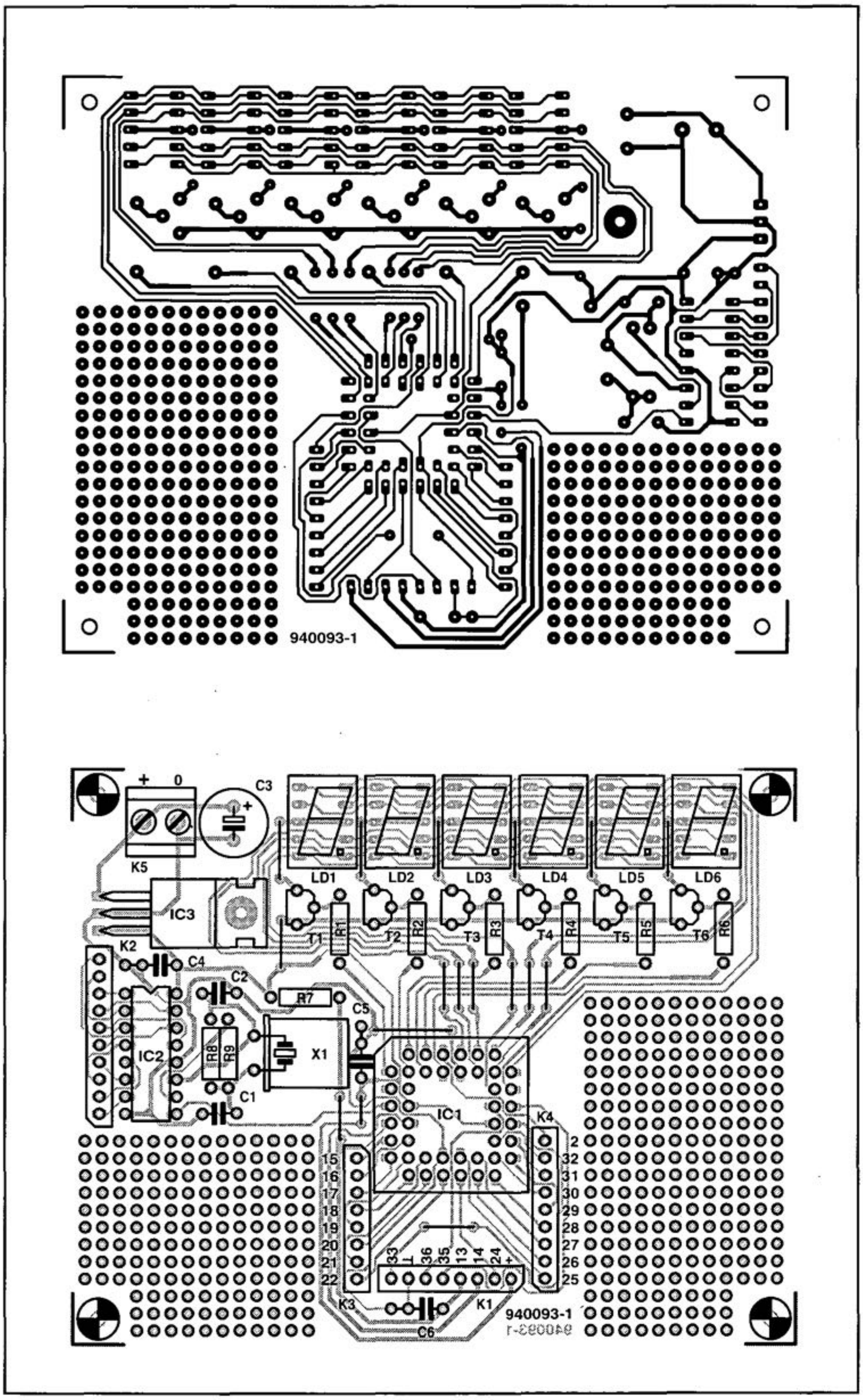


Fig. 2. Track layout and component mounting plan of the printed circuit board designed for the ispLSI experimenting board. Board available ready-made, see page 110.

COMPONENTS LIST

Resistors:
R_1-R_6 = 560Ω
R_7 = 10kΩ
R_8 = 10MΩ
R_9 = 1kΩ

Capacitors:
C_1,C_2 = 33pF
C_3 = 22µF 16 V radial
C_4,C_5 = 100nF
C_6 = 10nF

Semiconductors:
T_1-T_6 = BC557B
IC_1 = ispLSI1016 (Lattice Inc.)
IC_2 = 74HCT4060
IC_3 = 7805

Miscellaneous:
K_1-K_4 = 10-way header, single-row
K_5 = 2-way PCB terminal block
X_1 = 16MHz crystal
LD_1-LD_6 = HD1105-O (Siemens)
PCB and software on disk, set order code 940093.
PCB (separately): order code 940093-1.
Software (separately): order code 946204-1.
Prices and ordering information on page 110.

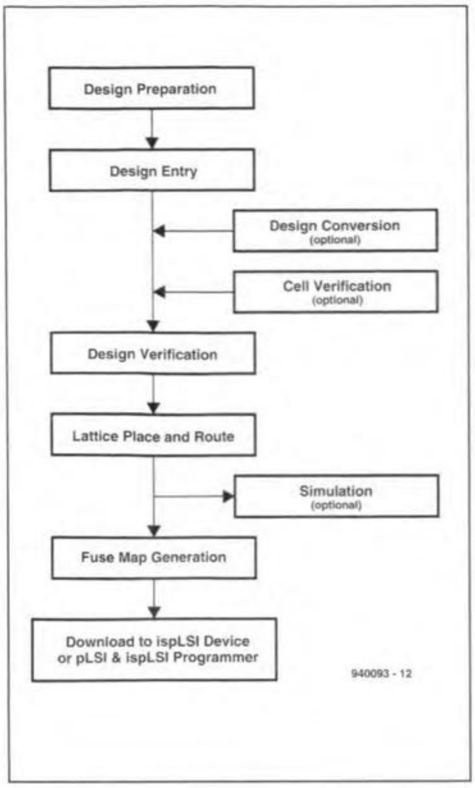


Fig. 3. These are the steps you have to go through if you want to program an ispLSI device.

entered in Boolean notation, or as macros. Macros can be picked from a library which forms part of the development system, and come complete with the indispensable comment on their function.

Alternatively, the device can be configured with the aid of a Lattice design file (*.LDF). Once you know the syntax and conventions, such a file can be produced using almost any word-processor capable of outputting ASCII text. Interestingly, a design produced via the PDS software is capable of exporting the PDS software in that format.

A 'cell verification' utility is available to test individual functions for design errors. The utility is extremely useful as it reveals any structural error contained in the set of logic functions. If possible, logic functions are minimized, while the software also checks if it is at all possible to realize the desired logic functions on the basis of the components selected. Next, the entire design is taken through a 'design verify' operation.

Once it is for sure that the design is error-free, the software compiles the programming algorithm for the relevant ispLSI. This algorithm is used at a later stage by the FuseMap program. Furthermore, the netlist used by the Lattice Route and Map are updated.

The Router starts to work as soon as the software has decided that the design meets all requirements. On the basis of the information contained in the input files, a division is made, and the logic sub-components are interconnected. The window which appears on the computer screen while this operation is being performed allows the user to interact in the process. For instance, it is possible at this stage to tell the program which pins are to supply the different I/O signals. Finally, the configuration of the IC is established.

Next, FuseMap is started. This sub-program generates the JEDEC file which is to be sent to the ispLSI programmer. During programming, all device inputs may be provided with a pull-up resistor. That may be necessary to prevent undefined logic levels at the inputs, causing the device to latch up. The final option which is still open to the user is to actuate the device read-out protection. That may be done to prevent copies of the programmed device being made.

An important advantage of the ispLSI environment is that it is always possible to convert an IC design to another component. That may be useful if it appears that an improper (too small or too large) IC type has been selected for the purpose. In such cases, a more appropriate component can be selected without problems, and the 'design' copied to it. Note however that that is only possible with Lattice's full-blown development system which unfortunately is much more expensive than the ispLSI Starter Kit.

(940093)

Fig. 4. The ispLSI Starter Kit sold by Lattice Inc. and its distributors.

The ispStarter Kit mentioned in this article is available from Lattice distributor MicroCall Ltd., 17 Thame Park Road, Thame, Oxon OX9 3XD, England. Tel. (0844) 261939, fax (0844) 261678.

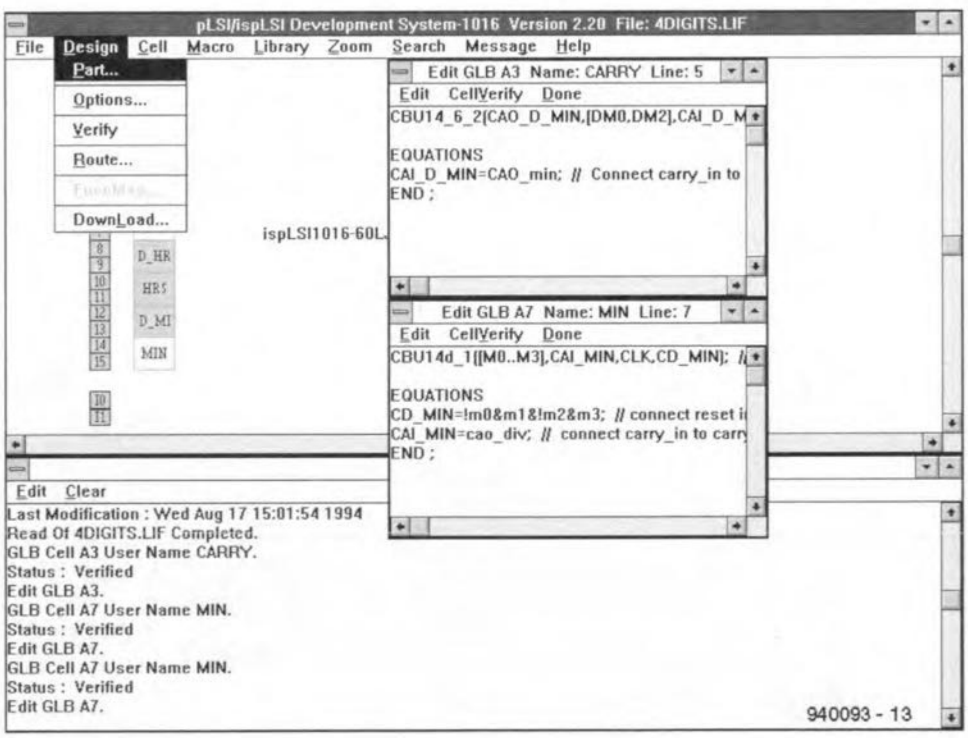


Fig. 5. This screendump shows how the ispLSI development software communicates with the user. Thanks to the Windows graphics user interface the software is simple to use.

PSD3XX CHIPS

By A. Rietjens

Microprocessors are powerful devices and it is, therefore, not to be wondered at that their popularity has grown so fast in such a short time. A disadvantage remaining until recently was the need of several more ICs to design a complete microprocessor system. Since more components require more space and result in higher costs, a demand arose for more powerful system components to simplify the design of a microprocessor circuit. US chip manufacturer Wafer Scale has responded to this need with the introduction of its PSD3xxx family.

The chips in the PSD3xxx family contain flexible I/O ports, a PLD, Page Registers, EPROM (choice of 32 kbyte, 64 kbyte, or 128 kbyte), 2 kbyte static RAM and some logic circuitry to make connection with the microprocessor—see **Fig. 1**. Because of the power of the new chips, the design of a microprocessor system is brought down to two components (since the latch for demultiplexing of the address and data buses is no longer needed). A similar function would until recently have required 8 to 12 discrete ICs. The chips have been designed to work with a variety of microprocessor ICs, including, for example, the 68HC11 as well as the Z80. The PSD3xxx family may be split into two branches: the PSD31x, intended for 8-bit processors, and the PSD3x for 16-bit processors.

A block diagram of one of the new chips, the 16-bit PSD30x, is given in **Fig. 2**. At the left are all the functions needed to make connection with a microprocessor; and at the right, the I/O functions. The memory banks are in the centre.

Inputs AD_0–AD_7 reach the nucleus of the IC via latches, which can be arranged to store data. This obviates the need of a separate register such as a 74HCT373 or 74HCT573. In the transparent mode, the latches function as buffers. However, the latched mode is more usual, for example, with the 8031. Data storage in the latches at the input of the circuit is then accomplished by an ALE instruction (in Motorola processors better known as Address Strobe or AS). The polarity of the inputs is determined by the operator as required. As long as the ALE or AS signals is valid, the input stage is transparent. When this signal becomes inactive, the data are stored in the latches.

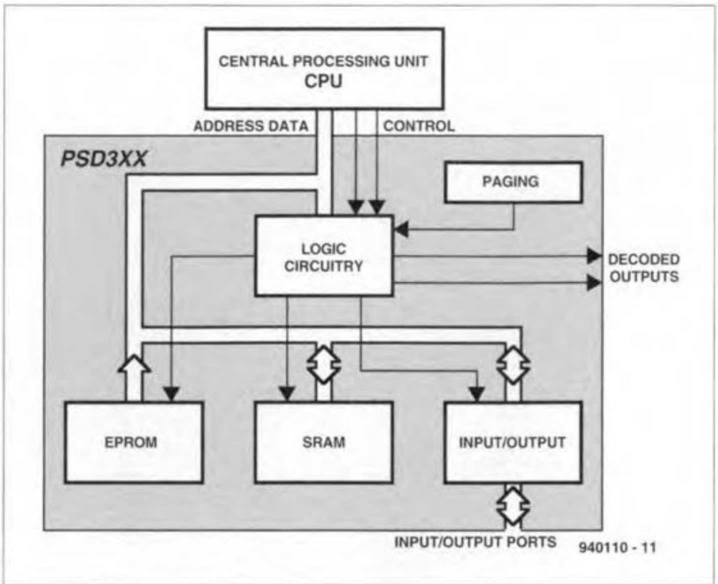


Fig. 1. Functions that can be assumed by a PSD3xxx chip.

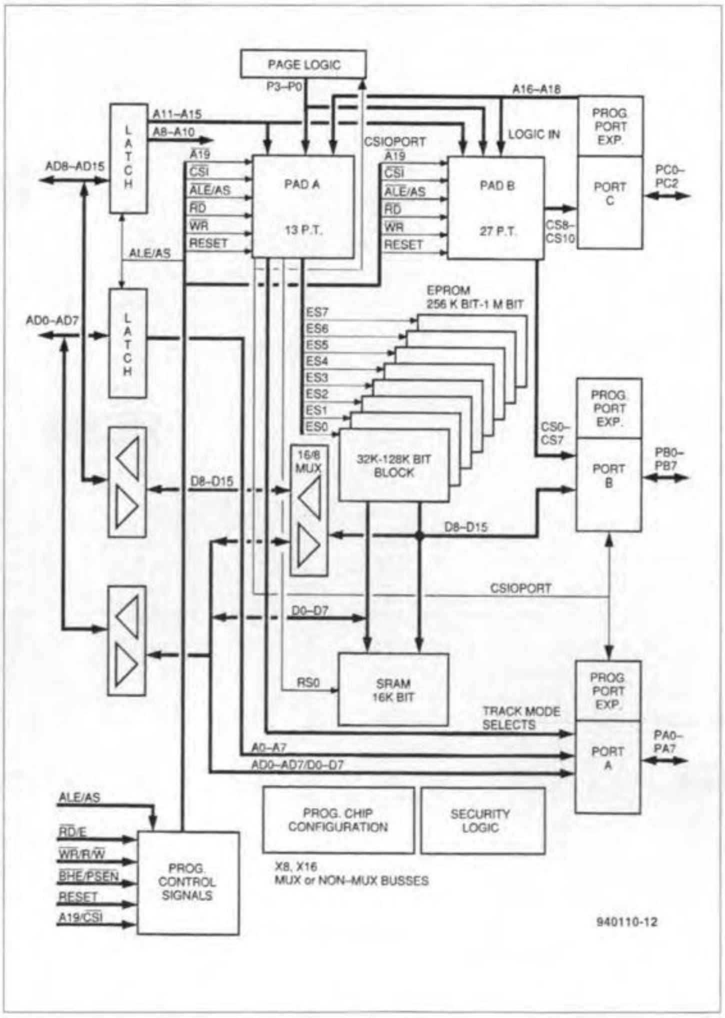


Fig. 2. Block diagram of the 16-bit PSD30x; in the 8-bit PSD31x the 16/8 demultiplexer and associated buffer are not used.

The Programmable Address Decoder (PAD) plays an important role. It will be seen in **Fig. 2** that address lines A_{11}–A_{15} and, if desired, A_{16}–A_{19}, are connected directly to the PAD. Other input signals applied to the decoder are RD(E), WR(R/W) and ALE (AS). Programming of the PAD gives the user the opportunity of selecting EPROM banks internally via the ES_0–ES_7 lines.

Also, there is a selection signal for the static memory, RS_0.

Port C is a 3-bit I/O port with two functions: it can take internal signal CS_8–CS_{10} outside, or it can receive address lines A_{16}–A_{18} and pass these to the PAD. Moreover, address line A_{19} can be passed directly to the PAD. This shows that the PAD is capable of arranging an entire address decoding up to 1 MHz without any external components.

Ports A and B are 8-bit ports that can be used by the operator as conventional I/O ports. **Figure 3** shows the special mode in which I/O ports A, B and C may be used. This is especially useful if an 8-bit processor system uses a multiplexed address and data bus (AD_0–AD_7). Address lines A_8–A_{15} are then available as normal and may be applied to the address inputs. Ports B and C are then available as I/O functions or for passing A_0–A_7 or AD_0–AD_7 lines (track mode) respectively.

Apart from functioning as an I/O port, port B can also take the chip-select signals of the PAD outside. Port C can then be used to obtain more inputs, or for writing additional address lines A_{16}–A_{18}, or for taking chip-select lines CS_8–CS_{10} outside.

Address lines

The 16 address lines of a multiplexed bus are stored in one or two 8-bit wide latches, depending on the type of processor. With non-multiplexed

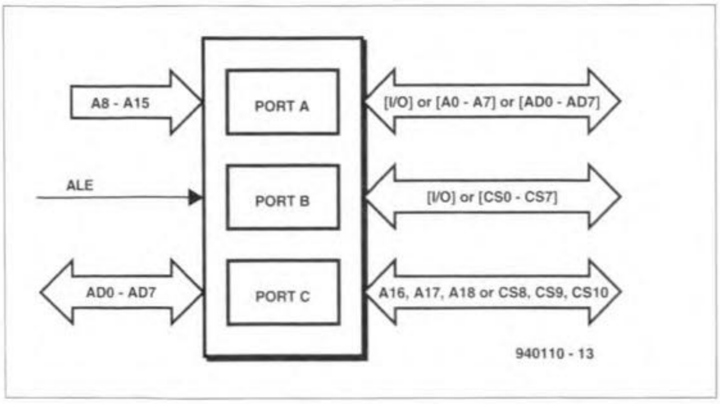


Fig. 3. The ports in the PSD3xx may have several functions.

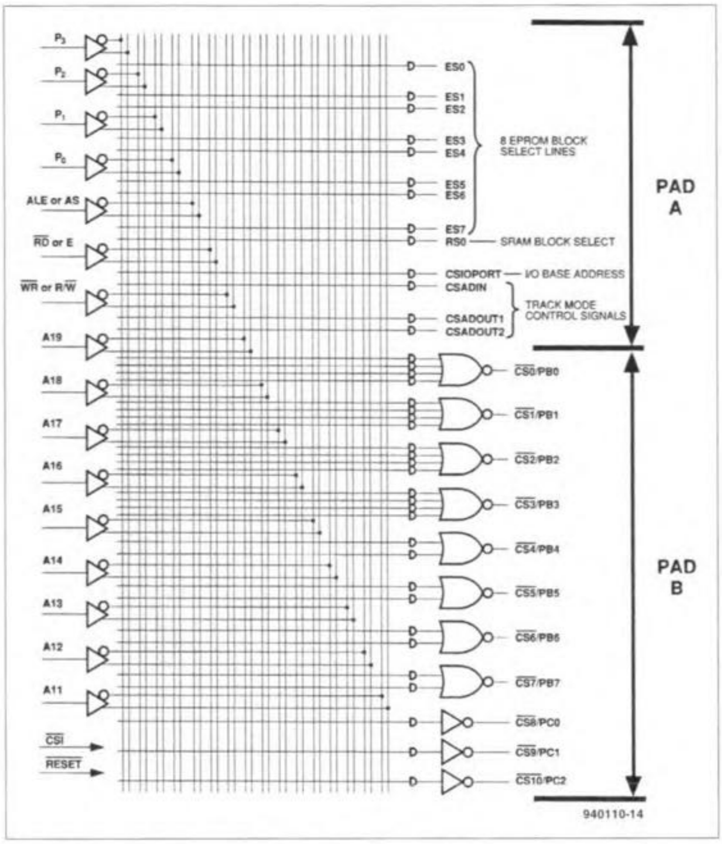


Fig. 4. The array with which the PAD is programmed.

buses, the inputs remain transparent. Address lines A_1–A_{10} go directly to the static memory, while address lines A_1–A_{11} are applied to the EPROM. The EPROM banks are selected via selection lines ES_0–ES_7, which originate in the PAD. The static memory is selected by the RS_0 signal that is also generated by the PAD. Address lines A_{11}–A_{15} and optional address lines A_{16}–A_{19} are used in the PAD array.

Internally, the memory banks are word wide (16 bits); provision is made by an isolation buffer for splitting this into two bytes. The operation of the buffer is enabled by the configuration of the PSD. If this is configured to work in the 16-bit mode, the data in the two blocks (D_0–D_7 and D_8–D_{15}) are buffered. If it is set to the byte-wide mode, the 8-bit wide data stream is controlled with BHE and A_0. Which of the two bytes is addressed depends on the level of A_0.

The EPROM section is divided into eight banks., which are selected by ES_0–ES_7. These signals originate in the PAD.

Figure 4 shows a detailed sketch of the construction of the PAD. Depending on the choice of the user, this section of the PSD generates the selection signals from the address signals and some control signals. The PAD is a reprogrammable fuse array with an EPROM-like structure. Used with Intel microprocessors, the PAD employs, apart from inputs A_{11}–A_{19}, signals ALE, RD and WR. With Motorola processors, signals R/W, AS and E are used. Inputs CS_1 and Reset are used to deselect the PAD, a state that is desired during power-down and initialization.

In the PSD301, signals ES_0–ES_7 are used to address

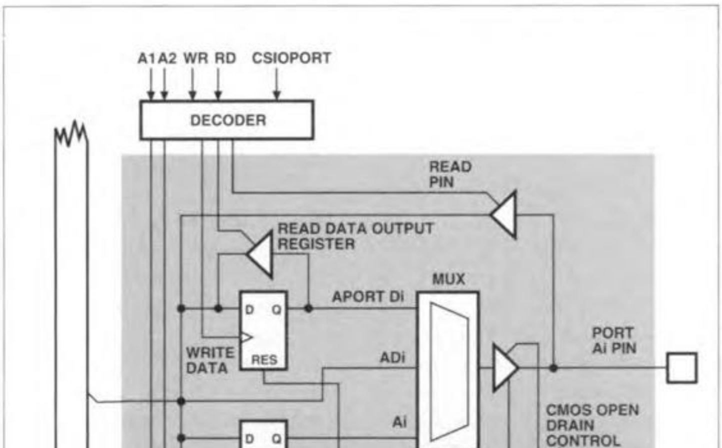


Fig. 5. Structure behind a single I/O line of port A.

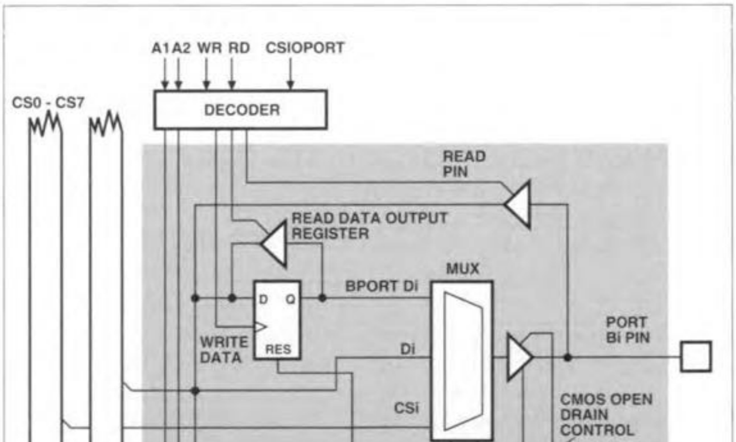


Fig. 6. Structure behind a single I/O line of port B.

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the EPROM banks. There is always one product term available per EPROM bank. A product term, RS_0, is also used for selecting the static memory.

The address and control lines of the EPROM are split into a block capacity of 4 to 16 kbyte. The exact location of the block in the address range of the processor can be decided freely by the user. The 2 kbyte available for the static RAM can be selected in a similar manner.

Other product terms provided by the PAD are CSIOPORT, CSADIN, CSADOUT1 and CSADOUT2. The single product term CSIOPORT determines the base address of ports A and B. An offset must be added to this base address to reach the various registers. Table 1 shows the structure of this division.

Port structure

The complete port section of the PSD3xx chips contains three registers: A (8 bits), B (8 bits), and C (3 bits). These registers support various I/O functions. For example, ports A and B may be arranged as 8-bit I/O ports that send data to, and receive data from, external components. **Figure 5** shows the structure of a single cell in port A, while **Figure 6** shows that of a single cell in port B. Writing data to a port is the same as writing data to a RAM location.

Although a port can not be addressed at bit level, it is possible to determine of each I/O line whether it is arranged as input or as output. Any combination of inputs and outputs (for example, PA_0–PA_5 input and PA_6 and PA_7 output) is thus possible.

Whether an I/O pin is arranged as an input or an output can be determined with the data direction register. Since this register operates dynamically, it is possible during the execution of a program to adapt the function of the I/O pins of port A and/or port B.

After a reset, all bits in the data direction register are low level, that is, the ports are set as inputs. If in the application only inputs are used, nothing more needs to be done. If outputs are wanted, the associated bits must be made high.

[940110]

Name of register	Offset w.r.t. base address
Pin register port A	+2 (byte)
Pin register port B	+3 (byte)
Direction register port A	+4 (byte)
Direction register port B	+5 (byte)
Data register port A	+6 (byte)
Data register port B	+7 (byte)
Pin register ports A and B	+2 (word
Direction register ports A and B	+4 (word)
Data register ports A and B	+6 (word)

Table 1. Addressing of the ports.

RF IMMUNE POWER SUPPLY

Transistorized RF power amplifiers and mobile transceivers require mains power supplies capable of supplying very high currents while remaining immune to strong RF fields. The PSU proposed in this article supplies an output voltage of 13.8 V with a capacity of up to 10 A.

MAIN SPECIFICATIONS

» Output voltage: 13.8 V (nom.)
» Output current: 10 A (max.)
» Immune to RF fields
» Short-circuit proof
» Thermal protection
» Mains filter
» Passive cooling

Design by K. Walraven

DESIGNING a power supply for a transmitter or RF power amplifier is by no means straightforward. The high frequencies produced in these circuits easily upset the operation of an ordinary power supply, with disastrous results. Stray RF signals which end up in the power supply can cause voltage fluctuations which in turn give rise to spurious transmitter output products like 'splatter' (SSB transmitters) and frequency drift (CW transmitters). Such spurious products are, obviously, undesirable, and may cause severe interference in other equipment. The matter of RFi (radio frequency interference) is now a problem area where radio amateurs, regulating authorities and manufacturers of audio/video equipment meet. In not a few cases, a badly designed power supply, rather than the transmitter proper, proved to be the cause of serious RFi problems.

A good transmitter power supply must meet a number of strict requirements: the output voltage must be adjustable to 13.8 V; an output current of more than 10 A is desirable; a short-circuit protection must be available; and the PSU must remain immune to very high RF signal levels as they may be present in a radio amateur's shack, or near an antenna.

Design considerations

From a point of view of economy and simplicity it is useful to build the power supply on the basis of one or more integrated voltage regulators. Unfortunately, regulators which meet all of the above requirements are virtually non-existent, mainly on account of the problem of cooling. High-power regulators are usually supplied in TO-220 packages. Assuming that a reasonably sized heat-sink is used with a thermal resistance of 1 K/W (max.), such an IC may dissipate about 30 W. Consequently, the voltage drop across the device must remain below 3 V at a load current of 10 A. That more or less forces one to use a pre-regulation device to ensure a regulator input voltage which is only a couple of volts above the desired output voltage. Obviously, the voltage regulator(s) used must then be low-drop types, because a difference smaller than 3 V is far too low for most normal voltage regulators to operate properly. Unfortunately, adding a preregulator adds considerably to the cost and complexity of the circuit.

An alternative solution

The fact that the dissipation of a typical TO-220 style regulator is limited to about 30 W forces one to think of alternative ways to design the supply. A solution has been found in parallel connection of a number of regulators. In the present supply, whose circuit diagram is shown in **Fig. 1**, four type LM350T regulators are used. Provided the output current is distributed equally across the four ICs, each of these supplies a current of 2.5 A for a total output current of 10 A. Consequently, the maximum dissipation is no longer a limiting factor because the maximum drop across each regulator is 12 V at 2.5 A output current, which gives plenty of headroom for the desired regulation span.

Connecting four voltage regulators in parallel is unfortunately not without problems. The biggest problem to cope with is device tolerance caused by inaccuracies in the manufacturing process (note, however, that the tolerance is accurately defined). Because of this tolerance, the IC output voltages will never be identical, so that equal current distribution among the four ICs will never

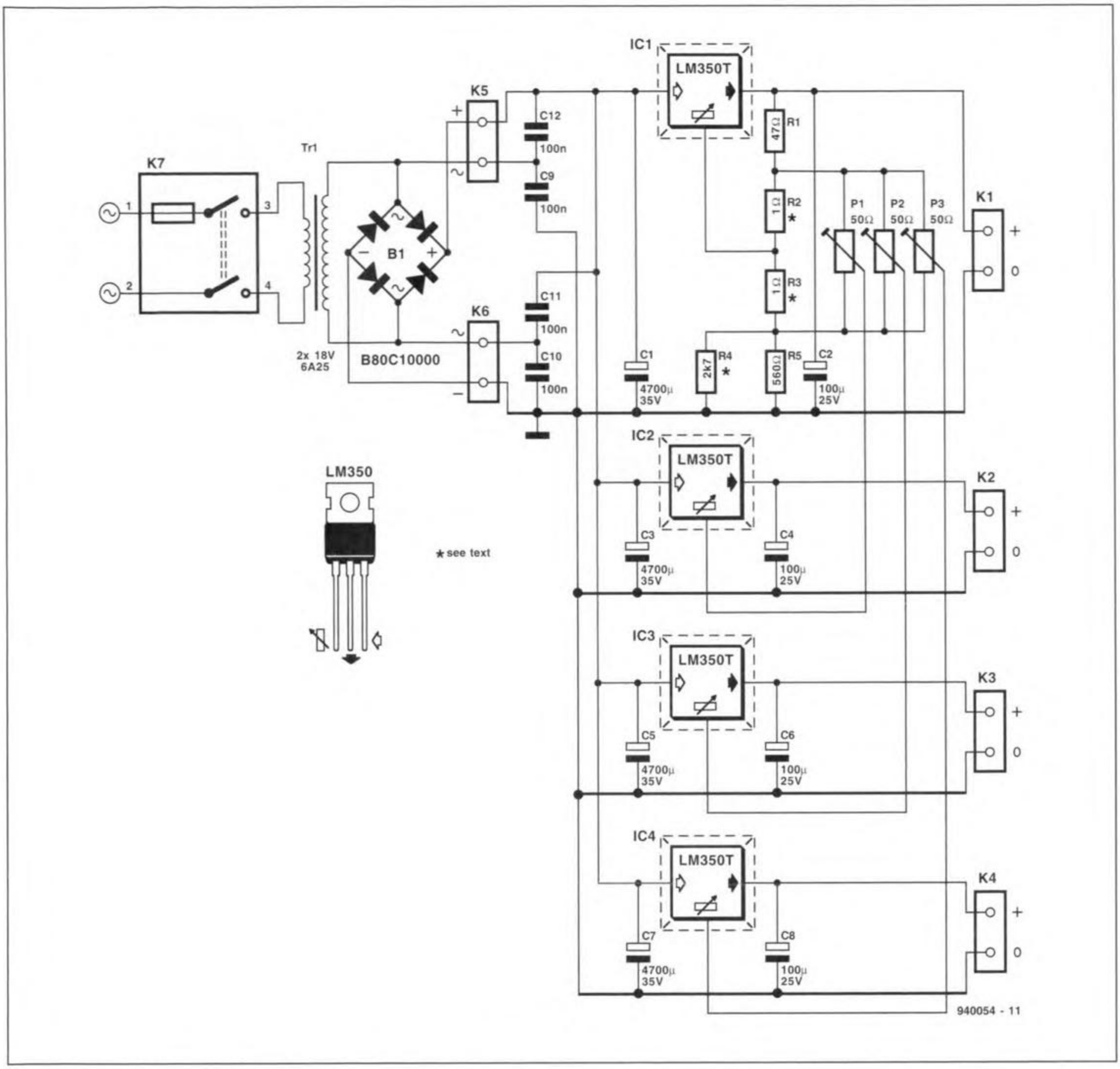


Fig. 1. Circuit diagram of the RF immune power supply.

come about just like that.

The present circuit has two provisions to combat the effects of output voltage differences between the regulator ICs. Presets P_1, P_2 and P_3 allow the output voltage of IC_2, IC_3 and IC_4 to be matched to that of IC_1. The output voltage of IC_1 is fixed at 13.8 V (theoretically) by a resistor network consisting of R_1 through R_5. The exact adjustment of the output voltage has to be done once only. The output voltage is not determined exactly by design and component values because the tolerances of the components are fairly large. For instance, the LM350 stabilizes the voltage between its output and its adjustment input to a nominal value of 1.25 V. According to the datasheets, however, this voltage can vary between 1.20 V and 1.30 V. This sort of tolerance simply means that manual calibration can not be avoided. The present supply is calibrated by careful adjustment of the value of R_4 (default: 2.7 kΩ). Assuming IC_1 has its reference voltage at the centre of its tolerance window, the default value results in an output voltage of 13.6 V. If the value of R_4 is lowered, the output voltage is lowered also. If it is increased, the output voltage increases. At a reference voltage of 1.25 V, a value of 2.94 kΩ (E96 1% series) yields the correct output voltage.

Once IC_1 has been adjusted to give the desired nominal output voltage of 13.8 V, presets P_1, P_2 and P_3 are used to match the output voltages of the other three regulators to that of IC_1. The span of the presets is normally sufficient to compensate the tolerance on the voltage regulators. Should the span be too small, the value of R_2 and R3 may be increased to 2.2 Ω.

The equal current distribution across the regulators is further aided by inserting small resistances, R_s, in the regulator output rails. Do not bother to scour for these resistors on the printed circuit board, because they are not there. Since the value of R_s must be very small, it was decided to exploit the resistance of the output cable for that purpose. Each regulator output is connected to the load via its own pair of wires. The four ground

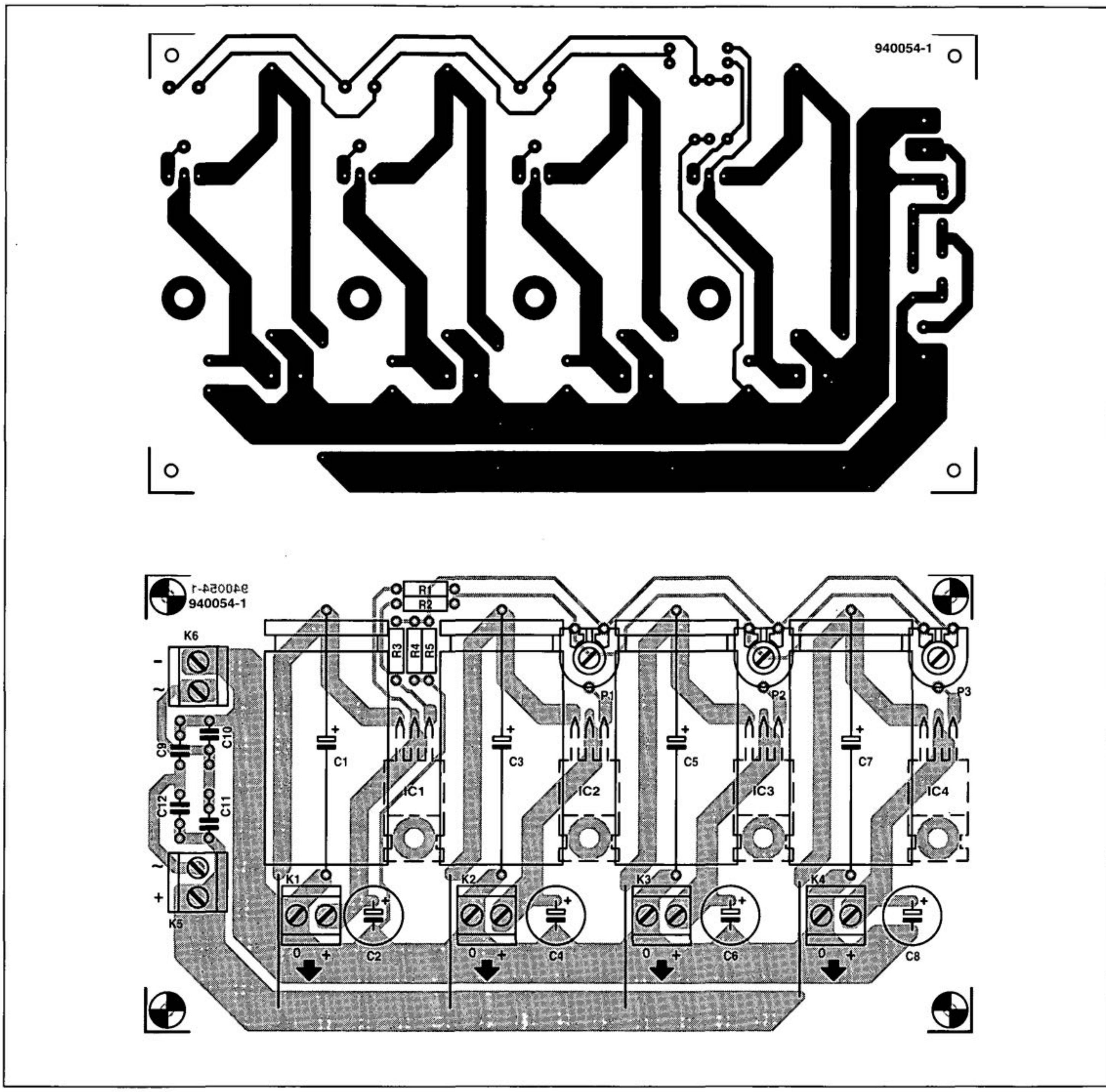


Fig. 2. Track layout (direct reading) and component mounting plan of the printed circuit board designed for the power supply.

wires must be as thick as possible. Remember, the series resistance is only required in the 'positive' wire. This should have a length of about 30 cm, and consist of ordinary flexible, insulated wire with a diameter of 0.7 mm. That produces a resistance of about 30 mΩ, which is ample for the desired equalizing effect. If a shorter connecting wire is desired, it must be thinner. Accordingly, at larger distances between the supply and the transmitter or booster, thicker wire must be used. At a current of 3 A, the wire should drop about 100 mV. Unfortunately this voltage loss can not be compensated because it is introduced 'behind' the regulation system. In practice, most transmitters have no problems with a maximum variation of 0.75% on the supply voltage, under 'full load' conditions.

For the best possible RF immunity, the output wires must be bundled into a cable. The 'positive' and 'ground' wires are best twisted.

Capacitors C_9 through C_{12} shunt the diodes in the bridge rectifier, and serve to afford extra suppression of RF signals. Finally, the usual 100-nF decoupling capacitors are not present in the supply because it is assumed that they are fitted as close as possible to the load.

Construction

The power supply is best built on the printed circuit board shown in **Fig. 2**. This board is available ready-made through our Readers Services. The passive parts, such as the electrolytic capacitors and the resistors, are fitted at the component side of the board, while the four regulators and the associated heat-sink are mounted at the solder side.

The mains transformer and the bridge rectifier are mounted as external parts. The board is secured on to the heat-sink with the aid of four 20-mm PCB mounting pillars.

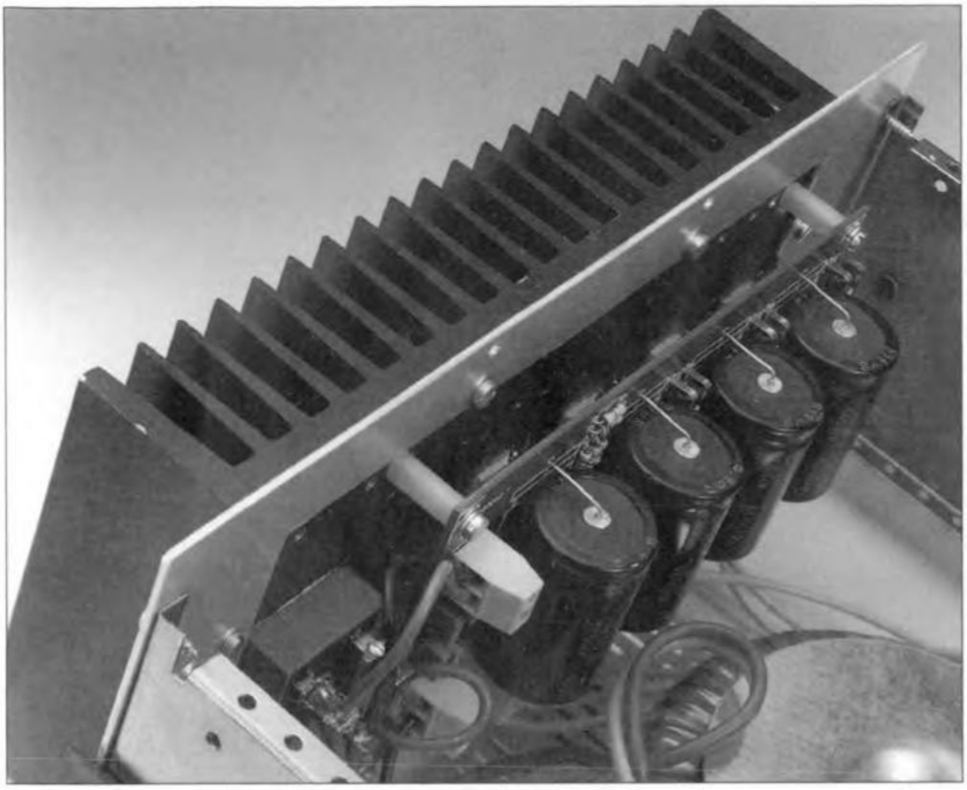


Fig. 3. Illustrating the mounting of the PCB on to the heat-sink.

COMPONENTS LIST

Resistors:
R_1 = 47Ω
R_2,R_3 = 1Ω
R_4 = 10kΩ
R_5 = 560Ω
P_1,P_2,P_3 = 50Ω preset H

Capacitors:
C_1,C_3,C_5,C_7 = 4700µF 35V
C_2,C_4,C_6,C_8 = 100µF 25 V radial
C_9-C_{12} = 100nF

Semiconductors:
B_1 = bridge rectifier B80C10000 or SB352
IC_1-IC_4 = LM350T

Miscellaneous:
K_1 = mains filter, 3 A, with socket and integral fuse holder (fuse 3.15 AT).
Fuse, 1.25 AT (slow).
K_1-K_6 = 2-way PCB terminal block, raster 5 mm.
Tr_1 = toroid mains transformer 2x18V @ 6.25A.
Metal enclosure (eg. Telet LC1050).*
Heatsink Fischer type SK49/100 (1K/W) or SK47/100 (0.5 K/W).**
Isolation material for IC_1-IC_4.
Printed circuit board 940054 (see page 110).

* C-I Electronics, fax (+31) 45 241788.
** Supplier info via Dau Components Ltd., Tel. (01243) 553031.

Fitting the passive parts on to the board is all plain sailing. Fitting the four regulators, however, requires a bit of dexterity, because they are mounted on the heat-sink, while their terminals are connected to the tracks at the copper (solder) side of the printed circuit board. The simplest solution is to insert solder pins into the respective PCB holes. These pins are fitted at the copper side of the PCB, and protrude about 10 mm towards the heat-sink. If you mount the regulators on to the heat-sink in the positions indicated by the PCB overlay, their terminals need to be bent only to enable them to be soldered to the solder pins.

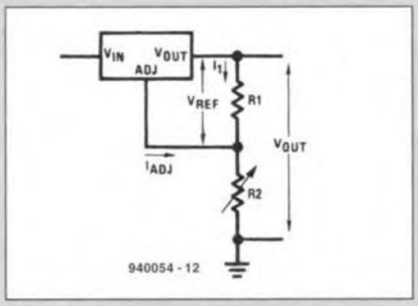
The choice of a suitable heat-sink should be given ample consideration. If the supply is expected to deliver 10 A quite often, a heat-sink with a thermal resistance of 0.5 K/W must be used. A good choice is a length of Fischer SK47 heat-sink with a height of 10 cm. If about half the maximum current is supplied, a heat-sink with a thermal resistance of about 1 K/W will be adequate. In such cases, turn to a length of type SK49 heat-sink with a height of 10 cm. This will ensure a thermal resistance of about 0.75 K/W.

The holes to secure the regulators on to the heat-sink are drilled with a 2.5-mm drill. Next, the holes are threaded with a 3-mm (6BA) tap.

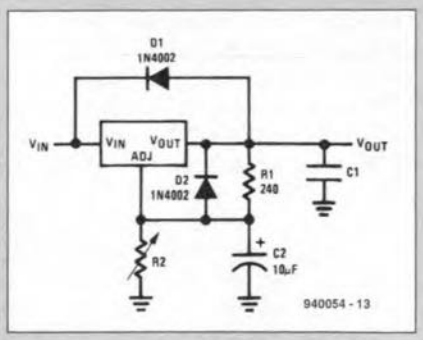
The ICs should be mounted on to the heat-sink with the aid of mica or ceramic washers, and associated insulating bushes. Also be sure to use a

LM350

The LM350 is a standard voltage regulator manufactured by Texas Instruments. The IC is designed for use in power supplies with an output current of up to 3 A. The LM350 has internal over-current and over-temperature shut-down circuits.
The reference voltage between the output and the adjust input of the regulator is 1.25 V at a minimum load current of 3.5 mA. The minimum voltage difference between the input and the output is 3 V. The maximum voltage drop is 35 V.
The output voltage of the device is defined by incorporating the adjust input into a network consisting of two resistors between the output and ground. The basic circuit is given below.



The output voltage, U_o, is calculated from
$U_o = 1.25\ (1+R_1/R_2)$ [volts]
The formula disregards the adjust current, I_{adj}, because that is negligible at 50 µA.
According to the datasheets, the LM350 must supply a minimum output current of 3.5 to 5 mA. At higher currents, the IC is always in its normal operating range. This condition is simple to fulfill by giving R_1 a minimum value of, for instance, 240 Ω. That results in an output current of 5 mA even if the supply is not connected to the actual load. Although not strictly necessary, an additional protection against discharge currents can be provided by fitting two diodes as shown below.



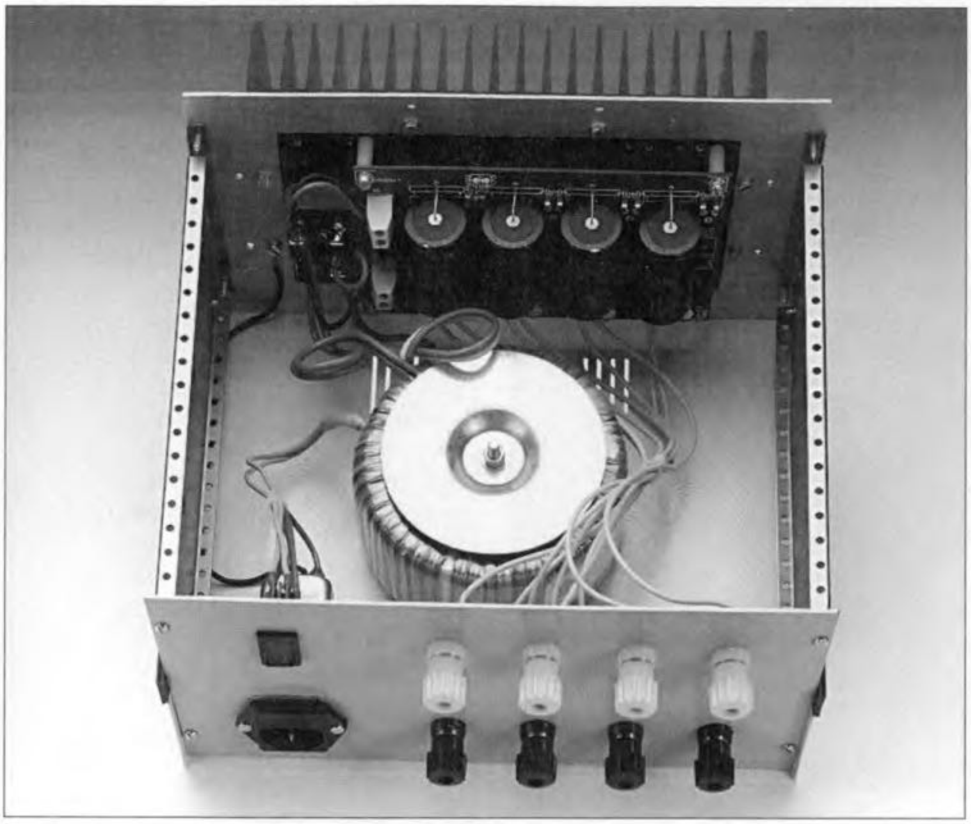


Fig. 4. Completed prototype of the power supply. Note the use of four pairs of connecting posts. These are joined at the load, not at the supply!

liberal amount of heat-conducting paste. It is important to first secure the regulators flat on to the heat-sink, and then solder them to the pins inserted into the board, not the other way around.

The bridge rectifier has a tough job at the maximum load, and needs to be mounted on to the heat-sink as well. Here, too, a good thermal contact must be ensured to prevent the device from overheating and breaking down. Do not skimp on heat-conducting paste, it is cheaper than almost any replacement semiconductor!

The transformer's secondary is connected to the a.c. terminals of the bridge rectifier via short pieces of heavy-duty wire. Next, run wires from the a.c. terminals on the bridge to the corresponding connections on terminal blocks K_5 and K_6 on the board. Next, connect the + and – terminals of the bridge to the respective screw terminals in K_5 and K_6.

At the mains side of the transformer, it is best to use a mains switch with a built-in filter. This affords good protection against interference on the mains. Likewise, the filter also prevents mains pollution by noise generated in and around the power supply. Finally, for the best possible screening the entire supply must be built into an earthed, metal enclosure.

Practical notes

A prototype of the power supply was tested for RF immunity by monitoring its output voltage in the very close presence of a transmitting 2-metre band handheld with an RF power output of 1.5 W. Nothing happened. Next, the power level was stepped up to about 10 W from a 2-metre FM mobile transceiver. During this test the antenna, a magnetic 5/8-λ car-roof type, was at a distance of less than 1 metre from the power supply. Although the digital voltmeter and ammeter connected to the supply went 'haywire', the power supply proper remained totally immune to the strong RF field.

If you happen to use a power amplifier which is known to radiate, that problem must be solved first, because the radiation is then bound to be on the supply wires also. Start by fitting feed-through capacitors to carry the supply voltage into the power amplifier — this is much more effective than fitting decoupling capacitors in the power supply. In addition to this measure, the supply cable may be wound two or three times on a ferrite ring core, close to the PA case.

Unless you are using a handheld, it is bad operating practice to have an antenna in the immediate vicinity of a transmitter, a power supply and, most importantly, yourself. At present there is a hot debate between several parties regarding the alleged dangers of high RF power levels near the human body. Since at present it is not at all clear what the effects of this type of radiation are, it is best to play it safe, and locate the antenna at a safe distance. More information on this matters may be found in recent issues of several ham radio magazines.

Adjustment

Change R_4 by trial and error until the output voltage of IC_1 at terminal K_1 is 13.8 V (or any other value you may require). Next, adjust the output voltages of IC_2, IC_3 and IC_4 to give the same value. This is done by carefully turning the respective presets, P_1, P_2 and P_3. Connect a wire pair to each output, and join all pairs at the load. A 'dummy' load which draws about 5 A should be used, although this is not critical. Now check if all regulators supply roughly the same amount of current by measuring the voltage drop across their positive output wires. The drop should be of the order of 100 mV at an output current of 10 A, and 50 mV at a current of 5 A. On the prototype, the voltage was between 70 mV and 80 mV. The voltages may be made equal by careful adjustment of the presets.

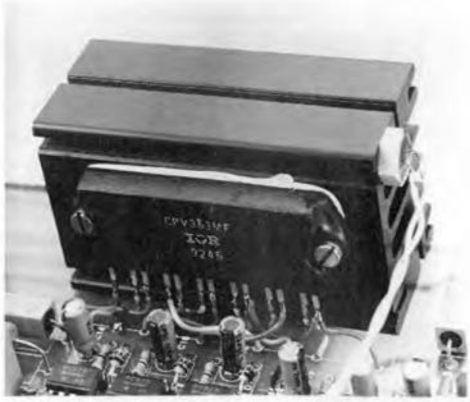
(940054)

For further reading:

EMC testing of PMR equipment, *Elektor Electronics* July/August 1993.

Electromagnetic compatibility, *Elektor Electronics* May and June 1993.

1-TO-3-PHASE CONVERTER



Design by B. Yahya

The unit described in this article converts the single-phase mains supply to a 3-phase voltage, which can be used to control small three-phase motors rated at up to 725 W, irrespective of whether these are synchronous or induction types.

The rotary speed of a three-phase motor depends on the frequency of the applied voltage. A very good way of controlling the motor speed, therefore, is by pulse-width modulation of the applied voltage. **Figure 1** shows the block schematic of the present converter which uses this type of modulation. The three-phase motor is connected at the centre of the star circuit. The signals ensuing or used during the operation of such a motor are shown in **Fig. 2**.

The rectifier converts the 240 V mains supply into a direct voltage of about 340 V. The motor is linked to the converter by six power transistors that are connected to the outputs of the digital section of the converter. These transistors and the logic circuits controlling them ensure that the currents flowing through the windings of the motor are sinusoidal, or nearly so, and that the three voltages have the correct phase relationship.

Figure 2b, c and d show that a sinusoidal waveform is simulated by varying the pulse/pause ratio of a rectangular voltage with a relatively (w.r.t. 50 Hz) high frequency. The inductance of the motor windings acts as an integrator that converts the pulses of varying widths into a sinusoidal signal. The converter ensures that the phase shift between the three generated voltages is 120°.

Circuit description

The circuit of the converter in **Fig. 3** is based on a Type 80C535 microprocessor (IC_4). This device provides the timing and control of the three power transistors. Because of the digital design, the system is stable and its properties are constant over a wide range of control.

Since IC_4 has pulse-width modulation (PWM) outputs, ensuring a 120° phase shift is merely a matter of software loops.

The main program ensures that the correct registers in the processor are regularly loaded, taking into account the required voltage level and frequency of the output signals.

Reversing the rotational direction is also effected by software, so that there is no need of switches and relays. This way of operation ensures a longer useful life of circuit and motor, particularly if its rotational direction is changed regularly.

The analogue inputs of the processor are used to enable the user setting certain parameter, such as the rotational speed.

The power transistors are contained in a Type MP6750 module (Toshiba). The CPV363MF from International Rectifier may also be used, but this requires some modifications. The transistors used in the module are Isolated Gate Bipolar Transistors (IGBT), which have a collector and an emitter, but are controlled by a positive voltage between gate and emitter. In other words, it can be driven by a voltage, not a current. This ensures fast, low-loss control. The module also contains the associated freewheeling diodes that are necessary in a three-phase system.

Although the module is not cheap, it does not cost much more than the sum of the discrete components required if it were not used. Also, it obviates the complexity of a discrete design.

The generation of a sinusoidal voltage necessitates the division of the waveform into a large number of steps. Each step is the result of a clock pulse and has a value that corresponds with the level of the sinusoidal voltage at the location of that step. The relevant voltage level can be generated with a certain pulse/pause ratio. For a good sinusoidal waveform to ensue, the PWM signal frequency needs to be many times higher than that of the sine wave.

Assuming a fixed clock frequency for the PWM signal, the frequency of the generated voltage rises in direct proportion to the number of steps a period of the sinusoidal signal is divided into. Stringing a number of identical steps, Z, together can effect an apparent lowering of the clock frequency (which does not change, however). This can be formulated as follows:

$$f_0 = f_c / 6SZ,$$

where f_0 = output frequency
f_c = clock frequency
S = number of steps in a period
Z = number of successive, identical steps
6 = divisor necessary for the

Number of phases	3
Rating	725 W
Frequency	0–50 Hz
Input voltage	240 V
Output voltage	0–220 V (between 2 phases)
Rotational direction	reversible
Starting time	presettable
Slow down time	presettable
Emergency stop	provided
Temperature protection	provided
Overload detection	provided

Some properties of the converter.

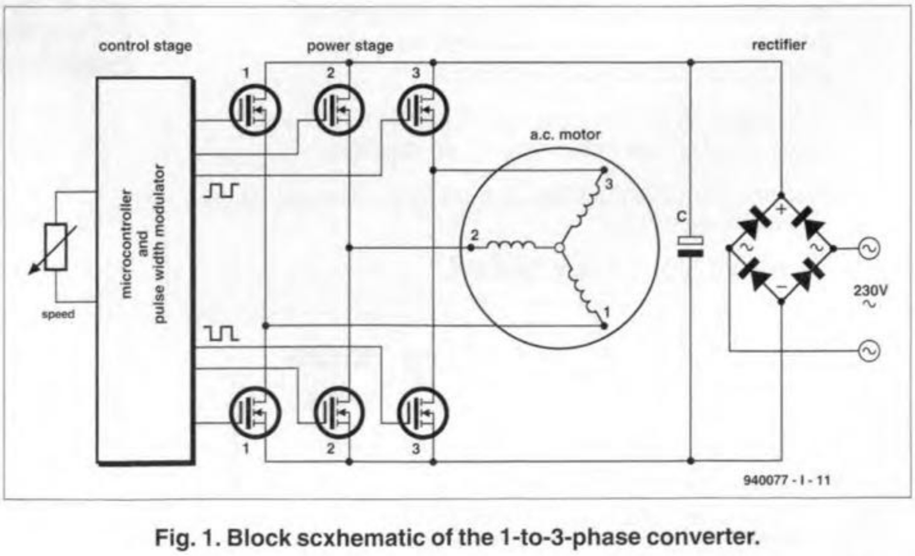


Fig. 1. Block scxhematic of the 1-to-3-phase converter.

generation of a three-phase signal.

The formula shows that generating a continuous variation of the signal frequency is possible only if S and Z are variable, and this is ensured by the microcontroller. Unfortunately, the artihmetical power of the controller is not adequate for the task, which makes a table containing the parameters necessary for generating a given output current essential.

The output current is strongly dependent on the output voltage and the frequency thereof. The many kilobytes of data contained in the table were generated on the hand of a given characteristic; they can not be changed by the user. On every command of a timer-2-interrupt, the CPU takes the next PWM value from the table and places this in the CCx register of the processor.

Measures have been taken to ensure that the couple of the motor remains constant. This is effected by a lowering of the average output voltage when the frequency is lowered. This ensures that the current through the motor is (nearly) constant. If the voltage were not lowered, the motor current would become so large at low frequencies that the motor would burn out.

The microprocessor uses a 12 MHz crystal to generate the 1 MHz clock frequency. When its supply voltage drops below a predetermined value, IC_3 generates a signal to reset the processor. This prevents the processor running amok when there are spurious signals on the supply lines. If it were allowed to do so, the damage to the output transistors would be severe. The reset pulse is also used to disable the outputs of IC_5, which generates the six control signals for the power transistors. This in turn causes all the output transistors to be switched off. The input to IC_5 consists of the three PWM signals generated by the processor.

In the switching of the output transistors, a dead time of 1 μs is used. This is adequate to allow two series-connected transistors to be switched off briefly and so avoid a short circuit resulting from the two transistor conducting simultaneously.

Control of the gates of the power transistors is effected by optoisolators IC_6–IC_{11}, which are specially designed for this purpose. The output stage in the optoisolators consists of two series-connected transistors that operate as a change-over switch. They are capable of rapidly removing the charge from the gate capacitance of the IGBTs, which ensures very fast switching of the transistors.

Since IGBTs are switched on by a positive voltage at its gate-emitter junction, the upper IGBT of each series circuit needs a bootstrap circuit. The operation of this circuit is shown in simplified form in **Fig. 4**. Output transistor T_1 in **4a** is off, while T_2 conducts because of the positive voltage (about 10 V) at its gate.

Since T_2 conducts, junction T_1-T_2-C_1 is at earth potential. Capacitor C_1 is then charged to 12 V via D_1. The electronic switch between the +ve terminal of C_1 and the gate of T_1 is formed by the output stage of the optoisolator. As soon as T_2 is switched off (**Fig. 4b**), the switch between the gate of T_1 and the +ve terminal of C_1 is closed, which causes a positive voltage on the gate, so that T_1 begins to conduct. Because of the switching off of T_2, the potential at junction T_1-T_2-C_1 rises to about 325 V. Since the –ve terminal of C_1 is connected to this junction, the potential at the gate of T_1 is always 12 V higher than the voltage at its emitter. This potential is used as power supply for the output transistor. As soon as the optoisolator is on, the gate voltage of T_1 rises to about 337 V. In this situation, D_1 is reverse biased. This diode must be able to withstand the high reverse bias voltage and to cut off rapidly.

The 33 Ω resistors in series with the gates of the IGBTs ensure that the gate capacitance of the IGBTs is charged and discharged rather more slowly. This causes the IGBT

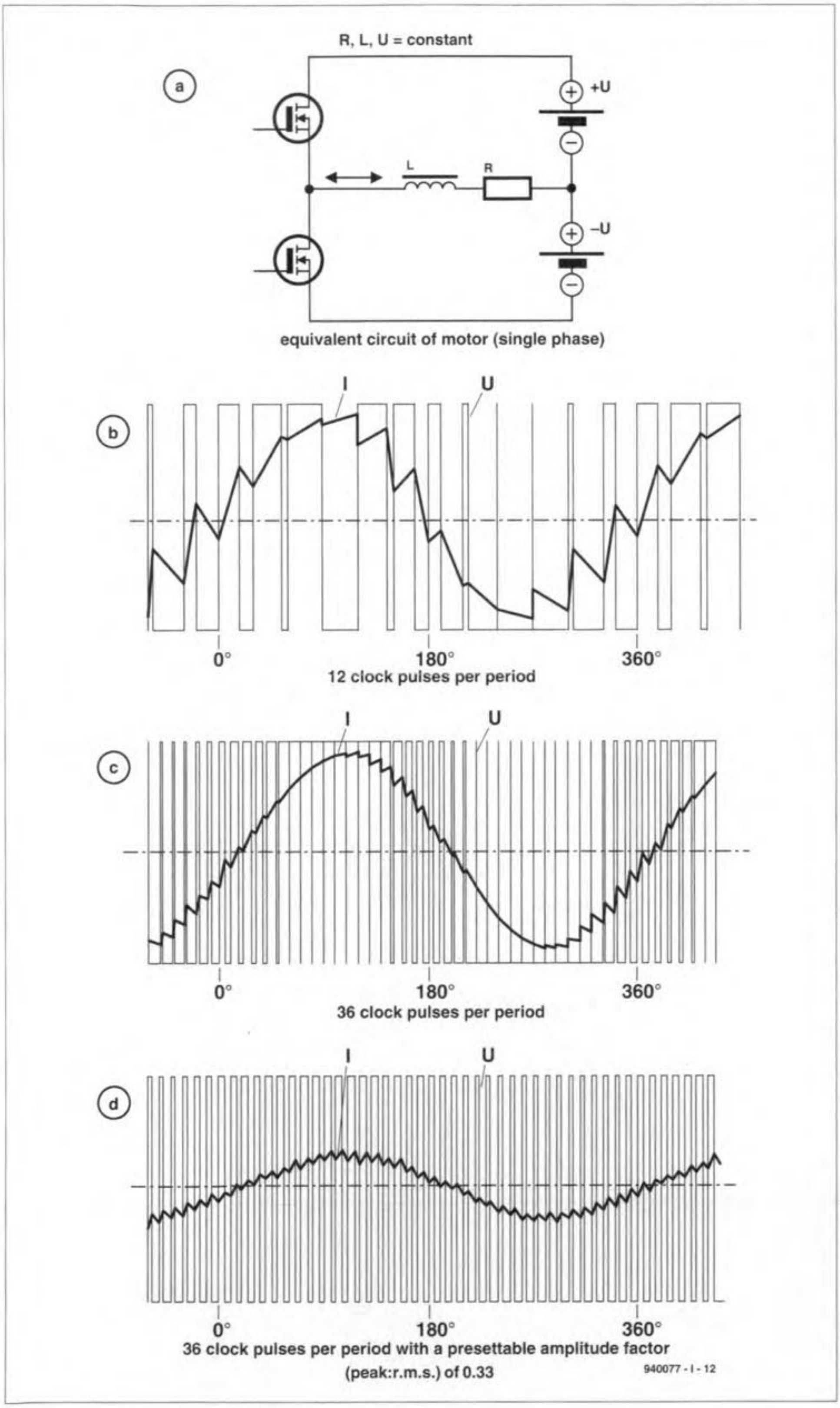


Fig. 2. Illustrating how PWM signals generate sinusoidal signals.

to switch more slowly so that the over-current protection circuit can come into operation if necessary.

Resistors R_{33} and R_{35} in series with the module are used to monitor the current through it. If the current exceeds a given value, say, 20 A, caused, for example by the interconnection of two phases or one of the phases being connected to ground, IC_5 gets an error message via optoisolators IC_{14} and/or IC_{15}. The output transistors are then immediately switched off.

The correct operation of the protection circuit depends on the rise time of the current remaining within normal limits. Short circuits arising on the printed circuit board or immediately following the input terminals can cause a lot of damage because there is no self inductance inherent in cables. This is the reason that the switching times of the IGBTs are limited by 33 Ω resistors.

The temperature of the module is monitored by R_{38}, which has a negative temperature coefficient (NTC). When the module gets too hot, a signal is passed to the microprocessor by IC_{13}.

Because of the important function optoisolators IC_{13}, IC_{14} and IC_{15} fulfil in protecting the circuit, it is imperative that the specified types are used. Alternative types may not be sensitive enough to guarantee effective protection.

Circuit IC_1 demultiplexes the combined address/data bus, while IC_2 stores the program of the converter.

Since the user has a number of controls at his disposal for setting the voltage and speed of the motor, the processor has several terminals for connecting preset potentiometers. Since small spurious signals may have serious consequences, these inputs are provided with the necessary protection.

All inputs that are accessible from outside via connectors K_1–K_4 make use of the unregulated +15 V supply lines. This minimizes the risk of the processor supply

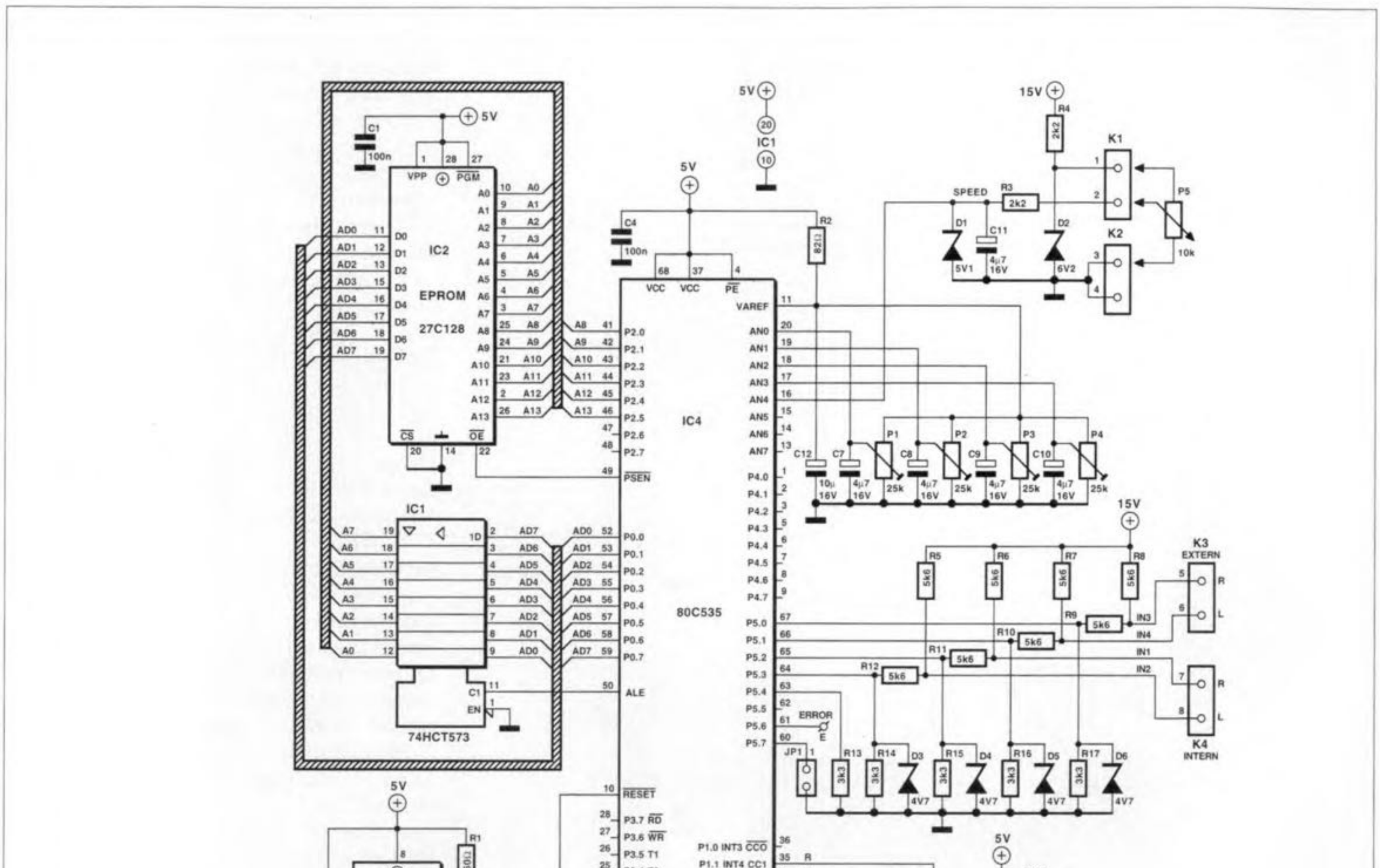
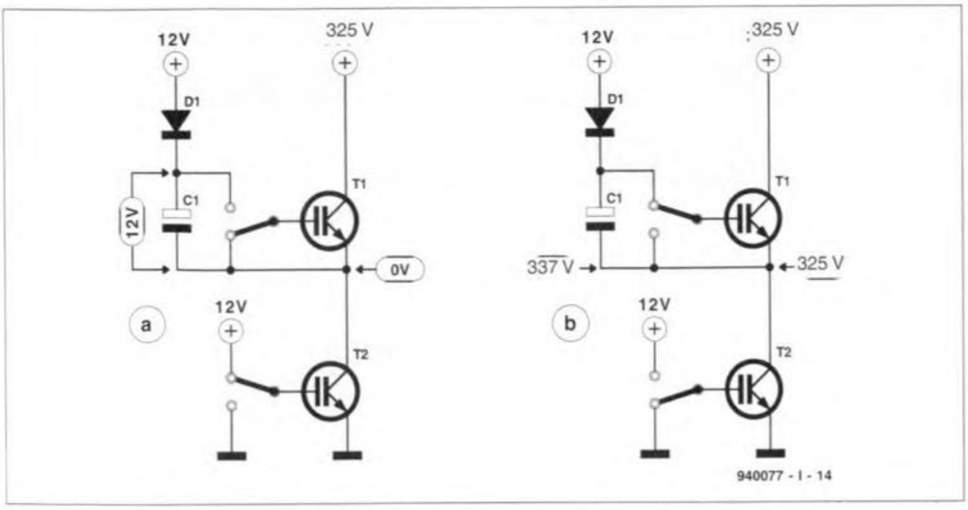


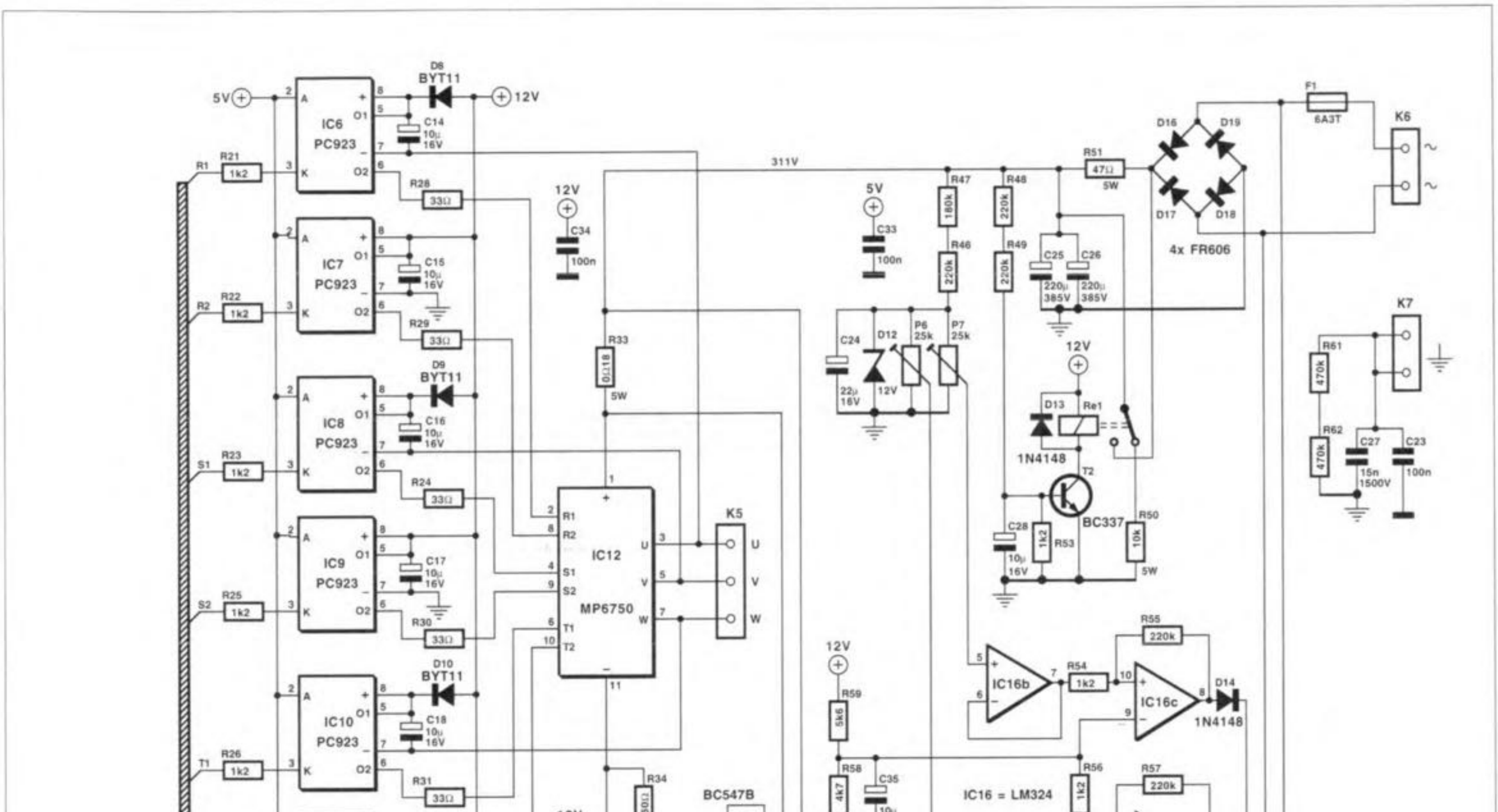
Fig. 3. Circuit diagram



being contaminated by operations at these inputs. The inputs are connected to the supply lines via a resistor, which makes them short-circuit proof. Zener diodes D_1–D_6 limit the voltage to the required level: 5.1 V for the analogue input on K_1, and 4.7 V for the digital inputs.

Presets P_1–P_4 and potentiometer P_5 are connected to one of the many analogue inputs of the processor. The motor speed is set with P_5; any changes are made with the presets. More will be said about this later.

Fig. 4. This bootstrap circuit may be used to generate the correct control voltage for the IGBT.



of the 1-to-3-phase converter.

The inputs on K_3 and K_4 are switching inputs: when input 5 on K_3 is linked to ground, the motor turns clockwise and the speed can be changed with P_1. When input 6 on K_3 is linked to ground, the motor runs anticlockwise, and its speed can be changed with P_2. When input 7 on K_4 is at ground potential, the motor runs clockwise and the speed may be varied with P_5. When input 8 on K_4 is linked to ground, the motor runs anticlockwise and its speed may be varied with P_5.

Preset P_3 determines the rate at which the motor accelerates to full speed and P_4 the rate at which the motor comes to a standstill. If, however, JP_1 is closed, P_4 has no effect and the motor is deenergized instantly.

An error is indicated by the lighting of D_7. Resetting is then possible only by switching off the converter and eliminating the cause of the error. Resetting occurs by not using, that is, by not connecting to ground, inputs 5–8. About 6 seconds after the inputs have become inactive, the converter is reset and the LED goes out.

Power supply

The mains voltage is rectified by bridge D_{16}–D_{19} to a potential of some 325 V. As long as the voltage across C_{25} and C_{26} remains below a given level, relay Re_1 is not energized. Resistor R_{51} is then in series with C_{25} and C_{26}, which keeps the starting current within limits. As soon as C_{28} has been charged (or very nearly so) via R_{48} and R_{49}, the relay is energized via T_2 and R_{51} is short-circuited.

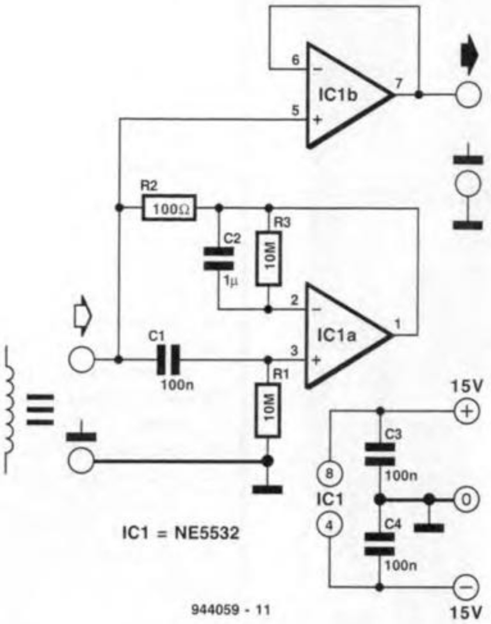
The level of the rectified mains voltage is monitored by discrete Schmitt triggers IC_{16a}–IC_{16d}. As soon as it rises above, or drops below, the predetermined levels set by P_6 and P_7 respectively, the output stages of the converted are switched off via IC_5.

When the converter is switched off, relay Re_1 is denergized instantly since the 12 V disappears suddenly. Resistor R_{50} is then in parallel with C_{25} and C_{26}, so that the residual charge on these capacitors is removed quickly. In a similar manner, C_{23} and C_{27} are discharged quickly via R_{60} and R_{61}, when the mains is switched off.

[940077-I]

GYRATOR

The gyrator, or electronic inductor, is based on a Type NE5532 operational amplifier and can be used in filters and other applications at frequencies of up to 1 kHz. The circuit is particularly useful for applications where large values of self inductance are required. It is not good design practice to use a gyrator of only a few mH, but winding an inductor of, say, 100 mH is a sufficiently cumbersome task to use a gyrator instead. The present circuit can simulate inductances of 1 mH to well over 100 H.



In the diagram, the actual gyrator is IC_{1a}, while IC_{1b} is a buffer that may prove useful for test purposes or for some other applications.

The series resistance of the gyrator is formed by R_2 and the parallel resistance by R_1. The self inductance, L is:

$$L = R_1 \cdot R_2 \cdot C_1.$$

The best inductance is obtained when R_1 is as large as feasible and R_2 as small as possible. There are, of course, practical limits to this. For instance, with most opamps, R_2 must not be smaller than 100 Ω (but, when it is greater than 10 MΩ, the noise level is exorbitant). Large values of R_1 cause the offset at the output to increase appreciably, although this effect can be partly negated by giving R_3 the same value as R_1.

Experiments showed that the opamp used in the prototype yielded better results than a number of other types. For the tests, resistors between 10 kΩ and 100 kΩ were connected in series with the input whereupon the frequency characteristics were measured at the output of IC_{1b}. Capacitors between 1 nF and 100 nF were used in the C_1 position. These tests showed that the gyrator gives excellent performance from very low frequencies up to about 1 kHz.

The current drain from a ±15 V supply was ≤8 mA.

Design: T. Giesberts
[944059]

CHAOS

intern

The present circuit offers, via an oscilloscope, a visualisation of the chaos theory. Roughly, this theory states that all elements around us are by nature disordered, stubbornly refuse to accept externally dictated order, and behave unpre-dictably. An example of this is the weather. Although its mechanics are well known and meteorologists have large quantities of data at their disposal, it has so far proved impossible to produce accurate, long-term weather forecasts.

The present circuit may be seen as 'chaos electronics', in that it has various states, each of which in itself is stable and predictable, but it can not be said with certainty which of these states it will assume. This produces the most interesting images on the screen of an oscilloscope in the X-Y mode.

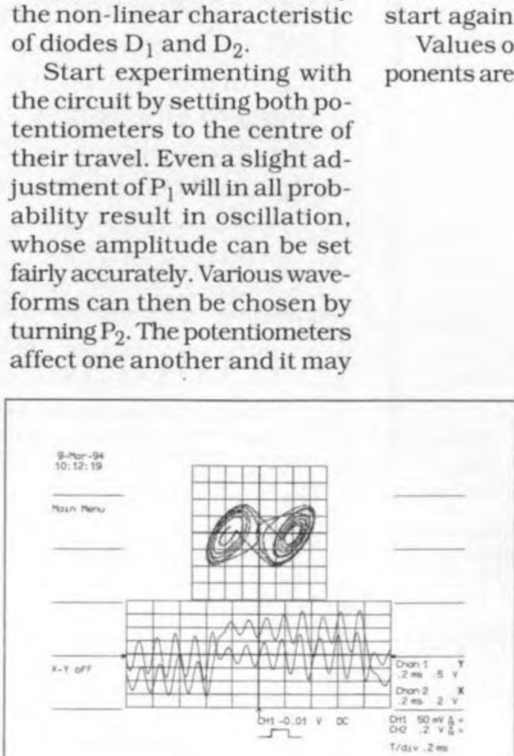
In the diagram, IC_1 forms a negative impedance which interacts via P_2 with a positive impedance formed by network L_1-C_4. The consequent oscillation, although inherently stable, is rendered 'chaotic' by the non-linear characteristic of diodes D_1 and D_2.

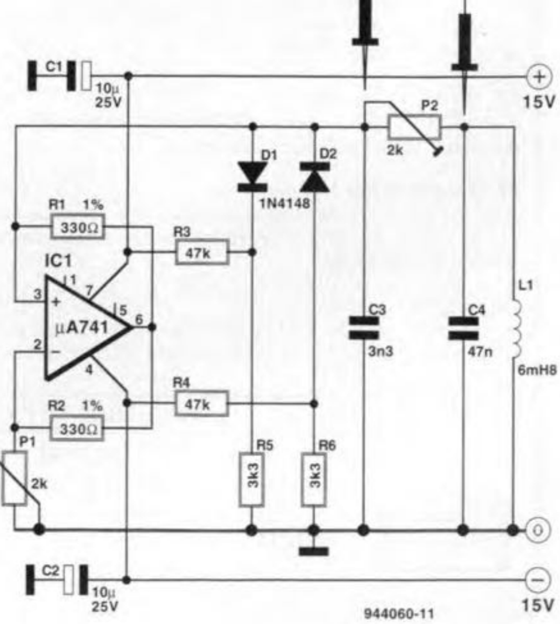
Start experimenting with the circuit by setting both potentiometers to the centre of their travel. Even a slight adjustment of P_1 will in all probability result in oscillation, whose amplitude can be set fairly accurately. Various waveforms can then be chosen by turning P_2. The potentiometers affect one another and it may well be that at certain settings oscillations cease. Merely turning either or both back a little will make the oscillations start again.

Values of the various components are not critical: changing that of L_1 merely alters the frequency.

The current drain from a ±15 V supply is only a few mA.

Design: Leon O. Chua
(University of California)
[944060]





STEP-UP NiCd BATTERY CHARGER

A solar panel rated at 6 V, 50 mA, provides enough energy to charge four small NiCd batteries. The problem is, however, that the voltage rating of most affordable solar panels is appreciably lower than 6 V. Semiconductor producer Maxim has available a series of integrated circuits that are suitable for converting direct voltages. One of these, the Type MAX631 (**Fig. 2**), is used in the present circuit (**Fig. 1**) to charge four NiCd cells rated at 500 mAh from one 3 V solar panel. The only drawback is that the panel's efficiency drops by about 20%.

In the diagram, the IC is arranged as a switching step-up converter which, basically, transforms a 1.5–5.6 V input into a steady 5 V output. Since that is not enough for charging four NiCd cells, the internal voltage divider is modified by R_1, which results in an output voltage of 7.5 V. This voltage is buffered by C_1. With this modification, and an input voltage of 3 V, the IC can sink about 30 mA. With higher input voltages, the current may be greater than 50 mA.

NiCd cells should not be charged with currents over 50 mA for other than short periods of time. If, therefore, the input voltage from the solar panel is expected to be higher than 4 V for long periods of time, the network shown between the two dashed lines (T_1 and R_2–R_4), must be added. This protects the batteries and also keeps the dissipation of the IC within specificied limits. When the output current exceeds 50 mA, the potential drop across R_3 and R_4 will cause T_1 to conduct. Since the transistor is in parallel with the potential divider, the output voltage is reduced to a level at which the output current does not exceed 50 mA.

The integral current limiting stage lowers the efficiency of the circuit to some degree. It should also be borne in mind that the voltage drop across R_3 and R_4 is about 0.5 V.

Inductor L_1 can be any standard type of coil that can handle the current flowing into the IC.

Finally, it should be noted that pin-compatible ICs, such as the MAX632 (12 V) and the MAX633 (15 V) may also be used.

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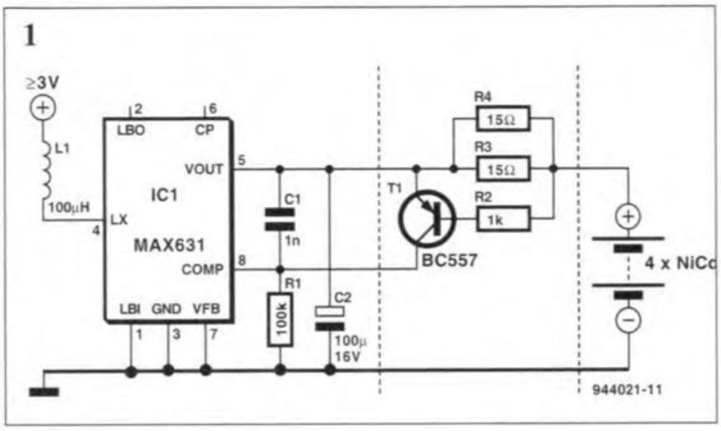

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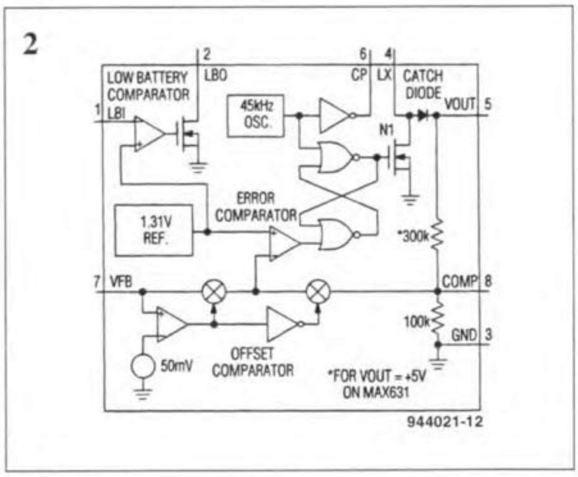


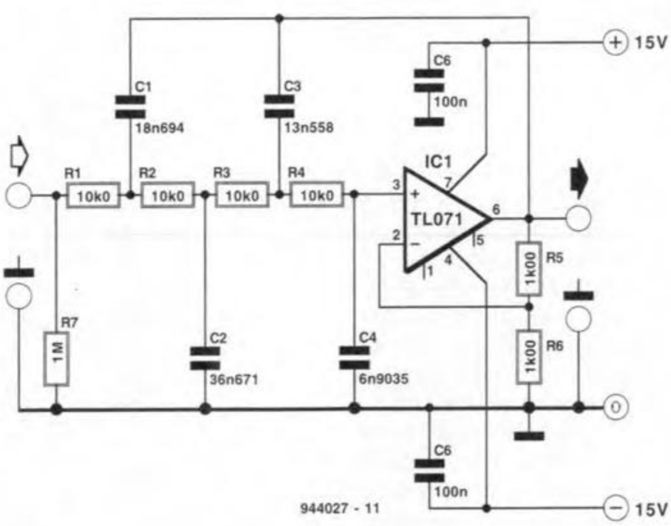
Design: K.M. Walraven
[944021]

4TH ORDER BUTTERWORTH FILTER

The present filter has an amplification of ×2, so that R_5 and R_6 in the feedback loop can be equivalent. In other words, their ratio is not an awkward one as is so often the case.

The values specified for C_1–C_4 give a cut-off frequency of 1 kHz. Other frequencies are possible by appropriate changes in the values of the resistors or capacitors. It should be borne in mind that the filter forms a load for the operational amplifier, particularly at high frequencies. With values as shown, this load is 10 kΩ:4 in parallel with R_5+R_6.

Replacing the calculated values of C_1–C_4 by the nearest E12 values (18 nF, 33 nF, 12 nF and 6.8 nF) does not affect the frequency characteristic all that much. The prototype showed a 0.4 dB peak around 435 Hz, while the cut-off frequency dropped to around 900 Hz. The whole characteristic shifts into the direction of a Chebyshev filter, which gives a slightly steeper slope at the cut-off point.



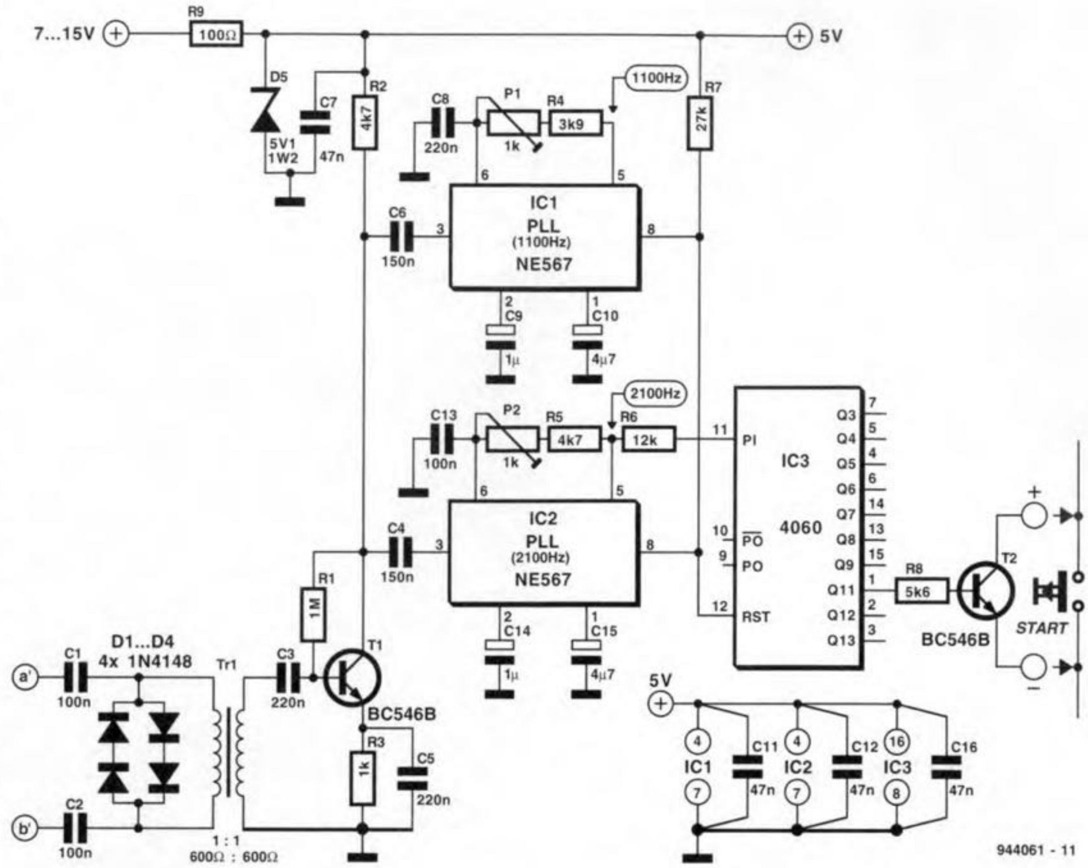
The choice of opamp is not critical, since most of the noise in the filter is caused by resistors R_1–R_4. At low frequencies, it amounts to about 26 nV Hz^{-1}.

Resistor R_7 ensures continuity of bias current to the opamp when the input of the filter is open circuit.

The TL071 draws a current of about 2 mA.

Design: T. Giesberts
[944027]

AUTO START FOR OLDER FAX MACHINES



When an older fax machine and a telephone, which share a line, are located in different rooms, an annoying situation may arise. When the telephone rings and the receiver is taken from the hook, it is found that it is a fax message. By the time the fax machine has been reached and the start knob on it is pressed, the caller has given up.

The present circuit, whose action is similar to that in modern fax machines, obviates this irritating deficiency. It amplifies any signal on the line, which is then fed to two different, parallel-connected phase locked loops (PLLs), IC_1 and IC_2. One of the PLLs locks on one of the two possible fax recognition frequencies (1100 Hz or 2100 Hz), and the second PLL to the other frequency. Thus, when the calling fax station puts its calling signal on to the line, one of the PLL outputs becomes low. Since the outputs are open-collector ones, they can simply be interconnected. If neither of the outputs is enabled, counter IC_3 is reset.

Counter IC_3 obtains its clock from the 2100 Hz PLL. After a delay of about half a second, a pulse appears at the output (pin 1) of the counter. This delay is necessary to prevent the fax machine being actuated by a spurious signal.

The output pulse of IC_3 (width about 0.5 s) switches on T_2 which thereupon short-circuits the start knob of the fax machine.

The present circuit is electrically isolated from the telephone line by a 600 Ω line transformer (1:1). The alternating input voltage is limited to a safe value by C_1, C_2 and diodes D_1–D_4.

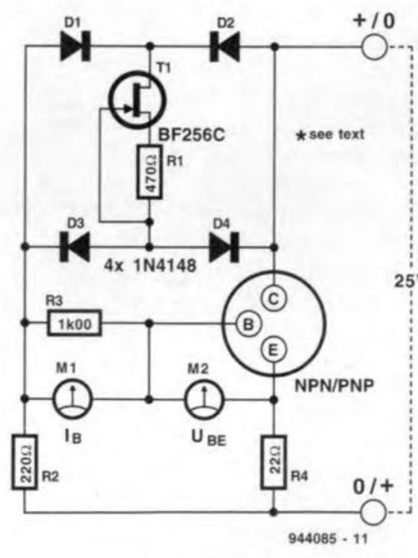
Thecircuit may be powered by any mains adaptor that provides a voltage of 5–15 V and a current of 20–30 mA. If it is built into the fax machine, there is almost certainly a suitable supply available there. A non-smoothed direct voltage may be regulated with a resistor, R_9, and a zener diode, D_5.

The PLLs must be calibrated carefully with the aid of a precision generator and a frequency meter. The 1100 Hz and 2100 Hz signals are available at pin 5 of IC_1 and IC_2 respectively. The capture range is 180 Hz for IC_1 and 270 Hz for IC_2.

The connections from T_2 to the fax machine will depend on the type of machine.

Design: R. Ratke
[944061]

N-P-N / P-N-P SELECTOR



The push-pull driver transistors in output amplifiers must be as closely complementary as possible. The selector matches pairs of complementary transistors by their base/emitter current. The design allows the polarity of the supply voltage to be reversed from n-p-n to p-n-p.

Assuming that an n-p-n transistor, T_x, is connected to the test socket, a current flows through current source T_1-R_1 and R_2. Since T_1 has been set to 5 mA, the voltage drop across R_2 is about 1.1 V. This potential is applied to the base of T_x, which then conducts, resulting in a voltage drop across R_4 of about 0.4 V. This means that the collector current of T_x is about 20 mA. The exact level of the base-emitter voltage of T_x can be read on M_2. When T_x is a p-n-p type, the current flows through R_2 and diodes D_1, D_4 via the same current source. This arrangement ensures that the level of the voltage across R_2 is the same, irrespective of whether T_x is an n-p-n or a p-n-p type, as long as these are truly complementary.

Resistor R_3 and M_1 enable the exact base current of T_x to be measured. This facility makes it possible for the h_{FE} of the transistor pair to be matched. These components can thus be omitted if this facility is not needed.

For best results, M_2 should be read immediately on switching on the supply voltage. That is, T_x must not be allowed to warm up, because every °C change in junction temperature causes a measurement error of 2 mV.

When first a number of n-p-n transistors are tested and then a number of p-n-p types, and their U_{BE} noted in all cases, several good complementary pairs will soon be found. This is provided the transistors are all from the same maker. In practice, similar types of transistor from different manufacturers normally show quite different levels of U_{BE}.

The supply voltage should not exceed 30 V. With values of components as specified, the current drain is about 25 mA.

It is advisable to check the voltage across R_1 before taking the selector into use: this should be about 2.35 V. Owing to the relatively large spread inherent in JFETs, it may be necessary to change the value of R_1 to some degree to ensure that the current through T_1 is about 5 mA.

For the test socket, half a six-pin IC socket or a single-row socket, or simply suitable terminated test leads may be used.

Design: T. Giesberts
[944085]

30-M QRP CW TRANSMITTER

THE joy of every QRP enthusiast is to be able to cover very long distances 'on a shoestring', that is, with a low-power (<10 W) transmitter running from a battery, and operating in one of the HF amateur radio bands. A QRP transmitter, which greatly adds to the excitement inherent in the radio hobby, is easy to build. The one described here is designed for CW (morse) use in the 30-m (10-MHz) band, and forms a perfect complement to a portable general-coverage receiver already stowed away in the holidays luggage(which should also contain a length of antenna wire,say, 30 m or so).

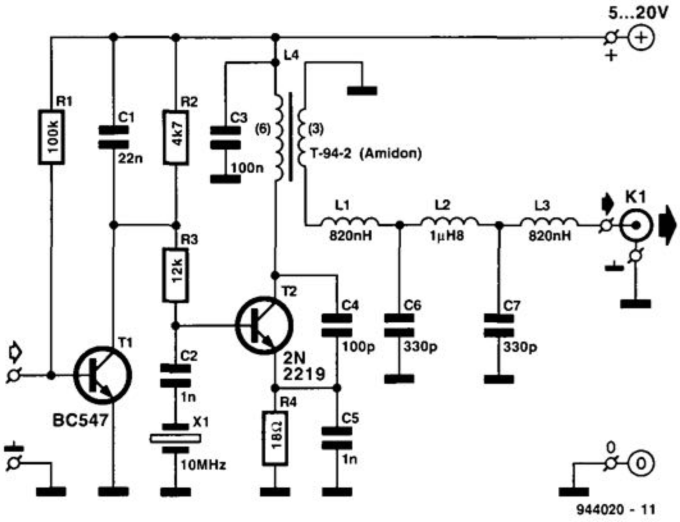
The transmitter is a Colpitts power oscillator based on the inexpensive and familiar Type 2N2219 or 2N2219A transistor. The r.f. output power will be 100–500 mW, depending on the supply voltage. The transmit frequency is held stable by a quartz crystal, which can be tuned slightly, if needed, by adding a 150 pF trimmer capacitor in series with C_2.

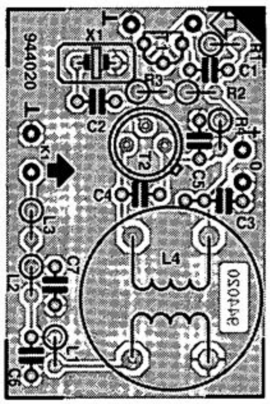
The r.f. signal is coupled inductively from the collector of T_2 into a 7-pole Chebychev low-pass 'pi' filter made from standard E12 series inductors and capacitors. The oscillator is keyed by T_1 which pulls the base of T_2 to ground when the morse key is opened. The stand-by current of the transmitter is only about 2 mA, while the keyed-up current is about 55 mA (at a supply voltage of 10 V).

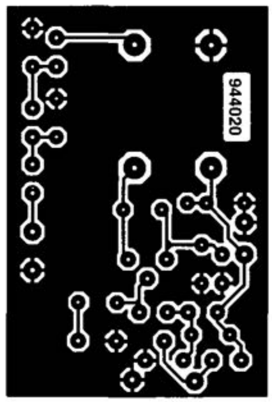
The transmitter is best constructed on the printed circuit board shown. Inductor L_4 consists of six turns of 0.5 mm dia (24 SWG) enamelled copper wire at the primary side (collector of T_2), and three turns of the same wire at the secondary. The core is a T-94-2 type from Amidon, which has an outside diameter of 24 mm (0.94 in.) and an inside diameter of 14 mm (0.56 in.). The A_L value of this core is about 84 µH per 100 turns; its relative permeability is stated as 10. T_2 must be fitted on to a heat-sink. The completed PCB should be built into a metal enclosure.

Harmonics suppression, measured inthe prototype, was 40 dB at 20 MHz, and 50 dB at 30 MHz.

Note: this transmitter may only be built and used by radio amateurs having the appropriate licence, in countries where CW in the 30m band is permitted.







Parts list

Resistors:
R_1 = 100 kΩ
R_2 = 4.7 kΩ
R_3 = 12 kΩ
R_4 = 18 Ω

Capacitors:
C_1 = 22 nF
C_2; C_5 = 1 nF
C_3 = 100 nF
C_4 = 100 pF
C_6; C_7 = 330 pF

Inductors:
L_1; L_3 = 820 nH
L_2 = 1.8 µH
L_4 = T-94-2 ring core (Amidon Associates Inc.); see text for winding details

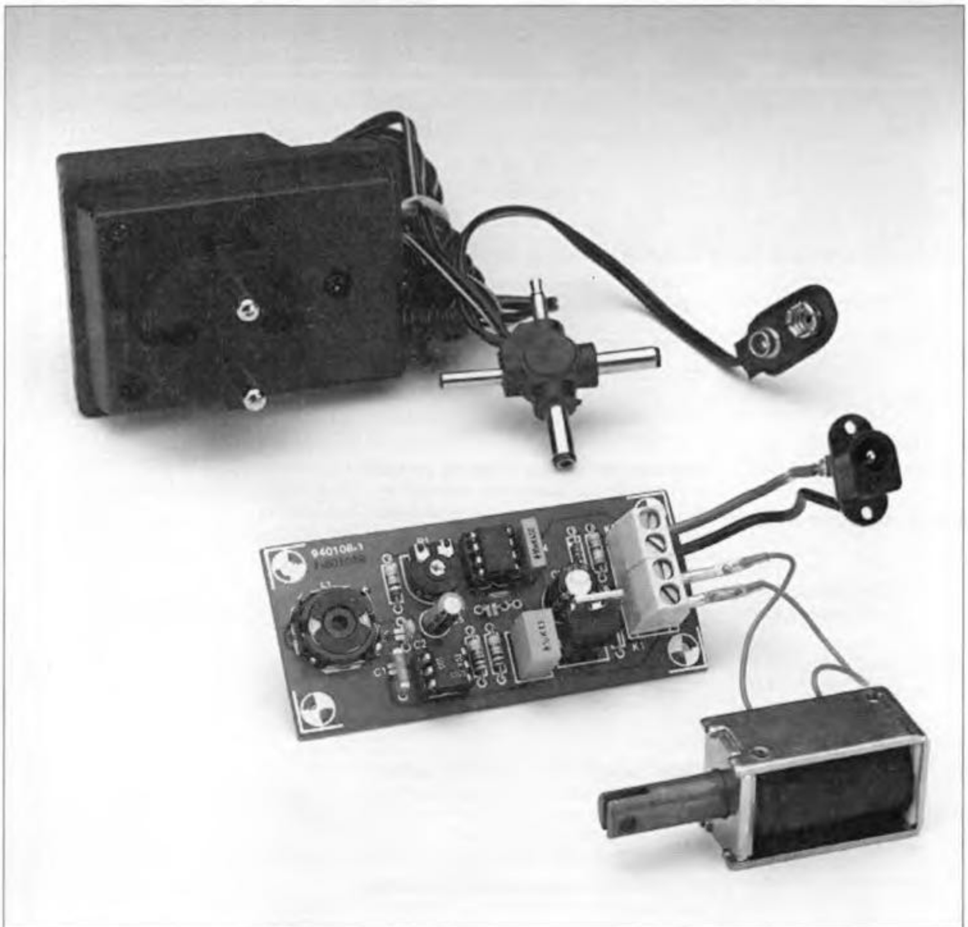
Semiconductors:
T1 = BC547
T2 = 2N2219

Miscellaneous:
K1 = BNC socket.
X1 = 10 MHz quartz crystal.
Heatsink for TO5 transistor.

Design: P. Wyns, ON7WP
[944020]

PROXIMITY DETECTOR

One tiny integrated circuit and a handful of standard components allow you to build a multi-purpose proximity switch. The circuit can be used as a hidden latch as part of a burglar alarm, or, in a more playful way, as an electronic cat door.



Design by J. Bosman

THE heart of the circuit is formed by a TCA105 integrated which contains, among others, a threshold switch. The IC is housed in a 6-pin DIP enclosure. It is specifically designed for use in proximity switches, light barriers and other contactless switching devices. Advantages of the TCA105 include a wide voltage range (4.5 V to 30 V), a fairly large output current capacity (50 mA), and TTL compatibility.

The internal structure of the TCA105 is given in **Fig. 1**. At the left we find the input stage, which is interesting because inputs 1 and 2 are formed by the base and emitter of the first transistor. Next comes a kind of double differential amplifier which, together with its surrounding components, forms the actual threshold comparator. Then follow a double-emitter transistor and an output circuit which consists of two complementary transistors.

The fact that the IC can be triggered by direct voltages creates many applications. A voltage monitor, for instance, is easily produced by connecting the voltage to be measured to pins 2 and 3, via a voltage divider, and connecting two LEDs directly to the outputs, pins 4 and 5. The LEDs then indicate whether the voltage is over or under the set threshold. Another example is a light-controlled switch: simply connect the input pins to a phototransistor. Undoubtedly the TCA105 can be used for many more switch-like functions based on direct voltages. However, a special feature of the IC is its ability to process alternating input voltages also. The input stage of the IC may be turned into an oscillator, and that is the design principle used for the present application, an inductive switch.

Inductive coupling

As shown by the circuit diagram in **Fig. 2**, an *LC* tuned circuit is connected to the input pins, rather than a direct voltage source. The circuit consists of L_{1a} and C_2, and is tuned to about 1 MHz. An oscillator is created by extending L_{1a} with a coupling inductor, L_{1b}. Referring back to Fig. 1, the oscillation condition is satisfied by coupling the emitter of the input transistor back to the base via a tap on the inductor.

As long as the input stage oscillates, IC_1 is in a kind of 'fixed' state. Output 1 (pin 4) is then low, and output 2 (pin 5) is high. That changes if the oscillator voltage disappears, or is significantly reduced. That can be brought about, of course, by inserting a switch in series with C_1, but that defeats the purpose of the circuit (after all, we want the switch to operate quasi-automatically). The aim is, evidently, to control the oscillation voltage without breaking any links in the circuit.

The solution is given by mutual inductance. By holding a piece of metal near the oscillator inductor, the changing magnetic field causes eddy currents in the metal. The energy needed for this to happen is drawn from the tuned circuit, causing the oscillator voltage to be damped considerably. When the damping exceeds a certain level, the outputs of IC_1 toggle. Consequently, pin 4 goes high, and pin 5, low. The threshold at which this happens lies at about 0.35 V. The outputs do not toggle again until a level of 0.5 V is reached, i.e., the hysteresis is about 0.15 V. Returning to Fig. 1, that hysteresis is created by a feedback line between the threshold switch and the input, where input 1 (pin 2) is pulled down virtually to the ground level by a transistor. Consequently, the level at this input is lower still than as a result of the damping. The upshot is that an extra high voltage is needed to get the system going again.

At this point we seem to have reached our goal already. Evidently, something has to be done about the switching output of IC_1, but the switch can be 'operated', in principle, simply by holding a piece of metal close to the pick-up coil. Although the coil will induce eddy currents into nearly every type of metal, there will be marked differences in the amount of energy drawn from the oscillator coil. Iron gives the largest damping because it has large hysteresis losses in addition to eddy losses. That is easily noted from the detection distance. While a copper bar must actually be pressed against the inductor for the switch to work, an iron bar of similar size is detected at a distance of a few millimetres.

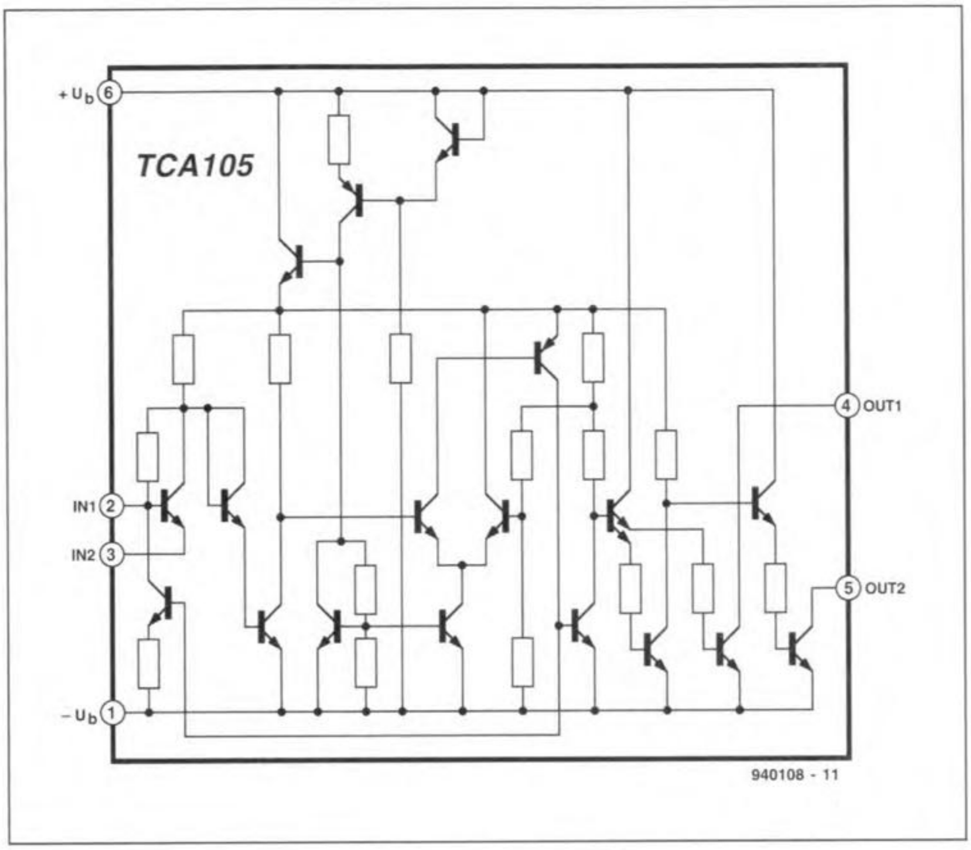


Fig. 1. Internal diagram of the TCA105 integrated threshold switch.

Switching detector

Two functions are yet to be realized at this stage. Firstly, some means has to be provided to prevent the switch being actuated erroneously. Secondly, the actuated state of the switch has to be retained long enough for, say, a door to be opened to let somebody in. All this boils down to delaying and lengthening the switch pulse supplied by IC_1.

Since one output is sufficient for the present purpose, only pin 5 is used; pin 4 of IC_1 is simply not connected. A pull-up resistor, R_1, is used to ensure that the output is high as long as no metal object is detected. Next, the switching signal is fed through a low-pass filter, R_2-C_3, which introduces a delay of about 0.1 second. This delay is effective against accidental switching actions, and also reduced the risk of noise and spurious pulses triggering IC_2.

IC_2 has a far less original character than IC_1. The TLC555 used here is an old faithful. It is wired as a monostable multivibrator and serves to lengthen the switching pulses. Each switching pulse applied to the trigger input causes a fixed 'high' period at the Q output of IC_2. The length of this period is determined by the *RC* time introduced by network R_3-R_1-C_5. The component values shown result in an adjustable period between one and ten seconds. Longer times are possible by increasing the value of C_5.

The Q output of IC_2 drives MOSFET

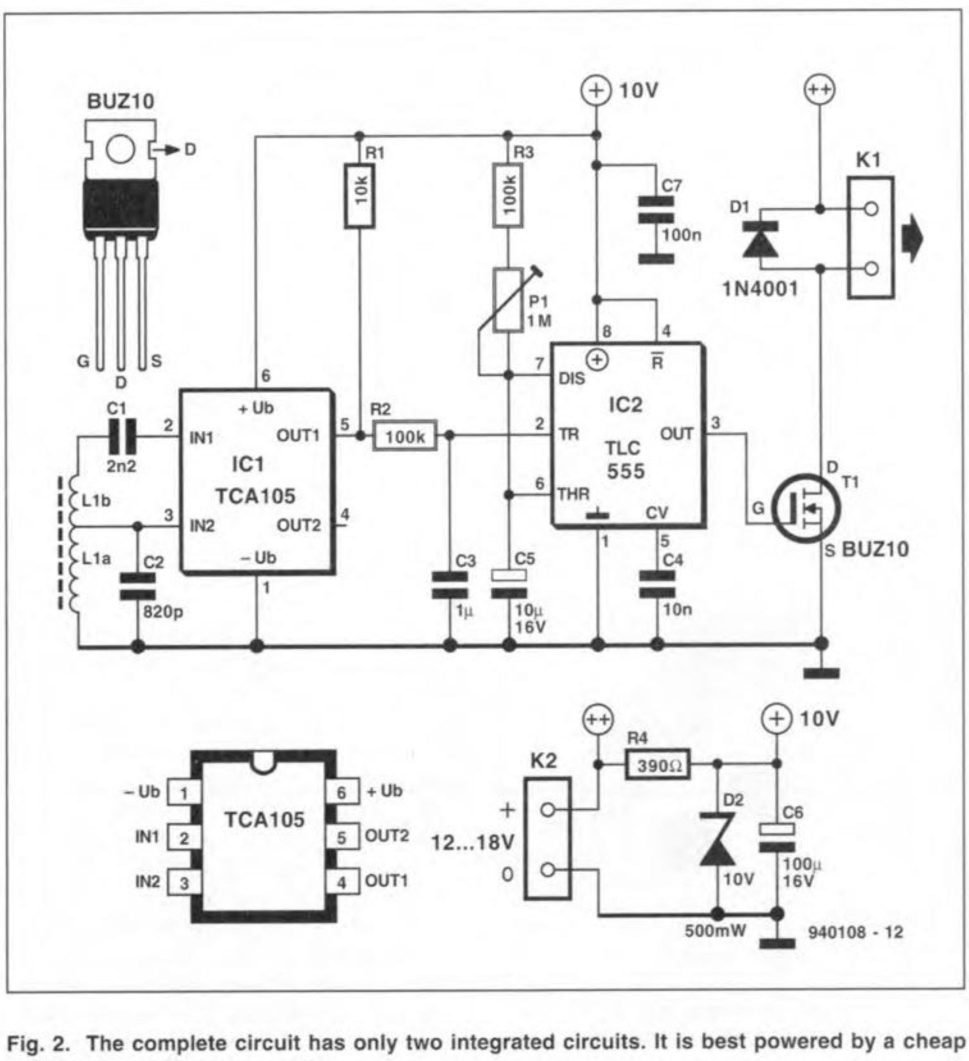


Fig. 2. The complete circuit has only two integrated circuits. It is best powered by a cheap mains adapter hooked up to K2.

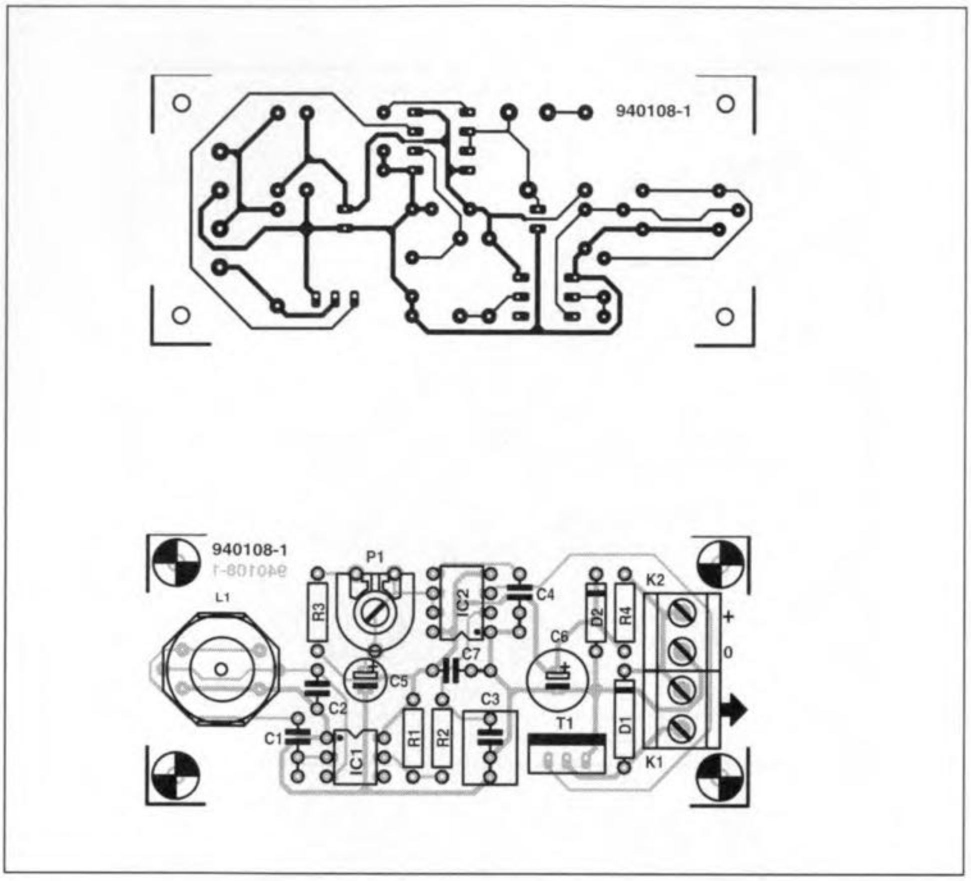


Fig. 3. The miniature PCB gives the proximity detector a 'matchbox' size (PCB not available ready-made through the Readers Services).

COMPONENTS LIST

Resistors:
R_1 = 10kΩ
R_2;R_3 = 100kΩ
R_4 = 390Ω
P_1 = 1MΩ preset H

Capacitors:
C_1 = 2nF2 MKT*
C_2 = 820pF ceramic
C_3 = 1µF MKT*
C_4 = 10nF
C_5 = 10µF 16V radial
C_6 = 100µF 16V radial
C_7 = 100nF Sibatit*

Semiconductors:
D_1 = 1N4001
D_2 = 10V 400mW zener diode
T_1 = BUZ10*
IC_1 = TCA105*
IC_2 = TLC555

Miscellaneous:
K_1;K_2 = 2-way PCB terminal block.
L_1 = pot core: B65541T400A48*;
bobbin: B65542BT2*;
mounting set: B65545B10*;
core material: B48, A_L=400nH.
outside diameter: 14.3mm;
bobbin diameter: 6mm;
height of half bobbin: 4.25mm.
Enamelled copper wire 0.3mm dia (SWG30).

* Siemens Components. UK distributor: ElectroValue Ltd. Tel. (01748) 442253, fax: (01784) 460320.

T_1, which, depending on the application, actuates a relay or a magnetic bolt via PCB connector K_1. Diode D_1 serves to suppress the back-e.m.f. generated by the relay coil.

The circuit is best powered by a ready-made 12-V d.c. mains adaptor, which is connected to K_2. The choice of the adaptor is uncritical, although you must keep in mind that it also powers the relay or the magnetic bolt. Some of these may draw a few hundred milliamps! Since most mains adaptors have a fairly wide spread on their output voltage, the IC supply voltage is limited to 10 V by means of a current limiting resistor, R_4, and a zener diode, D_1. Capacitor C_6 functions as a reservoir device, while C_7 eliminates noise and high-frequency interference.

Fig. 4. Only half a pot core is used to make inductor L1. The double-section bobbin is cut in two with a sharp hobby knife.

Etching and winding

The artwork designed for the printed circuit board is given in **Fig. 3**. Unfortunately, this board is not available ready-made through the Readers Services, so you have to produce it yourself, or have it produced. If you think that is too much work, consider building the circuit on a piece of veroboard. Since there are a handful of components, that should not be too difficult.

Inductor L1 is a home-made type. Fortunately, it is fairly easy to make. A small pot core is used with a diameter of 14 mm (see parts list). Only half the device is used, because L_1 may not be completely screened, and because the metal object to be detected must come as close as possible to the inductor. For this reason, the two-section bobbin has to be cut in two (see **Fig. 4**).

The inductor is wound on the half that remains. It consists of 42 turns of 0.3-mm (SWG30) enamelled copper wire, with a tap at 40 turns. The inductor must be wound neatly, or the total amount of wire will not fit into the available room, **and** the Q (quality) factor will be reduced. Next, the core is fixed on to the base plate with a drop of glue, and the inductor wires are connected to the base pins as shown in **Fig. 5**. The drawing may also be used as a guide to mounting the pot core assembly on to the board.

The photograph in **Fig. 7** shows the finished prototype of the proximity detector, complete with the pick-up coil. The board has two additional connecting points for L_1, so that the inductor may be mounted at the copper side also (note that it must be turned 180°

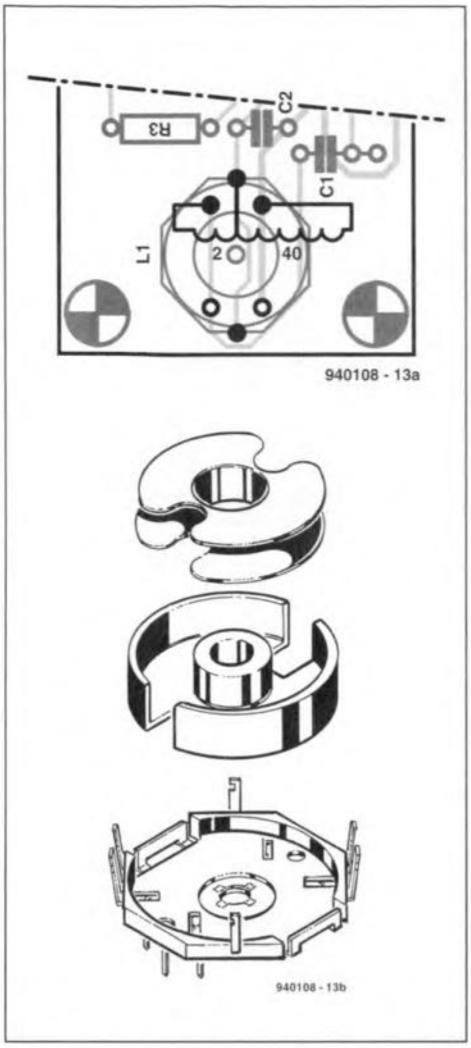


Fig. 5. Showing how the inductor, consisting of 40 and 2 turns, is connected and mounted on to the printed circuit board.

since the connections remain the same).

Practical use

Creative constructors will have plenty of applications in mind for the proximity switch. One of these could be a cat door which opens automatically when the cat rubs its (metallized) collar against the pick-up coil. We have reasons to doubt whether that would work, however, since it requires quite a bit of training of the cat. The most obvious application of the circuit is some kind of door opener, with the circuit fitted in an ingeniously hidden location. In all cases, the oscillator coil has to be easily accessible from the outside, since the detection distance is rather small. In most cases, L_1 is best mounted at the copper side of the board, so that it rests flat against the inside of the enclosure. The enclosure, finally, must be made of plastic. The material must not be too thick, else the maximum detection distance will go down.

(940108)

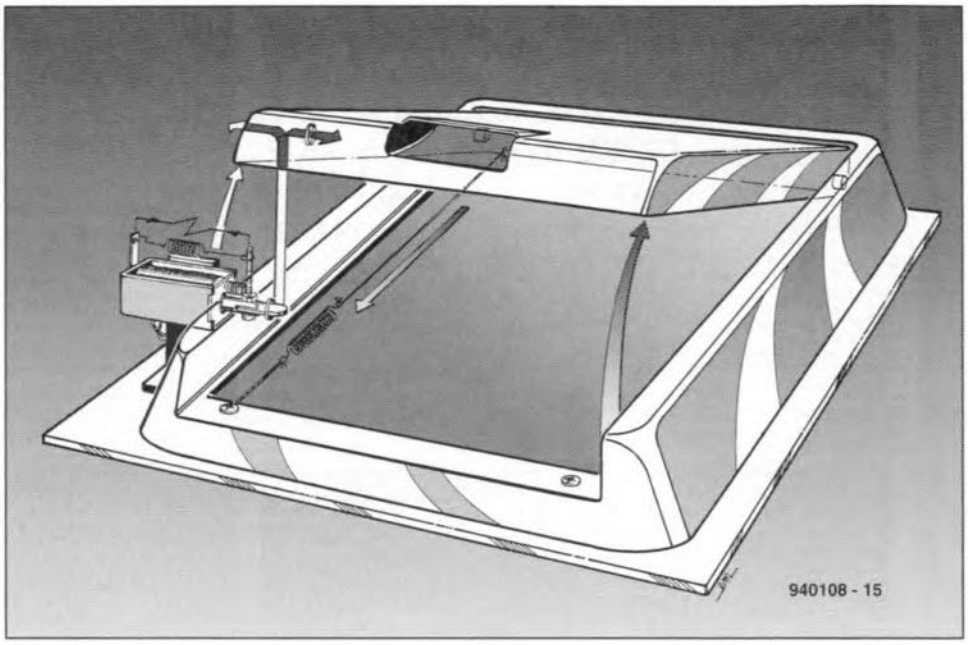


Fig. 6. Suggested construction of an automatically controlled lid.

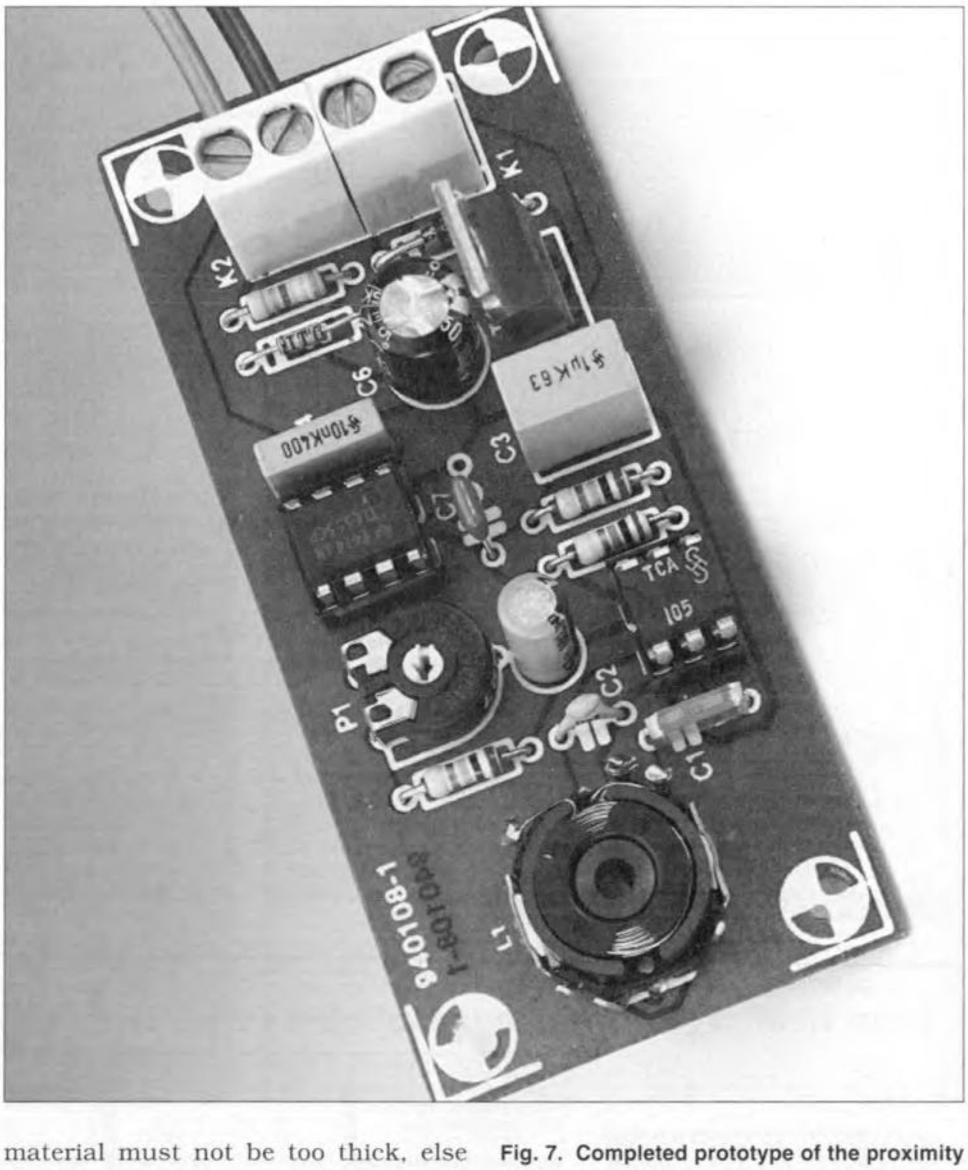
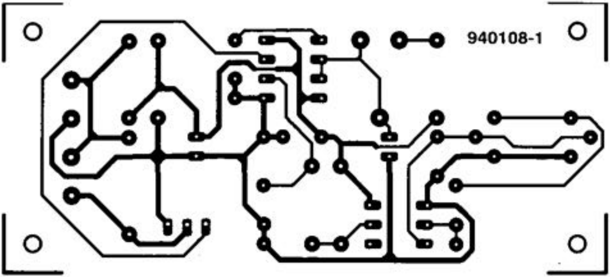
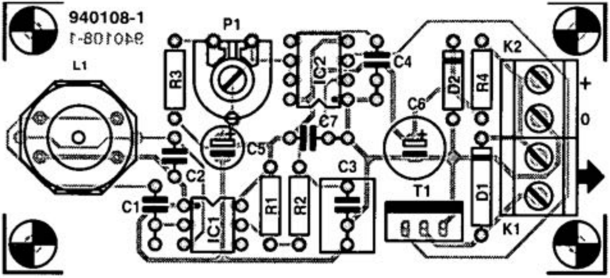
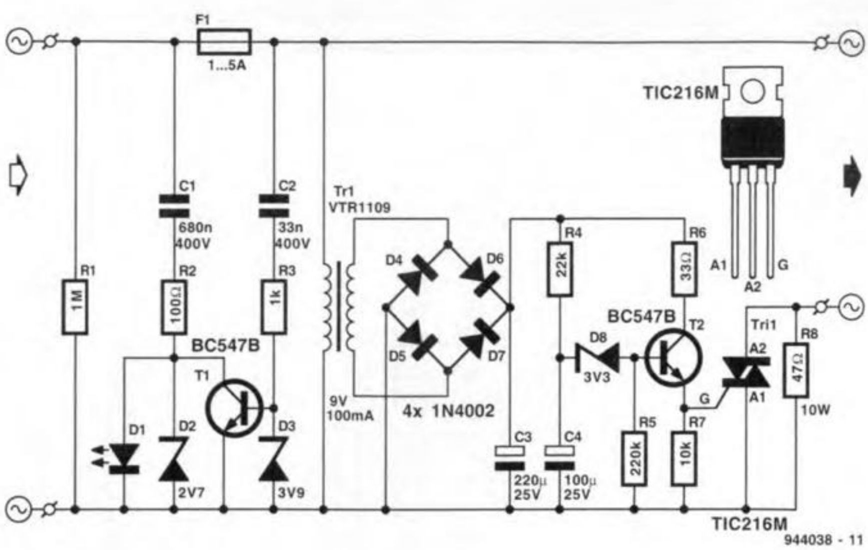


Fig. 7. Completed prototype of the proximity switch.


940108-1


940108-1
P1
L1
R3
IC2
C4
C6
D2
R4
K2
+
0
C7
C5
C2
C3
T1
C1
R1
R2
IC1
D1
K1

FUSE PROTECTOR



Although a fuse is primarily intended to melt when the current through it becomes too large, there are times when this should not happen as, for instance, when a circuit containing a large transformer is switched on. The present circuit protects the fuse against a premature demise.

The circuit offers protection in the first instance by connecting the load to the mains via series resistor R_8 (in UK in the Neutral line). About half a second later, the resistor is short-circuited by triac Tri_1, so that the full mains voltage is then present across the load. This half second delay allows the transformer core to become sufficiently magnetized.

Immediately upon switch-on, the secondary voltage of Tr_1 is rectified by D_4–D_7 and smoothed by C_3. Capacitor C_4 is then charged via R_4 to a potential of about 4 V. When this voltage is reached, zener diode D_8 conducts, whereupon the voltage across R_5 is sufficient to switch on T_2. This results in the triac being started, which short-circuits R_8.

If the fuse, F_1, is defect, there is sufficient voltage across D_1 for it to light. If the fuse is intact, however, T_1 is on and short-circuits the LED. Zener diode D_2 ensures that the potential across the LED can not become too high; at the same time, the reverse voltage is limited to 0.7 V.

With values of components as specified, the circuit can be used for currents up to 5 A.

Design: A. Rietjens
[944038]

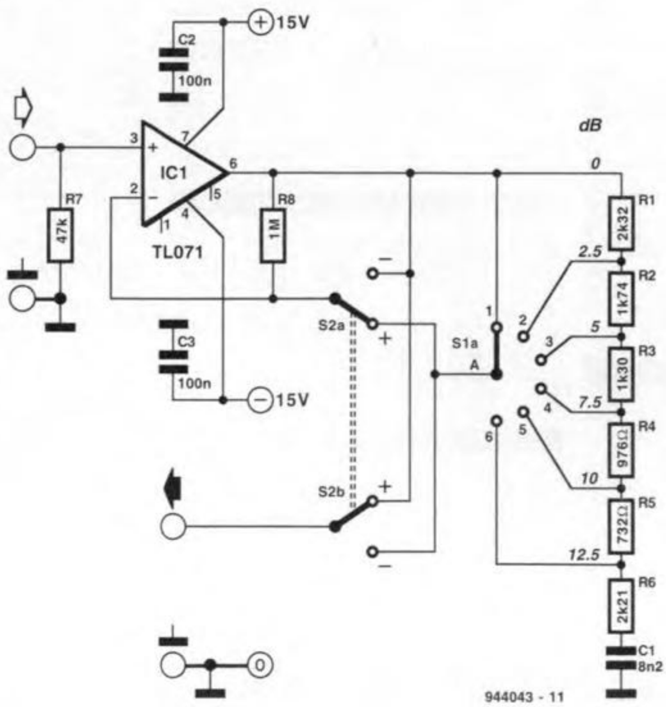
SWITCHED TONE CONTROL

Most readers will be acquainted with a tone control as found on many audio amplifiers. Such a control normally consists of a potentiometer. The one described here uses a rotary switch and a change-over switch to select attenuation or amplifcation. For stereo applications, this requires a two-wafer, six position rotary switch and two small change-over switches (or small relays). For clarity's sake, only one of the channels is shown in the diagram.

With S_{2a} and S_{2b} in position + (amplification), the circuit functions as an active tone control. Frequency-dependent potential divider R_1–R_6-C_1 determinesthe magnification of the higher frequencies, depending on the position of S_1.

With S_{2a} and S_{2b} in position – (attenuation), the potential divider functions as a passive attenuator. The operational amplifier then operates as a voltage follower and buffers the input signal. This arrangement has a drawback in that the output impedance is no longer constant. It may, therefore, be necessary to follow the present circuit by a buffer stage if a low-impedance load is used.

The switches are arranged to give an amplification or attenuation of exactly 2.5 dB per step. Thus, eleven positions are available: –12.5 dB, –10 dB, –7.5 dB, –5 dB, –2.5 dB, 0 dB, +2.5 dB, +5 dB, +7.5 dB, +10 dB and +12.5 dB. Linear steps with this accuracy are hardly possible with the use of potentiometers.



The change-over point of the circuit is at about 2.5 kHz, which may be altered slightly by changing the value of C_1.

Resistor R_8 ensures that the opamp is provided with bias current when S_2 is operated.

If an opamp with a JFET or MOSFET input is used, the click of the switch will not (or hardly) be audible. The Type TL071 used in the prototype may be be replaced by a better type, such as the OPA627, without any problem.

With values as specified, the circuit draws a current of about 2 mA per channel.

Design: T. Giesberts
[944043]

RELAY-CONTROLLED POTENTIOMETER

Standard potentiometers have some unfortunate properties, such as crackling when they get older and unequal tracking in stereo types. A worthwhile alternative is a relay controlled resistive divider.

The present configuration is capable of providing 64 levels of attenuation. With values as specified, each step is 1 dB so that the whole range covers 0–63 dB.

Each of the six attenuator sections has an input and output impedance of 10 kΩ. This means that any one section may be omitted without affecting the input or output impedance of the whole unit.

In the quiescent state (all relays unenergized), the unit provides maximum attenuation. The design is such that the section nearest to the output has the largest attenuation of the six. This arrangement has proved to give the best possible signal-to-noise performance. Consequently, point A forms the lowest significant bit (LSB) and point F the most signicifcant bit (MSB).

The unit is driven by a 6-bit up/down counter, which may need to be followed by a buffer IC, depending on the relays used. The drive may also be obtained from a microprocessor.

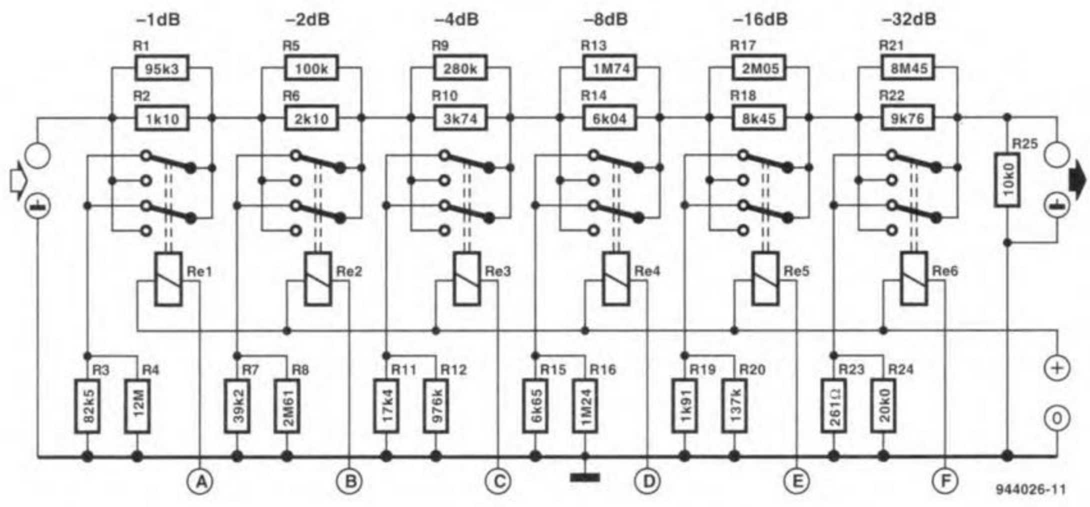
Since errors have a cumulative effect, the calculated values of the resistors have been approached as closely as possible by the use of two parallel-connected 1% resistors. Greater accuracy is obtained by the use of 0.1% types.

When the the unit is built into a preamplifier, it should be borne in mind that the output impedance across which the unit is to be connected is low (otherwise, the values of R_1 and R_2 must be altered as relevant). The load across the output, shunted by R_{25}, must be 10 kΩ (the specified value of R_{25} is, therefore, correct only if the load has a very high impedance).

The choice of relays depends on individual requirements.

The unit draws a current which depends entirely on the relays.

Design: T. Giesberts [944026]



MOTOR SPEED CONTROLLER

The controller was designed to stabilize the speed of small d.c. motors within certain limits. It holds constant the counter-e.m.f. produced by the motor, so that the rotational speed is also constant.

The controller uses a motor speed regulator IC from SGS Thomson, in which the control function is integrated. The IC has three terminals: pin 1 for a reference voltage; pin 2 for the regulator motor current; and pin 3 for ground connection.

The motor is connected between the +ve supply voltage and pin 2 of IC_1. The value of R_1 is about 20× the d.c. resistance of the motor. The current through R_1 is thus $^1/_{20}$ of that through the motor. This compensates for the loss caused by the resistive element of the motor.

The voltage at pin 2 is adjusted by the IC to a value at which the current into pin 1 is $^1/_{20}$ of that into pin 2.

The voltage across the motor, and thus the rotational speed, depends on the setting of P_1.

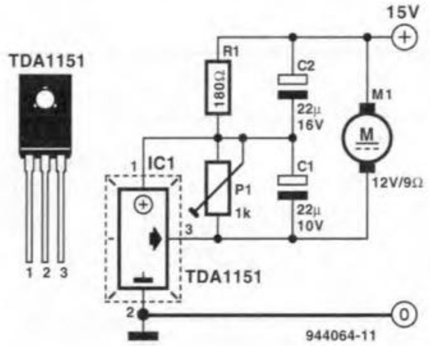
Capacitors C_1 and C_2 ensure that brief voltage variations do not affect the control.

When during operation the rotational speed, and thus the counter-e.m.f., alters, the ratio of the currents at pins 1 and 2 is no longer 1:20. The IC then adjusts the output voltage until that ratio is obtained again.

The maximum current drawn by the motor must not exceed 300 mA.

It is advisable to mount the IC on a small heatsink of about 10 K W^{-1}.

Design: H. Bonenkamp [944064]



8TH ORDER BESSEL FILTER

The filter consists of two fourth-order sections in series. Various components in the second section have been given two different values. The non-bracketed values are used if IC_{1b} functions as a voltage follower; R_{11} is then replaced by a link and R_{12} is omitted. The values in brackets are for use if IC_{2b} is used as a ×2 amplifier; R_{11} and R_{12} are then equivalent.

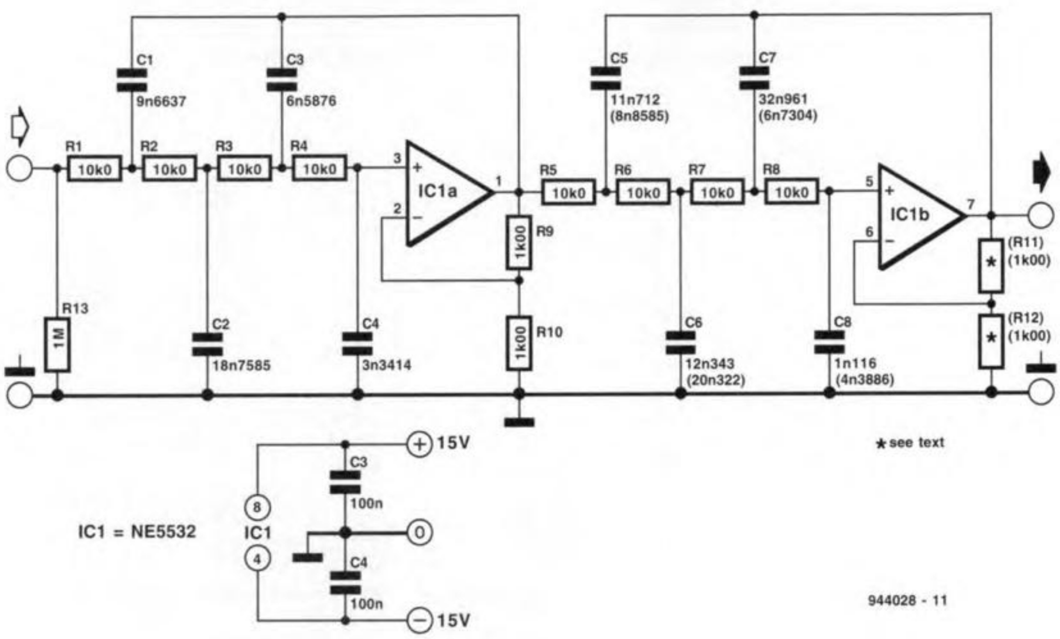
When IC_{1b} functions as a voltage follower, the ratio of C_7 to C_8 is about 30, which is rather high. This means that, particularly at high cut-off frequencies, the value of C_8 becomes so small that it is no longer available as a polyester type. It will then have to be a ceramic or a polystyrene type, which is awkward, because ceramic types should not really be used in audio applications, while polystyrene types tend to be costly. When IC_{1b} functions as a ×2 amplifier, these drawbacks do not exist. The ratio C_6:C_8 is then about 4.6.

The chip may be an NE5532 or a TL071, which draws only half the current of the 5532.

Bear in mind that this eighth-order filter produces rather more noise than a fourth-order one, if only simply because there are two fourth-order sections in series. An eighth-order filter using only one opamp might produce rather less noise. But then, perhaps, the different value resistors may generate more noise.

In general, the present design will prove satisfactory in most applications.

Design: T. Giesberts
[944028]



MOVEMENT ALARM

This tiny circuit can be used as a movement alarm in many situations: in a suitcase or on a bicycle to name but a few.

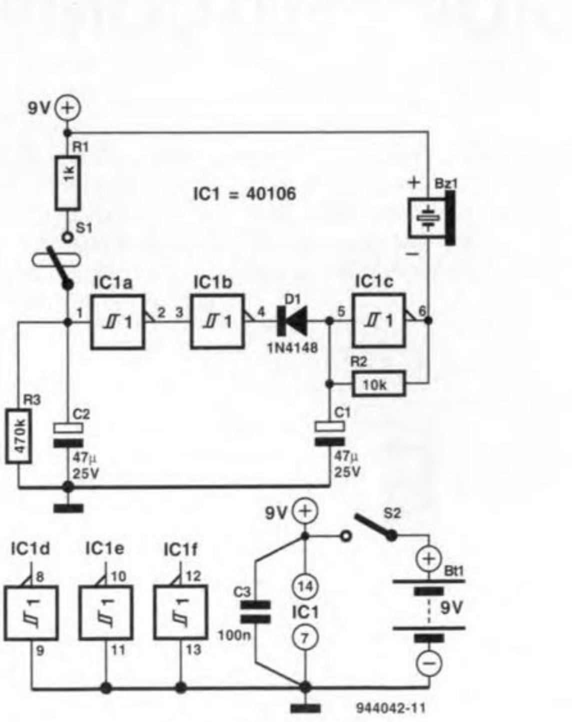
The movement sensor is a mercury switch, S_1. This should be fitted in a suitcase or on a bicycle in such a way that it is open when the suitcase or bicycle is at rest. At the slightest movement, however, it will close. Provided the alarm has been set with S_2, capacitor C_2 will be charged rapidly via R_1 and S_1. Almost immediately, oscillator IC_{1c}, is enabled, whereupon buzzer Bz_1 will sound.

Even if the mercury switch has opened again, the alarm signal will continue until C_2 has been discharged via R_3, which will take about 30 seconds.

Gates IC_{1a} and IC_{1b} function as buffers between the switch-on circuit and the oscillator. Diode D_1 prevents C_1 being discharged via R_3, which would disable the oscillator prematurely.

Since the current drawn by the buzzer does not exceed about 10 mA, a 9 V battery (dry or rechargeable) will last quite a while.

Design: Amrit Bir Tiwana
[944042]



BATTERY ADAPTOR FOR PORTABLE CD PLAYERS

Portable CD players have anannoying drawback: their current drain is so high that even rechargeable NiCd batteries, let alone dry types, last not much longer than two hours. This is not just expensive (if dry batteries are used), but it also means that the user must always have a spare set of new or fully charged batteries.

Measurements show that most portable CD players draw a starting current of about 600 mA and an operating current of 400–500 mA. After the player has been used continuously for about two hours, the internal battery is flat or nearly so. If no spare battery is available, the only way to use the player for a little longer (independently of the mains) is to draw the power from a 6–12 V lead-acid battery (perhaps a car battery). This, however, requires an adaptor like the one described here.

The converter has two main tasks: to reduce the external battery voltage to that required by the player (3.4 V or 6.0 V) and to provide auto switch off after the player has been used for a predetermined period of time. Moreover, it is proof against polarity reversal, both at its input and its output.

If the external voltage (8–16 V) is to be lowered to 6 V, a Type 7806 regulator is needed; if to 3.4 V, a Type 7805.

If a 6 V external battery is used, a low-drop regulator, such as the Type LM2940T-5.0, must be used. To avoid any potential drop before the regulator, diode D_1 is then best omitted.

If a supply of 3.4 V is required, two Type 1N4001 diodes must be connected in series with the output of the regulator. Since the forward drop across each of these diodes is 0.8 V (or nearly so), the 6 V regulator output is reduced to 3.4 V across the load.

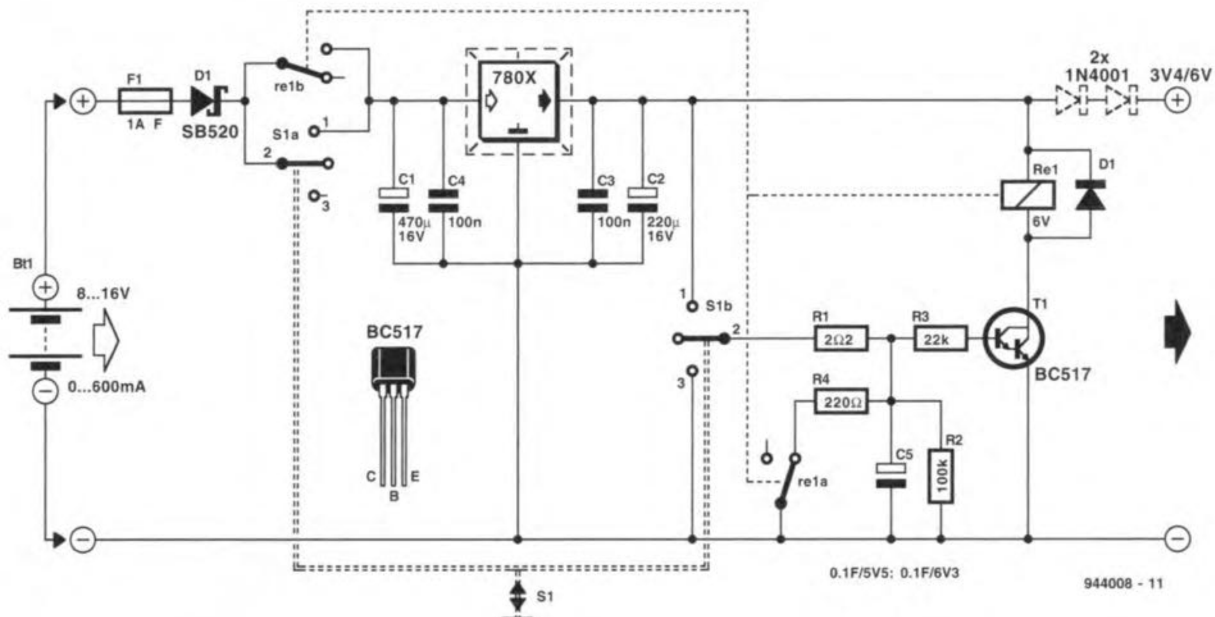
Whatever configuration is used, the regulator should be mounted on a small heatsink.

Diode D_1 protects the converter against incorrect polarity of the input voltage.

Automatic switch-off is provided by T_1 in conjunction with relay Re_1 and double-pole switch S_1. When S_1 is pressed (to position 1), a current flows from the external battery via the regulator to the CD player. At the same time, S_{1b} connects the base of T_1 to the regulator output via R_1 and R_3. The transistor then conducts, so that the relay is energized. Relay contact Re_{1b} is in parallel with S_{1a}, while contact Re_{1a} removes the earth connection from R_4. Capacitor C_5 is then charged to full capacity within about a second.

When in this condition S_1 is released (whereupon it returns to to its centre position), C_5 is discharged via R_2, and R_3 and the base-emitter junction of T_1. When the potential across the capacitor has dropped to about 1.1 V which, with values as specified, takes about an hour, T_1 switches off, the relay is no longer energized and the adaptor is disabled.

Design: W. Zeiller
[944008]



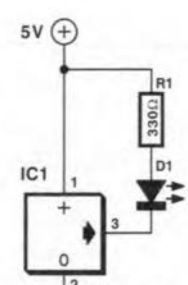
MAGNETIC-FIELD DETECTOR

The Type UGN3140 Hall sensor from Sprague enables a magnetic-field detector to be constructed that is conspicuous by its simplicity.

The IC contains the sensor proper and also a differential amplifier and a compararor with hysteresis. Moreover, its open-collector output is buffered, so that an LED can be driven directly.

The UGN3140 switches at magnetic fieldstrengths of 4.5–27 mT. The hysteresis is about 2 mT. The maximum supply voltage is 24 V. If higher supply voltages than 5 V are used, the value of the series resistor for the LED must, of course, be changed as required.

Design: H. Bonekamp
[944100]



IN-CAR AUDIO AMPLIFIER PART 3 (FINAL)

Design by T. Giesberts

Protection circuits

As described in Part 1, the temperature sensor (T_{23} in **Fig. 2**) is linked to the protection circuits in the inverter via K_1 (pins 2, 4, 6 and 8). Its output is compared with a reference value, set by P_2, in IC_{3b}. Any noise on the connections is filtered out by R_{32}-C_{16} to prevent spurious actuation of the circuit. Hysteresis is provided by R_{31}, which means that the heat sink must cool down by about half a degree before the amplifier is switched on again. In practice, the temperature protection circuit should be set to switch off the supply when the temperature of the heat sink reaches 65–70 °C.

When setting P_2, note that the potential at pin 5 of IC_3 is about 47 mV higher than that across the sensor (because of the drop across R_{32}). If, therefore, the ambient temperature is 25 °C, and it is desired to set the onset of the protection at 70 °C, the potential difference between pins 5 and 6 of IC_3 must be 90 mV (pin 6 negative with respect to pin 5 when the heat sink is at ambient temperature).

The detector for too high a battery voltage is formed by IC_{3a}. This compares de battery voltage via potential divider R_{27}-R_{28} with a reference voltage of 5.1 V. The design is such that the inverter is switched off when the battery voltage reaches 15 V.

Since the outputs of IC_3 are open-collector types, they can be interlinked without any difficulty. The combined output signal is applied to the circuit based on T_2 and T_3. The latter stage applies the comparators' output to the shutdown input of IC_1.

The combination T_2-T_3 acts as a Schmitt trigger which converts the signals from the optoisolators in the amplifier into a switching signal for the shutdown input of IC_1. When the current in the amplifier output stages is large enough, the optoisolators cause the Schmitt trigger to change state via R_{13}. Because of the hysteresis provided by R_{17}, any error states will result in either the switching off of the supply voltage or the lowering of the duty factor. If the supply is switched off, error indicator D_{23} is actuated by T_2.

The circuit based on T_4 has two functions: it switches the inverter on and off, and it reacts to too low a battery voltage. With component values as specified, the inverter will be switched off when the battery voltage drops to 10.8 V. At that level, it must be assumed that the battery is nearly flat. If input 'S' is connected to the switched battery voltage in the car, the likelihood of the inverter draining the battery completely is avoided. With the inverter at 'standby', it draws only 35 mA. In this context, it is strongly recommended not to operate the inverter for longish periods when the car engine is off to avoid the situation that the car can not be started again.

The circuit based on T_5, T_6 and T_7 controls the output relay in the amplifier. This relay has two functions: it provides a soft start and it disconnects the loudspeakers immediately it is actuated during an error state.

The soft start is provided by the time taken (C_1-R_{25} = 4–5 s) by C_{12} to become charged after the supply has been switched on. When C_{12} is charged (or very nearly so), T_6 and T_5 begin to conduct, whereupon the relay is actuated via pins 9–13 of K_1.

When the shutdown input of IC_1 is enabled, T_7 begins to conduct via R_{23}, which causes C_{12} to be discharged instantly. This results in T_6 and T_7 being switched off at once, whereupon the relay is deenergized. Diode D_{25} ensures that there can be no charge on C_{12} when C_1 is not charged.

Construction

Building the inverter on the printed circuit board shown in **Fig. 17** is comparable to constructing an audio amplifier using heavy-duty materials. The two transformers are not available as commercial products and will thus have to be wound by the constructor.

Winding the transformers

Transformer Tr_2 is wound on a Type EDT29 former with N67 core material. This material can be used up to 300 kHz (owing to the double phase rectification, the frequency is 220 kHz).

Each of the two windings consists of 12 turns of 1.5 mm dia. enamelled copper wire. Their winding sense is the same. The former has 13 numbered

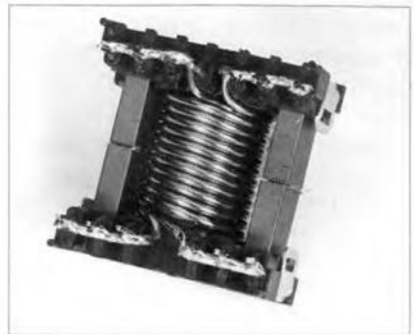


Fig. 9. Winding Tr_2 on an EDT-29 former is straightforward.

Fig. 10. This is how Tr_1 looks after the first secondary winding has been laid.

Fig. 11. ...and like this after the fourth has been laid.

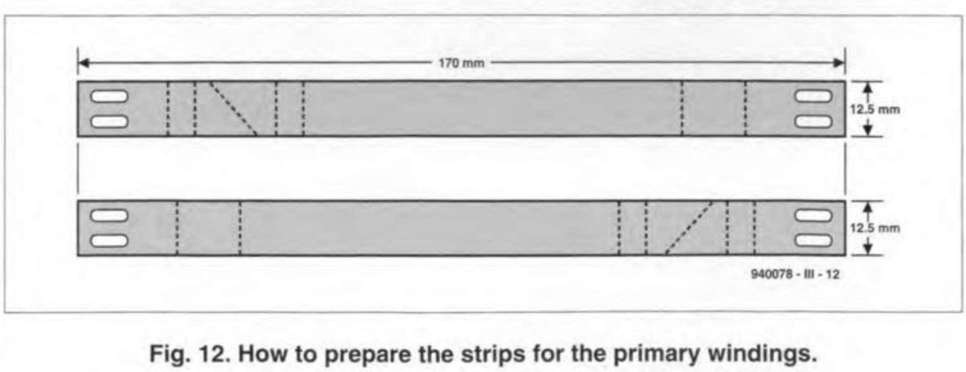


Fig. 12. How to prepare the strips for the primary windings.

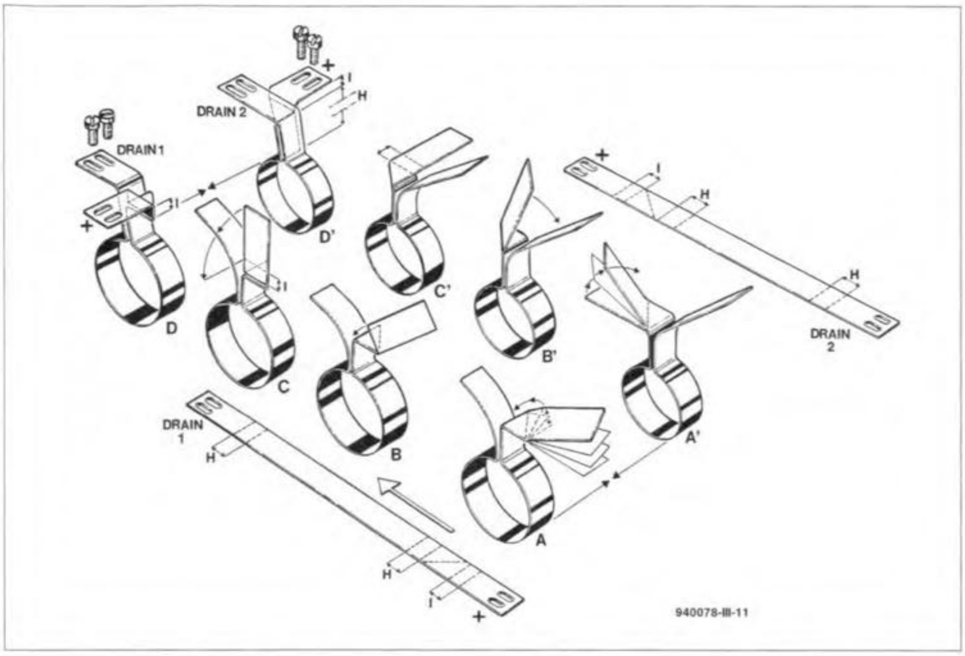


Fig. 13. These are the primary windings for Tr_1.

terminals of which number 10 is not used.One winding runs from terminals 1, 2, 3 to terminals 7, 8, 9; the second from 4, 5, 6 to 11, 12, 13.

The present application requires an air gap of 1.2 mm, which is obtained by placing two pieces of 0.6 mm thick cardboard between the arms of the two core halves.

A transformer constructed to the specifications stated (see **Fig. 9**) can handle currents up to 20 A before becoming saturated.

Transformer Tr_1 is wound on a Type EDT49 former, also with N67 core material. The primaries are wound from copper foil, 12.5 mm wide and 0.3 mm thick. The secondary windings are made from litz wire (four cores each of 30 strands 0.1 mm thick). This kind of wire reduces the skin effect, that is, the concentration of the current at the surface of the wire. It also ensures a much better distribution of the four windings on the former. Each winding is made from two parallel connected wires (eight cores).

Because of the very heavy currents, all connections at the primary side must be made with screw terminals, except in the case of those to the switching transistors, which are soldered (but note that here the current is divided over six solder joints).

In contrast to usual practice, the secondary windings are laid on the former first. Each winding is made by laying the two parallel-connected wires tightly on to the former as shown in **Fig. 10**. Then, place a layer of insulating foil across the winding and lay the second, third and fourth windings on top in an identical manner. The first winding is connected to pins 7 and 14, the second to pins 8 and 13, the third to 9 and 12, and the fourth to 10 and 11. It is imperative that all 240 strands are burnt clean before they are soldered in place. This stage of the construction is illustrated in **Fig. 11**.

Each of the primaries consists of one turn of 0.3 mm thick copper foil (cut to size as shown in **Fig. 12**). Actually, each turn is made of two strips of foil laid on top of each other and connected in parallel. The two primaries must be wound in opposite sense. The windings should be insulated from one another (for which Teflon tape as used by plumbers is excellent: use a few layers since this tape is fairly thin). **Figure 13** shows how the two windings should be folded.

The terminals of the two windings should aligned ready to be pushed through the Z-shaped slot in the PCB. **Figure 14** shows the transformer at this stage of the construction. The slot starts adjacent to pins 3 and 4 and ends next to pins 17 and 18 (see **Fig. 15**). Note that terminals 1–6 and 15–20 of the former are not used. In fact, pins 1, 2, 3, 15, 16 and 17 must cut off down to the copper to make place for the connections to the 12 V supply.

It is advisable to place the transformer temporarily on the board and bend the terminals to locate the exact position for the 3 mm screw holes. Drill these holes and place the core material on to the transformer, which is then ready for soldering on to the board with pins 4–14 and 18–20.

The ancillary windings for the ±46.6 V supply consist of two turns of 0.5 mm dia. insulated wire on the arm of the former at the side of pins 10 and 11. Twist the wires and connect them to the adjacent soldering eyelet—see **Fig. 16**.

Completing the board

To begin with, the printed circuit board in **Fig. 17** must be cut into two. The smaller part is for housing the control and protection circuits. Populating it is straightforward, but note that although C_{11} is a radial capacitor (because of its superior HF performance) it must be mounted lying down. Also, C_5, C_7 and C_9 are surface-mount components that

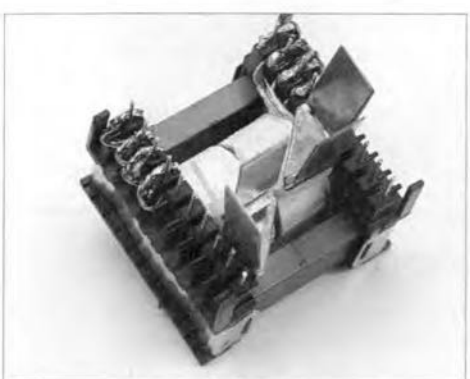


Fig. 14. How to lay the primary windings on the former.

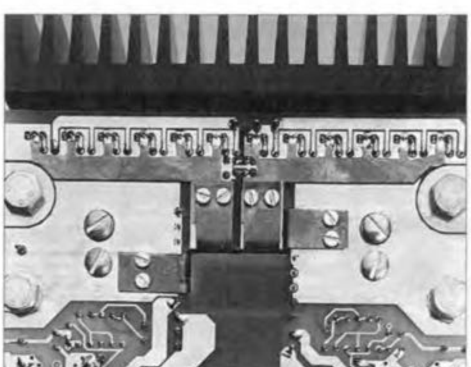


Fig. 15. The ends of the primary windings are pushed through the Z-shaped slot on the PCB and then screwed into place.

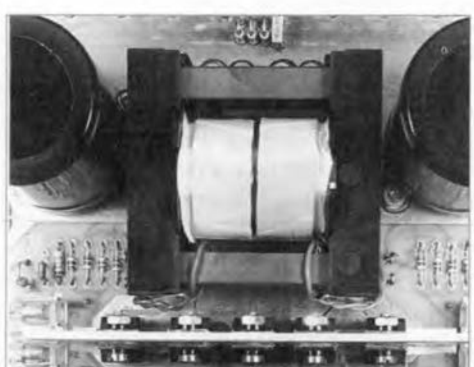


Fig. 16. The ancillary windings for the 46.6 V output consist of just one turn.

are soldered directly underneath or next to IC_1.

The copper on the larger part of the board is only 35 μm thick, which is not sufficient for the large primary currents. Additional copper conductors must therefore be soldered between the drains of power transistors T_{12}–T_{23} and the screw connections on Tr_1. These conductors are cut from 0.3 mm thick copper foil: it is advisable to use a photocopy of the board layout for cutting to their correct shape.

High-grade electrolytic capacitors C_{18} and C_{19} have screw fittings for which suitable holes are provided. The buffer capacitors are radial types with soldering terminals. Do not fit these until the circuit has been aligned.

Fit the diodes on a DIY heat sink (95×62 mm) made from 2 mm thick aluminium sheet. Isolate them from the heat sink by ceramic or mica washers. To prevent the board being damaged by the aluminium, place insulating (Teflon) tape between the it and the heat sink. Create a slight double bend in the terminals of the diodes to the lessen mechanical stress.

Use the drain connections of the FETs on the board as a template for drilling the required fixing holes in the heat sink. Create an S-bend in the terminals of the FETs to lessen the effect of thermal and mechanical stressThe FETs must be isolated from the heat sink by ceramic washers.

Note that MOSFETs have a positive temperature coefficient, whereas diodes have a negative one. For efficient operation it is thus desirable that the heat sink of the FETs remains cool, while that of the diodes may become relatively hot.

The completed board is shown in **Fig. 18**.

Mount the board with the control and protection circuits on insulated spacers to the underside of the larger board directly underneath the FETs: this ensures the shortest possible connections between the driver outputs and the gates of the FETs. Interconnect points G1, G2, FB+, FB–, +12 V and ⊥ on the two boards with normal insulated circuit wire. The completed assembly is shown in **Fig. 19**.

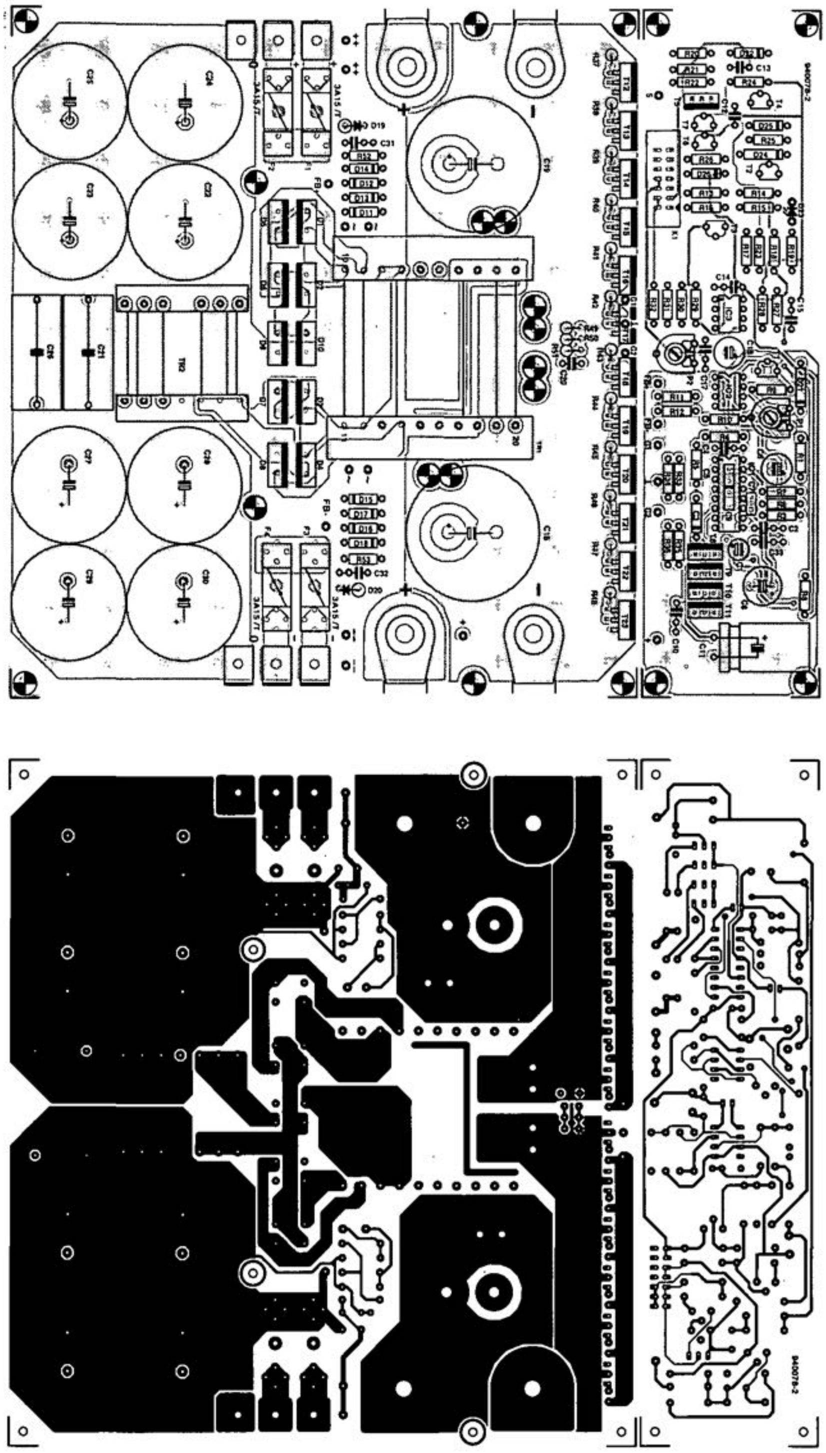


Fig. 17. Printed circuit board for the in-car audio amplifier (scale 1:2).

Wiring

Each part of the inverter has its own terminals for connection to the ±12 V supply. These connections must be made in ≥10 mm² cable terminated in M8 size eyelets. Combine the four cables in a gold-plated junction block capable of linking the four to a ≥25 mm² cable connecting the inverter to the car battery.

The 60 A primary fuse must be fitted in an in-line fuse holder in the +12 V cable close to the battery.

Link the ±43 V power lines to the amplifier via 2.5 mm² cable. Normal insulated circuit wire may be used for the ±46.6 V power lines to the amplifier.

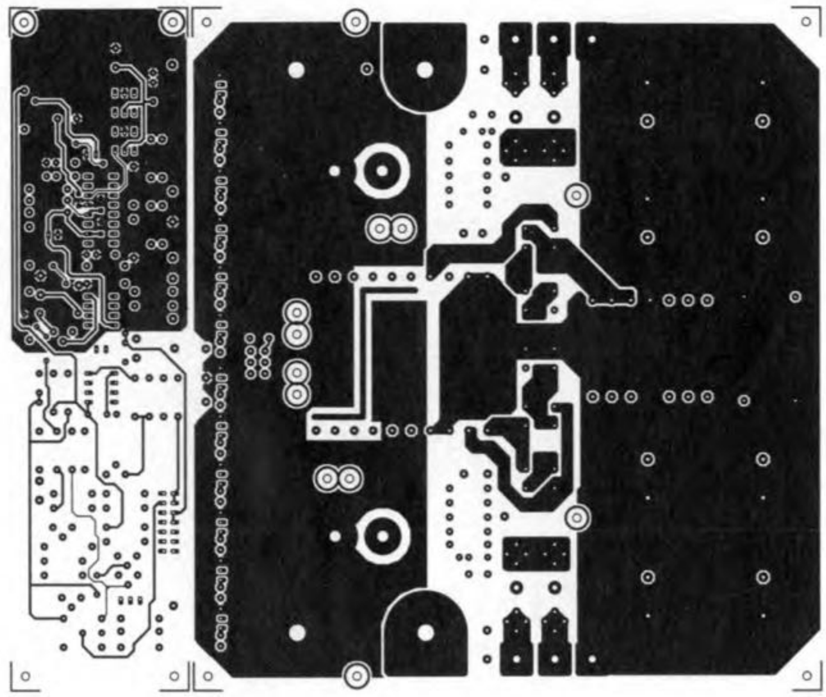
Interconnect the K_1 connectors on the inverter and amplifier with flatcable.

Connect point 'S' to the switched battery voltage in the car.

A wiring diagram of the amplifier and inverter is given in **Fig. 20**.

Alignment

The amplitude of the ±43 V supply is set with pulse width control P_1. As stated earlier, it is best to do do this before C_{22}–C_{25} and C_{27}–C_{30} are fitted. Load the +43 V lines with a 390 Ω, 10 W resistor to earth and shunt this with a voltmeter. Adjust P_1 until the voltmeter shows 43 V ±0.5 V. If that



value can not be obtained, it may be necessary to alter the values of R_9 and P_1 to some degree. Too high a voltage is not recommended for the output amplifier.

Suspend the temperature sensor in a bowl of water at 65–70 °C (keep the terminals dry!) and adjust P_2 until the protection circuit comes into operation as indicated by the lighting of D_{23}.

Fix P_1 and P_2 in position with some nail varnish.

Fit electrolytic capacitors C_{22}–C_{25} and C_{27}–C_{30}.

Next, in the amplifier verify with an ohmmeter that P_2 is set to maximum resistance: this is important. Then, connect the ±43 V and ±46.6 V lines to the amplifier. Connect a millivoltmeter across the amplifier output and adjust P_1 until the meter reading is zero.

Connect the millivoltmeter across one of the emitter resistors of the output transistors and adjust P_2 slowly until the meter reads 22 mV.

Enclosures

Because of the forced cooling necessary in the amplifier (and, perhaps, individual factors), it is best to use separate enclosures for the amplifier and inverter made from aluminium sheet. If, nevertheless, one enclosure is used for both, it is important to separate the two units by an aluminium partition. Whichever is used, the important factor is the adequate cooling of the amplifier.

Parts list

Resistors:
R_1 = 1.2 kΩ
R_2 = not used
R_3 = 470 kΩ
R_4, R_{14} = 1 kΩ
R_5 = 5.6 kΩ
R_6–R_8 = 2.2 Ω
R_9, R_{13}, R_{18}, R_{23}, R_{29} = 4.7 kΩ
R_{10}, R_{16} = 10 kΩ
R_{11}, R_{12} = 6.81 kΩ, 1%
R_{15}, R_{21}, R_{30} = 56 kΩ
R_{17} = 15 kΩ
R_{19} = 2.2 kΩ
R_{20} = 39 kΩ
R_{22} = 820 kΩ
R_{24} = 47 kΩ
R_{25} = 10 MΩ
R_{26} = 8.2 kΩ
R_{27} = 20.0 kΩ, 1%
R_{28} = 10.0 kΩ, 1%
R_{31} = 220 kΩ
R_{32} = 47 Ω
R_{33}–R_{36} = 6.8 Ω
R_{37}–R_{48} = 4.7 Ω
R_{49}–R_{51} = 150 Ω
R_{52}, R_{53} = 1.8 Ω
P_1 = 4.7 kΩ (5.0 kΩ) preset
P_2 = 10 kΩ preset

Capacitors:
C_1 = 4.7 μF, 63 V, radial
C_2 = 68 nF
C_3 = 1 nF, 160 V, polystyrene
C_4, C_{10}, C_{31}, C_{32} = 100 nF, ceramic
C_5, C_7 = 100 nF, SMD, ceramic
C_6 = 100 μF, 25 V, radial
C_8 = 220 μF, 25 V, radial
C_9 = 10 nF, SMD, ceramic
C_{11} = 1000 μF, 25 V, radial
C_{12}, C_{13} = 1 μF, polypropylene, pitch 5 mm
C_{14}, C_{17} = 100 nF
C_{15} = 220 nF
C_{16} = 220 μF, 10 V, radial
C_{18}, C_{19} = 10,000 μF, high-grade, 20 V
C_{20} = 4.7 nF
C_{21}, C_{26} = 10 μF, 100 V, polypropylene
C_{22}–C_{25}, C_{27}–C_{30} = 10,000 μF, 50 V, radial, for PCB mounting
C_{33} = 10 nF, ceramic

Semiconductors:
D_1–D_{10} = BYW29-200
D_1–D_{18}, D_{22}, D_{25}, D_{26} = 1N4148
D_{19}, D_{20} = zener 3.6 V, 1.3 W
D_{21} = zener, 3.9 V, 0.5 W
D_{23} = LED, red, low current
D_{24} = zener, 5.6 V, 0.5 W
T_1, T_2, T_7 = BC547B
T_3, T_4 = BC557B
T_5, T_8, T_{10} = BD139
T_6 = BC517
T_9, T_{11} = BD140
T_{12}–T_{23} = BUZ11

Integrated circuits:
IC_1 = SG3525A
IC_2 = ILD55
IC_3 = LM393

Miscellaneous:
K_1 = 14-way box header, male, vertical
Tr_1 = see text
Tr_2 = see text
F_1–F_4 = 3.15 A fuse, slow, with holder
6 off PCB connector with screw fitting
10 off mica washer for D_1–D_{10} (TO-220)

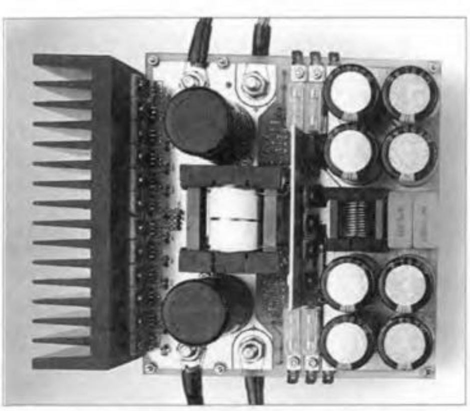


Fig. 18. The component side of the completed board.

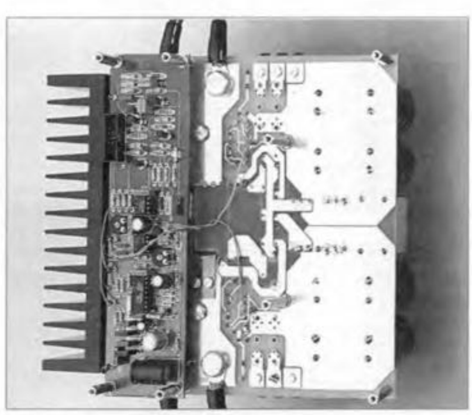


Fig. 19. The location of the control/protection board ensures the shortest possible connections to the gates of the MOSFETS

12 off ceramic washer for T_{12}–T_{23}
2 off gold-plated junction block — see text
4 off gold-plated cable eyelet, M8 size for 10 mm^2 cable
4 off short M8 bolts and nuts
10 mm^2 cable as required
25 mm^2 cable as required
1 off in-line fuse holder with 60 A fuse
1 off heat sink Type SK85 SA*
1 off aluminium heat sink — see text
14-way flatcable
14-way flatcable with 3 14-way flatcable sockets
1 off PCB Order No. 940078-2

[940078-III]

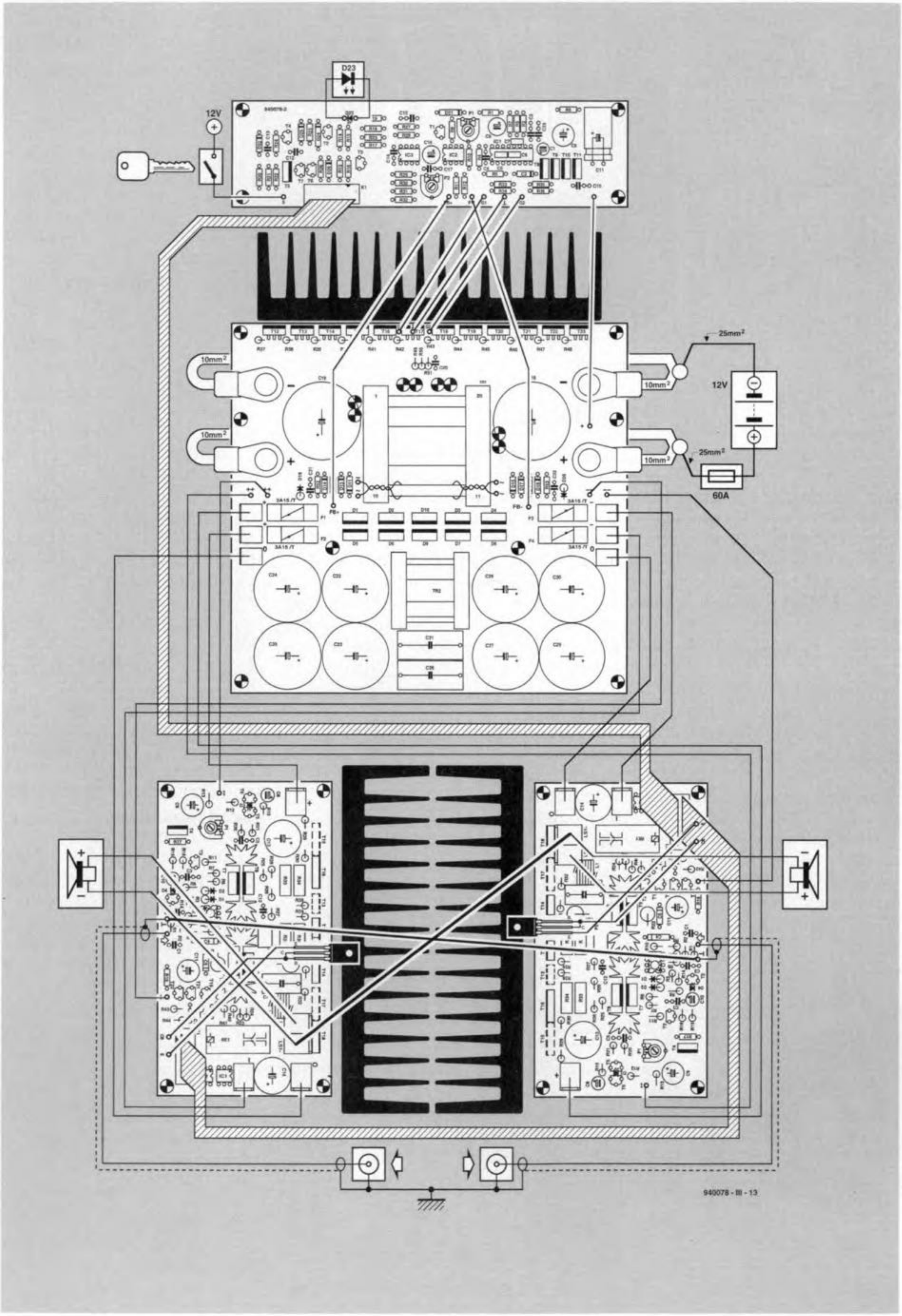
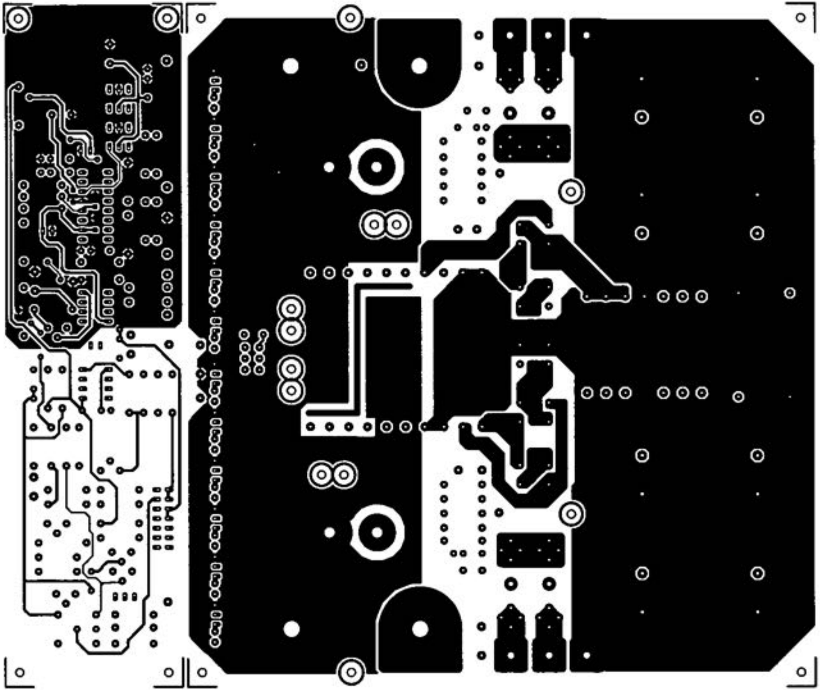


Fig. 20. Wiring diagram of the complete in-car audio amplifier.



APPLICATION NOTE

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

TLC247x DIFFERENTIAL AUDIO FILTERED AMPLIFIERS

Source: Texas Instruments

The TLC247x family from Texas Instruments is a series of integrated circuits specifically designed to make driving a loudspeaker with an analogue or a pulse-width modulated signal as simple as possible. Practically everything is contained in these ICs to supply a power of about 500 milliwatts (peak) to a miniature 8-ohm loudspeaker. Apart from the IC, only a handful of passive parts is required.

THIS series of integrated circuits from Texas Instruments comprises four members: the TLC2470I, TLC2471I, TLC2472I and TLC2473I. These are audio amplifiers with built-in low-pass filters, designed to drive miniature loudspeakers. The chip integrates input buffers, filters, power stages and control electronics. The only external parts which are normally needed are one capacitor and one potentiometer. The possibility to apply audio as well as PWM signals to the inputs makes these ICs eminently suited to applications in speech synthesis systems. In many cases, a TLC247x allows a separate D-A converter to be omitted, which saves board space and money. Furthermore, the TLC247x has on-board logic which switches the IC into stand-by mode when there is no input signal. This is done to save power, which is a must for battery-powered equipment.

Block diagram

The internal 'architecture' of the TLC247x is shown in **Fig. 1**. The input section consists of a modified instrumentation amplifier with balanced (symmetrical) inputs. These inputs are compatible with four different types of input signal. The inputs have internal resistor networks which ensure that an unused input is automatically held at about half the supply voltage. To keep the distortion as low as possible, Texas Instruments recommends decoupling the unused input with a 0.22-µF capacitor to ground. If you use an analogue, unbalanced, input signal, input IN– remains open (0.22-µF capacitor), and the signal is applied to input IN+ (via a coupling capacitor, or else related to $V_{dd}/2$). The maximum input voltage for full drive is 1 V_p with the TLC2470 and the TLC2741, or 2 V_p with the TLC2472 and the TLC2473. If balanced (symmetrical) signals are used, the stated levels

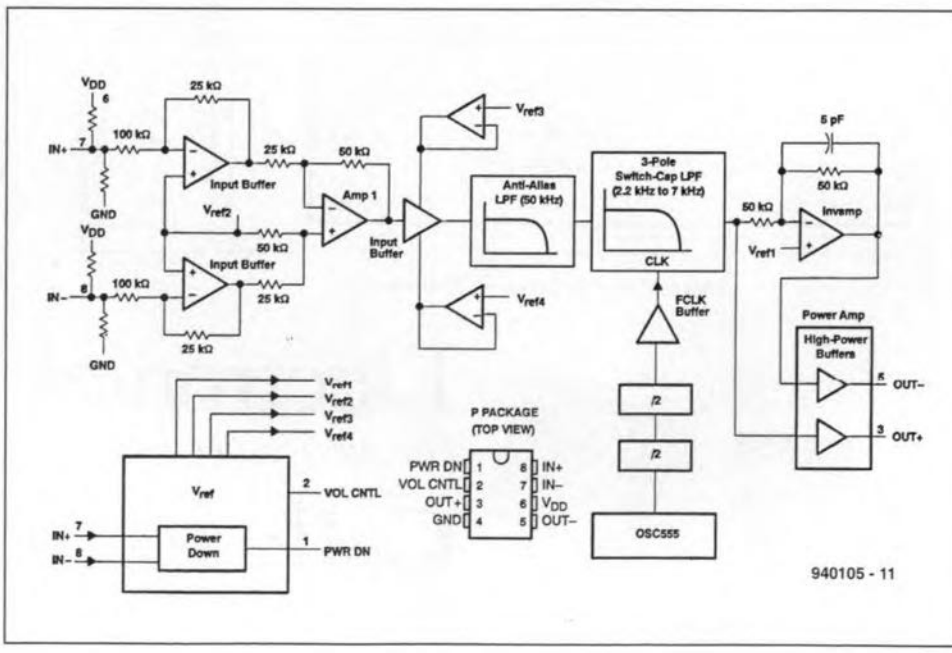


Fig. 1. The block diagram shows the internal structure of TI's TLC247x series of low-power differential input IC amplifiers.

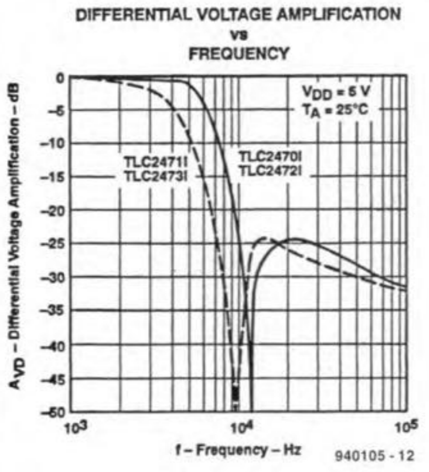


Fig. 2. Frequency characteristic of the internal switched-capacitor filter.

absolute maximum ratings over operating free-air temperature range (unless otherwise noted)

Supply voltage, V_{DD}	6.5 V
Differential input voltage (PWM modes), $\lvert V_{IN+} - V_{IN-} \rvert$	V_{DD}
Input current, I_I (each input)	200 µA
Output current, I_O (differential configuration)	175 mA
Supply current, I_{DD}	300 mA
Duration of short-circuit current at (or below) 25°C	10 ms
Continuous total dissipation	See Dissipation Rating Table
Operating free-air temperature range, T_A	–40°C to 85°C
Storage temperature range	–65°C to 150°C
Lead temperature 1,6 mm (1/16 inch) from case for 10 seconds	260°C

DISSIPATION RATING TABLE

PACKAGE	T_A = 25°C POWER RATING	DERATING FACTOR ABOVE T_A = 25°C	T_A = 70°C POWER RATING	T_A = 85°C POWER RATING
P	1000 mW	8.0 mW/°C	640 mW	520 mW

recommended operating conditions

	MIN	MAX	UNIT
Supply voltage, V_{DD}	4	6.5	V
Input voltage, V_I	–0.3	V_{DD} +0.3	V
Operating free-air temperature, T_A	–40	85	°C

electrical characteristics, V_{DD} = 5 V, PWR DN = 0 V, VOL CNTL = (2/3) V_{DD}, T_A = 25°C (unless otherwise specified)

PARAMETER		TEST CONDITIONS		MIN	TYP	MAX	UNIT
V_{OM}	Maximum differential peak output voltage swing (see Note 1)	$V_{DD} \leq 6.5$ V,	R_L = 8 Ω		2.1	2.2	V
V_{OM+}	Maximum single-ended positive peak output voltage swing	R_L = 8 Ω			3.6		V
V_{OM-}	Maximum single-ended negative peak output voltage swing				1.4		
A_{VD}	Large-signal differential voltage amplification	Single ended,	TLC2470I, TLC2471I		2		V/V
		Output differential	TLC2472I, TLC2473I		1		
I_{DD}	Supply current	V_{DD} = 4 V, Inputs open	No load,		10	15	mA
		V_{DD} = 6.5 V, Inputs open	No load,		13	20	mA
		Power down			50	100	µA
		V_{DD} = 6.5 V,	Power down		50	120	µA
		V_{DD} = 6.5 V, Inputs open	R_L = 8 Ω,		30	50	mA
	Output offset voltage	V_{DD} = 4 V to 6.5 V, Inputs open or both Inputs = V_{DD}	R_L = 8 Ω,	–300		300	mV
	Corner frequency	Gain = –3 dB	TLC2470I, TLC2472I		5		kHz
			TLC2471I, TLC2473I		3.5		

NOTE 1: At V_{DD} > 5 V, limit the maximum differential output voltage to 2.2 V max by reducing the PWM volume via the volume control pin or by reducing the amplitude of the analog input to prevent excessive power dissipation.

apply to the difference voltages between the inputs.

If PWM signals are used, there are three possibilities:

(1) Push-pull mode using two inputs. The 'off' state is selected by making both inputs logic high. By dropping one of the inputs low, the voltage across the loudspeaker is driven to the positive or negative supply level.
(2) One input is not connected (IN–), while the other accepts the PWM signal (a duty factor of 0.5 yields a loudspeaker output voltage of 0 V).
(3) One input (IN–) not connected, and IN+ held at $V_{dd}/2$. Positive pulses then produce a positive output voltage, and negative pulses, a negative output voltage.

The complete input stage has an attenuation of ×2. The buffer that follows it limits the signal to a value of $V_{dd}/2$ ±1 V. That causes the voltage swing of the output amplifier to be limited with PWM signals, while analogue signals smaller than 1 V are passed.

Next comes an anti-aliasing filter which serves to prevent interference between the PWM input signal and the clock signal of the (switched) filter behind it. The signal is subsequently fed through a third-order low-pass filter built from a switched capacitor network. This unit removes all unwanted frequency components from the PWM signal. The roll-off frequency of the network depends on the IC type, and is set to 3.5 kHz or 5 kHz (see **Fig. 2**). The oscillators and dividers needed for this filter are, of course, on board the IC; no external parts are needed. The filtered signal is then applied to two power buffers, one of which is driven via an inverter. That creates a push-pull output for the loudspeaker, which obviates an electrolytic coupling capacitor, and guarantees a relatively high power at a low supply voltage.

With PWM signals, the output volume is varied between 'off' and maximum by applying a voltage between 0 V and 2/3 V_{dd} to pin 2 (VOL CNTL). A higher control voltage, or no voltage at all, results in maximum amplification.

An automatic power-down switch is integrated in the chip for battery-powered applications. This function requires a 3.3-µF capacitor to be connected between the PWR-DN pin and ground. If half the supply voltage is applied to both inputs, the IC switches off after 0.5 s, and the current consumption is reduced to about 50 µA. The power-down mode is also actuated if both inputs are open-circuited, or capacitively coupled to the signal source (which supplies no signal). If an input signal appears, the TLC247x wakes up within 2 ms.

It is also possible to switch on the TLC247x by hand (or keep it switched on) by tying the PWR DN pin to ground. If the PWR DN input is connected to V_{dd} to switch the IC off, no input signal may be applied any more. As a result of the automatic switch-on operation of the IC, that would cause a relatively high current (about 4 mA) to flow.

Applications

Two applications of these interesting devices are shown in **Fig. 3**. Fig. 3a shows a small PWM amplifier, and Fig. 3b, a sketch for 'regular' AF signals. The 3-µF capacitor may be omitted if the automatic power-up /power-down feature is not required.

With the PWM amplifier, the volume is controlled by the 1-MΩ potentiometer. Here, a push-pull PWM signal is applied to the inputs. If you have an 'ordinary' PWM signal (for instance, one produced by a musical greeting card), pin 7 is simply not used, and de-

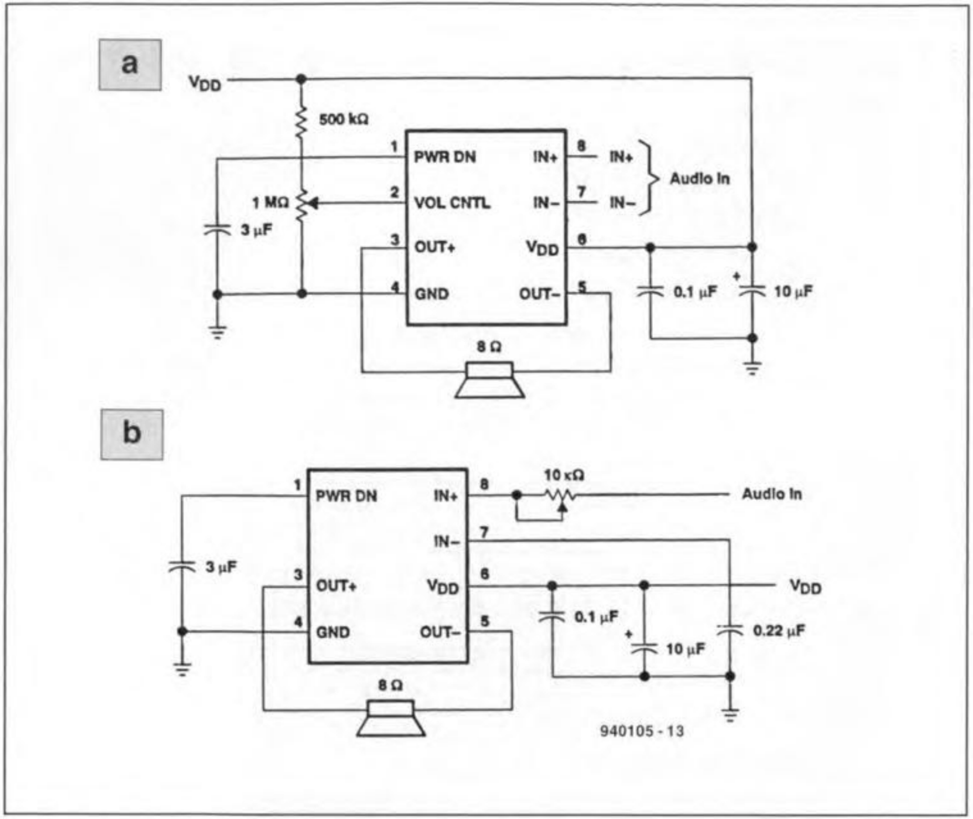


Fig. 3. Two applications: (a) shows an amplifier for pulse-width modulated signals, and (b) one for ordinary (analogue) audio signals.

coupled to ground via a 0.22-µF capacitor. The analogue AF amplifier circuit has a 10-kΩ potentiometer in series with the + input to enable the volume to be controlled. In both circuits, the power supply is decoupled with an electrolytic capacitor (10 µF) and a film capacitor (100 nF).

The loudspeaker is connected between pins 3 and 5. Obviously the loudspeaker terminals may not be connected to ground, for instance, via the cone frame and the amplifier case, since that would short-circuit one of the IC outputs.

(940105)

Source:
Texas Instruments data sheet: TLC2470I, TLC2471I, TLC2473I differential audio filtered amplifiers.

ELECTRONICS FAIR MUNICH 1994

The 'Electronica' (electronics) fair to be held in Munich, Germany, from 8 to 12 November 1994 attracted an estimated 80,000 to 90,000 professional electronics users from over 70 countries.

The main trends this year were ASIC (application specific integrated circuits) and DSP (digital signal processor) devices, both of which are currently experiencing an enormous market growth. Once an ASIC is bought and used by several customers, it is often called an ASSP (application specific standard product). ASICs, however, look poised to become the absolute technology leaders in the entire digital electronics market. That goes for the tools necessary to design them (CAD programs), as well as for the enclosures and the possibilities to test them. ASICs, like the latest microprocessors, are fabricated using 0.5-micron structures. In the top-15 of ASIC manufacturers, the only European competitor, GEC Plessey, is in 13th position. Market leader Fujitsu is followed by manufacturers like LSI Logic, NEC, AT&T, Toshiba and Texas Instruments. ASICS are found in a wide variety of applications, including notebook and laptop computers, mobile telephones and pagers. Until 1998, the market for ASICs is expected to grow by 9% on average. This year, the market represents a total value of about 9.55 billion dollar.

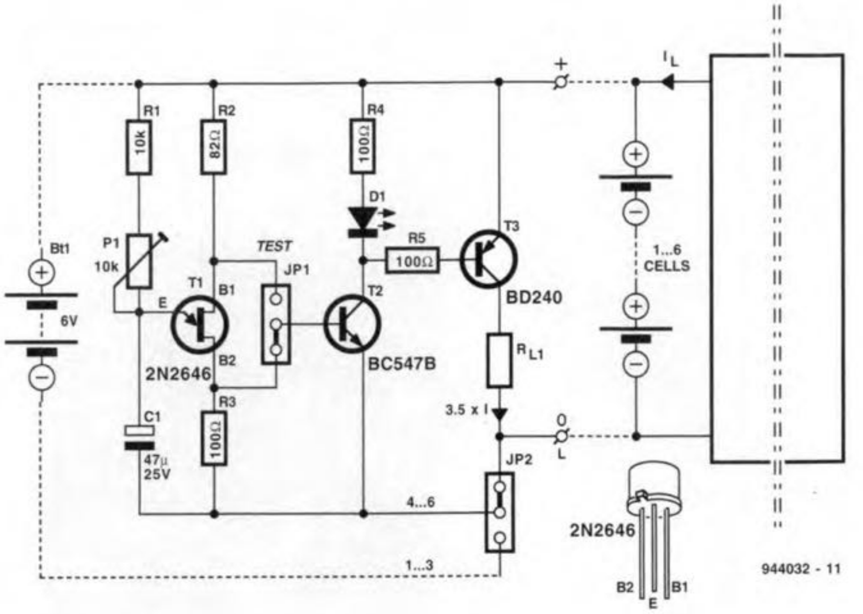
®electronica 94

DSPs were marked this year by an explosion of computing power. They are now among the fastest growing segments of the semiconductor industry, and insiders are convinced that that is likely to remain so for a considerable period, since DSPs are said to be the processor technology of the nineties. In fact, by 1996, at least 90% of all new PCs are expected to have at least one DSP. Not surprisingly, leading manufacturers of DSPs like Texas Instruments, AT&T, Motorola and others, are seeing more and more microprocessor manufacturers entering the DSP market as rivals. The 'new' faces are Intel, National Semiconductor, IBM and Siemens, all of whom have started to develop DSP functions which are integrated in ... ASICs!

Also new is the trend to make ASICs work at 3 V, obviously with the huge market for portable computers in mind.

(940041)

REFLEX-CHARGING ADAPTOR



The adaptor enables almost any NiCd charger to work on the reflex principle. It is simply connected in parallel with a battery when this is being charged.

Transistor T_1 is a unijunction type, which has the property that the emitter-base 2 junction is blocked when the emitter voltage is low. When this voltage rises to a certain value, the emitter-base 2 and base 1-base 2 junctions become low-impedance. When, in the diagram, T_1 is off, the base of T_2 is at ground level. Capacitor C_1 is then charged via R_1 and P_1 to the value at which T_1 begins to conduct. The capacitor is then discharged, whereupon the base voltage of T_2 rises, so that this transistor also conducts. Because of the reducing voltage across C_1, the emitter voltage of T_1 drops below the value at which T_1 conducts, so that this transistor is cut off. Thereupon C_1 will be charged again, and the cycle repeats itself.

In this way, T_1 produces 40 ms pulses with a pulse spacing of 1 s. The pulse rate is determined by R_1-P_1-C_1, and the pulse width by R_2-C_1-base-emitter resistance of T_2.

Transistor T_2 amplifies the pulses, which are then fed to power transistor T_3. Since T_3 conducts briefly during each pulse, relatively high current pulses are drawn from the battery being charged, which is typical of the reflex principle. The flashing of D_1 shows whether the circuit is working correctly.

The operation described assumes the battery being charged to consist of 4–6 cells (6–9 V). If the battery consists of only 1–3 cells, its voltage (1.5–4.5 V) is not high enough to function as the power supply for the present circuit. In that case, the circuit must be powered by a 6 V ancillary battery, Bt_1. Link JP_2 must then be placed the other way from that shown.

Load resistor R_{L1} must have a value which results in a current through T_3 that is about 3.5× the charging current. This value can be determined by placing JP_1 in position 'test', so that T_1 and T_2 conduct continuously. Then, insert an ammeter between the present circuit and the battery being charged to ascertain the current through T_3. Discharging in position 'test' must be kept short to prevent T_3 overheating.

Design: H. Junge [944032]

SINGLE-DIODE FULL-WAVE RECTIFICATION

For measurements on alternating voltage, an active rectifier is normally needed to convert that voltage into a direct voltage which is applied to a meter or an A-D (analogue to digital) converter. Although the present circuit uses only a single diode, full-wave rectification is effected. The output voltage, U_0, for positive inputs is

$$U_o = U_i R_p / (R_2 + R_p),$$

where U_i is the input signal and R_p is the parallel combination of R_3 and R_4.

For negative inputs,

$$U_o = [U_i R_p / (R_2 + R_p)] \times (1 - R_2R_5 / R_1R_p).$$

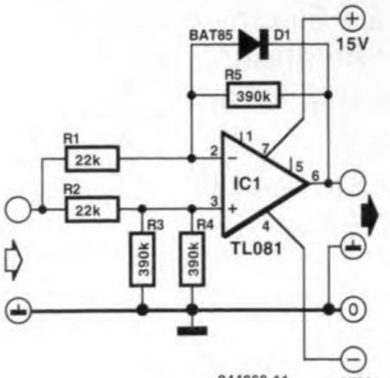
To enable use of the circuit as a full-wave rectifier, these two formulas must be equal.

Computing this gives:

$$R_2R_5 / R_1R_p = 2.$$

If $R_1 = R_2$ (here 22 kΩ), R_5 must be $2R_p$. In the prototype,

$$R_3 = R_4 = R_5 = 390\ \text{k}\Omega.$$



To keep any error in the positive output voltage to a minimum, D_1 is a Schottky type. For alternating voltages, this does not matter, because the output voltage is determined by r_D.

With values as specified, the amplification of the rectifier is ×0.9. The bandwidth is about 10 kHz, but this depends to a large extent on the diode and opamp used. The input impedance is around 20 kΩ. The output impedance is ≤100 Ω.

Design: H. Bonekamp [944068]

PUMP FOR OPERATIONAL AMPLIFIER

It is often necessary to provide power for an operational amplifier from a single supply. Instead of the usual potential divider with coupling capacitors or the rather more sophisticated transducer, the present circuit uses a pump as shown in **Fig. 1**. Operational amplifier IC_{1a} is arranged as the pump oscillator.

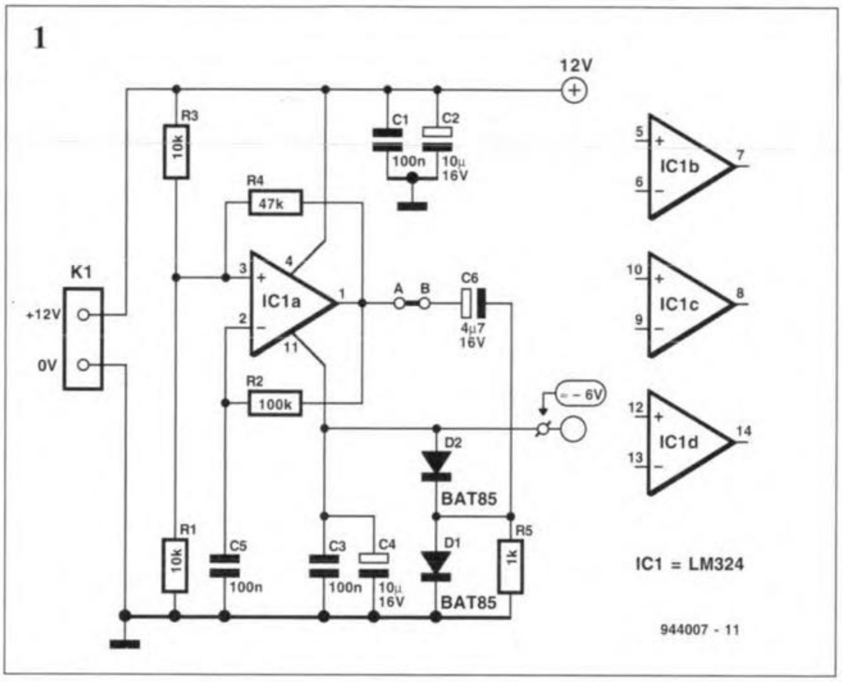
When the supply is switched on, the potential at the negative supply pin of the opamp is two diode forward voltages (D_1 and D_2) above earth potential. Capacitors C_4 and C_6 are then not charged.

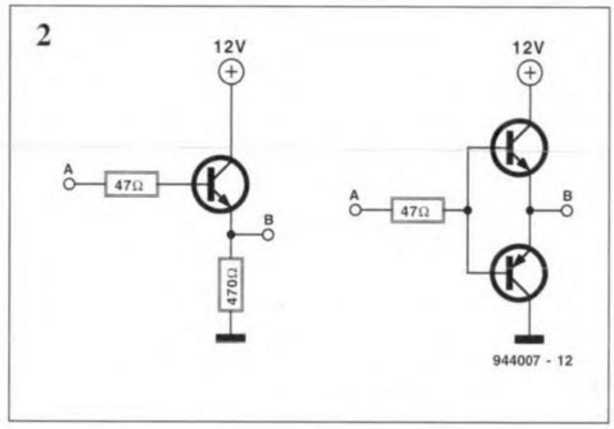
When the output of IC_{1a}, and thus the negative pin of C_6, drops below 0 V, D_2 conducts and C_4 charges to a negative potential. When the opamp chages state, D_2 is cut off and C_6 is discharged via D_1. Thus, the negative potential across C_4 stabilizes at a value that depends on the type of opamp. In the prototype, an LM324 was used, which resulted in a potential across C_4 of –6 V. A TLC274 did not give such good results.

It should be borne in mind that the maximum current from this negative supply must not exceed about 5 mA. If a higher current is required, a buffer stage as shown in **Fig. 2** should be inserted between A and B in **Fig. 1**.

The pump frequency must have a value that does not lie in the operating range of the powered circuit.

Design: E. Berberich
[944007]

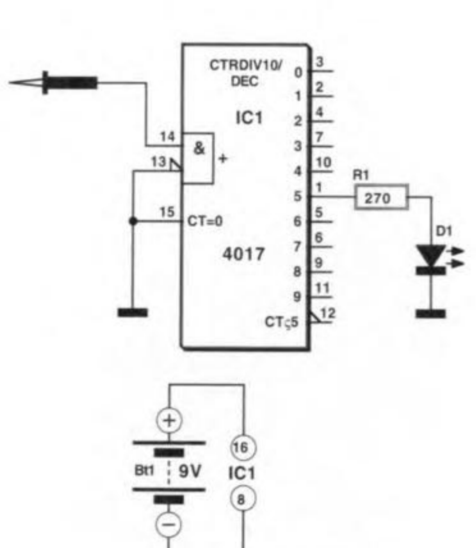




ELECTRIC-FIELD DETECTOR

A small detector to indicate the present of a pulsating or alternating electric field is easily made from an inexpensive IC, an LED, a resistor and a 9 V battery as shown in the diagram.

The IC is a Type 4017 decadic counter with ten decoded outputs. A 'probe', consisting of about an inch (couple of centimetres) of stout circuit wire, is connected to the clock input of the IC. If the probe is in an electric field, a tiny voltage is induced in it. Because of the high input impedance of the IC, this very small voltage is sufficient to act as a clock for the counter. This causes the



LED, which may be connected to any one of the outputs, to flash at a frequency which is a tenth of that of the inducing field. That is,. in the case of a mains voltage field (f = 50 Hz), the LED will flash five times a second.

The indicator draws a current of about 10 mA.

Design: M. Baireuther
[944012]

ELECTRONIC KNOW-HOW

Batteries

Just in case you had not noticed: the world is going portable. Camcorders, mobile telephones, drills, model planes, nothing works without batteries. In particular the spectacular developments in the portable PC and mobile telephone sectors have boosted innovations in the battery industry. In addition to a flood of battery charger ICs and battery management concepts, new battery technologies like nickel-hydride and lithium-ion have come to the fore.

By our editorial staff

IN spite of the remarkable variety found in the application of batteries, dry cells (the 'one-time use' type) still have a larger share than rechargeable batteries in the 'mains independent power supply' market. Unfortunately, some basic terms have got mixed up in the mean time. A battery is generally an energy source consisting of one or more cells, which are either of the primary or the secondary type. By convention, however, the terms primary battery and secondary battery are used. Despite some claims to the contrary, primary batteries are not rechargeable, and can only be used once. Secondary batteries are rechargeable. The latter have the additional advantage that there are now environmentally aware recycling systems with a good efficiency. This is in stark contrast with primary batteries, which require more energy and raw materials for their production, and are separately disposed of as chemical waste because a standardized (but still expensive) recycling method has not been accepted as yet.

Non-rechargeable vs. rechargeable

Despite the undeniable advantages of secondary batteries as regards raw material consumption and safe disposal/recycling methods, there are good reasons for using non-rechargeable batteries (also called 'dry cells') in some applications.

Table 1 and **Fig. 1** show the advantages at least with alkaline-manganese batteries: they have a much higher capacity than their rechargeable NiCd counterparts. To achieve roughly the same usable period as an alkaline-manganese type, a rechargeable battery has to be charged two or three times. So, when it comes to long periods of use without recharging, the alkaline primary cell has the edge. This is particularly important with equipment which has a low current consumption, or is used only briefly at long intervals. In these cases, rechargeable batteries are less economical, and have to be recharged in any case after a few months, to compensate the energy loss caused by their self-discharging. Alkaline batteries, on the other hand, are sure to last a couple of years under the same circumstances. Typical examples of such long-time use are infra-red remote controls, electronic clocks, digital scales, LCD pocket and desktop calculators, and so on.

In apparatus with a relatively high current consumption, on the other hand, rechargeable batteries are the best choice, not only from a point of view of the environment and cost, but also because they have a longer life than non-rechargeable batteries. Because of its higher internal resistance, the voltage supplied by the alkaline cell drops even faster than shown in Fig. 1. The output voltage of a NiCd (nickel-cadmium) battery, however, remains virtually stable during the entire discharging period, even at relatively high currents. The electrical behaviour of ordinary carbon-zinc batteries at high load currents is even worse, particularly at continuous discharging. Here, it really pays to replace them by NiCd batteries because the apparatus can be used longer. A cost comparison between alkaline-manganese batteries and NiCd cells may be based on the assumption that a single NiCd cell with a 'lifetime' of about 1,000 charge cycles replaces about 330 alkaline batteries of the same physical size. Assuming that a radio is powered by, say, four batteries, you will save a small fortune by not using ordinary batteries, even if you use the most expensive rechargeable batteries **plus** a 'deluxe' charger.

NiCd and NiMH

As illustrated in **Fig. 2**, the construction of NiMH (nickel-metal-hydride) cells is very similar to that of NiCd cells (more precisely: cells with an alkaline-nickel-cadmium system). The essential differ-

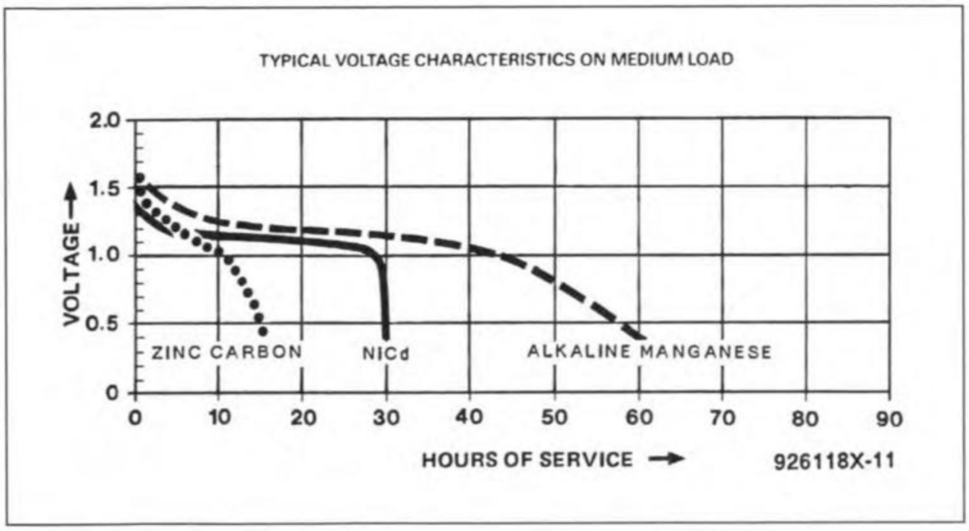


Fig. 1. Comparing the discharging behaviour of primary and secondary cells. Although secondary cells do not achieve the capacity of primary cells (here, an alkaline-manganese type), they do have a much better high-current behaviour, as indicated by the curve.

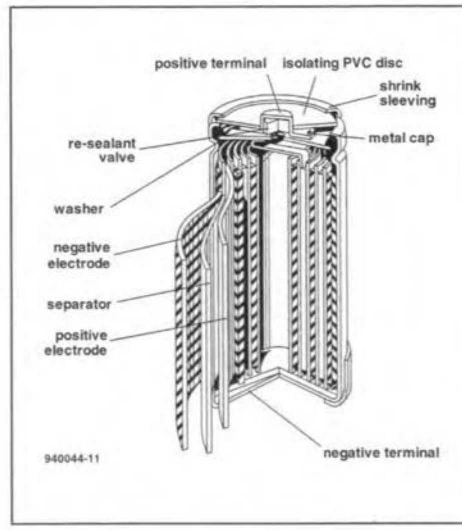


Fig. 2. Round NiMH cells look very much like the familiar NiCd 'UM3' cells. Instead of cadmium, however, special alloys capable of storing hydrogen (almost without additional pressure) are used to make the negative terminal.

ence is the structure of the negative electrode. The metallic cadmium (in charged state) is replaced by a metal alloy capable of storing large amounts of hydrogen, without additional pressure. While the chemical processes at the positive electrode are the same for both battery types, things are different at the negative electrode. In a NiMH battery, hydrogen atoms collect in an alloy mesh, where metal hydride is produced as the charging product. The atoms travel out of the mesh again during discharging, leaving the 'bare' metal alloy as the discharging product. In NiCd cells, cadmium is turned into cadmium-hydroxide during discharging. In both battery types, an 'oversized' negative electrode (with respect to the positive electrode) prevents damage to the cell system when deep discharging or overcharging occurs.

Last year, NiCd cells had a (value) share of about 70% in the global market for secondary batteries, corresponding to a production volume of over 1 billion pieces per year. The share of NiMH cells (of which high-volume production started only recently) is roughly 5% for the year 1993. The market share of these cells is expected to rise to over 40%, however, in the next five years. This hopeful expectation is based on a couple of rather elementary advantages of these normally green batteries:

- free from heavy metals (no cadmium, lead or silver);
- high energy density (up to 1.2 Ah for 'Mignon' (HP7/UM3) types);
- no 'memory' effect.

Table 2 shows that NiMH cells have certain distinctive features which give them the edge over other battery systems. That is, apart from the cost factor. That disadvantage is caused mainly by the hydrogen capturing metal alloys being more expensive than cadmium. Be that as it may, the price of NiMH cells is expected to go down once mass production is under way.

It is advantageous for the proliferation of NiMH cells that their characteristics are very much like those of NiCd cells: a life expectancy of 500 to 1,000 charging cycles; a discharging voltage of 1.2 V with a virtually flat curve; cell voltage rising to about 1.55 V when charged (see **Fig. 5**); normal charging current equal to 0.1 times the nominal capacity at a charging period of 12 to 14 hours; and overcharging allowed for up to 100 hours at the nominal charging current. As far as voltage is concerned, that means that it is perfectly all right to replace a NiCd battery by a NiMH type. The discharge curve of the latter is almost identical to that of a NiCd type. Its capacity, however, is almost twice as high as that of a standard NiCd cell, as illustrated in **Fig. 3** for 'Mignon' size batteries. Interestingly, the comparison for 'Mignon' cells is not even the most convincing, looking at the performance of the latest high-capacity NiCd cells of this size. Panasonic, for example, recently introduced a 'Mignon' size NiCd battery with a nominal capacity of 1,000 mAh, which is claimed to have an even higher typical capacity of 1,100 mAh (see **Fig. 4**).

The discharge curve shown in Fig. 3 applies to a current of 1 A. At higher currents, however, NiCd batteries surpass NiMH batteries, whose capacity drops faster. High-current discharging with more than 3*C* A (i.e., 3 A with the NiCd battery in Fig. 3) is not allowed on a NiMH battery, while NiCd types have no problems with even higher currents.

While the discharge characteristics of NiCd and NiMH batteries are practically the same at small and average load currents, there are marked differences in the voltage characteristics during charging. In general, the charging voltage of a NiMH cell is a little lower than that of a NiCd cell. The voltage peak at the end of

Table 1

Capacity in [mAh]

Cell type	Alkaline-Manganese dry cell	Nickel-Cadmium rechargeable battery
HP7/AA/UM3/Mignon	1,500 - 2,000	500 - 1,100
HP11/C/UM2/Baby	5,000 - 6,000	1,200 - 2,500
HP2/D/UM1/Mono	10,000 - 12,000	2,200 - 5,000
PP3/6F22/1604	400 - 600	70 - 120

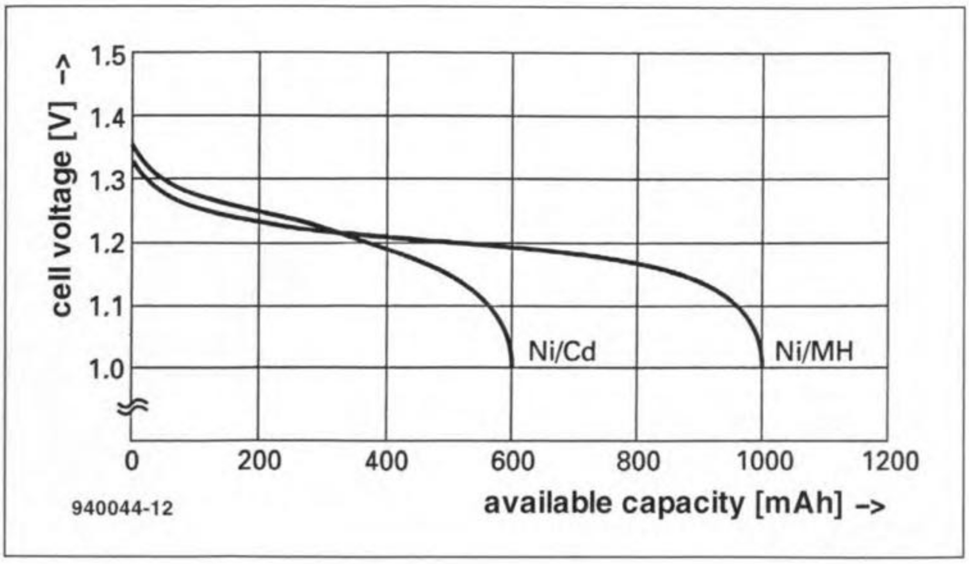


Fig. 3. Discharging voltage plotted as a function of available capacity with NiMH and NiCd 'UM3' size cells. The discharging current is 1 A.

Table 2

Characteristic	NiCd	Lead	NiMH	Li-ion
Energy density (vol. related)	–	–	++	++
Cycle behaviour	++	–	++	++
Self discharging	+	+	+	++
Fast charging	++	–	+	–
High-current loading	++	+	+	–
Reliability	+	++	+	–
Cost	+	++	–	–
Voltage compatibility	++	–	++	—
Environment aspect	—	–	++	+
Discharging voltage stability	++	–	++	–

Legend:
++ excellent
\+ good
– adequate for many applications
— essential disadvantages
* projected data

a high-current charging period on a NiCd cell is less marked on a NiMH cell, as illustrated in **Fig. 5**. Because the peak almost disappears at low charging currents as well as at higher temperatures, automatic switch-off according to the delta-U method is not feasible just like that.

Charging techniques

The standard charging operation for NiCd as well as for NiMH consists of a charge with a constant current of 0.1*C*, where a 'long' overcharging period of up to 100 hours is allowed (but only at 0.1*C*). It is recommended, however, to stay on the safe side, and stop the charging period on reaching between 150% and 160% of the nominal capacity (140% for NiCd cells), with the aid of a timer. The corresponding charging periods are then 15 to 16 hours (NiCd: 14 hours). The standard charging operation is only allowed at temperatures between 0 °C and +45 °C. At lower temperatures, the charging current must be decreased to 0.05*C* A (t<0 °C) for NiCd, and 0.03*C* A for NiMH batteries.

Charging at currents higher than 0.1*C* A ('fast' charging) is only allowed at room temperatures if the battery is not overcharged at this current. The less frequently a battery is overcharged, the longer its lifetime — that goes for both systems. To prevent overcharging, pre-discharging should be used to make sure that no partially discharged batteries are given a fast charge. It is then sufficient to use a timer to limit the charging time, which should take no more than five hours (0.3*C* A), or four hours for NiCd cells at the same charging current. In addition, a temperature monitor should be used which switches off the charger at +45 °C (max. 50 °C) for NiCd batteries, and +55 °C (max. +60 °C) for NiMH types. Charging at 0.3*C* A should not be done at temperatures below +10 °C and above +45 °C.

Even faster charging is allowed on NiMH batteries using currents of between 0.5*C* A and a **maximum** of 1*C* A. Here, too, do not rely on a timer only, even if the batteries are known to be fully discharged. It is recommended to use a battery charger IC, which uses two switch-off conditions: (1) a decreasing charging voltage after reaching the maximum (minus delta U), and (2) the rate of rise of the battery temperature, related to minimum and maximum temperature levels. In addition, a timer may be provided to limit the charge time. To be able to recognize the delta-U factor, it is essential for the battery voltage to be measured at regular intervals (during a short interruption of the charging current). On finishing the fast charging period, the charger should switch to voltage retention charging at 0.03 to 0.05*C* A. The following switch-off conditions are recommended for NiMH batteries:

» thermal shutdown at temperatures <10 °C and >60 °C;
» minus delta-U switch-off at voltage decrease rate of <10 mV per cell (NiCd: 10 to 20 mV per cell);
» temperature rate of rise shutdown at >1 °C per minute (NiCd: 0.5 °C per minute).

A battery charger designed on the basis of the above principles may be used for NiMH **and** NiCd batteries. That is the case with the chargers shown in **Fig. 6** and **Fig. 7**.

Ultra-fast charging is not allowed with present-day NiMH batteries. By contrast, all round NiCd cells are suitable for **pulsed charging** at currents between 4*C* A and 6*C* A, so, for instance, with 4 to 6 A at a cell capacity of 1 Ah. In this process, the charger is switched off under voltage control, just before the full charge is reached. At a cell temperature of 20 °C, the switch-off level is 1.55 V. This value is subject to a rate of –4 mV per degree of temperature rise, to be regulated by a reliable temperature compensation system. The cell voltage may only be measured while no current flows. That is achieved by inserting 30 to 50 ms long measurement periods which occur every 1.2 s. As soon as the peak value is detected a number of times during these periods, the system should change to charge retention mode, which is usually achieved by trickle charging. Additionally, a thermal shutdown circuit set for a cell temperature of 45 °C (max.

50 °C), and/or a temperature rate of rise greater than 0.5 K/min, should be provided. Since ultra-fast charging is not allowed at low temperatures, the temperature guard should also have a lower limit set to a value between 10 and 15 °C.

In addition to the above charging methods, there is also a process known as pulse charging, which means that a voltage gradient (for instance, the first-order integral of the charging voltage characteristic) is defined as a switch-off condition, and monitored by a microcontroller sporting an A-to-D converter. As an additional safety measure, time and temperature guards are also integrated. This process is generally referred to as 'reflex' charging: each charge pulse is followed by a brief high-current discharging pulse.

Whatever charging method is used, faulty batteries must never be charged. Pulse charging allows you to determine whether all cells in the battery are actually storing energy, simply by measuring the cell voltage after the test. When several cells are being charged at a very high current (charging time less than 15 minutes), each cell must be fitted with its own temperature sensor.

Lithium-Ion batteries

Sony have been producing light-weight rechargeable lithium-ion batteries for about two years already. Other manufacturers have followed suit, including Sanyo, AT&T Battery, Matsushita (Panasonic) and, quite recently, NEC. The electrodes in these batteries are made from a special lithium compound. The Li-ion battery is activated by moving ions between the electrodes, as result of charging and discharging. By virtue of the special lithium compound, the battery is simple to maintain. The Li-ion battery is marked by high efficiency, which gives it a wide range of applications. Sony's camcorders CCD-TR1/TR3/TR8, and the SC series, have been supplied with Li-ion battery packs for more than two years.

Li-ion batteries can be charged more than 1,000 times, and are free from the so-called 'memory' effect. Also noteworthy is their high energy density, which is nearly three times that of a similarly sized NiCd battery. The difference is actually almost a factor four if the weights of the two battery types are compared. Li-ion batteries take overcharging remarkably well. The almost continuous charging of the batteries in a cordless telephone, for instance, presents no problems for a Li-ion pack, while many a NiCd set gives up the ghost after a few months. The self-discharging loss of these new batteries is about 50% less than with NiCd and NiMH types, and automatically results in much longer storage times. On the down side, Li-ion batteries have a high cell voltage of 3.6 V, which means that they are not voltage-compatible with dry cells or NiCd batteries. Their main application will, therefore, be in battery packs, where each Li-ion cell replaces three NiCd cells. **Table 3** shows a comparison between the data of a Li-ion video recorder battery and a NiCd pack with the same capacity. Because of the materials used in the Li-ion battery, it can be disposed of as chemical waste without serious problems. Although the battery is environmentally 'safe', recycling is possible.

Apart from the cell voltage, the discharge curve is also quite different from that of a NiCd or a NiMH battery. As the battery is drained, its voltage drops gradually, almost like a dry cell. The fact that the cell voltage is not stable will hardly present problems, though, because the higher cell voltage allows (switched) regulators to be used. Also, the voltage drop, which is proportional to the amount of current drawn from the battery, allows the capacity of the battery to be checked rapidly and easily. The Li-ion cell is, therefore, an excellent choice for all mobile applications, because it allows battery use to be accurately scheduled, avoiding unnecessary charging cycles. It also enables the capacity of batteries which have been stored for some time to be measured accurately and reliably, and that is nearly impossible with NiCd batteries.

Capacitor instead of battery

The so-called GoldCap double-layer capacitor developed by Matsushita in the 1970s has very high capacitance values (in the farad range). The device is classified, as far as its electrical characteristics are concerned, between an aluminium electrolytic capacitor and a secondary battery. The main disadvantage of the GoldCap, its lower current capacity as compared to NiCd and NiMH button cells, is made good by the fact that it is a long-life, short-circuit resistant, and rapidly charged device. A charging circuit is not required — actually, not even a series resistor — because the internal resistance of the Goldcap limits the charging current to acceptable (high) levels. Evidently, there is no memory effect, either, and the device is free from materials like heavy metals which are difficult to recycle and a threat to the environment.

Thanks to their remarkable characteristics, Goldcaps have found many applications as back-up batteries in video recorders, telephones and computers.

Fig. 4. The capacity of a NiMH battery is almost twice as high as that of a standard NiCd cell. The heat is on again, however, with the introduction of Panasonics' new 1,000-mAh NiCd 'UM3' size cells.

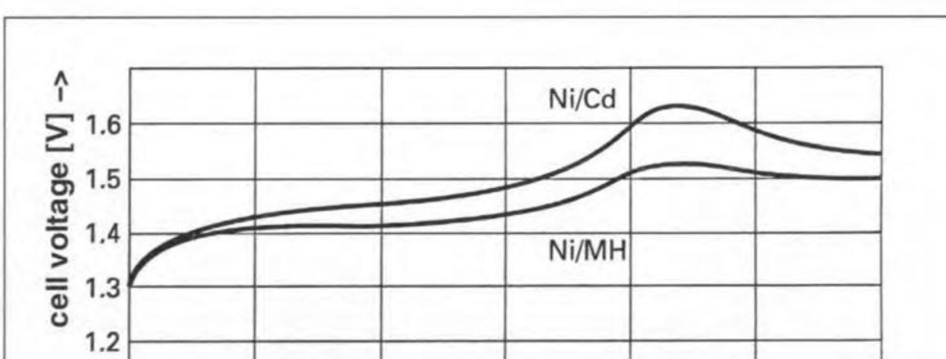


Fig. 5. Charging voltage characteristics of NiCd and NiMH batteries at a temperature of 20 °C and a charging current of 1*C*.

Terms and definitions

Capacity *C*
The capacity, C, of a battery is the product of the discharging current, I, and the discharging time, t:

$C = I\,t$ [Ah]

t = time from start of discharging to reaching lowest discharging voltage; unit: hours.
I = constant nominal discharging current; unit: ampère [A].

Currents
Charging and discharging currents are written in relation to the nominal battery capacity, C. For instance, for a battery with a nominal capacity, C, of 1 Ah:

$0.1C$ = 100 mA
$3C$ = 3 A.

Nominal charging current
With NiCd cells, the nominal charging current for full charging within 14 to 16 hours is $0.1C$ A.

Nominal discharging current
With NiCd cells, this is defined as the current which drains the nominal capacity from the battery in a period of 5 hours. For instance, for a battery with a nominal capacity, C, of 1 Ah:

$I = C/t$ = 1 Ah/5 h = 0.2 A

Nominal capacity measurement
The nominal capacity, C (in Ah), refers to the total amount of energy which is supplied by the battery at a nominal discharging current of $0.2C$, over a period of 5 hours. A NiCd cell is discharged to the minimum discharging level of 1.0 V at a temperature of 20 °C ±5 °C.

Table 3

Product	NP-500H	NP-55H
Battery type	Lithium-Ion	NiCd secondary
No. of cells	2	5
Nominal voltage	7.2 V	6.0 V
Capacity	1,200 mAh	1,200 mAh
Self-discharging after 6 months	30%	60%
Operating temperature	0°C - 50°C	0°C - 50°C
Weight	95 g	143 g
Size (wxhxd)	38.4x20.6x70.8mm	45.5x18x89mm

They are also used as battery replacements in bicycle lights, solar powered watches and calculators, tooth brushes and electric shavers.

The latest development in this area is the **PowerCap**. Based on 'ultra-capacitor' technology, this device offers even higher capacitance than a Goldcap (typically between 470 and 1,500 farad at 3 V). An application example: to improve the composition of car exhaust gases, a 12-V module containing PowerCaps is used to heat the catalytic converter to 800 °C in about 10 seconds. The PowerCap is then recharged within 40 seconds at a current of 150 A. The advantage over a direct 'cat' heating from the car battery is mainly that this energy source need not be made larger than strictly necessary. In the electric and hybrid cars of the future, PowerCaps can 'help' the lead-acid battery when a high current is suddenly required, for instance, during acceleration. The development of new electrode materials aims at a further increase of the energy density of PowerCaps. The ultimate aim is, of course, to replace the lead-acid battery altogether. For obvious reasons, automotive industries have shown considerable interest in these developments.

Batteries of the future

For quite some time now, a number of manufacturers, including Varta and BASF, have been working on the so-called **polymer** battery, in which the positive electrode is formed by a polymer foil. Cells have been developed in which a conductive polymer called Polypyrrol is used. This particular polymer holds great promises for the future because it has been found to give results comparable to those of a Li-ion battery when used in conjunction with a lithium electrode and an organic electrolyte, both in round and flat constructions (yes, a 'stamp' size battery).

Meanwhile, researchers in the U.S.A. have reported on rechargeable **zinc-air** systems. Computer manufacturer Zenith Data Systems, for instance, has revealed that it co-operates with AER Energy Resources in the development of a battery which, they hope, is capable of pow-

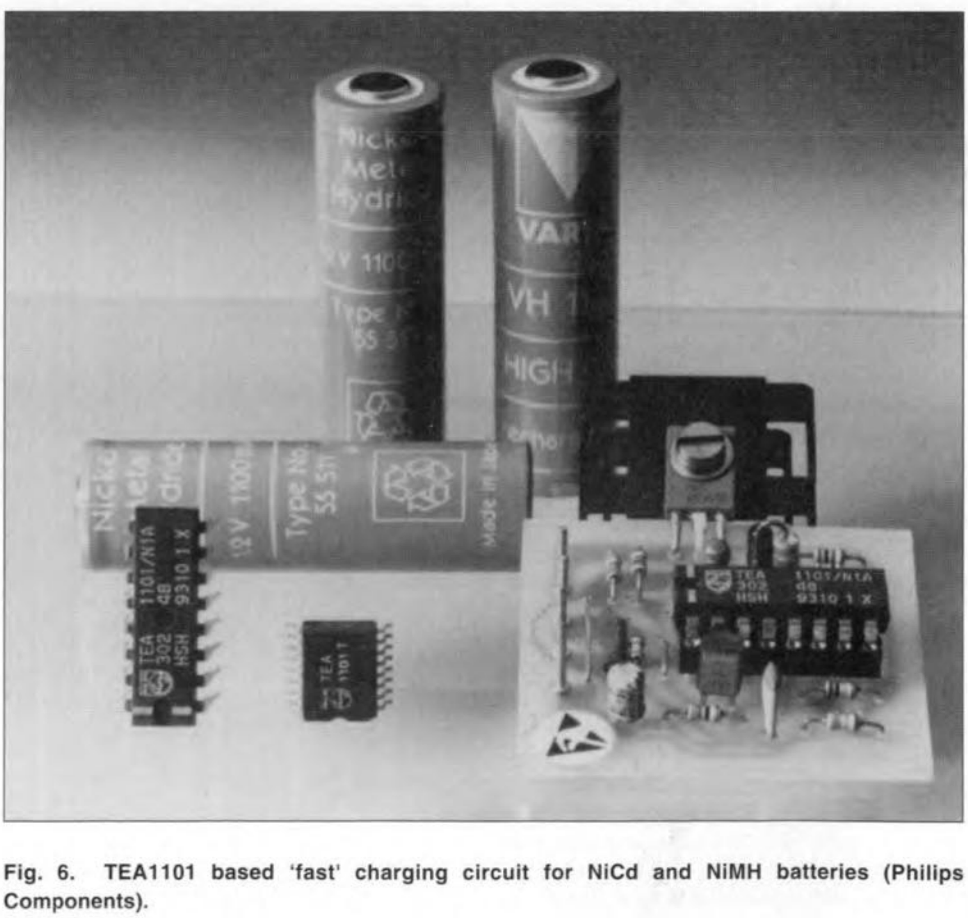


Fig. 6. TEA1101 based 'fast' charging circuit for NiCd and NiMH batteries (Philips Components).

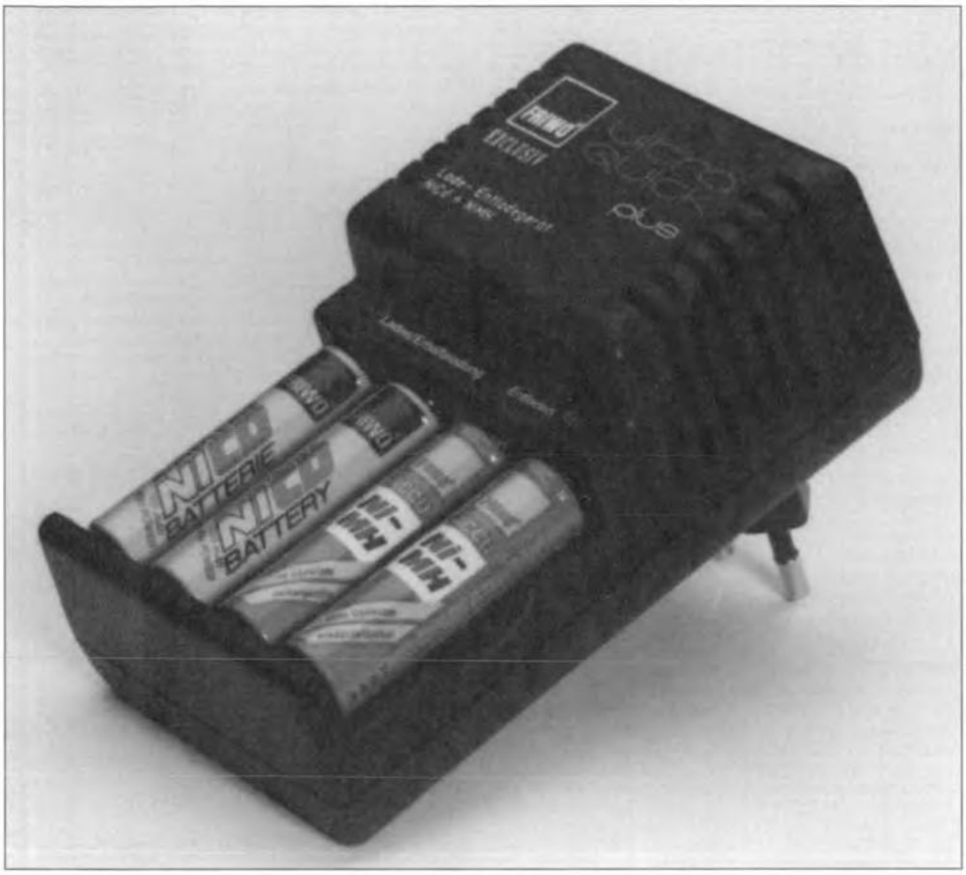


Fig. 7. Example of a multi-purpose, 1-hour charge, intelligent battery charger from Friwo. This NiCd and NIMH compatible charger features fast charging, automatic discharging, minus delta-U shutdown and an automatic charging time limiter.

ering a portable computer for 10 to 20 hours on a single charge. The energy density of this new battery type is claimed to be two to four times that of a NiMH type.

Another development which aims to improve battery management rather than battery technology is the **in-battery microchip**. The NiMH based 'smart battery' will be capable of communicating with a processor, via a bus, in order to achieve more operating hours and get the most out of the available capacity. Furthermore, the smart battery can convey information as regards its state to an (intelligent) charger, which then acts appropriately.

The main thought behind the development of the smart battery is to combat the staggering variety of battery types that has come into existence since the introduction of mobile phones and computers. A standard for batteries in mobile computers and communication equipment developed by Intel and Duracell is slowly but gradually being endorsed by a growing number of manufacturers in the respective fields. To mention a few: Phoenix Technologies (yes, the PC BIOS specialists), Maxim, VLSI Technologies and, of course, Intel. The specifications for battery control comprise a communication protocol called SDB (smart battery standard) and an SMBus which gives access to the power management system. As you can see, the future of batteries has only just begun!

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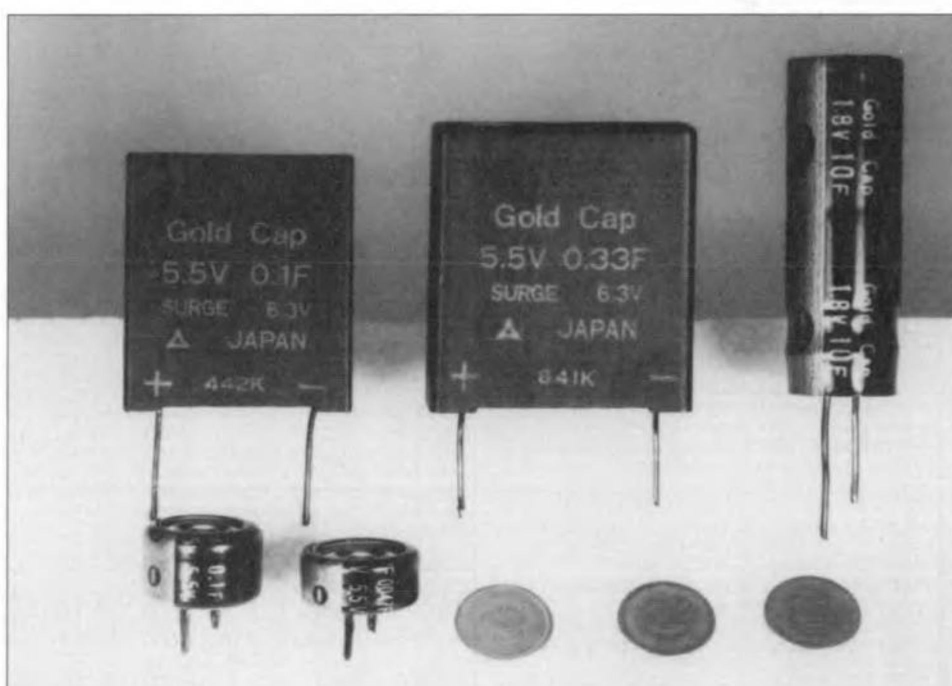


Fig. 8. GoldCaps and their successors, PowerCaps, look poised to replace batteries in a number of important applications. Their advantages are an utterly simple charge/discharge behaviour, long life (>100,000 charge cycles) and their ability to supply very high peak currents.

For further reading:
Battery reference book, by T. R. Crompton; published by Butterworths, ISBN 0-408-00790-7.

THE DIGITAL SOLUTION

Part 1 – Digits and Data

By Owen Bishop

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

The word 'digit' comes from the Latin *digitus*, meaning finger. The connection between fingers and electronics derives from the ancient (and still current) practice of *counting* on one's fingers. Counting fingers is still the easiest way to add 2 and 2. Digital electronics provides a faster and more reliable way of counting *without* using fingers. Because of the use of fingers for counting, the word 'digit' has been adopted as the name for the symbols used in counting. Most often we use ten such symbols, the digits 0 to 9. This makes possible a system of counting known as the *decimal system*, which presumably originated from our having ten digits on two hands. Any number in the decimal system may be written by using just the ten decimal digits. There is no logical reason why we can not use other systems, based on other numbers. It is just a matter of convenience. In some ways, 10 is not a convenient base, for it factorizes only into 2 and 5. People often find that a system based on 12 is more useful since it can be divided equally by 2, 3, 4 and 6. A system based on 16 is advantageous for similar reasons. For electronic circuits, the most convenient system is the *binary* system, based on the number 2. In this system we need only digit symbols, 0 and 1, to express any number.

Magnitude or state?

A decimal digit may take any one of ten distinct values. If a decimal digit is to be represented in an electronic circuit, we need to be able to distinguish ten different levels of some quantity. There might be ten different voltages, for example, the ten levels 0 V to 10 V in steps of 1 V. The *magnitude* of the voltage represents the value of the voltage. Alternatively, we could represent values by the magnitude of a current, perhaps from 0 mA to 10 mA in steps of 1 mA. Precision amplifiers are called for that can generate accurate output voltages, so why not go one step further? Let the voltage represent *any* value on the scale 0 to 10? Let 4.5 be represented by 4.5 V; let 6.731 be represented by 6.731 V. Perhaps this reaches the limits of reproducibility, but it is sufficiently precise for many applications. For values greater than 10, introduce a scaling factor. Thinking along these lines, we abandon digital electronics (temporarily) for the complementary field of *analogue* electronics. Voltages can take any *intermediate* value within a given range; they are not stepped. The magnitude of the voltage represents, or is an *analogue* of, the magnitude of some physical quantity, such as temperature, light intensity, or position. We can input a value to an analogue circuit by using a rotary potentiometer wired as a potential divider (**Fig. 1**). The voltage output of the divider is the analogue of the position of the shaft of the potentiometer or of anything mechanically attached to that shaft. The output of a temperature-sensing or light-sensing circuit is the analogue of the quantity sensed. In an audio system, the voltage waveforms are the analogue of the original local fluctuations in air pressure which we generally refer to as sound. On the output side of an analogue circuit we may read a moving-coil voltmeter, hear sound from a loudspeaker, note the changes in position of a smoothly moving actuator, or observe the variations in the speed of a motor or the brilliance of a lamp. Such are the characteristics of analogue circuitry.

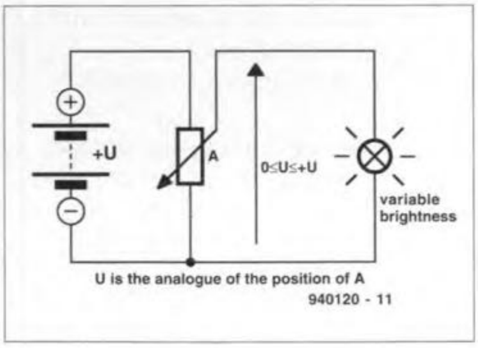


Fig. 1

The input and output sides of an analogue circuit are linked by stages which process the input and produce the output. In simple systems, such as an intercom or a motor speed control, only a few stages are involved, perhaps only one stage. More complex systems, such as analogue computers, have many processing stages. Their circuitry is based in a range of different subcircuits designed specially for handling analogue data. These include the operational amplifier, which wholly owes its existence to analogue computing. Other circuit building blocks include adders, multipliers and integrators. Most of these subcircuits are available as integrated circuits. Analogue computers are often used in industrial control systems, or in modelling complex natural system, such as the water flow in a river basin. In general, analogue systems are set up once and for all for a particular purpose. They can perform highly complex calculations, and they give almost instant results, but they tend to lack flexibility. In contrast, an average digital microcomputer is able to control an industrial process, model a wide range of systems (including modelling digital systems, as we shall see in a later part in this series), as well as compute and print out the firm's accounts, receive and send communications, provide on-line information in the form of an encyclopaedia on CD-ROM, and beat the operator at chess, virtually all at the same time. It is easy to see why the digital solution has gained ascendancy over the analogue approach. Analogue computers depend mainly on hardware; digital computers gain flexibility because they depend on software.

The key to digital systems is that data is not represented by an analogue of its *magnitude* but by the *state* of one or more circuits. For simplicity and reliability, digital systems recognize only two states. For comparison with **Fig. 1**, we have drawn **Fig. 2** to represent the digital system. Here, the input is represented by a switch, which is in one of two states, either open or closed. There is no in-between state. The output is represented by a lamp, which is in one of two states, either on or off. There is no in-between state.

High and low

When describing the action of a digital

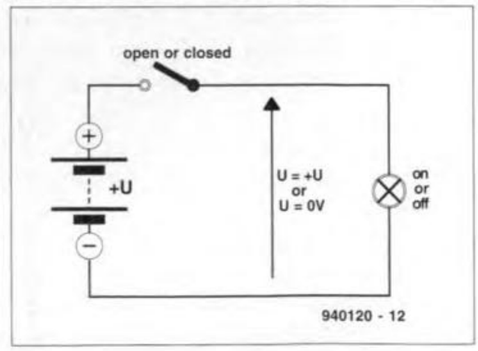


Fig. 2

circit, we can refer to the state of a transistor or an LED, but most often we refer to voltage levels. The input stages of a digital subcircuit deliver only two voltage levels, low and high, recognizable as such by the inputs of other digital subcircuits. Thus, many subcircuits may be linked to produce a complex logic circuit. In any ystem we have to specify what is meant by 'low' and what is meant by 'high'.

A low voltage is usually one close to 0 V. A high voltage is normally one close to the supply voltage. A low voltage is usually taken to represent the digit '0' in the binary system. A high voltage is normally taken to represent the digit '1'.

The voltage levels are defined differently for input and output. For example, the 7400 series of integrated circuit (which we shall look at in more detail later) operates on a 5 V supply. A low output is specified as a voltage between 0 V and 0.4 V. A low input voltage is between 0 V and 0.8 V. Thus, there is a margin of error, allowing for tolerances in manufacture. A low output from any individual circuit will certainly be read as low by any other circuit. Similarly, a high output is any voltage between 2.4 V and the maximum permissible supply voltage, 5.5 V. A high level input is anything between 2.2 V and 5.5 V. Note the gap between the maximum low input (0.8 V) and the maximum high input (2.2 V), a region that gives indeterminate results.

Processing problems

Processing analogue signals is beset with several problems. There is almost certain to be a stage at which the signal has to be amplified. If this is done by a single transistor, there arise the problems of biasing and linearity. These may lead to distortion of the original signal, and there may also be cross-over distortion and clipping. The precise shape and amplitude of the waveform are usually of critical importance in analogue circuits, and these are too easily lost. Even if the circuit is made more precise by using operational amplifiers, no opamp is ideal and there are errors caused by slew rate, input offset voltage and limited bandwidth. In a digital circuit, transistors are switched fully off or fully on (saturated) so these errors can not arise. The waveform is a square wave and its exact amplitude does not matter, provided that the levels corresponding to 0 and 1 are easily distinguishable. In **Fig. 2**, the circuit works for as long as the battery provides enough power to illuminate the lamp, even though it may only be dimly.

Noise is another unavoidable difficulty with analogue circuits. The original analogue signal will have noise superimposed on it, not only from interference by outside sources, but noise inherent in the conduction of discrete electric charges and their behaviour as they pass through semiconductor devices. As a result, a signal that originally varied smoothly (**Fig. 3a**) becomes clothed in irregular spikes (**Fig. 3b**) The same amount of noise on a digital signal is of no consequence, for it is still possible to distinguish high (1) and low (0) levels. In logic circuits, the specified high and low levels for input and output voltages (see High and low above) give immunity to relatively large amounts of noise.

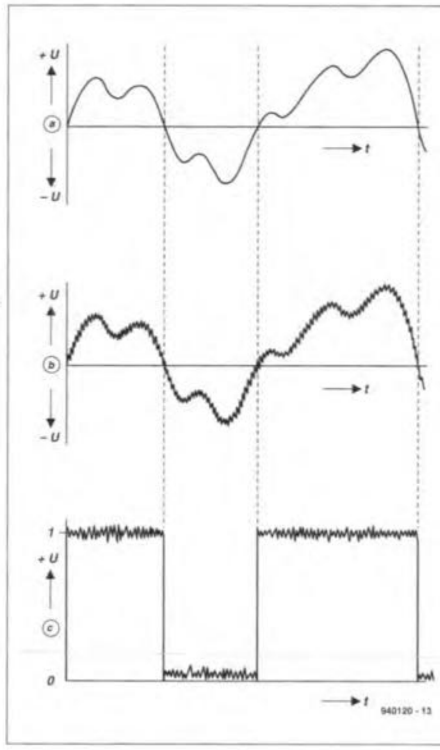


Fig. 3

Representing facts

Data may be numerical or it may consist of statement of fact. Leaving aside numerical data for the moment, let us look at how a fact may be represented digitally. Consider a simple statement, such as

A = the door is open.

This statement may be true or it may be not-true (false). We do not allow the door to be half-open. Statement A is given a value to indicate whether or not it is true. Since there are only two possible states of the door, there are only two possible values of A. If it is true, A = 1; if it is false, A = 0.

The state of the door is easily communicated to an electronic circuit, simply by placing a microswitch so that this is open when the door is open and closed when the door is closed. The circuit of **Fig. 2** uses a lamp to indicate remotely the state of the door. When the lamp is on, we know that A = 1, and the door is open; when it is off, A = 0, and the door is closed. This is, perhaps, one of the simplest possible of digital circuits, but it illustrates the way in which a fact is processed electronically.

Representing consequences

Given certain facts, other facts or actions may follow from them. We are entering the realms of logic, reasoning from one set of facts to another set. The most common line of reasoning is an argument which goes like this

IF... THEN...

If the first statement is true THEN it follows that a second statement is true. For example, IF the door is open, THEN the wind blows in. Here we have a statement Z, where Z = the wind blows in. Having defined statements A and Z, we can express their relationship in short form

IF A THEN Z.

We can introduce numerical values into this statement:

IF A = 1 THEN Z = 1,

and summarize the situation with an equation:

A = Z.

Logical statements may also include statements that are false. In everyday life, the word 'false' is associated with lies and deception, but there are no such overtones in logic or digital electronics. 'False' simply means 'not true'. A false statement is merely one that has zero value. In the case of the door, where A = Z, we have said that A = 0 means that the door is NOT open: it is closed. In this case, Z = A =0. The wind does NOT blow in. Negation is an example of a *logical operator*.

Naturally, Z does not necessarily follow from A unless A is the *only* condition for the truth of Z. It may be a windless day, or the door may be the door of an inside room. If there are several relevant conditions, these must also be true. For instance, B = it is a windy day, C = the door is an outside door, and so on. We can either take their truth for granted or, if there are reasons for doing so, write them into the statement. The earlier statement could be expanded to include all relevant information:

IF A AND B AND C AND...THEN Z

We have linked together statements A, B and C by another logical operator. This is the operator AND, which is used when two or more statements must be simultaneously true for a consequence to be true.

A third logical operator is OR, which is used when any *one* or more (but not necessarily all) statements must be true for a consequence to be true. For example, given the statement

D = the window is open,

we might deduce that

IF A OR D THEN Z.

All possible logical operations can be represented in terms of the three operators AND, OR and NOT.

This brief overview of logic serves to show how facts and their relationships can be represented by simple logical statements, all of which can be translated into a set of states of a suitable electronic circuit.

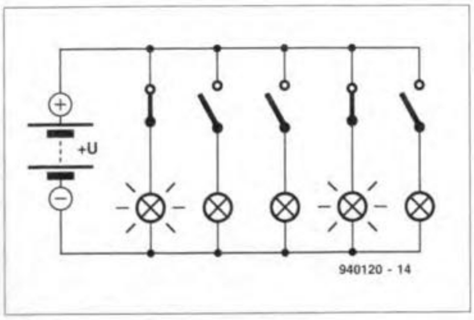


Fig. 4

Representing numbers

Any number, no matter how large or how small, is represented in the decimal system by writing out a string of one or more decimal digits. A single digit covers all integral values from zero to 9. Beyond 9 we make use of the *position* of the digit within the string to give further information about the value it represents. In the number 475, for example:

5 represents $5\times10^0 = 5\times1 =$ 5
7 represents $7\times10^1 = 7\times10 =$ 70
4 represents $4\times10^2 = 4\times100 =$ 400
total 475.

As we move to the left, each digital represents the value of that digital multiplied by an increasing power of 10.

The same applies in the binary system, except that we are restricted to only two digits, 0 and 1. For example, in the number 10110, and reading from right to the left:

0 represents $0\times2^0 = 0\times1 =$ 0
1 represents $1\times2^1 = 1\times2 =$ 2
1 represents $1\times2^2 = 1\times4 =$ 4
0 represents $0\times2^3 = 0\times8 =$ 0
1 represents $1\times2^4 = 1\times16 =$ 16
total 22

Here we are multiplying by increasing powers of 2. The table shows that 10110 in the binary system is equal to 22 in the decimal system. Unless the context makes it clear that we are dealing exclusively with one system, we specify which system a number is written in by writing a suffix after the number:

$10110_2 = 22_{10}$.

This example illustrates the technique for converting binary numbers to decimal. Usually we do not write out the lines in which the multiplicand is 0. It was done in the example to clarify the principle of the technique.

The advantage of the binary system is that it lends itself to representing numbers in electronic form. Given a row of five switches,we can represent the number 10010_2 by turning three off and two on as in **Fig. 4**. These five switches can be used to represent any 5-digit binary number from 00000 (0) to 11111 (31_{10}). We may use not only switches but any other device or circuit that can exist in one of two complementary states. For example, a transistor may be on or off, a voltage may be present or absent, a bistable (flip-flop) may be in one state or the opposite state, a capacitor may be charged or not charged, a light-emitting diode may be on or off. Each of the five lamps in **Fig. 4** may be on or off, so here is a circuit with digital input and digital output.

Other number systems

The main disadvantage of the otherwise simple binary system is that it needs more digits than other systems to represent a given number. As shown earlier, the number 22 needs only two decimal digits, but its binary equivalent needs five digits. We need more circuits (more switches, more lamps) to count in binary than in decimal. The circuits are simpler because, as we shall see, it is easier to build reliable circuits for binary than for decimal representation. But we need more of them. Or, if we deal with the digits one after another (serially), it takes longer to process them. Another practical problem is that a string of binary digits is much more difficult for the (normal) human mind to comprehend (which is presumably another reason why we adopted decimal and not binary as a counting systems in the days before electronics). For example, the number

0100111010011010

is virtually incomprehensible. Not only that but, when working on paper with long strings of 0s and 1s, errors are more likely to occur. We need a number system with a larger number base (but not 10) that is compatible with the binary system. The system most often used is the *hexadecimal* system, based on the number 16. It takes 16 different number symbols to count in sixteens. Rather than invent six new symbols to represent the decimal numbers 10 to 15, we use existing letters of the alphabet: A to F (see **Box 1**). The hexadecimal system permits large numbers to be written with relatively few digits. It is easily convertible to the binary system, because 16 is a power of 2. To see how this works, look again at the binary number given earlier, but now divide the digits into groups of four:

0100 1110 1001 1010.

Take each group to be a 4-digit hexadecimal number and write its value (Box 1) beneath it, in hexadecimal

0100 1110 1001 1010
4 E 9 A.

This gives the hexadecimal equivalent of 0100111010011010 as 4E9A. To confirm this result, see **Box 2**.

Decimal	Binary	Hexa-decimal
0	0	0
1	1	1
2	10	2
3	11	3
4	100	4
5	101	5
6	11	6
7	111	7
8	1000	8
9	1001	9
10	1010	A
11	1011	B
12	1100	C
13	1101	D
14	1110	E
15	1111	F
16	10000	10
17	10001	11
31	11111	1F
32	100000	20
64	1000000	40

Box 1

Binary codes

The binary system is used in several different ways, one of which is to express values directly. For example, the value 13 in decimal is 1101 in binary. In an analogue circuit, this value would possibly be represented by the *magnitude* of a voltage, say, 13 V or perhaps 1.3 V. In a digital circuit, this would be represented by the *state* of four replicated parts of the

Box 2. Converting binary and hexadecimal to decimal. Example: 0100 1110 1001 1010_2 = $4E9A_{16}$

Binary to decimal

$1\times2^1 = 1\times2 =$	2
$1\times2^3 = 1\times8 =$	8
$1\times2^4 = 1\times16 =$	16
$1\times2^7 = 1\times128 =$	128
$1\times2^9 = 1\times512 =$	512
$1\times2^{10} = 1\times1024 =$	1024
$1\times2^{11} = 1\times2048 =$	2048
$1\times2^{14} = 1\times16384 =$	16384
Total	20122

Hexadecimal to decimal

$A\times16^0 = A\times1 = 10_{10}\times1 =$	10
$9\times16^1 = 9\times16 =$	144
$E\times16^2 = E\times256 = 14_{10}\times256 =$	3584
$4\times16^3 = 4\times4096 =$	16384
Total	20122

circuit. Perhaps there would be four switches or four LEDs or four transistors in the states on-on-off-on, in that order. Or there could be four bistables on the states set-set-reset-set.

Another way of representing numbers is by a *code* in which the number is not directly expressed in binary form. Instead, we convert the number into its coded form by applying a fixed set of rules. One such code, which is often used when interfacing digital circuits to decimal devices such as keyboards and numeric displays, is known as *binary coded decimal*, or BCD for short. We take the digits of a decimal number and convert each of them into its four-digit binary equivalent. For example, the number 572 is converted as follows:

decimal digits	5	7	2
binary equivalents	0101	0111	0010
BCD	010101110010		

Taken as a an ordinary binary number, 010101110010_2 equals 1394_{10}. Obviously, it is essential to recognize when a number is BCD and when it is not.

There are several other codes commonly used in digital circuits, including the Gray code, the Excess-3 code, machine codes and the ASCII code. Numbers, instructions, and facts (expressed in logical form) are all processed by digital circuits in binary form. The electrical signals representing these different kinds of informations are identical in structure. They are all *digital*. The ways in which they are created, processed and interpreted form the subject of the remainder of this series.

Test yourself

1. Express as a binary number:
   a) 19_{10}
   b) 45_{10}
   c) $A7_{16}$

2. Express as a decimal number
   a) 1100_2
   b) 100111_2
   c) $B51_{16}$

3. Express as a hexadecimal number
   a) 22_{10}
   b) 1011111_2
   c) 110011010111

4. Express 682_{10} in BCD

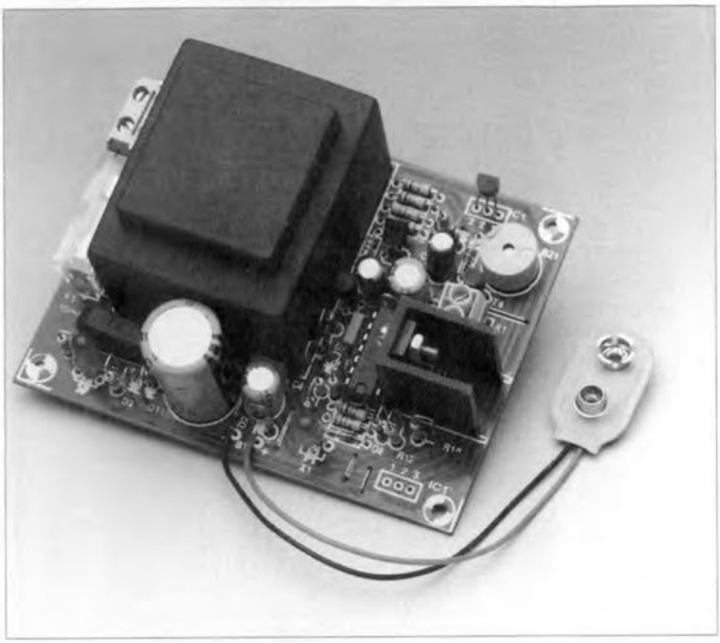
5. Express BCD 100101010001 in decimal.

[940120-I]

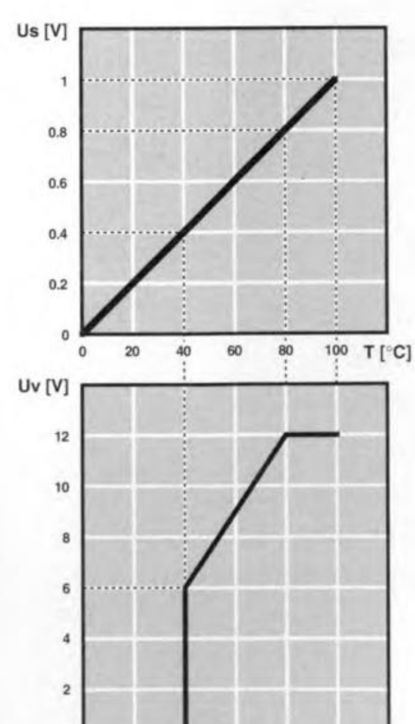
TEMPERATURE PROTECTION FOR RESISTIVE HIGH-POWER LOADS

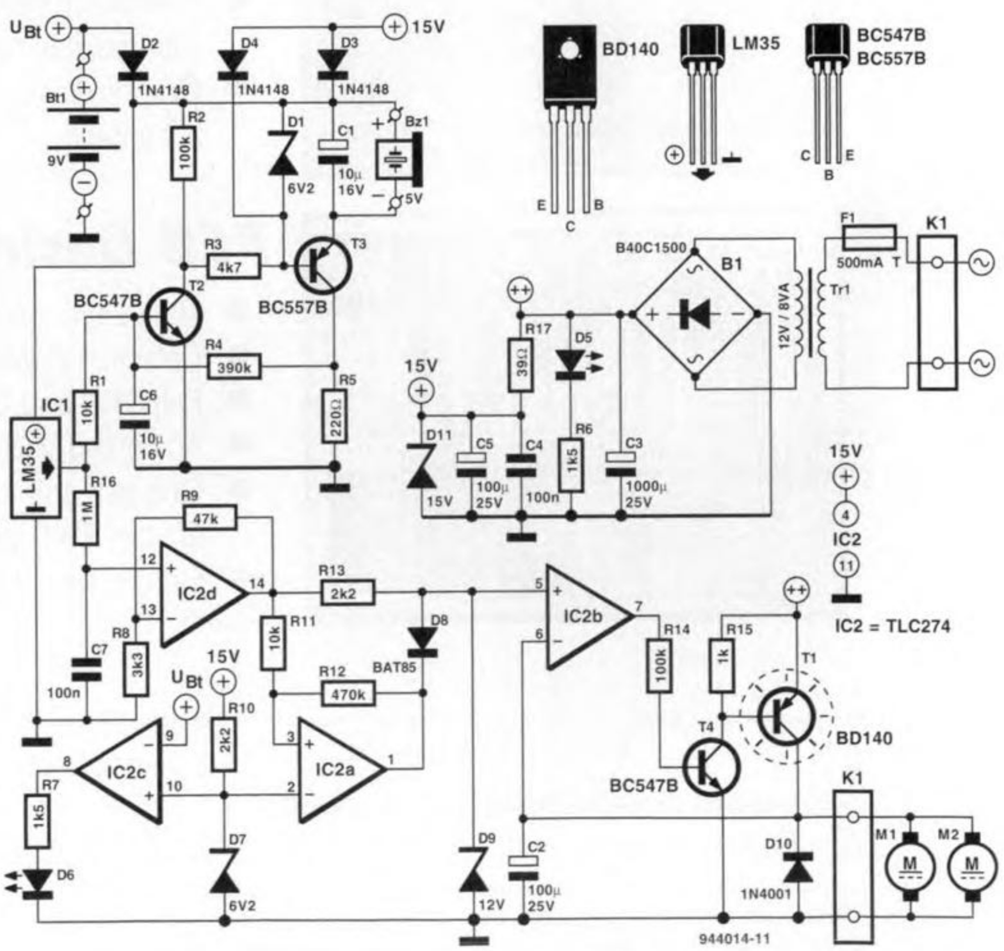
THE circuit shown in **Fig. 1** has been designed to protect (expensive) high-power resistors against overheating. Very high power resistors (of the order of 50 watts or more) are used in electronics laboratories and workshops to test audio amplifiers, power supplies and the like. To assist in their cooling, high-power resistors often come fitted on a heatsink. Since they are not cheap, it stands to reason that damage as a result of overheating must be prevented. This can be achieved at a small outlay by the present circuit, which sounds a buzzer if the heatsink temperature reaches a critical level. Provided the guard is supplied by the mains, a pair of fans is then switched on to assist in the cooling of the power resistors.

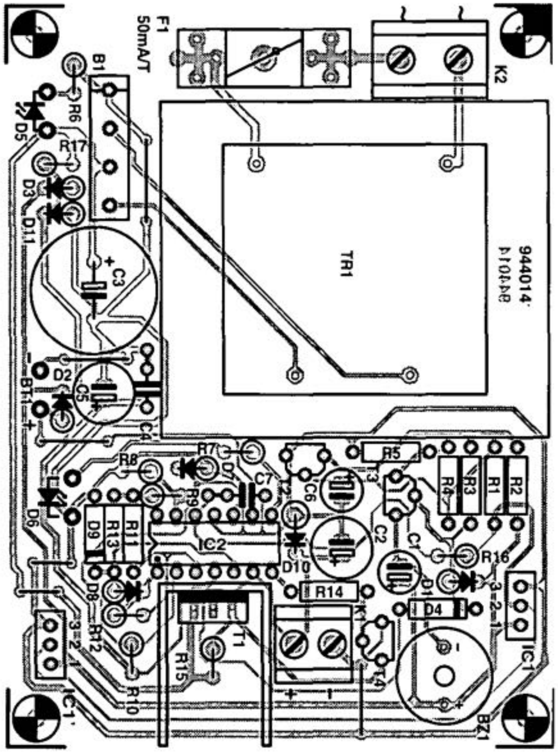
The circuit features automatic switchover between a 9-V battery and a mains supply. An LM35 semiconductor temperature sensor is used to keep the stand-by current at about 60 μA. Consequently, a 450-mAh PP3 battery can be expected to last about one year. The alarm is switched on at a sensor output voltage of about 0.5 V, which corresponds to a heatsink temperature of 50 °C. A small amount of positive feedback via R_4 ensures that the buzzer receives the full operating voltage (approx. 5 V) when the critical temperature is reached. At the same time, the feedback creates a hysteresis of about 2 °C.

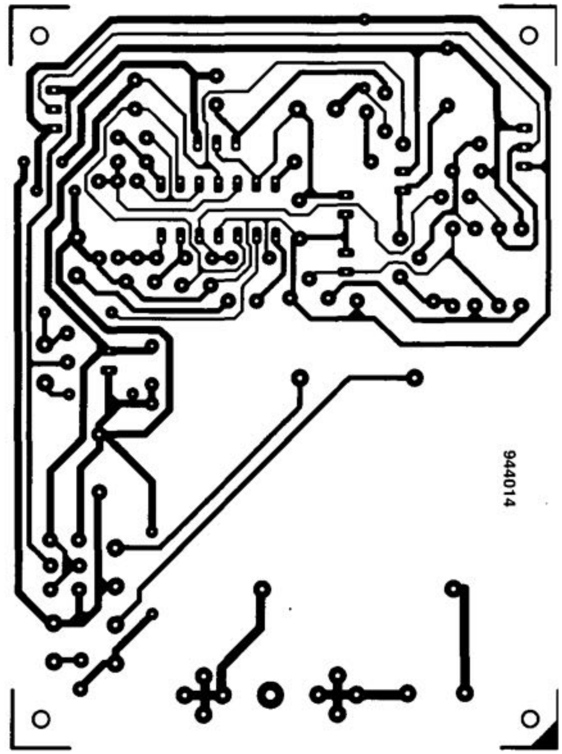


Once the mains supply is switched on, the alarm driver is disabled. The sensor voltage is amplified 15 times by opamp IC_{2d}. The resultant output voltage reaches the fan driver, IC_{2b}-T_4-T_1, only if it is greater than 6 V (minimum fan voltage). Voltages exceeding 12 V (maximum fan voltage) are limited to that value. The sensor voltage amplification is kept relatively low to ensure a smooth behaviour of the circuit. Although the resulting absolute error is fairly large, that is not a problem because the aim of the circuit is to keep the heatsink temperature below the safe limit, not to keep it constant in any way. The two graphs show the sensor voltage, U_S, and the fan voltage, U_V, as a function of the heatsink temperature.









Diode D_6 lights if the mains supply is switched on, and the battery voltage has dropped below 6 V.

The maximum current that can be supplied to the fans is about 500 mA.

As a matter of course, the board must be fitted in a manner which ensures good thermal contact between the flat side of the LM35 (IC_1) and the heatsink.

Parts list

Resistors:
R_1, R_{11} = 10 kΩ
R_2, R_{14} = 100 kΩ
R_3 = 4.7 kΩ
R_4 = 390 kΩ
R_5 = 220 Ω
R_6, R_7 = 1.5 kΩ
R_8 = 3.3 kΩ
R_9 = 47 kΩ
R_{10}, R_{13} = 2.2 kΩ
R_{12} = 470 kΩ
R_{15} = 1 kΩ
R_{16} = 1 MΩ
R_{17} = 39 Ω

Capacitors:
C_1, C_6 = 10 µF, 16 V, radial
C_3 = 1000 µF, 25 V, radial
C_4, C_7 = 100 nF
C_2, C_5 = 100 µF, 25 V, radial

Semiconductors:
B_1 = B40C1500
D_1, D_7 = 6.2 V, 500 mW
D_2–D_4 = 1N4148
D_8 = BAT85
D_9 = 12 V, 500 mW
D_{10} = 1N4001
D_{11} = 15V 1.3W
D_5, D_6 = LED, red
T_1 = BD140
T_2, T_4 = BC547B
T_3 = BC557B

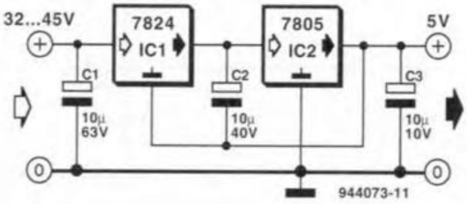
Integrated circuits:
IC_1 = LM35C
IC_2 = TLC274CN

Miscellaneous:
K_1,K_2 = 2-way PCB terminal block,pitch 7.5 mm.
BT_1 = 9 V (PP3) battery.
M_1,M_2 = miniature fan, 12 V, 200 mA d.c.
BZ_1 = 5 V d.c. buzzer
F_1 = 50 mA slow fuse
TR_1 = mains transformer 12 V, 8 VA
Heatsink for T_1, e.g. Fischer SK12 (14 K W^{-1}).

Design: H. Bonekamp)
[944014]

5 V REGULATOR ON 45 V

The Type 7805 voltage regulator cannot handle input voltages higher than 35 V. If only voltages higher than 35 V are available, it is, of course, possible to shunt the input with a zener diode. However, since prices of standard voltage regulators are low, there are advantages in using a second regulator: the stability of the output voltage as well as the hum rejection will be improved appreciably.



In the present circuit, the 7805 is preceded by a 7824. To obtain as high an input voltage rating as possible, the centre pin of the 7824 is not connected to ground, as is usual, but to the output of the 7805. The maximum input voltage of the combination is then 5 V higher than the permissible 40 V for the 7824.

The minimum input voltage must satisfy the rule that each of the ICs must have a difference between its input and output of not less than 3 V. The output of IC_1 is 24 + 5 = 29 V, which is more than enough for IC_2. Since IC_1 must also obey the rule, its input must be 3 V higher than its output, that is, 29 + 3 = 32 V.

Owing to the large difference in input voltage and output voltage, the maximum output current is somewhat restricted.

82

The 7805 has a voltage of 24 V across it , so that it dissipates 2 W when the output current is 80 mA. Without a heatsink, the trigger point of the thermal protection of this IC would then have been reached. This means that the IC must be fitted on an ample heatsink.

Design: H. Bonekamp
[944073]

I^2C REAL-TIME CLOCK

The PCF8583* is a real-time clock calendar with I^2C interface. The addition of the components shown in **Fig. 2** gives a standard application of the IC. If the components are surface-mount types, a very compact design results that can be built into, say, a mini DIN plug. A drawback is that, without battery back-up, all data and settings of the clock are lost when the supply of the PCF8583 fails.

The IC, whose internal circuit is shown in **Fig. 3**, can be read or written to at address 101000AX, where X is the read-write bit.

The IC contains a256×8 bit static RAM with an automatically increasing word address register, a 32.768 kHz oscillator, a :256 divider, a power on reset, and an I^2C bus interface.

The first eight bytes of the RAM (00–07) are used for the clock/calendar and divider functions as shown in **Fig. 3**. A full description of the content of all registers cannot be given here; the reader is referred to the data sheet*.

Figure 1 shows how the construction may be carried out. First, cut the pins of the IC to a length of about 1.5 mm. Then, fix resistors R_1 and R_2 at right angles to pins 5 and 6 at the underside of the IC with fast-drying glue, and solder them to these pins. Solder short lengths of thin (0.1 mm!) wire to the other terminals of the resist-ors (later, these are soldered to pins 2 and 5 of the mini DIN plug). Next, glue C_1, R_3, and C_2, in that order, exactly in the middle between pins 1-2 and 7-8. Using lengths of thin (0.1 mm!) wire, solder C_1 to pin 1, R_3 to pin 3, and C_2 to pin 4. Solder the other terminals of these components to pin 8 and also (later) to pins 3 and 4 of the DIN plug. Finally, solder the crystal lengthways between pins 1 and 2.

Space is made in the plug by shortening the metal housing to about 15 mm ($^5/_8$ in) and removing the rear part of the plastic casing. Glue a two-way header in the cable entry in a manner that allows a jumper to be inserted into it from outside the plug. This jumper determines the free bit of the address. Finally, solder the wires outlined before to the pins of the plug and reassemble the plug.

Design: W. Hackländer [944074]

* Philips Components.

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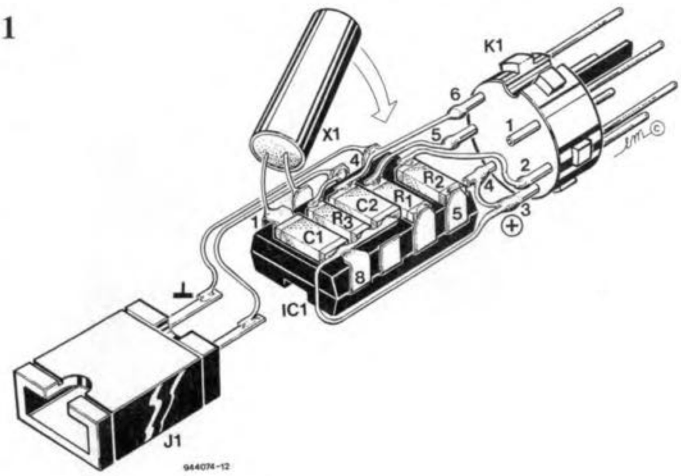

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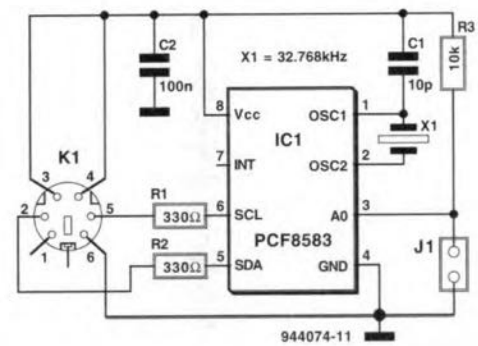

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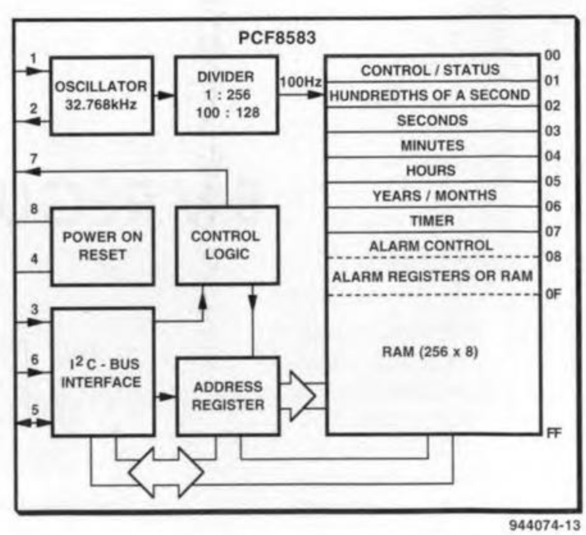
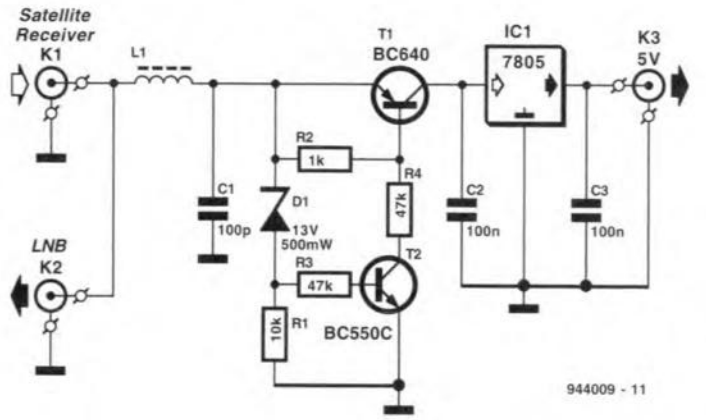


SWITCHING VOLTAGE FOR POLARIZER



Most satellite receivers provide a direct voltage at their coaxial output which, depending on the plane of polarization of the set program, is either 13/17 V or 14/18 V. This voltage is applied to the external unit via the coaxial cable to enable the polarizer to be switched over. This functions very well as long as the polarizer is integrated in the LNB. An external polarizer (between LNB and feedhorn) is switched via a separate line by a voltage that is either 0 V or 5 V. If such a unit is connected to the receiver, the LNB is powered, but the polarizer cannot be moved. To do this, a 5 V mains adaptor, a change-over switch and a separate control line to the polarizer are needed. Apart from the expense, such a setup has the disadvantage that the polarizer is manually controlled.

A more sophisticated way of solving the problem must meet the following requirements:

- A direct voltage must be available at the coaxial output of the receiver that does not affect the r.f. signal in the 950–2050 MHz band.
- The 13/17 V or 14/18 V direct voltage must be converted to 0–5 V.
- A simple adaptor circuit for insertion between the coaxial cable and LNB input must be provided.

The r.f. signal from the LNB is applied to the adaptor via F-connector K_2. This connector is linked directly to the F-socket K_1, to which the satellite receiver is connected.

The centre connector of K_1 also carries the switching voltage for the polarizer. Inductor L_1, in conjunction with decoupling capacitor C_1, ensures that the r.f. signal cannot pass to the adaptor circuit and also prevents this signal being loaded by the adaptor circuit. The switching voltage is not affected by this arrangement.

The switching voltage is applied via L_1 to a voltage detector circuit based on zener diode D_1 and T_2. The diode must be a 13 V type when the switching voltage is 13–17 V and a 15 V type when the voltage is 14/18 V.

Transistor T_2 cannot conduct until the switching voltage exceeds the zener voltage plus its base-emitter potential (0.7 V).

As long as T_2 does not conduct, T_1 is also cut off, so that voltage regulator IC_1, does not receive an input potential. The output voltage to the polarizer (K_3) is then 0 V. As soon as T_2 conducts, the switching voltage is applied to IC_1 via T_1. Consequently, the voltage at K_3 is then a steady 5 V.

The adaptor thus fulfils the requirements stated earlier: the external polarizer is switched between horizontal and vertical polarization by the (converted) switching voltage.

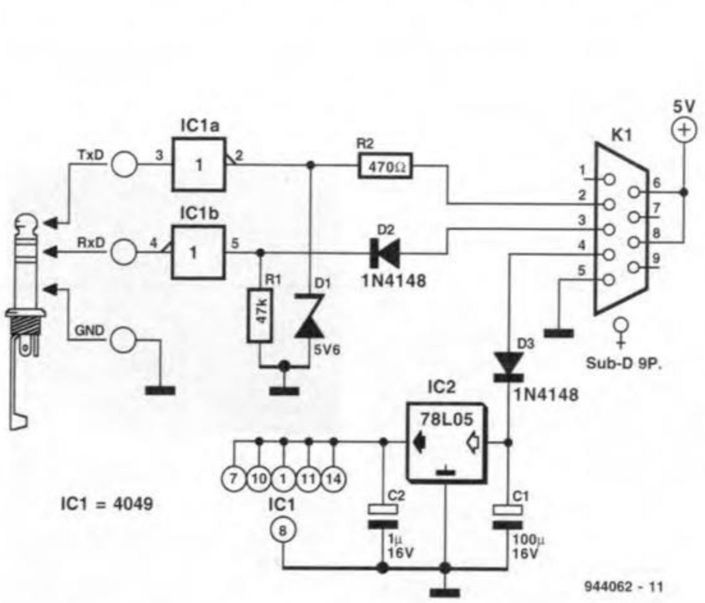
This assumes, of course, that the voltage provided by the receiver can sink enough current for the LNB and the exernal polarizer. In the prototype, the polarizer drew 80 mA and the LNB, 160 mA (at 5 V). The receiver was an Amstrad SRX200, which provides 300 mA at 13/17 V, which is ample. With a PACE receiver, which provides 250 mA, operation becomes marginal, whereas with a Hirschmann receiver, which provides 500 mA, there is nothing to worry about whatsoever.

If the adaptor is to be mounted in the open, it must, of course, be fitted in a watertight enclosure.

Design: A. van den Driesche

[944009]

PC INTERFACE FOR CASIO ORGANISER



The interface provides a link between a Casio organiser (Type SF5100, SF5300, or SF9300) and a personal computer.

In the diagram, inverters IC_{1a} and IC_{1b} buffer the TxD and RxD line respectively. Note that the interface works with TTL levels only; virtually all PCs can handle these.

Diode D_1 and R_2 protect the interface against reversal of the TxD and RxD connections. This may happen, for instance, if a wrong cable is used. The interconnecting cable is virtually proof against short-circuits by virtue of R_2.

As soon as a negative level appears on the TxD output of the PC, D_2 is cut off, while R_1 pulls the input of the inverter to ground.

The inverters are internally protected against supply voltages higher than 5 V.

The power supply may be taken from the RS232 output of the PC, since the DTR connection of the PC can sink sufficient current. Low-drop voltage regulator IC_2 converts the potential at the DTR output to a stable 5 V supply.

Design: A. Schiefen

[944062]

COMPONENTS SELECT

NPN power transistors

Available from Philips Components are two new n-p-n power transistors, the BU2522AF and BU2527AF, which, because of their short switching times, can appreciably reduce the power dissipation in the horizontal deflection of a colour monitor. Also, the tolerance of their current amplication factor is small, thanks to a carefully monitored diffusion process and 100% testing of the RBSOA (Reverse Bias Safe Operating Area).

A list of Philips Components distributors in the UK was given on p. 19 of our September 1994 issue.

Components for Elektor Electronics projects

Many components for projects published in this magazine may be obtained from Viewcom Electronics, a small but flexible organization. They also stock a very wide range of components for all sorts of application. See their advert on pages 14 and 15 of this issue.

New program simplifies filter design

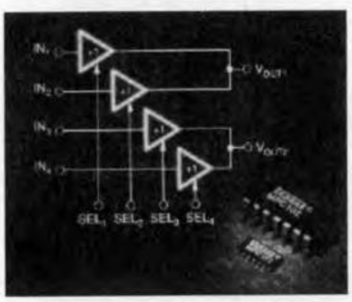
For many electronics engineers, designing active and passive filters is a tedious and error-prone process. They must either perform many repetitive and complex calculations or look up dozens of normalized coeeficients in tables. Number One's filter design program, FILTECH, gives the practising engineer quick and painless answers to filter design problems while encouraging experimentation and learning in an interactive manner. FILTECH can analyse the synthesized filter circuits independently and display a graphic plot of the calculated frequency response. The program can synthesize both active and passive filters up to sixth order with a frequency range from fractions of a hertz to over a gigahertz. Number One's advert appears on page 4 of this issue.

World's largest TFT LCD

Sharp has recently introduced the world's largest liquid crystal display. The screen has a diagonal of 53 cm (21 in) and is only 2.7 cm (just over one inch) thick. This type of display will undoubtedly find application in television sets and computer monitors. Because of its distortion-less colour image, the screen will also be eminently suitable for use in multimedia applications.

Each pixel of a TFT screen is formed by an active transistor cell. The main difficulties in producing large TFT screens are achieving a homogeneous distribution of the liquid crystal and obviating timing errors.

Dual video multiplexer



Burr-Browen has introduced a dual video multiplexer Type MPC102. This device has been designed especially for wideband systems (250 MHz, 1.4 V_{pp}) such as television and transmitting equipment. Bipolar complementary buffers provide one-way transmission with very high output-to-input isolation. The MPC102 has four identical monolithic open-loop buffer amplifier with switching elements. The differential gain and phase deviations are typically 0.02% and 0.02° respectively. Cross-talk at a frequency of 30 MHz is –68 dB, switching transients amount to +6 mV to –8 mV and the slew rate is 500 V μs^{-1}.

Printed-circuit boards

Badger Boards in Sutton Coldfield can provide PCBs for many of the projects published in this magazine, which are not available elsewhere. They can also provide one off prototype boards to individual requirements. See their advert on page 31 of this issue.

P.C.Bee of Bolton also provide a PCB service, but since their minimum charge is £120·00, this is probably of more interest to small firms than to private individuals.

Phone or fax Ian Beeby on 1204 24218.

New virtual instruments

Pico Technology have added three products to their PC-based instrument series. Like the rest of the range, the units offer the performance of desktop instruments with all the benefits that PC conmnection offers (clear displays, on-screen instructions, disk and printer support). The SLA16, probably the smallest and most economical PC-based logic analyser, offers 16-channel operation with an 8 k trace buffer. The supplied software provides state listings and waveform display.

The ADC 100 with the supplied PicoScope software allows a computer to be used as a dual channel digital storage scope, specturm analyser, frequency meter and voltmeter. Since it is powered from the PC's parallel port, it is ideal for portable use with a notebook PC.

The ADC22 is a data logging unit designed for users who require a large number of input channels (22). Its other specification are similar to the existing ADC11.

Pico Technology's ads appear on pages 8 and 30 of this issue

SPI bus compatible fast EEPROMs

After their introduction of I²C bus compatible EEPROMS, SGS Thomson has now brought out a a series of fast EEPROMSs that are compatible with Motorola's SPI bus. Type-coded ST95P02C, ST95P04C and ST95P08C, the new devices are organized in matrices of 256×8, 512×8 and 1024×8 respectively. Protection against undesired writing of data to the memory is available in hardware or software form. The memories contains, 16, 32 or 64 pages, each with a capacity of 16 bytes. One write command enables all 16 bytes of a page to be written.

Two-channel high-power optoisolator

Hewlett Packard has recently introduced the world"s smallest two-channel, high-power optoisolator. The surface mount device has the same specifications as the 8-pin DIL packaged version, but takes up only one third of its space on a board. The new family, comprising the HPCL-0530, -0630 and -0730, has been designed specially for applications where space is at a premium.

Hewlett-Packard, 308–314 Kins Road, Reading RG1 4EJ.

5 V boomer from National Semiconductor

National Semiconductor has introduced its Boomer™ audio power amplifier which offers an audio solution that doubles the output power at low supply voltage compared with conventional audio amplifier. The LM4860/1 brings high fidelity to portanle and desktop computers, cellular phones and electronic games. The new amplifiers deliver 500 mW (LM4861) or 1 W (LM4860) of continuous average power (sine wave) into an 8 Ω speaker with less than 1% total harmonic distortion plus noise (THD+N) from a 5 V supply.

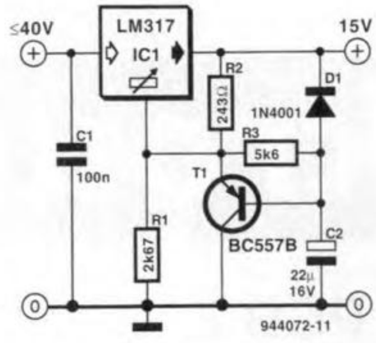
SOFT START REGULATOR

The regulator ensures slow turn on of the supply voltage. Starting at 1.25 V it takes 3 s before the full 15 V is available. The maximum input voltage may be 41.25 V.

Resistors R_1 and R_2 form part of the standard application of an LM317; they enable the output voltage, U_o, of the regulator to be set. The computation of that voltage is fairly simple (ignoring R_3, C_2, D_1 and T_1):

$$U_o = 1.25(1 + R_1/R_2) \approx 15\ \text{V}.$$

However, R_1 is now shunted by the collector-emitter junction of T_1. Since at switch-on



C_2 is uncharged, the base of T_1 is at ground potential. This means that the transistor conducts, so that R_1 is short-circuited. The output voltage is then equal to the drop across R_2, that is, 1.25 V. Capacitor C_2 is then being charged via R_2 and R_3, which causes the base voltage of T_1 to rise gradually. That is, the transistor is moving slowly to the non-conducting state and the short-circuit of R_1 is removed gradually.

If the switch-on time of 3 s is considered too slow, the value of C_2 may be reduced. If, however, a longer time is required, the value of R_3 may be increased (but not by much: how much can be found only by trial and error).

Diode D_1 ensures that C_2 is discharged rapidly via the load when the supply is switched off.

The minimum and maximum input voltages are 18 V and 41.25 V respectively. The LM317 is short-circuit proof and can sink up to 1.5 A.

Design: W. Hackländer
[944072]

SPEED CONTROL FOR D.C. MOTORS

The circuit presented is suitable for use with d.c. motors operating from 5–24 V. It can deliver an output current of up to 10 A, depending on the type of output transistor used.

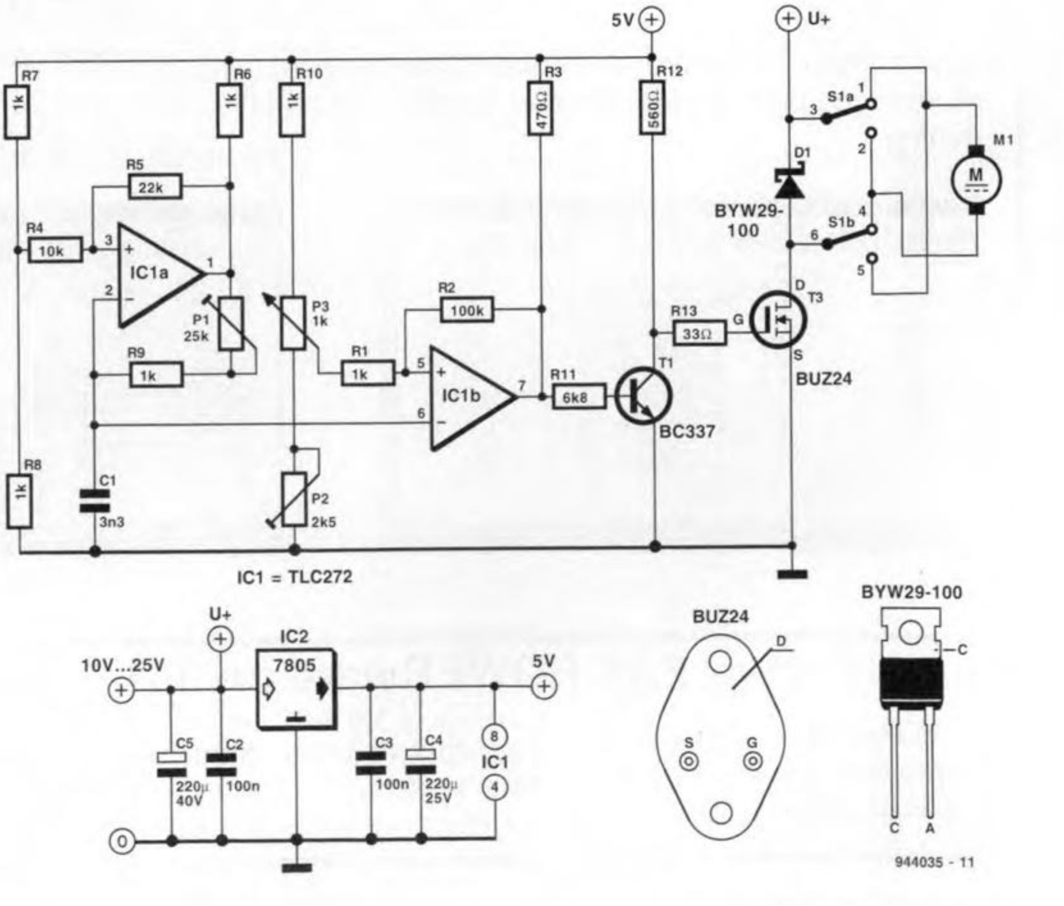
Operational amplifier IC_{1a} operates as an oscillator with a frequency that may be set between 6 kHz and 100 kHz. It provides a sawtooth voltage at a level of about 1.5 V at the inverting (–) input of comparator IC_{1b}. The non-inverting input of the comparator is connected to a reference voltage.

The comparator provides a rectangular voltage whose duty factor, and thus the speed of the motor, depends on the position of P_3.

The range over which the speed of the motor can be controlled is set with P_2, which is preset as follows. Set P_3 to minimum resistance (wiper towards P_2) and adjust P_2 so that the voltage at the wiper of P_3 is the same as the minimum value of the sawtooth voltage.

The maximum power that can be provided depends on the type of output transistor, which may be a darlington transistor or a field-effect transistor (FET). In the prototype a BUZ24 was used. This is a FET with a drain-source on-resistance ($R_{DS\text{-}on}$) of 60 mΩ and a maximum drain-source voltage (U_{DS}) of 100 V. Other suitabe types are given in the table.

The drain-source voltage needs to be high, since owing to the inductive load high counter-e.m.f.s occur on switch-on. Even the use of a varistor or a standard reverse-biased diode between drain and source does not always protect the transistor. High counter voltages are also



generated during lulls in the control signal, since the motor then behaves as a generator. In normal circumstances, the Type BYW29-100 fast recovery diode suppresses any of the generated counter voltages.

The circuit draws a quiescent current of about 25 mA.

Design: W. Zeiller

[944035]

Suitable MOSFETs

Type	U_{DS} (V)	I_D (A)	$R_{DS\text{-}on}$ (Ω)	Case
BUZ10	50	20	0.08	TO220
BUZ11	50	30	0.04	TO220
BUZ11A	50	25	0.06	TO220
BUZ14	50	39	0.04	TO3
BUZ15	50	45	0.03	TO3
BUZ18	50	37	0.03	TO220
BUZ21	100	19	0.1	TO220
BUZ23	100	10	0.2	TO3
BUZ24	100	32	0.06	TO3
BUZ25	100	19	0.21	TO3
BUZ27	100	26	0.06	TO220
BUZ71A	50	13	0.12	TO220
BUZ72	100	10	0.2	TO220
BUZ72A	100	9	0.25	TO220
BUK416 (100AE)	100	55	0.13	SOT227B

IONISATION METER

Modern central heating boilers have no pilot light but electronic ignition. Checking whether ignition has taken place can be carried out by measuring the ionisation current caused by the flame. When the ionisation current is too low, protection circuits come into action. The ignition circuit then tries to ignite the burner again. If after a few attempts the burner still does not come on, an error signal is given.

The present meter enables the ionisation current to be measured. It is capable of withstanding high ignition voltages and is suitable for measuring currents between 1 μA and 100 μA. Its control switch has an offset adjustment range and four metering ranges (0.3–3 μA; 1–10 μA; 3–30 μA; 10–100 μA) enabling it to be used with most kinds of boiler.

The current is ascertained by measuring the voltage drop across R_1. This resistor is shunted by two anti-parallel connected diodes that protect the opampt against too high input voltages. The diodes should not have too high a leakage current because of the sensitivity of the circuit.

The amplification of IC_1, depending on the setting of control switch S_1, is ×1000 (offset adjustment); ×1000, ×300, ×100 or ×10.

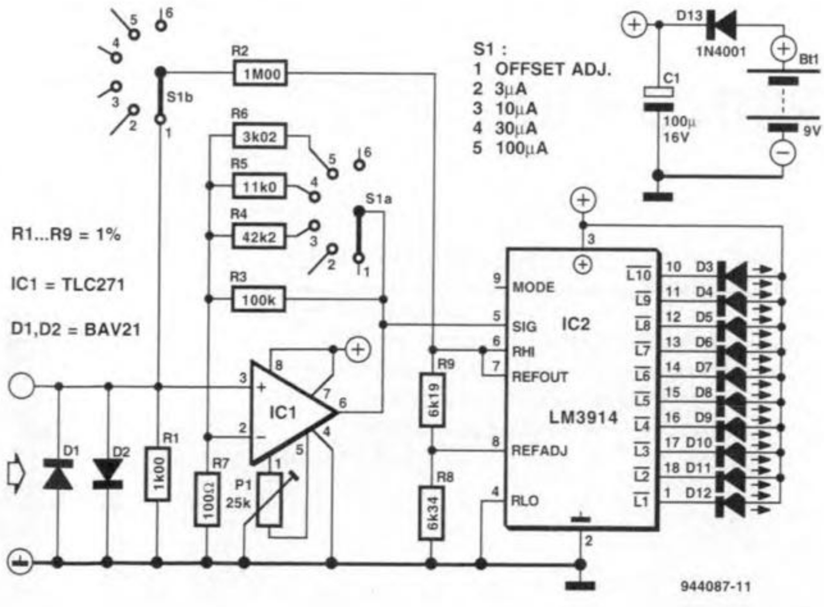
The output of IC_1 is applied to IC_2, which indicates the measured current on an LED scale. The reference voltage for this IC is set to 3 V.

Calibration of the circuit must be carried out with its input open. With S_1 in position 1 (as shown in the diagram), the reference voltage of IC_2 is linked to the input of IC_1 via R_2. Thus, a current of 3 μA flows through R_1. Then, adjust P_1 till the upper two LEDs just light.

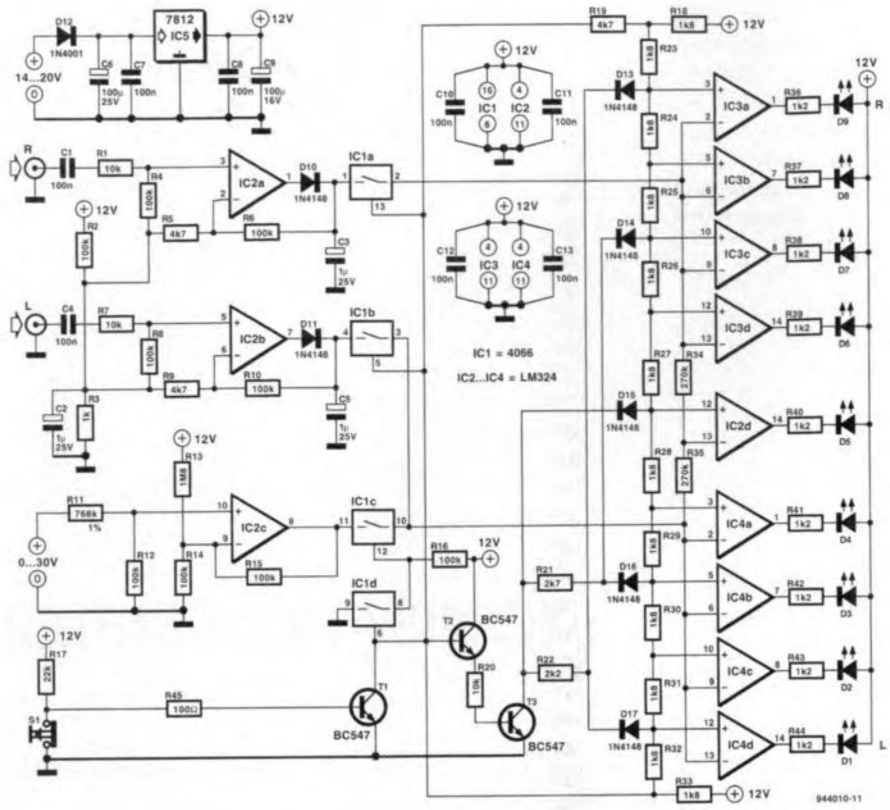
A 9 V battery suffices for the power supply, since the circuit draws a current of only 10 mA.

Design: H. Bonekamp

[944087]



DUAL-PURPOSE LED DISPLAY



The display is intended as a tuning indicator or a VU meter on a stereo FM radio. The first function is selected by pressing S_1 (which may, of course, be replaced by a normal on/off switch). Depending on the mode selected, the circuit acts as a single linear scale (for the tuning voltage), or a stereo VU meter with a quasi-logarithmic scale.

Opamps IC_{2a} and IC_{2b} act as peak detectors which charge capacitors C_3 and C_5 to a potential that depends on the level of the R(ight-hand) and L(eft-hand) audio input signals taken from the radio's line outputs. Assuming that S1 is not pressed, the peak rectified voltages are fed to the –ve inputs of the comparators contained in two LM324s, IC_3 and IC_4. Transistors T_2 and T_3 conduct, and the +input of each comparator is held at a certain reference level to create a quasi-logarithmic scale with the aid of R_{21}, R_{22} and D_{13}-D_{17}. When there is no input signal, no LED lights. At the lowest sound level, only the centre LED lights. When the volume increases, the LEDs at both sides of the centre light. Resistors R_{34} and R_{35} ensure that the centre LED responds to the average volume on the L and R channel.

The circuit is switched to tuning indicator mode by pressing S_1. Because T_1 is then switched on, T_2 and T_3 are switched off, removing the forward bias from diodes D_{13}-D_{17}. At the same time, switches IC_{1a}, IC_{1b} and IC_{1d} are opened, so that the peak rectified signal voltages are no longer fed to the LM324s. The receiver's tuning voltage (usually 0–30 V) is first attenuated about ×9 by R_{11}-R_{12}, and then amplified ×2 by IC_{2c}. The output signal of this opamp is fed to the inputs of IC_{4a}–IC_{4d} via IC_{1c}. This causes the LEDs to light from the left to the right as the tuning voltage increases. The value of R_{13} gives IC_{2c} a small off-set voltage which causes the tuning scale to start at about 1 V. Similarly, the value of R_{11} causes all LEDs in the bar to light at a tuning voltage of about 29 V. Resistors R_{11}, R_{12} and R_{13} will need to be given different values from those shown if the tuning voltage in your FM tuner differs from the usual 0-30 V range.

The current drain of the circuit is smaller than 150 mA.

Design: V. Mitrovic
[944010]

REVERSING INDICATOR

Just a few components suffice to provide a car with a visible as well as an audible indication when it is being reversed.

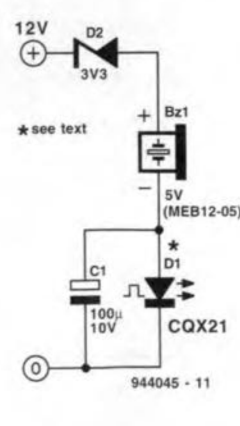
In the diagram, D_1 is a flashing LED, that is, a type which has some integral electronics that arrange the flashing. The diode is in series with a d.c. buzzer, also a type with integral electronics to arrange the production of a squeak.

When the LED lights, its current drain increases to about 10 mA, which causes the buzzer to sound. When it does not conduct, the current through it is so small that the buzzer can not operate.

The capacitor shunting the LED provides buffering of the voltage across the diode. Since the potential drop across the series combination of LED and buzzer is about 9 V, a zener diode, D_2, has been added to make the circuit suitable for operation from a 12 V car battery.

In the car, connect the circuit in parallel with the reversing light, when it will work automatically as soon as reverse gear is selected.

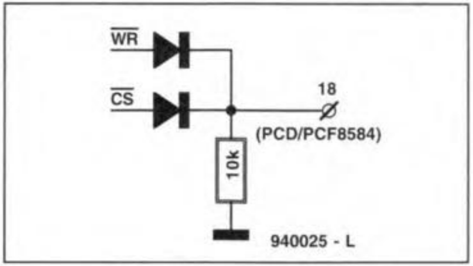
Design: L. Lemmens
[944045]



CORRECTIONS AND UPDATES

80C535 Extension card

June 1994, p. 8-11
The PCD8584 may be switched to '6800' mode if a WR signal arrives without a CS signal. The problem may be solved by combining WR and CS in a diode-AND gate as shown below. Pin 18 if IC_4 is taken out of the IC socket and connected to ground via a 10-kΩ resistor. The WR signal is found on socket pin 18, and CS on pin 10 of IC_5. Also note that the PCD8584 is currently supplied as the PC**F**8584.



Dual-purpose LED display

December 1994, p. 90
Resistor R_{33} should be connected to ground, not to +12 V as shown in the circuit diagram.

Experimentation board for PICs

July/August 1994, p. 74.
In the circuit diagram, the signals on pins 7 and 8 of both connectors K_1 should be swapped. MCLR is on pin 8, and RTCC on pin 7. The relevant PCB is all right.

Mains signalling system (2)

May 1994, p. 10-14.
The instructions for command 'T' should read: 'T' must be followed by the address in ASCII, and terminated with a semicolon (';').

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

Electronic fuse

March 1994, p. 56.
To prevent transistor T_2 from burning out when the reset switch is pressed during an overload condition, switch S_1 should be connected between the collector of T_1 and the base of T_2.

GENERAL PURPOSE FILTERS

The active filters in the two diagrams are intended for low-frequency application. Both are based on the relatively inexpensive, low-noise Type TL072.

The circuit in **Fig. 1** is a band-pass section, whose passband is shown in **Fig. 2**. The frequency is determined by R_1, R_2 and C_1. Components with the same number have the same value. Interchanging R_1 and C_1 at the input yields a high-pass filter.

The circuit in **Fig. 3** is a band-stop filter, whose characteristic is given in **Fig. 4**. Omitting R_2 parallel with C_1 at the input yields an all-pass section.

The current drawn by the circuits is about 4 mA.

If a different type of operational amplifier is used, make sure that it is frequency compensated for unity gain. In, for instance, the NE5534 that is not the case.

Design: K. Kraus
[944066]

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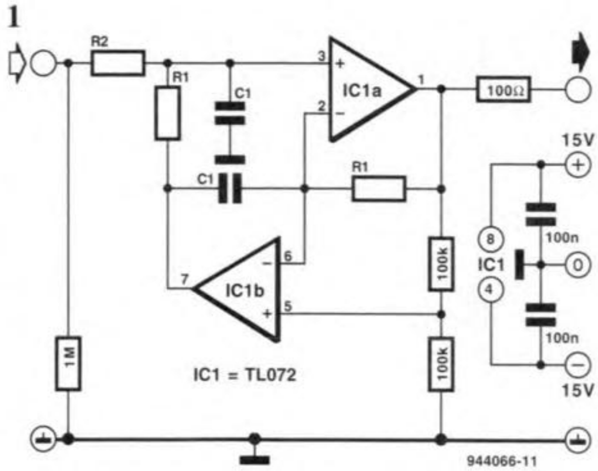

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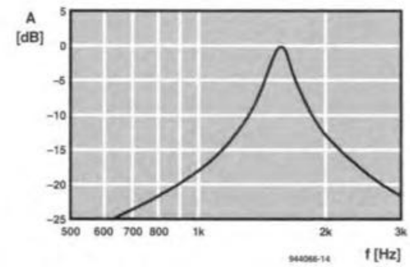

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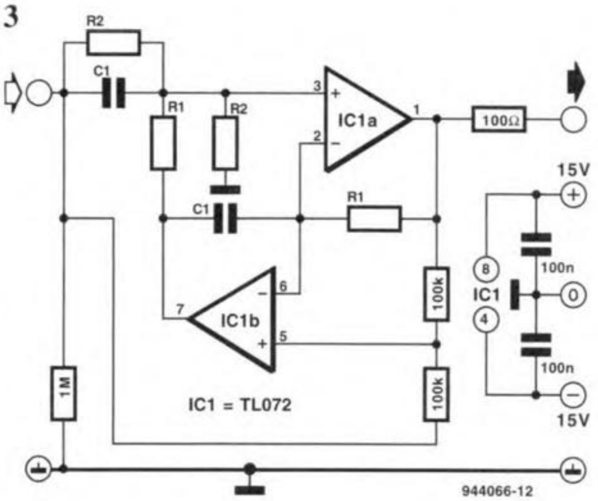

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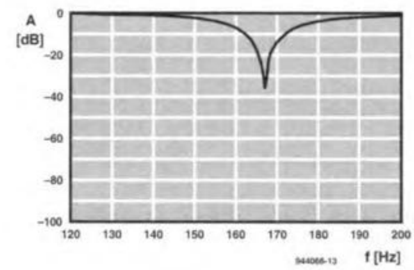


LOW-FREQUENCY TONE CONTROL

The control, which makes use of a 6-position rotary switch, provides stepped amplification or attenuation of low audio frequencies.

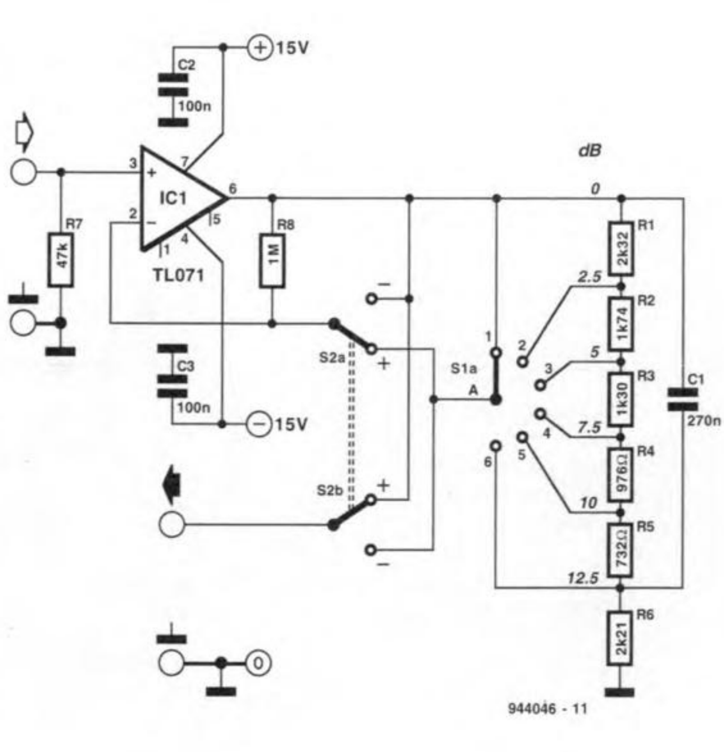
The attenuator consists of resistors R_1–R_6 and capacitor C_1. Since the capacitor is in parallel with the resistors, at high frequencies the circuit functions as a voltage follower.

Control is provided in 11 steps: −12.5 dB, −10 dB, −7.5 dB, −5 dB, −2.5 dB, 0 dB, +2.5 dB, +5 dB, +7.5 dB, +10 dB, and +12.5 dB. The degree of amplification or attenuation is selected with S_1, while the position of S_2 determines whether the circuit amplifies (+) or attenuates (−).

With values as specified, the cut-off point is roughly at 350 Hz; this may be lowered or raised to some extent by altering the value of C_1.

It is advisable to connect an a.c. coupled buffer at the output of the circuit, since at maximum amplification of the low frequencies, the d.c. offset is also amplified ×4. With a capacitive load, such as screened cable, it is recommended to insert a 100 Ω resistor in series with the output.

Design: T. Giesberts
[944046]

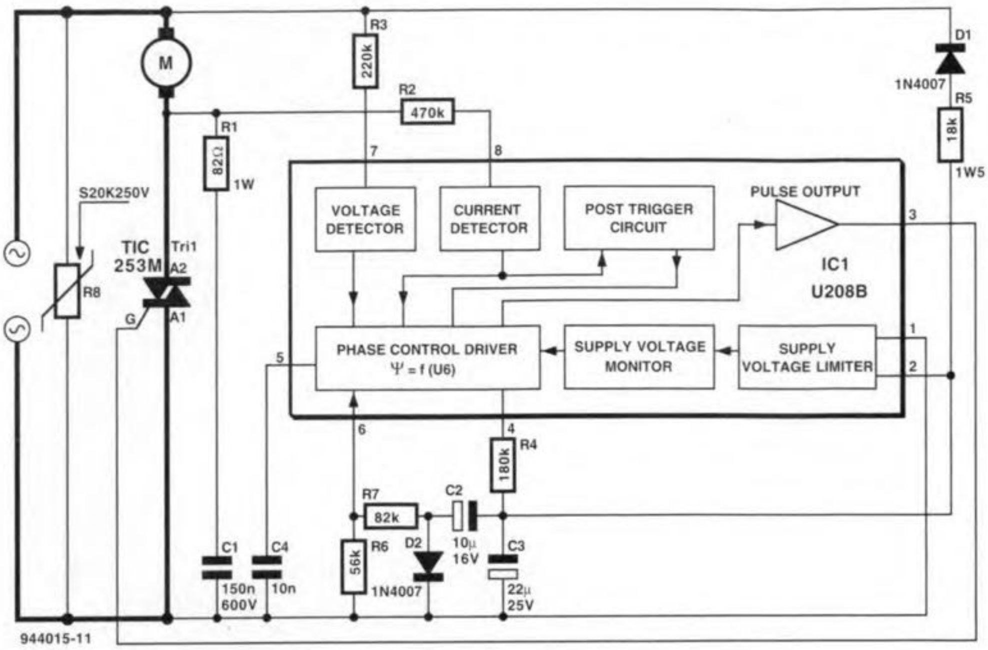


SOFT START FOR MOTORS

When medium to large electric motors are started, no counter-e.m.f. is as yet induced, so that very high currents are drawn that often result in a blown fuse (or two). The circuit described here can switch on resistive and inductive loads up to 4.5 kW in a manner which obviates this occurring.

The circuit, which is connected directly to the mains, uses phase proportional control to eliminate current peaks at switch-on. These peaks are also suppressed to some extent by varistor R_8.

The IC, a Telefunken Type U208B, internally regulates ('supply voltage limiter') the supply voltage applied via rectifier D_1, R_5 and smoothing capacitor C_3.



Current and voltage synchronization is effected by the phase control driver. This stage delivers a start pulse for triac Tri_1 via the pulse output and pin 3. The phase angle of the pulse is determined by the ramp voltage at pin 5 and the potential at pin 6. The slope of the ramp voltage is fixed at 180° by the charging current of C_4, which flows through R_4. When the potential at pin 5 is equal to that at pin 6 (reference), a start pulse with a length t of 80 μs [$t = 8\times10^{-6}\times C_4$ (nF)] is produced.

When the mains is switched on, capacitor C_2 is discharged, so that the control voltage at pin 6 has a maximum value of 7 V (via R_6 and R_7). The capacitor then begins to charge and reaches the level at pin 1 within 4 s. Consequently, the phase angle reduces from 180° to 0°, and the current flowing through the load (resistive or inductive) rises to its maximum level, in 4 s.

The current detector at pin 8 prevents a start pulse being generated while there is still a current flowing resulting from the previous half period.

The post-trigger stage prevents the triac blocking if for any reason the pulse has no effect.

The triac is protected against voltage peaks by network R_1-C_1.

When the mains is switched off, C_2 is discharged via D_2, so that at the next switch-on the phase angle is back at 180°.

A few hints for constructors. When a printed-circuit board is designed, the distances between C_2, pin 5 and pin 1 must be kept as short as feasible. The load current must in no circumstances be allowed to flow via the connection to pin 1.

Capacitor C_4 should have a temperature coefficient that is as small as possible.

The triac must be fitted on a heatsink whose capacity depends on the load: 5 K W^{-1} for a 4.5 kW motor and 12 K W^{-1} for a 2 kW motor.

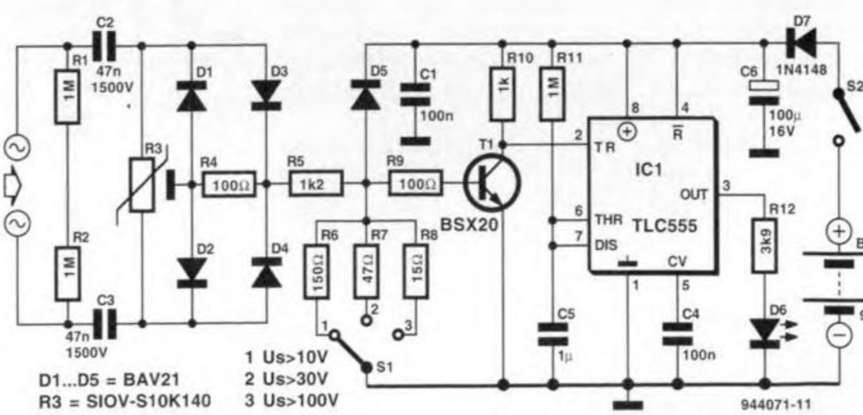
Design: J. Kircher [944015]

SPIKE DETECTOR

The domestic mains voltage, even at the best of times, does not have a clean waveform. For all sorts of reason, there are almost always spikes superimposed on the voltage and these adversely affect consumers' equipment. These spikes normally last for between 0.1 μs and 10 μs. Their amplitude varies considerably, but can be as high as 1000 V.

It is, of course, often very useful to know the duration and amplitude of the spikes, so that equipment may be protected against them. The present circuit measures spikes on the mains supply and indicates their relative level by an LED.

With a rotary switch, S_1, the user can choose one of three



different voltage levels by which the level of the mains voltage differs from the nominal level: 10 V, 30 V, or 100 V.

A bridge rectifier, D_1–D_4, is connected across the mains via C_2 and C_3. The bridge is loaded by a 100 Ω resistor, R_4. The capacitors and R_4 form a sort of high-pass filter, which attenuates the mains voltage appreciably; the drop across R_4 is a mere 230 mV_{pp}. Because of its steep edges, any spike will pass through C_2 and C_3 without any loss. This voltage is applied to a potential divider formed by R_5 and R_6, R_7 or R_8, depending on the position of S_1. If the voltage at the junction of the divider exceeds about 0.6 V, T_1 will conduct. This triggers monostable multivibrator IC_1, which causes D_6 to light. This diode is a high-efficiency type because IC_1 can not provide more than a small current.

Diode D_5 ensures that the peak at the junction of the divider does not exceed the potential across R_4 plus 0.6 V, so that the transistor can not be damaged.

Capacitors C_2 and C_3 are followed by a varistor, R_3, which protects the diodes in the bridge rectifier against too high voltage levels (the full level of any voltage peaks is across the non-conducting diodes). Resistors R_1 and R_2 ensure that the capacitors are discharged rapidly when the circuit is disconnected from the mains.

Since 1500 V capacitors are used at the input of the circuit, it may be considered electrically isolated from the mains supply. In spite of this, it is advisable to fit the capacitors in a plastic case.

The circuit can be powered by a 9 V battery; it draws a current of only 160 μA (LED out) or 2 mA (LED on).

Design: H. Bonekamp

[944071]

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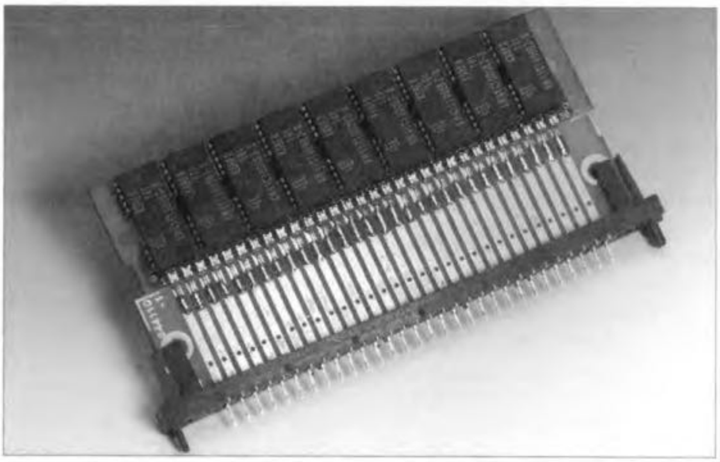
Design: H. Bonekamp
[944071]

SIP-TO-SIM ADAPTOR BOARD



For some reason or other, SIM type memory modules have become more popular than SIP types. The difference between these is the way they are inserted into the memory expansion sockets on the PC motherboard: a SIP module has wire pins, while a SIM module has contact fingers etched on a printed circuit board which carries the memory ICs. Although in first-generation PC-ATs both types of memory module can be used, and, indeed, mixed freely, an increasing number of PCs, particularly the newer 386/486 machines, allow SIM modules to be used only.

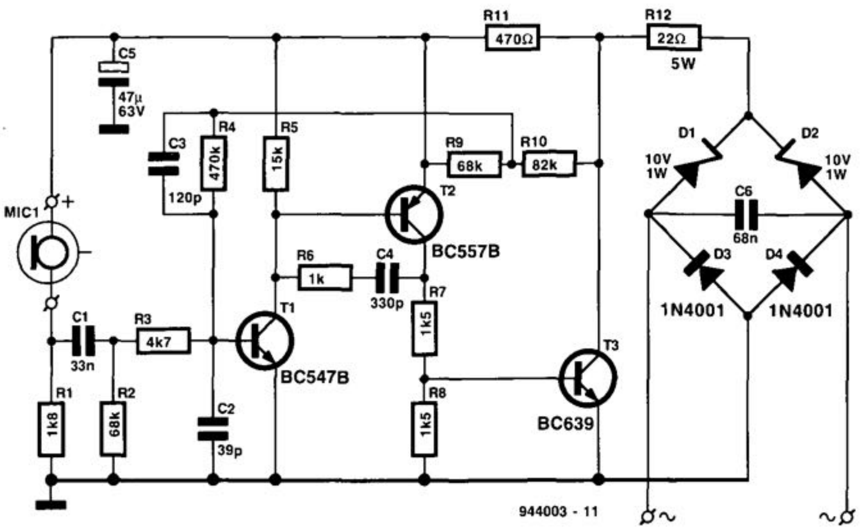
The printed circuit board shown here is for those of you who have SIP modules lying about which can not be used because the computer accepts SIMs only. The board is cut into eight for an equal number of adaptor boards. The SIP pins are carefully soldered to the copper tracks on the adaptor board. In some cases, it will be necessary to reduce the height of the PCB a little so that it does not get in the way of any hardware above it.

For proper contact with the PC's memory expansion slot contacts it is recommended to use a tin or gold layer on the copper tracks.

The printed circuit board shown here is not available ready made. If you wish to make your own board, use 1.2 mm thick material, not the more usual 1.6 mm, since that is too thick for the present application.

Design: P. Verhoosel
[944110]

ELECTRET MICROPHONE FOR TELEPHONE MOUTHPIECE



Although many older telephone sets are electrically and mechanically perfectly sound units, their speech quality is poor compared with that of modern, all-electronic, sets (except the cheap types used in domestic intercoms). The reason for this deficiency is the carbon microphone in the mouthpiece. Here, an up-to-date replacement is discussed for the carbon microphone. It takes the form of an electret microphone and an amplifier with a special band-pass characteristic.

The circuit diagram shows a conventional three-stage direct-coupled transistor amplifier whose output signal is superimposed on the supply voltage. In this way, the amplifier is fully compatible (electrically, that is) with a carbon microphone. Only the sound is much better.

Since an electret microphone has a virtually straight frequency response, the function of pass-band shaping is transferred to the amplifier. Here, the circuit is laid out to give a frequency response suitable for telephony, i.e., about 500 Hz to 4.2 kHz. The microphone signal is first sent through a high-pass filter, C_1-R_2. The high-frequency roll-off is achieved with the aid of capacitor C_3 and resistor R_4 in the feedback circuit between T_2 and T_1. Capacitors C_2 and C_5 serve to suppress r.f. signals which may be picked up by the telephone line, the receiver cord, or the electret microphone. R_6 and C_4 improve the amplifier's stability.

The d.c. behaviour of the amplifier is such that it behaves like a carbon microphone, i.e., as a non-linear resistance. Diodes D_1-D_4 at the amplifier output form a full-wave rectifier which provides an amplifier supply voltage which is sufficiently independent from the telephone line current (which can vary between 15 mA and 150 mA depending on the telephone system, line length, and other factors). Also, the rectifier ensures the correct supply polarity in all cases. For the audio signal, the rectifier is simply not there since the diodes conduct as a result of the line current which flows when the receiver is lifted. The two zener diodes, D_1 and D_2, are included as protective devices — they behave like ordinary diodes as long as the voltage on the line terminals remains below the zener voltage. If a higher voltage occurs on the line, the zeners still conduct, keeping the amplifier supply voltage within safe limits with the aid of resistor R_{12}.

The amplifier is built on the board shown, so that it can actually replace the carbon microphone, which is carefully removed from the mouthpiece. Since many different types of telephone exist, the best way of doing this will have to be figured out carefully. In most cases, it will be necessary to solder wires to the spring terminals provided for the original carbon transmitter. The electret microphone is secured at the solder side of the board, and connected with short wires to the copper tracks that form the amplifier inputs. After trimming it to size, the completed board is mounted upside down into the mouthpiece, and glued into place. The solder side of the board should be sprayed with protective lacquer, or covered with a potting compound to protect it against the heavily corrosive effect of breath. In some cases, you may also use the thin disc originally used to cover the carbon microphone. Every care should be taken to ensure that the amplifier and the electret microphone are securely mounted in the mouthpiece. If they are not, lifting the receiver and moving it about will cause noise, which defeats the use of the circuit because mechanical noise is an inherent disadvantage of the old carbon microphone!

Parts list

Resistors:

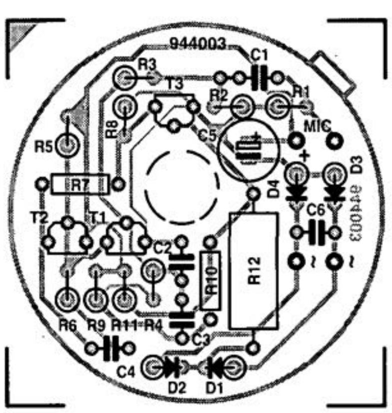
R_1 = 1.8 kΩ
R_2;R_9 = 68 kΩ
R_3 = 4.7 kΩ
R_4 = 470 kΩ
R_5 = 15 kΩ
R6 = 1kΩ
R_7;R_8 = 1.5 kΩ
R_{10} = 82 kΩ
R_{11} = 470 Ω
R_{12} = 22 Ω, 5 W

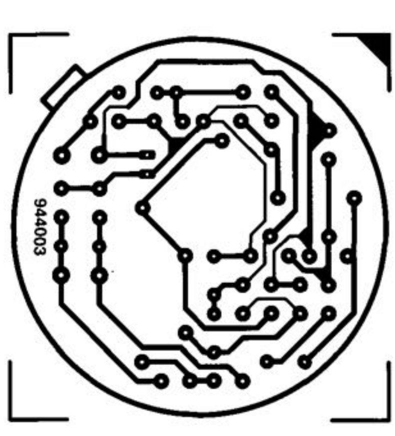
Capacitors:

C_1 = 33 nF
C_2 = 39 pF
C_3 = 120 pF
C_4 = 330 pF
C_5 = 47 µF, 63 V, radial
C_6 = 68 nF, pitch 5 mm

Semiconductors:

D_1;D_2 = 10 V, 1 W zener-diode
D_3;D_4 = 1N4001
T_1 = BC547B
T_2 = BC557B
T_3 = BC639





Miscellaneous:

Mic_1 = CM105-8 electret microphone (dia. 10 mm; Z_0 = 2 kΩ)

Design: F. Hueber
[944003]

PRESENTATION RUNNING LIGHTS

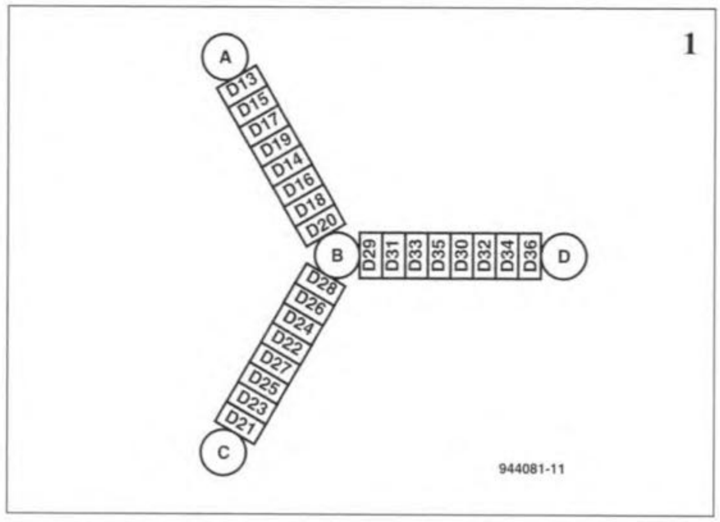
The running lights are particularly suitable to demonstrate in an original manner how two data streams arrive alternately at the same point. The display consists of a total of 24 LEDs, divided over three legs as shown in **Fig. 1**. Other patterns are, of course, also possible.

The design is aimed at giving an alternating indication of routes A-B-D and C-B-D. This means that leg B-D is on all the time and that legs A-B and C-B must be separated by an OR function.

In **Fig. 2**, scaler IC_2 is arranged as an eight-counter by linking its Q_8 output to the reset input. The time base for the counter is formed by IC_1, which has been connected as an astable multivibrator.

The LEDs are driven by the Q outputs of IC_2 via transistors T_1–T_4. The anodes of diodes D_{29}–D_{36} are linked to the +ve supply line via series resistors R_{22}–R_{25}. These diodes thus light every time IC_2 generates the right condition (relevant Q output high).

The series resistors of the other diodes are not linked to the +ve supply line directly but via T_5–T_7 or T_6–T_8 respectively. These transistor pairs thus form the OR function mentioned earlier. When one of outputs Q_4–Q_7 is high, T_5 and T_7 conduct so that one of LED pairs D_{13}–D_{20} lights. The collector of T_5 is low, so that T_6 and T_8 remain off. When Q_4–Q_7 are all low, T_5 does not conduct. Its collector is then high, whereupon T_6 and T_8 conduct and one of the LED pairs D_{21}–D_{28} lights.



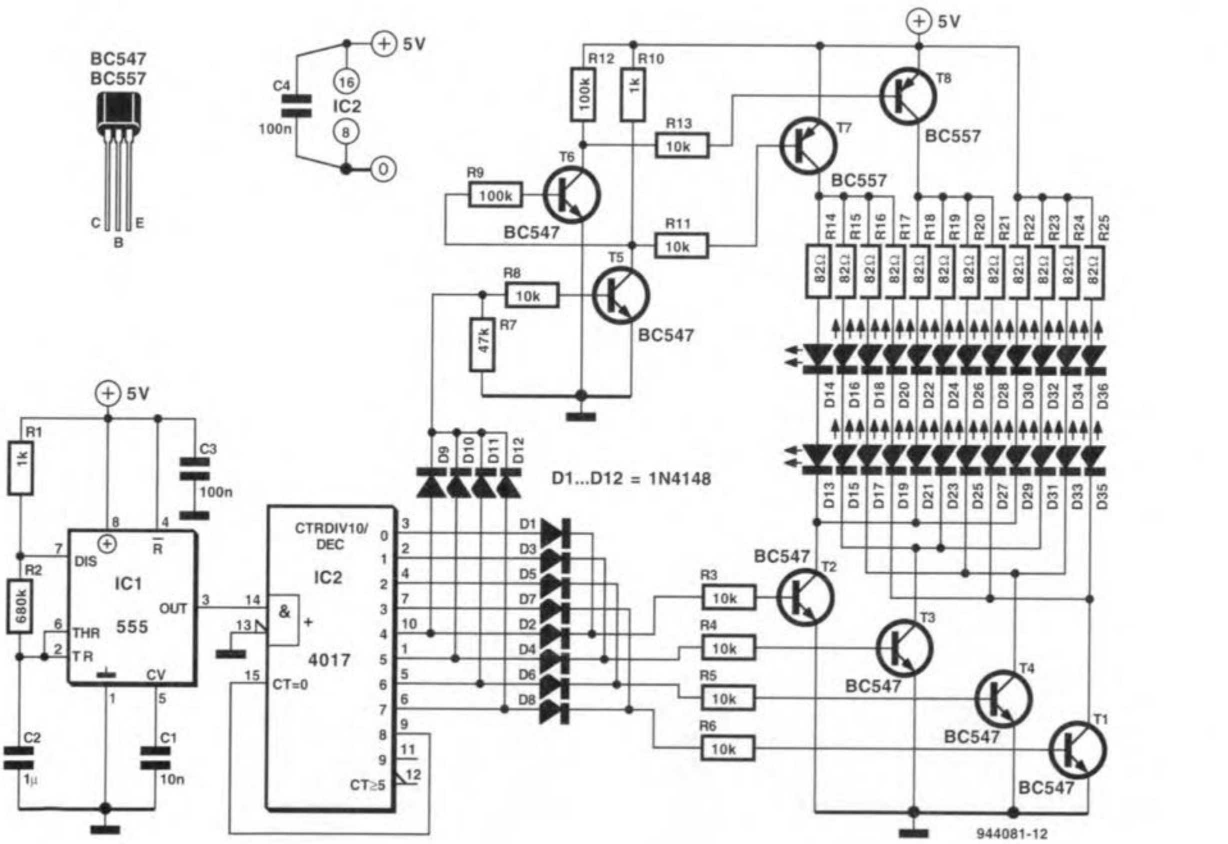
1

The number of LEDs may, of course, be increased by adding one or two to the twelve series connections. This means, however, that the supply voltage and the value of the series resistors must be changed accordingly.

The current drawn by the running lights is determined primarily by the LEDs, but does not exceed 50 mA.

Design: H. Bonekamp

[944081]



2

H-FIELD SENSOR

As earlier reported in this magazine[1], from 1 January 1996 all (new) electronic and electrical apparatus must meet the European standard for electr-magnetic compatibility (EMC). For home-constructed equipment, the sensor described here will be found useful, yet simple to make. It is in fact a small loop antenna, made from 50 Ω coaxial cable, that functions as an *H*-field (magnetic field) sensor. Although it is simple, it is capable of detecting a magnetic field caused by, for instance, a printed circuit board.

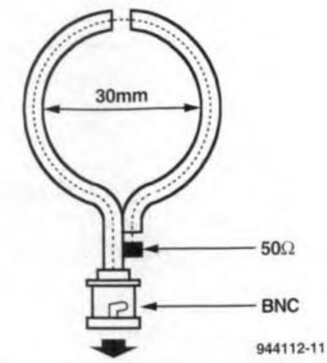
The *H*-field induces a small voltage in the screen of the cable, which is applied via a BNC connector to an oscilloscope (a spectrum analyser is better, but not many home constructors have one). Because of its symmetry, the sensor is not sensitive to the *E*(lectric)-field, which makes it possible to view at once what the consequences are on the spurious radiation of, in this case, a modification of the PCB.

However, to comply with legal requirements, an absolute measurement of the electric field at a distance of 3 m from the PCB (in this case) is also required. This is difficult without special equipment, but it can be computed from

$$E = 2.7\times10^{-17}\times f^2\times H,$$

where *E* is the electric-field strength in V m^{-1} at a distance of 3 m; *f* is the frequency in Hz; and *H* is the magnetic-field strength in A m^{-1} at a distance of 0.1 m. The formula can be used for frequencies of 16–477 MHz. In the case discussed, it should also be noted that the measurement is relevant only if the current loops on the board are much smaller than the wavelength. This will normally be so, unless ultra-high frequencies are used.

The sensor consists of a length of 50 Ω coaxial cable formed into a loop with a diameter of 300 mm. At the ends, solder the screens together and solder a 50 Ω surface-mount resistor to the junction. At the centre of the circumference, break the screen as shown. Take care not to damage the inner conductor.



Design: K.M. Walraven
[944112]

[1] May, June, 1993.

DISCRETE A-D CONVERTER FOR 8051

If an analogue-to-digital (A-D) converter is required for a particular application of an 8051 microcontroller design, this may be built from discrete components. The one described here has a maximum measuring error of 1% and an input voltage range of 0–2 V.

The important component in the diagram is capacitor C_1. This is discharged very rapidly by T_1 when the port connected to the gate of the MOSFET goes low. The collector of T_2 then carries the full supply voltage. When the level at port P*.* rises to 1 again, T_1 is inhibited.

Stages IC_{1a} and T_2 form a constant-current source, driven by the input voltage, which charges C_1 when T_1 is inhibited. The level of the potential across C_1 is compared with a reference voltage of 2.5 V, held steady by D_1, by IC_{1b}.

When the potential across C_1 drops below 2.5 V (with respect to earth), the comparator output changes state (from high to low) and the controller receives an interrupt. The interval between the leading edge at the P*.* port and the interrupt is inversely proportional with the applied input voltage. The interval is measured by the internal timer of the controller. This method is not very precise, because the processor does not react instantly to the interrupt. This is, however, not a problem when the interval is (relatively) long.

The input voltage, U_i, is:

$$U_i = 2K / N \quad (V),$$

where N is the counter status and K is a constant determined by a calibration measurement.

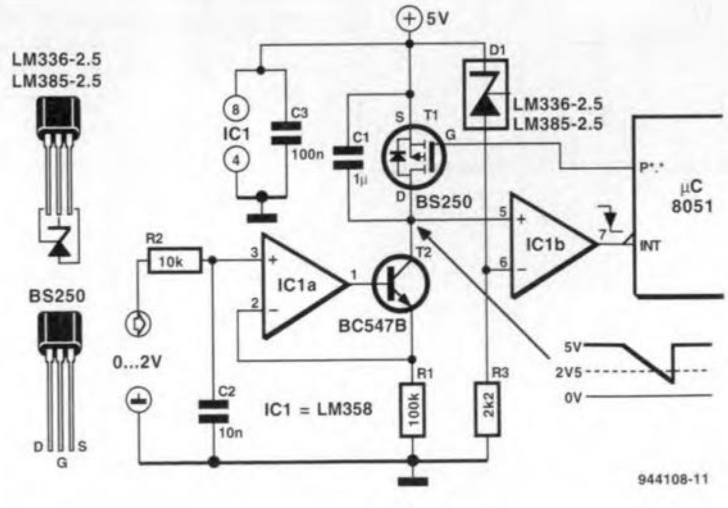
A rather more precise method is programming the software counter, but the drawback of this is that the processor cannot fulfil other functions during the measurement.

Another method is using the gate signal that controls the internal timers. This enables a fairly accurate assessment of the width of the input pulse without the CPU losing time.

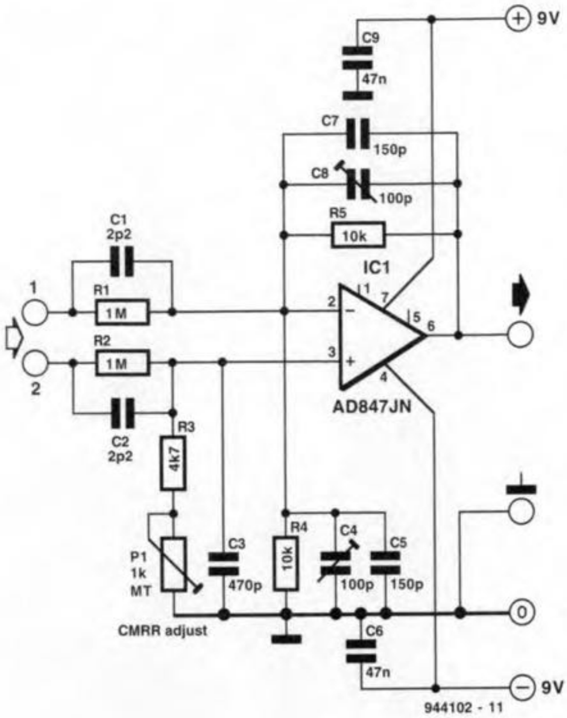
The values of R_1 and C_1 are chosen to give a conversion time of 125 ms with an input voltage of 2 V. For shorter conversion times, the value of either C_1 or R_1 must be lowered, but not by more than a factor 10, otherwise the opamp does not react sufficiently fast any more.

The converter draws a current of only a few milliamperes.

Design: G. Kleine
[944108]



DIFFERENTIAL PROBE



Normally, the standard probe supplied with an oscilloscope can be used only for measurements with respect to earth. This, occasionally frustrating, limitation can be nullified by a differential probe of which both prods 'float' with respect to earth. An accurate differential amplifier ensures that only the potential difference between the prods is registered.

The design also makes it possible to measure high voltages in a safe manner. The present probe can handle voltages up to 700 V_{pp}, provided, of course, that R_1, R_2, C_1, and C_2 are rated appropriately. It has a bandwidth of 1 MHz and a common-mode rejection that varies from 80 dB (direct voltage) to 20 dB at 1 MHz.

Since, owing to the type of operational amplifier used, the amplification must be ≥1, the signal at input 1 is attenuated a hundredfold by R_1-R_4 and then amplified ×1 in IC_1 (R_5/R_4). Frequency compensation is provided by C_8 in the feedback loop of the amplifier. Note that C_4 does not influence the compensation, because the inverting input of IC_1 forms a virtual earth for the signal.

The signal at input 2 is attenuated two-hundredfold by R_2-R_3-P_1 and then amplified ×2 in $IC_1(1+R_5/R_4)$. Frequency compensation for the attenuator is provided by C_3 and for the amplifier by C_4 and C_5; this effectively counters the effect of R_5-C_7-C_8.

To calibrate the probe, connect an oscilloscope or digital millivoltmeter to its output and note the offset. Next, short-circuit the two inputs and apply a direct or low-frequency voltage between the inputs and earth and adjust P_1 to obtain an offset that is equal to the one noted earlier. Then, remove the short from the inputs and apply a 1 kHz rectangular signal between input 1 and earth, and adjust C_8 to obtain as faithful a rectangular waveform as possible on the oscilloscope. Finally, apply the rectangular signal between input 2 and earth and adjust C_4 in an identical manner.

The probe draws a current of about 5 mA.

Design: H. Bonekamp
[944102]

FAX-PC INTERCONNECTION

The interconnection is of use to readers who have a fax machine and a PC with fax card and relevant software. The fax machine can then be used as printer and scanner for the PC. With older fax machines the resolution will be 200·100 d.p.i.; with modern ones 200·200 d.p.i.

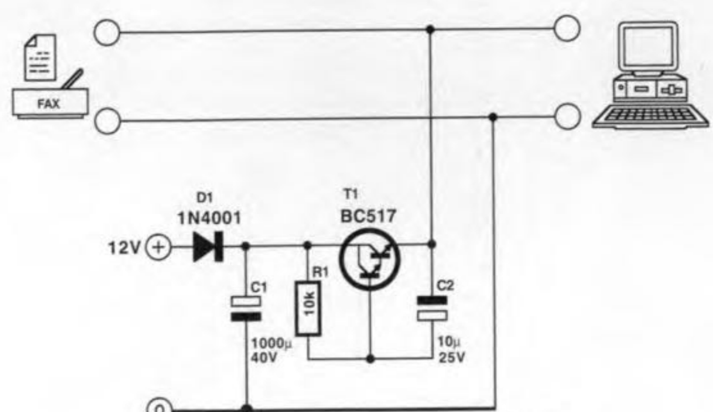
When a sheet of paper with text or graphics is inserted into the fax, the digitized image appears on the computer screen, which can then be stored on the computer disk. Conversely, text or graphics on the computer screen can be printed by the fax machine. An advantage is that compatibility questions do not arise.

There is, however, a drawback to this setup: unless a house telephone or two public telephone lines are available, there is no other communication between the PC and the fax machine.

Fortunately, the two units may be coupled via a two-wire connection, to which a 12 V supply (mains adaptor) and a current source (T_1) are added—see diagram. Each of the units draws a current of about 40 mA, although in some equipment this may be quite different. Polarity reversal of the 12 V supply is prevented by D_1.

It should be noted that the two-wire connection does not provide calling facilities. It is thus necessary that both the fax machine and the fax card in the PC can be started manually.

Design: K.M. Walraven
[944086]



CENTRONICS INPUT

Most computer users know that their computer has a parallel printer port, often colloquially called 'Centronics port'. Not many of them know, however, that this port offers more facilities than just driving a printer.

This articles describes how, with the aid of a small auxiliary circuit and auxiliary program in Pascal, 8-bit wide data can be written via this port. The circuit makes use of the fact that, apart from the eight data outputs, every printer port also has several control lines that are used to coordinate the exchange of data with the printer.

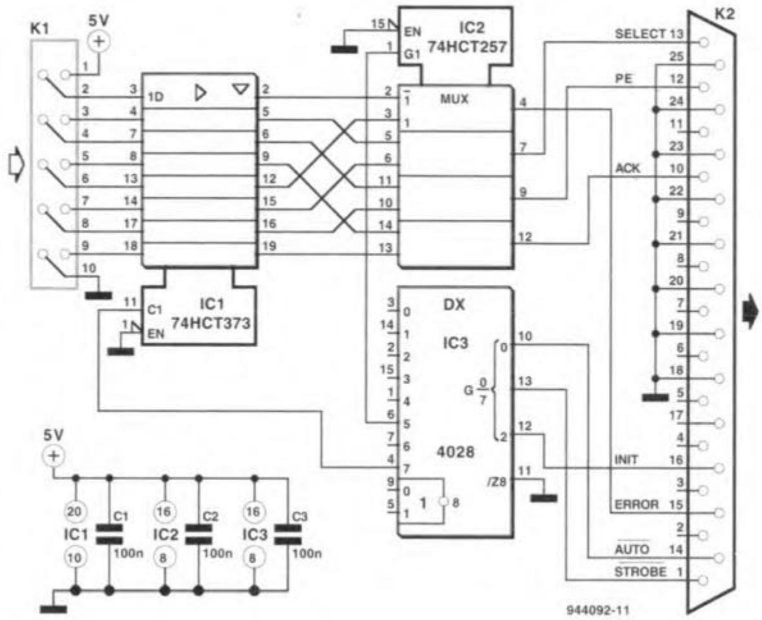
In the present circuit, use is made of outputs STROBE, AUTO and INIT, and inputs ACK, PE, SELECT and ERROR. These seven lines make it possible to make an 8-bit input. The 8-bit wide data are written in two steps, each of four bits. Switching from step to step is effected by IC_2, a Type 74HCT257 multiplexer.

IC_1 functions as a bus buffer. IC_3 ensures the enabling of the buffer, which in its quiescent state is in high-impedance mode via the G input.

IC_3 also arranges the switching between the most and least significant nibbles of the data via the A/B input of IC_2.

The program, whose listing is shown in the table, ensures the correct drive for these three components. It is written in standard Pascal and may thus be used without any difficulty with all Pascal compilers.

Design: R. van Steenis [944092]




```
program readlpt;

uses crt;
var LSB, MSB, Total_New, Total_Old : integer;
  Ctrl, Readbyte : integer;
const Base_Address = $378;

{         b7 b6 b5 b4 b3 b2 b1 b0
   readbyte X I4 I3 I2 I1 X  X  X }

begin
 {initialisation}
 clrscr;
 Ctrl    := Base_Address+2;   {Base address can be $3BC, $278 or $378}
 Readbyte := Base_Address+1;  {depending on printer port number used}
 Total_Old := Total_new +1;
 gotoxy(30,10); writeln('Value LSB  =');
 gotoxy(30,11); writeln('Value MSB  =');
 gotoxy(30,13); writeln('Total value =');
 gotoxy(30,14); writeln('Inputs     =');

 {repeat until key pressed}
 repeat
  port[Ctrl]:=0;
  port[Ctrl]:=4;              {latch Total_New value in buffer}
  port[Ctrl]:=0;
  LSB := port[readbyte];        {read 4 LSBs into memory}
  LSB := (LSB and $78) div 8;      {shift right three positions}
  port[Ctrl]:=5;              {select 4 MSBs}
  delay(1);                  {wait 1 ms}
  MSB := port[readbyte];        {read 4 MSBs into memory}
  MSB := (MSB and $78) div 8;      {shift right three positions}
  Total_New := LSB+MSB*16;
  if Total_New <> Total_Old then     {input changed?}
  begin
   gotoxy (45,10);
   write (LSB,' ');             {write 4 LSBs to screen}
   gotoxy(45,11); write(MSB,' ');   {write 4 MSBs to screen}
   gotoxy(45,13); write(Total_New,' ');{write decimal value to screen}
   gotoxy(45,14);                {and binary value}
   write((Total_New and 128)div 128);  {write bit7 to screen}
   write((Total_New and 64 )div 64);   {write bit6 to screen}
   write((Total_New and 32 )div 32);   {write bit5 to screen}
   write((Total_New and 16 )div 16);   {write bit4 to screen}
   write((Total_New and 8  )div 8);    {write bit3 to screen}
   write((Total_New and 4  )div 4);    {write bit2 to screen}
   write((Total_New and 2  )div 2);    {write bit1 to screen}
   write(Total_New and 1 );         {write bit0 to screen}
  end;
  Total_Old := Total_New;          {save new value as old}
 until keypressed;
end.
```

CURRENT PROBE

When it comes to measuring alternating currents in a safe manner, current probes have a number of significant advantages over series resistors. To begin with, the measuring circuit is electrically isolated from the a.c. supply line which carries the current to be measured. Secondly, the a.c. supply line need not be broken to insert a series resistor, or a cluster of series resistors with an associated range switch. Thirdly, voltage loss caused by the the current probe is negligible.

Current probes are based on inductive coupling only, and usually have a transformer at the input to couple on to the a.c. supply line. To ensure that the current which flows through the a.c. supply line is transformed down to a safe level, which can be handled by an opamp, the input transformer usually has a turns ratio of 1:1,000 or so. In practice, that means that the primary of the transformer has one turn, and the secondary, 1,000.

Here, two transformers with a turns ratio of 1:32 are cascaded to obtain a total current step-down ratio of 1:1,000 ($1:32^2$). This was done mainly to avoid the tedium of having to wind 1,000 turns of wire on to a ferrite ring core. Despite the fact that two transformers are used instead of one, the current probe still has an impressive frequency and current range.

Tr_1 and Tr_2 lower the current carried on the a.c. line by a factor of 32^2. The output current is applied to a current-to-voltage converter based on an AD847 opamp. Resistor R_1 gives the opamp a conversion factor of about 0.1. An off-set adjustment, P_1, is provided to keep the oscilloscope 'in the picture' without having to switch it to a.c. input mode.

The transformers are wound using ferrite ring cores type RCC26/10-3C11 (order code 4330 030 3752) from Philips Components. The inductance conversion factor, A_L, of this core is quite high at 5 µH per turn. The cross-sectional area is 55.9 mm², and the effective length is 60.1 mm. The core material is 3C11. The secondary windings consist of 32 turns of 1 mm dia. (20 SWG) enamelled copper wire; the primary windings of one turn of the same wire. For proper insulation, the primary of Tr_1 is best made from solid, insulated, wire. The PCB connection via terminal block K_1 must not be used at a.c. line voltages greater than 42 V, or currents greater than 5 A. In those cases, the measurement wire must be run through the core of Tr_1.

Finally, some test data obtained with a prototype of the current probe:

frequency range:
50 Hz to 100 KHz (relative error <1.5%)
30 Hz to 1 MHz (relative error < 5%)

current range at 50 Hz:
I_{max} = 15 A (relative error <1.5%)
I_{max} = 20 A (relative error <5%)

Parts list

Resistors:
R_1 = 102 Ω, 1%
P_1 = 10 kΩ multiturn preset

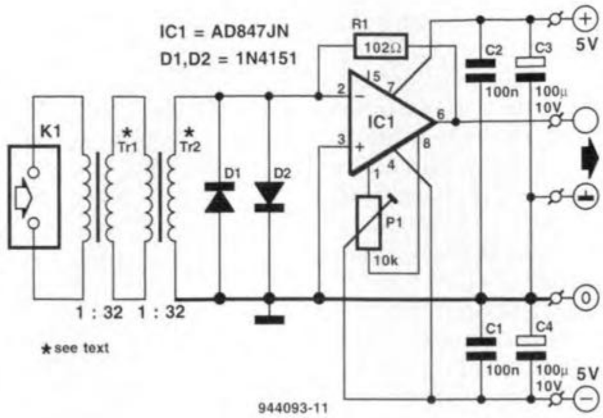
Capacitors:
C_1, C_2 = 100 nF
C_3, C_4 = 100 µF, 10 V, radial

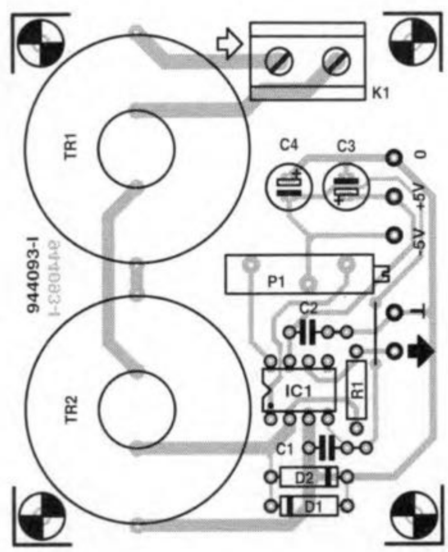
Semiconductors:
D_1, D_2 = 1N4151

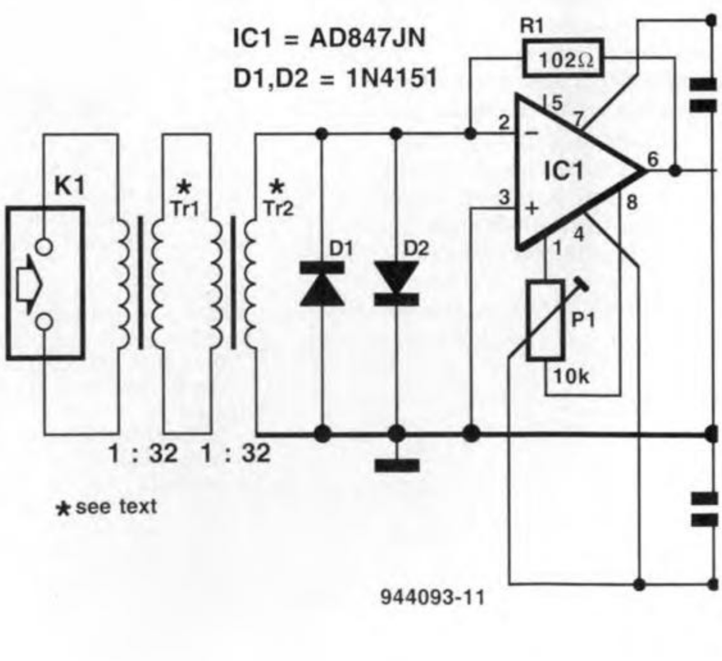
Integrated circuits:
IC_1 = AD847JN

Miscellaneous:
K_1 = PCB terminal block, pitch 7.5mm
Tr_1, Tr_2 = ferrite ring core RCC26/10-3C11 (Philips Components order code 4330 030)

Design: H. Bonekamp
[944093]







50 HZ BAND-STOP FILTER

The suppression of the centre frequency in this band-stop filter is not dependent on the capacitors used, but on the resistors and the properties of the operational amplifier. The resistors are 0.1% precision types and the opamp provides high common-mode rejection.

The suppression of the centre frequency provided by the prototype amounted to about 57 dB. This value hardly changes as long as the total resistance of R_4+R_5 = 20 kΩ. Changing the ratio of R_4:R_5 from that in the diagram does, however, alter the centre frequency, f_c, which is determined by:

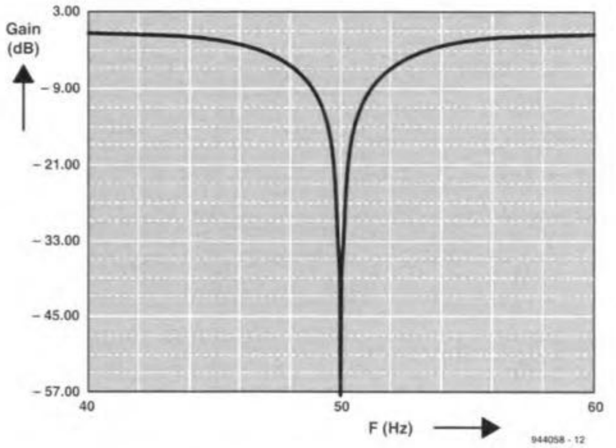
$$f_c = 1/2\pi\sqrt{(R_4 \cdot R_5 \cdot C_1 \cdot C_2)}.$$

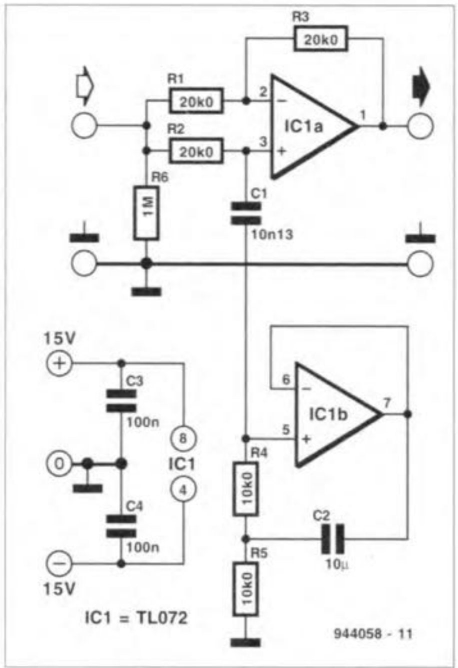
Normally, the resistances should be related asa follows:

$$R_1 = R_2 = R_3$$

and

$$R_4 = R_5 = {}^1/_2 R_1.$$





The Q of the filter depends on the ratio C_1:C_2 – the larger the ratio, the better the Q. With values as specified, the Q of the circuit should theoretically be greater than 8. However, owing to losses in the dielectrics, the practical value measured was only 7.6. It is thus advisable to use high-quality capacitors: polystyrene types for the smaller values. Unfortunately, these are not available in values >56 nF; above that value, MKT types should be used.

Since the internal magnification of the filter is fairly large, the level of the input signal should be kept below 1 $V_{r.m.s.}$.

Resistor R_6 ensures that in no-signal conditions the input of IC_{1a} is at ground potential.

With capacitive loads, it is advisable to connect a 100 Ω resistor in series with the output.

The current drain from a ±15 V supply is about 4 mA.

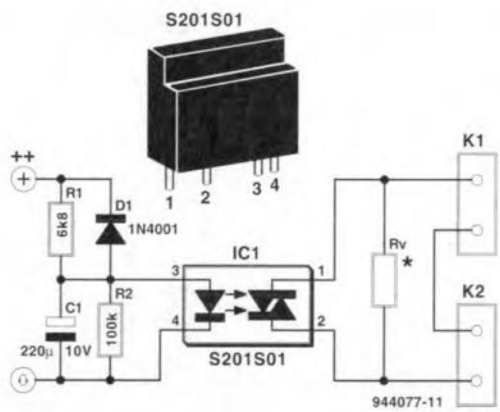
Design: T. Giesberts
[944058]

POWER-ON DELAY

Switching on inductive loads, such as transformers, causes a very high initial current, because at that instant, the load does not produce a counter-e.m.f. Consequently, measures must be taken to prevent the (perfectly correct) mains fuse from giving up the ghost.

One such measure is afforded by the present circuit. Headers K_1 and K_2 are connected in series with the mains switch of the relevant equipment. At power-on, there will be a fairly high drop across R_v, which limits the initial current appreciably.

A voltage is tapped from the supply of the equipment and taken to point ++. As soon as C_1 has been charged via R_1, and the threshold voltage of the LED in IC_1 has been exceeded, the triac in the solid-state relay will come on and short-circuit R_v. The triac remains on, irrespective of whether the load is resistive or inductive.



The value of R_v depends on the load; it will normally be 50–100 Ω. The rating of the resistor must be ≥25 W. It is, of course, possible to use a number of 5 W or 10 W resistors in parallel.

The value of R_1 must be $(U_{++}-1.5)$ / 200.

The value of C_1 must make the product R_1C_1 = 1 s.

The values specified in the diagram assume a supply voltage of about 40 V.

It should be borne in mind that the delay does not function if the load is switched on and off at short intervals, because C_1 cannot be discharged (via R_2) rapidly enough.

Also, remember that full power is not available until the triac has short-circuited R_v.

The solid-state relay used in the prototype can switch at any angle of the alternating voltage. Other types, such as the S101S02 andS101S04 (also from Sharp), have a zero crossing detector on board—the S101S03 and S101S05 have not. The Types S101S03 and S101S04 have an internal series resistor for the LED—the S101S01 and S101S02 have not. The maximum permissible voltage for the S101... series is 400 V; that for the S102... series is 600 V.

The isolating voltage between LED and triac is >4000 V. The leakage distance between the connections must be >6 mm.

Design: K.M. Walraven
[944077]

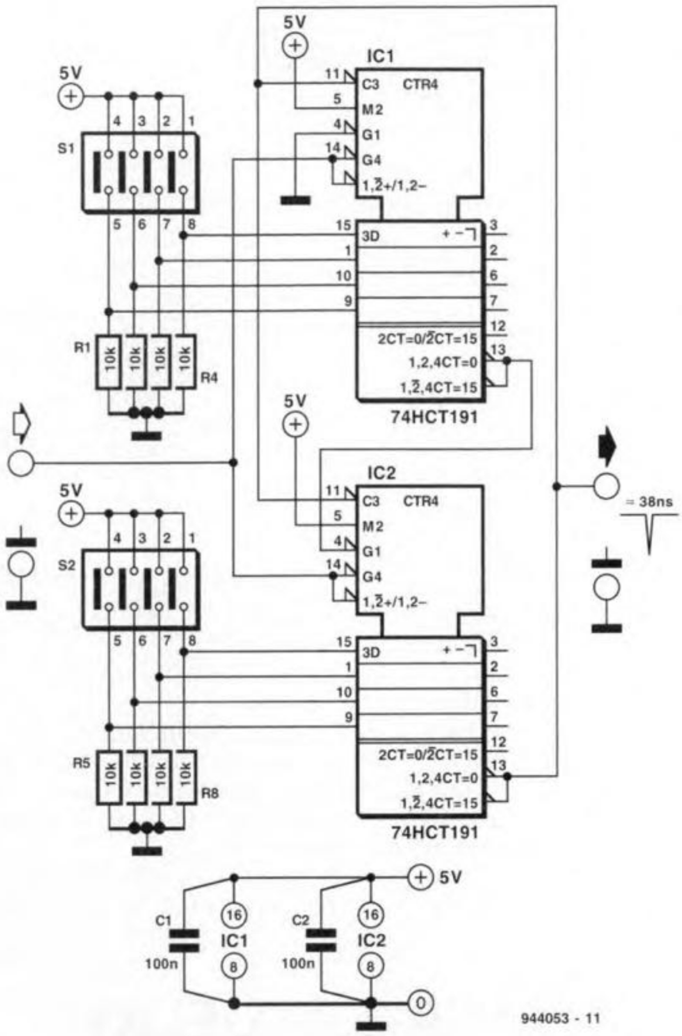
PROGRAMMABLE DIVIDER

Many counter ICs on the market, such as the 4059, 40102 and 40103, are specifically designed for use as programmable timers or dividers. Certain other ICs can, however, also be used for this purpose.

In the prototype, use is made of two presettable 4-bit up/down counters Type 74HCT191—see diagram. By using the 'ripple clock' for presetting the counters, the number of clock pulses needed during down counting is the same as the binary value applied to the data inputs. In this way, the division is equal to the set divisor. The diagram shows how two counters can be combined to an 8-bit divider. Three or more counters may be joined together. The load and clock pins of each IC are then linked, while the ripple clock output (pin 13) of an IC is connected to the enable input (pin 4) of the next IC.

The ripple clock of the MSB counters provides the load pulse and also functions as the output of the divider. When it becomes active, the counters are preset, whereupon the ripple clock is disabled.

The pulse width of the output signal is thus determined by the sum of the propagation delays of the counters. According to the manufacturers' data sheet, this is about 17 ns. In the prototype, which used two counters, a value of 38 ns was measured.



The current drain at low frequencies is determined almost entirely by the pull-down resistors at the data inputs. When the data inputs are all high, the current drain is about 4 mA.

Design: T. Giesberts
[944053]

IMPROVED OPTOISOLATOR

A recurring problem in the design of optoisolator circuits is arriving at the right compromise between insensitivity to spurious signals, component tolerances and long-term stability. One aid in minimizing this problem is replacing the usual collector resistor of the optoisolator by a current source as shown in the diagram. The resulting circuit is simple, inexpensive, uses readily available components and, last but not least, has good reliability.

In the diagram, IC_1 functions as a threshold switch and is also part of the current source, T_1-R_2. Since the NE555 has a TTL-compatible as well asn open-collector output, it can be connected directly to a microprocessor.

The value of R_1 depends on the maximum permissible current through the optoisolator. The value of R_2 is

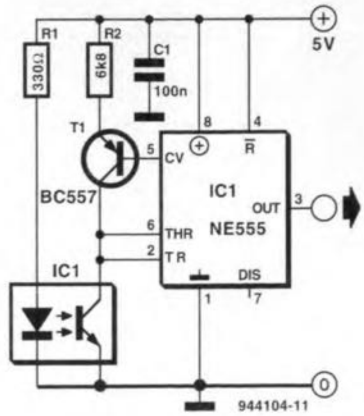
$$R_2 = [(U_b/3-0.6)]/[I_1+I_d)/2],$$

where U_b is the supply voltage, I_1 is the current when the optoisolator receives light, and I_d is the dark current.

The values specified in the diagram assume an optoisolator Type CNY17. If a different type is used, the value of R_2 may have to be altered.

Design: A. Rietjens
[944104]

1



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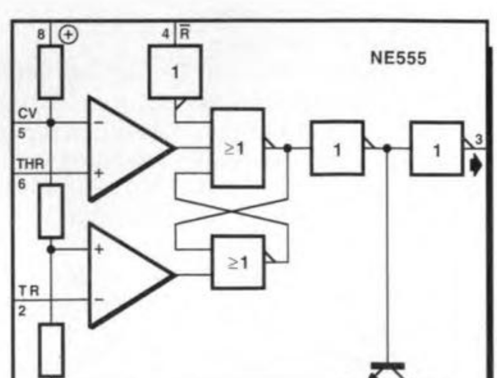


STAND-ALONE S/PDIF COPYBIT ELIMINATOR

The S/PDIF copybit eliminator described in Ref. 1 has to be incorporated into a DAT or DCC recorder, which means that the player has to be opened and, in some cases, modified. The small extension card described here enables the copybit eliminator to be used as an external device, that is, inserted into the coaxial or optical link between a CD player and a DAT recorder. This can be achieved by combining the following circuits:

1. the digital-audio enhancer (Ref. 2);
2. the copybit eliminator (Ref. 1);
3. the extension board described here.

The copybit eliminator so made is a stand-alone unit capable of recognizing all three sample frequencies, 32 kHz, 44 kHz and 48 kHz, and requires no modifications to existing equipment.

The copybit eliminator requires a clock frequency of 128 times the sample frequency. This clock is available at pin 11 of dual bistable IC_3 in the digital-audio enhancer. However the phase has to be reversed for the copybit eliminator. This function is accomplished by inverter IC_{2a} on the present extension board.

Since the header on the digital-audio enhancer can not be used to connect the present extension, a different solution has been found. The extension board is plugged into the socket from which IC_3 is removed. In this way, the copybit eliminator is inserted into the signal path ahead of IC_{3b} (whose function is taken over by IC_{1b} on the extension board). An extra advantage of this connection is that the digital-audio enhancer then also removes any spurious signals introduced by the copybit eliminator.

The socket for IC_3 on the digital-audio enhancer board supplies all signals necessary for the extension board to operate: the supply voltage, the recovered clock, and, of course, the S/PDIF signal at TTL level. The copybit eliminator is connected to 10-way boxheader K_1 on the extension board via a short length of flatcable.

Although the 74HC74 removed from the digital-audio enhancer board can be used again in position IC_1 on the extension board, it is better to replace it with a 74HC**T**74.

Remove IC_3 from the digital-audio enhancer board. Connector $K_{2'}$ on the extension board consists of 14 short, solid wires all cut to the same length (approx. 10 mm) and soldered at the track side. Once the extension board is complete, carefully align.these wires and insert them into the socket for IC_3. Alternatively, remove the socket and solder the wires on the digital-audio enhancer board also.

References:

1. Copybit eliminator, *Elektor Electronics* February 1994.
2. Digital-audio enhancer, *Elektor Electronics* February 1993.

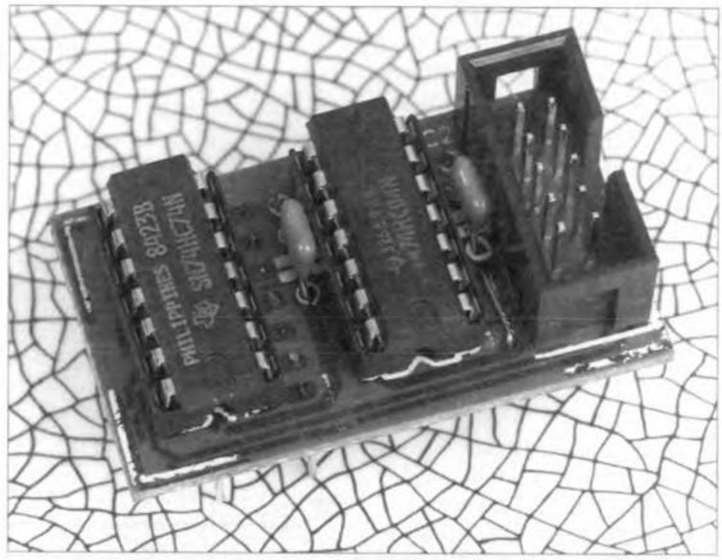
Parts list

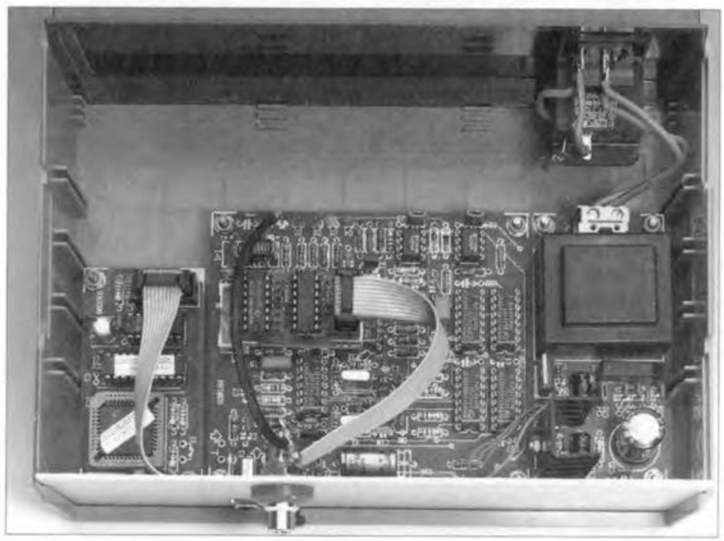
Capacitors:
C_1,C_2 = 47 nF ceramic

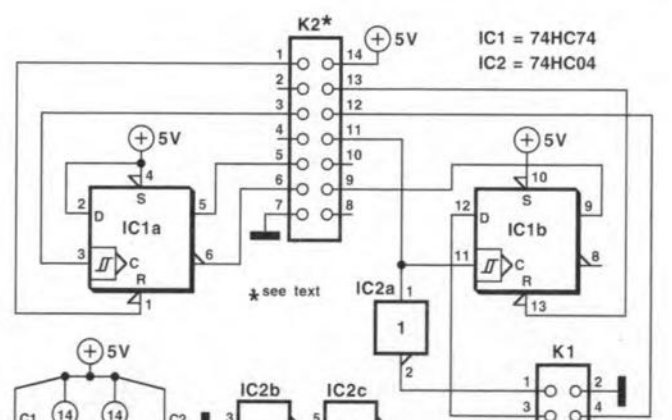
Integrated circuits:
IC_1 = 74HC74 or 74HCT74
IC_2 = 74HC04

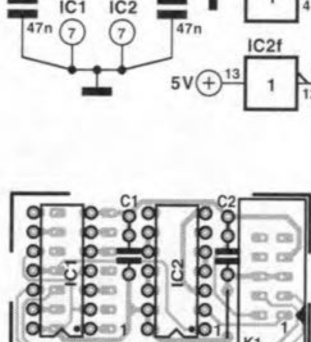
Miscellaneous:
K_1 = 10-way boxheader
K_2 = 14 wires (see text)

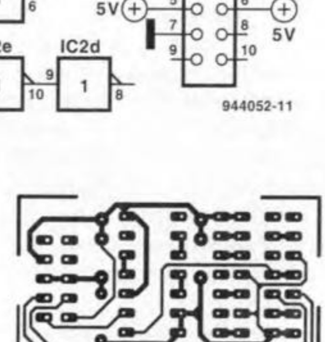
Design: T. Giesberts
[944052]

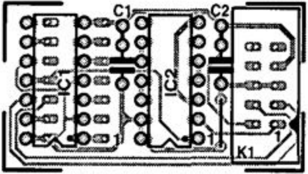


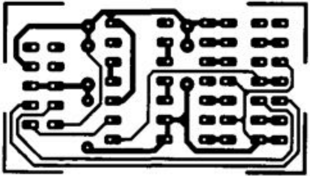








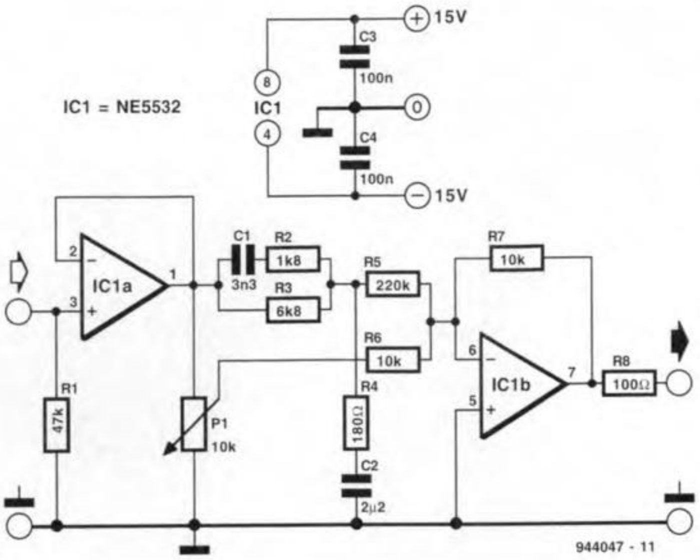

C1
C2
IC1
IC2
1
1
1
K1



LOUDNESS CONTROL

The sensitivity of the human ear vs frequency characteristic is not a straight line. Our hearing is much less sensitive to low frequencies, and slightly less sensitive to high frequencies, than to frequencies in the middle range. This is an evolutionary adaptation for the loudness at which these frequencies occur in nature.

At low levels of sound pressure, the sensitivity of our hearing decreases even more, particularly at low frequencies. This presents a problem in the reproduction of a.f. signals, since at low volume levels the relationship with respect to the original changes. This is the reason that in certain audio amplifiers a physiological volume control, also called loudness control, is incorporated. Usually, this consists of a frequency-dependent network in parallel with the volume control, which provides more attenuation at the middle frequencies than at the low and high ones.



An active type of such a frequency-dependent network is shown in the diagram. It consists of an input buffer, IC_{1a}, and a summation amplifier, IC_{1b}, which are linked by two different signal paths.

One path is via volume control P_1 and R_6, shunted by a frequency-correcting network that consists of C_1, C_2 and R_2–R_5. The network amplifies low frequencies up to 20 Hz by a maximum of 24 dB and high frequencies up to 20 kHz by up to 8 dB. The ratio R_3:R_4 determines the maximum bass amplification, while the value of C_2 determines the cut-off point. Resistor R_2 ensures that no amplification occurs at frequencies above about 20 kHz.

The circuit has a small drawback: because of the correcting network, it is not possible to turn down the volume completely. With the value specified for R_5, the maximum attenuation is 60 dB. If this is not sufficient, the value of R_5 may be increased, but then the frequency correction falls off rather too quickly when the volume is increased. A better solution is a second potentiometer, mechanically coupled with P_1, at the output. However, in a stereo version, that would mean finding a virtually unobtainable quadruple potentiometer.

The NE5532 may be replaced by a comparable type without any difficulty.

The ±15 V supply must be well regulated. It needs to supply a current of only a few mA.

Design: T.Giesberts

[944047]

COMPARATOR WITH UNILATERAL HYSTERESIS

Normally, a comparator has inherent hysteresis, that is, when the input voltage rises, the comparator will react only when the reference voltage is exceeded by a certain amount. When the input voltage drops, the comparator changes state only when that voltage is well below the reference potential. Hysteresis is essential to prevent a system clattering or vibrating around its reference level.

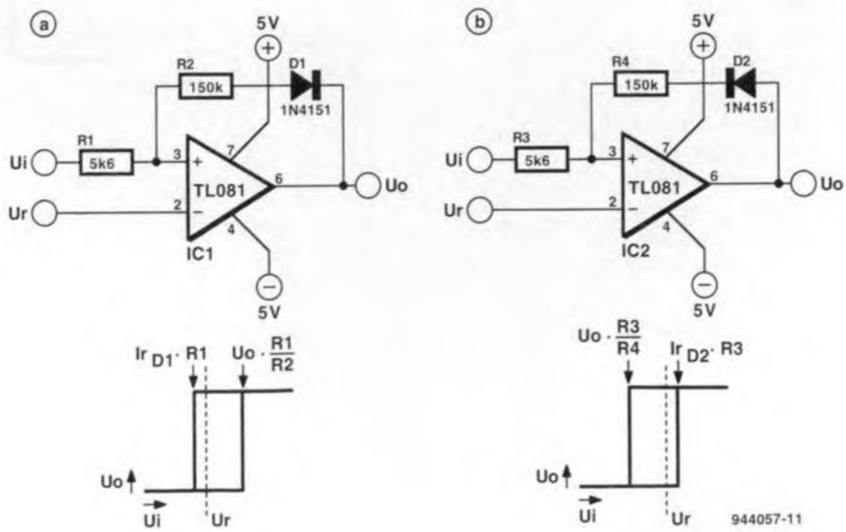
The hysteresis is normally symmetrical, but occasionally it is necessary to make one of the thresholds exactly equal to the reference voltage. This can be achieved by adding a diode as shown in the diagrams.

In **(a)**, the lower edge of the hysteresis window is equal to the reference voltageand in **(b)**, the upper edge.

Normally, feedback resistors R_1 and R_2 determine the size of the window. Diode D_1 causes the resistors to act in only one direction, which results in the window shifting at one side to (virtually) the reference voltage. There will be a slight difference owing to the reverse leakage current of the diode which results in a drop across the resistors. However, the input bias current of the opamp also plays a role. Therefore, if exact operation is required, a good-quality diode (with low leakage current) and opamp must be used.

Design:H. Bonenkamp

[944057]



FAST SWITCH-OFF FOR POWER AMPLIFIER

The 'Medium power a.f. amplifier'[1] uses a relay for the switch-on delay to suppress annoying switching plops. When the amplifier is switched off, however, the relay, owing to hysteresis, remains actuated for a short time, which results in switch-off clicks. Although these are harmless, they are annoying. The present circuit makes them things of the past.

In **Fig. 1** the connection between Re_1-D_6 (anode) and R_{56}-D_7 (anode) is broken and a transistor connected between the two open points. This transistor is controlled via an optoisolator by the rectified (D_1, D_2) and buffered (C_1) secondary voltage of the mains transformer.

When the power amplifier is on, the optoisolator is also on, and the relay is energized by T_1. When the mains is switched off, the optoisolator is disabled and the relay is denergized instantly.

A second solution is replacing resistor R_{59} by two soldering pins, to which the circuit shown in **Fig. 2** is connected. The action of this circuit is near enough the same as that described earlier. In some cases it may, however be necessary to replace C_{22} by a lower value capacitor.

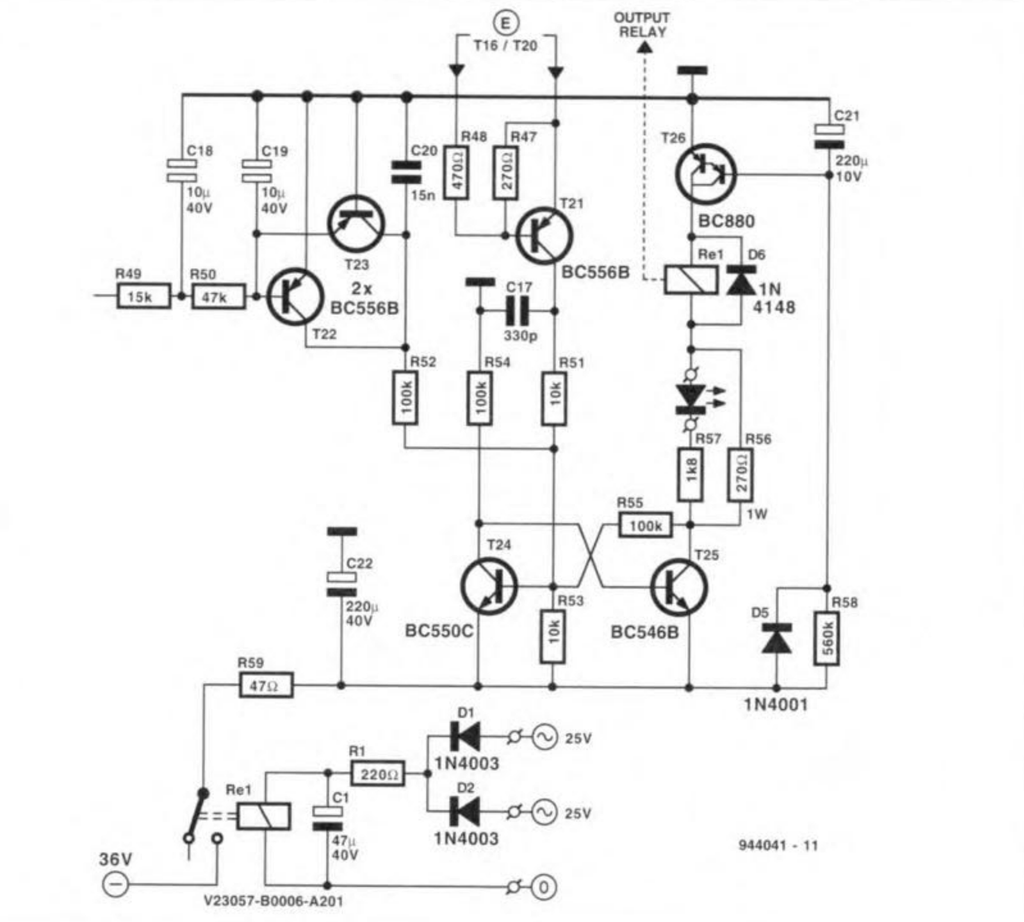
Figure 2 may also be used for yet a third solution, in which C_{20} is short-circuited by a relay contact. A contact of an auxiliary relay, which is closed when the relay is not energized, gives the protection circuit the impression that a direct voltage exists at the output of the amplifier. The protection circuit acts instantly and denergizes the output relay.

The auxiliary relay is protected against large discharge currents of C_2 by a small series resistor.

Design: R. Beck [944041]

[1] *Elektor Electronics*, October/November 1990.

1



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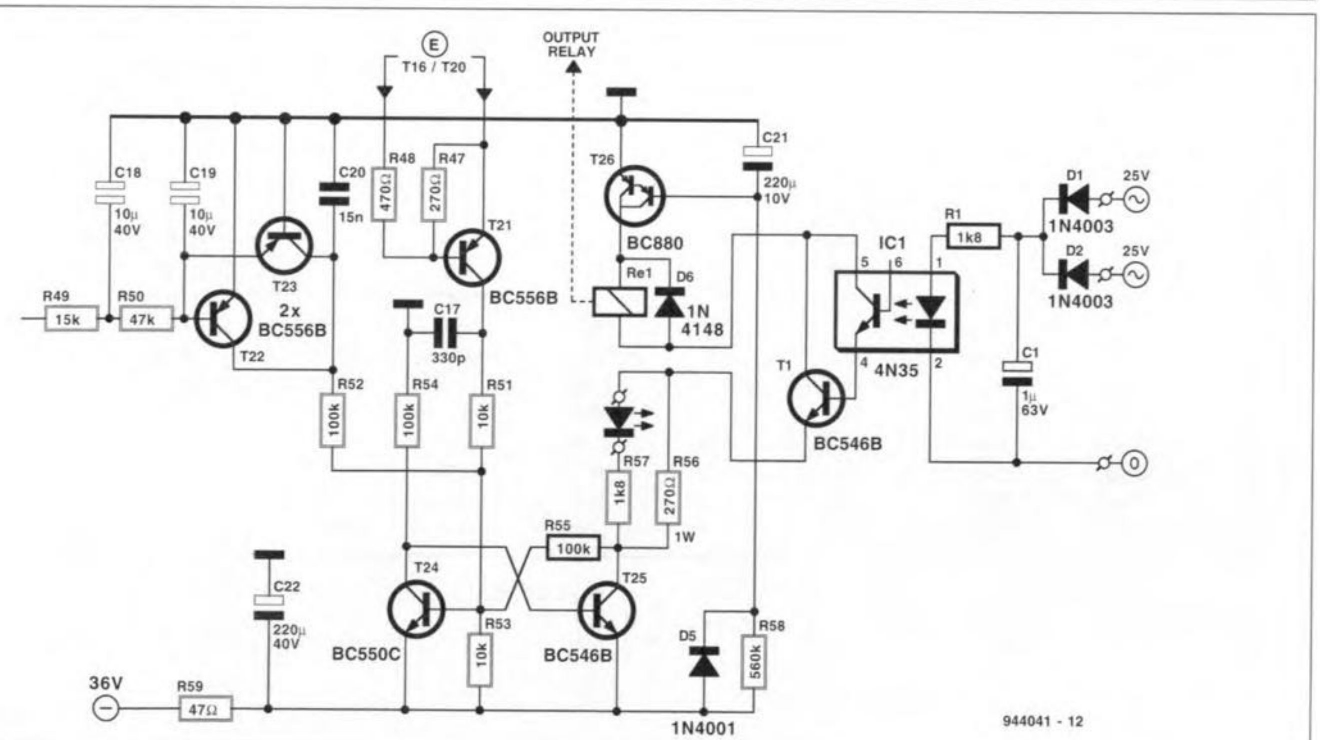


COMPOSITE OPERATIONAL AMPLIFIERS

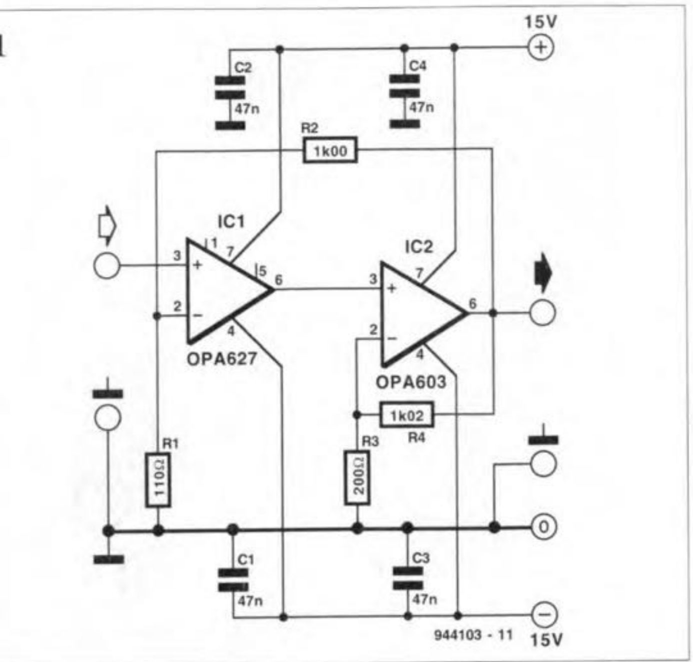
Classical (i.e., voltage-feedback) operational amplifiers, such as the OPA627, have excellent performance in applications where the required gain bandwidth is low compared with the gain-bandwidth product of the opamp. However, increasing closed-loop gain decreases the error-reducing loop gain. Furthermore, starting at relatively low frequencies, the loop gain rolls off at 20 dB/decade of signal frequency increase. In combination, these effects can produce significant errors, especially at higher frequencies where the loop gain can be very low.

Current-feedback opamps, such as the OPA603, have good dynamic performance at both low and high gains. This is because the feedback components set both closed-loop gain and open-loop gain, making loop gain and dynamic performance relatively independent of closed-loop gain. Unfortunately, the d.c. performance of current-feedback amplifiers is poor compared with classical opamps.

A composite amplifier using a classical amplifier and the OPA603 current-feedback amplifier can combine the best qualities of both amplifiers.

Figures 1 and **2** show non-inverting and inverting composite amplifiers. The table shows suggested component values for selected gains and measured performance results.

1



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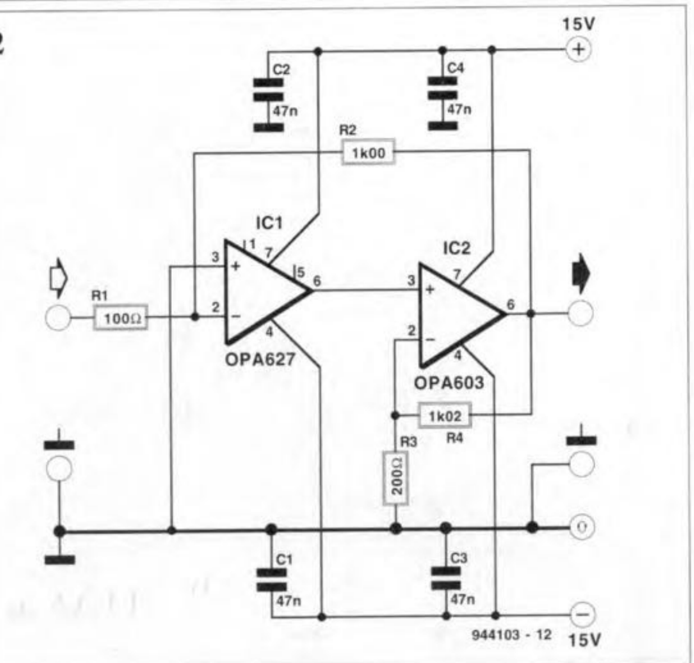

D.c. performance of the composite amplifier is excellent. Since the OPA603 is in the feedback loop of the OPA627, the composite amplifier retains the excellent d.c. characteristics of the OPA627. In fact, since the OPA627 does not drive the load directly, its d.c. accuracy can be better than the OPA627 alone. Thermal feedback within an amplifier driving large loads will cause errors owing to internal thermal gradients and package self-heating. The composite amplifier with an OPA603 can drive 150 Ω loads to ±10 V with no thermal feedback to the OPA627.

The gain of the composite amplifier is set by R_1 and R_2 alone. Errors caused by R_3 and R_4 do not affect the gain of the composite amplifier. The gain of the second amplifier, set by R_3 and R_4, should be within ±5% to ensure expected dynamic performance.

Slew rate and full-power response of the classical amplifier are boosted in the composite amplifier. Since the OPA603 adds gain at the output of the OPA627, the slew rate of the OPA627 is increased by the gain of the OPA603. For example, in the gain-of-100 composite amplifier, the slew rate and full-power response of the OPA627 are increased from 40 V µs^{-1} and 600 kHz to over 700 V µs^{-1} and 11 MHz respectively.

Burr-Brown Application
[944103]

OVERALL GAIN [V/V]	BW [MHz]	IC_1	OPA603 GAIN [V/V]	R_1(1) [Ω]	R_2 [Ω]	R_3(4) [Ω]	R_4 [Ω]	SLEW RATE [V µS^{-1}]	SETTLING (0.1%)(2) [ns]	SETTLING (0.01%)(2) [ns]
5	90	OPA627	3	255	1020	499	1020	100	265	520
10	180	OPA627	6	110	1000	200	1020	240	240	500
20	330	OPA627	12	52.3	1000	93.1	1020	620	200	520
50	750	OPA627	26	49.9	2430	40.2	1020	730	320	530
100	1500	OPA627	52	49.9	4990	20	1020	730	330	(3)
200	2500	OPA637	18	49.9	10 k	60.4	1020	580	350	(3)
500	6000	OPA637	42	49.9	25 k	24.3	1020	590	580	(3)
1000	10000	OPA637	85	49.9	50 k	12.1	1020	510	640	(3)

NOTES: (1) R_1 shown is for noninverting composite amplifier. For inverting amplifier, R_1 = gain/R_2. (2) Settling time for 10 V output step. (3) Output noise exceeds 0.01% at this gain. (4) For intermediate gains, use the higher value R_3.

B7	B6	B5	B4	B3	B2	B1	B0	
x	x	x	x	x	x	0	0	drive A:
x	x	x	x	x	x	0	1	drive B:
x	x	x	x	x	x	1	x	DOS selects number
x	x	x	x	x	1	x	x	18 sectors/track
x	x	x	x	x	0	x	x	36 sectors/track
x	x	x	x	1	x	x	x	read/write
x	x	x	x	0	x	x	x	write protect

x = don't care
1 = jumper not fitted
0 = jumper fitted

Table 2. Configuration bit functions.

board. A dash indicates that a jumper must be fitted.

Once the card address has been set up, the board is put aside for a moment. Do not yet insert into the PC.

Start the software configuration by informing the PC about an additional drive. Enter the structure of the drive into the CONFIG.SYS file. Use one of the following commands to do so:

```
driveparm=/D:00/F:07/H:2/S:xx/T:80
```

to assign the solid-state disk to drive A:, or

```
driveparm=/D:01/F:07/H:2/S:xx/T:80
```

to assign it to drive B:.

The parameter 'xx' indicates the number of sectors per track. It should be set to '18' to emulate a 1.44-MByte disk, or '36' for a 2.88-MByte disk.

If the computer already has two diskette drives, and you wish to add the solid-state disk, a DOS driver must be started. That can be achieved by adding the following line to the CONFIG.SYS file:

```
Device=\<DOS directory>\driver.sys
/D:n/F:07/H:2/S:xx/T:80
```

where n takes a value between 0 and 3.

Having added these lines, switch the computer off. Insert the solid-state disk card, secure the bracket, and close the case. The solid-state disk is not usable yet, because it has to be formatted first. That is done with the aid of the familiar 'FORMAT' command available under DOS. Despite the fact that the capacity of the drive has been fixed by the 'driveparm' instruction, it is still recommended to repeat the desired capacity. Do this by using the /F: parameter offered by the FORMAT command, as follows:

```
FORMAT B:/F:1440
FORMAT B:/F:2880
```

for a 1.44-Mbyte or 2.88-MByte emulation of disk drive B:, respectively. Evidently, the 'station' letter (B: in the examples) must be changed in accordance with the identification assigned to the solid-state disk.

If the solid-state disk is to be used in a PC which has no disk drives at all, it has to be formatted in another PC, and then loaded with the application software. Note that this requires the disk to be made 'bootable', which is achieved using FORMAT /S.

Finally, we repeat our warning that only 2 Mbyte of memory capacity is available even if the disk card is formatted for 2.88 Mbyte.

(940085)

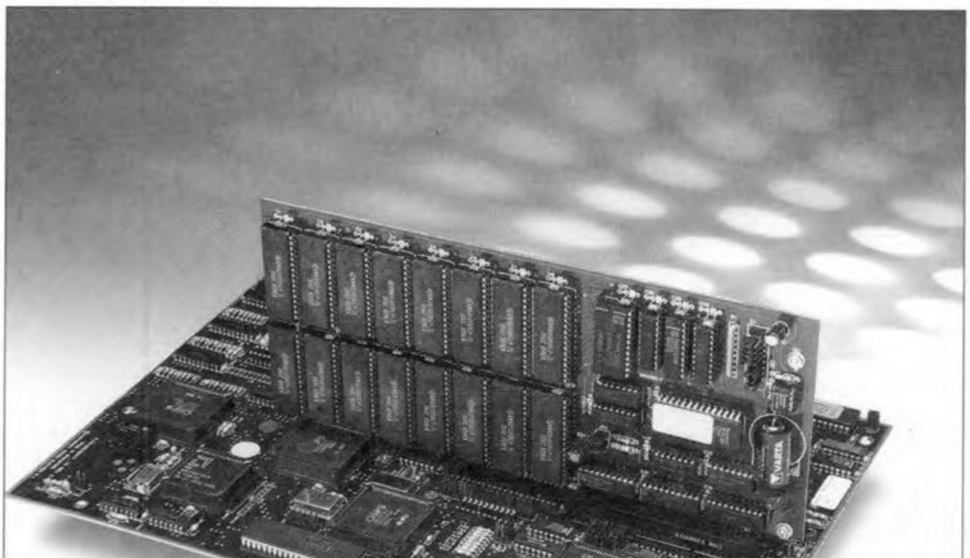

Fig. 3. Completed prototype of the solid-state disk card. mounted into an extension slot of a 386 motherboard. The key words are: no wear and tear, quiet operation, high speed and excellent reliability.