THE ELECTRONICS & COMPUTER MAGAZINE

DECEMBER 1998

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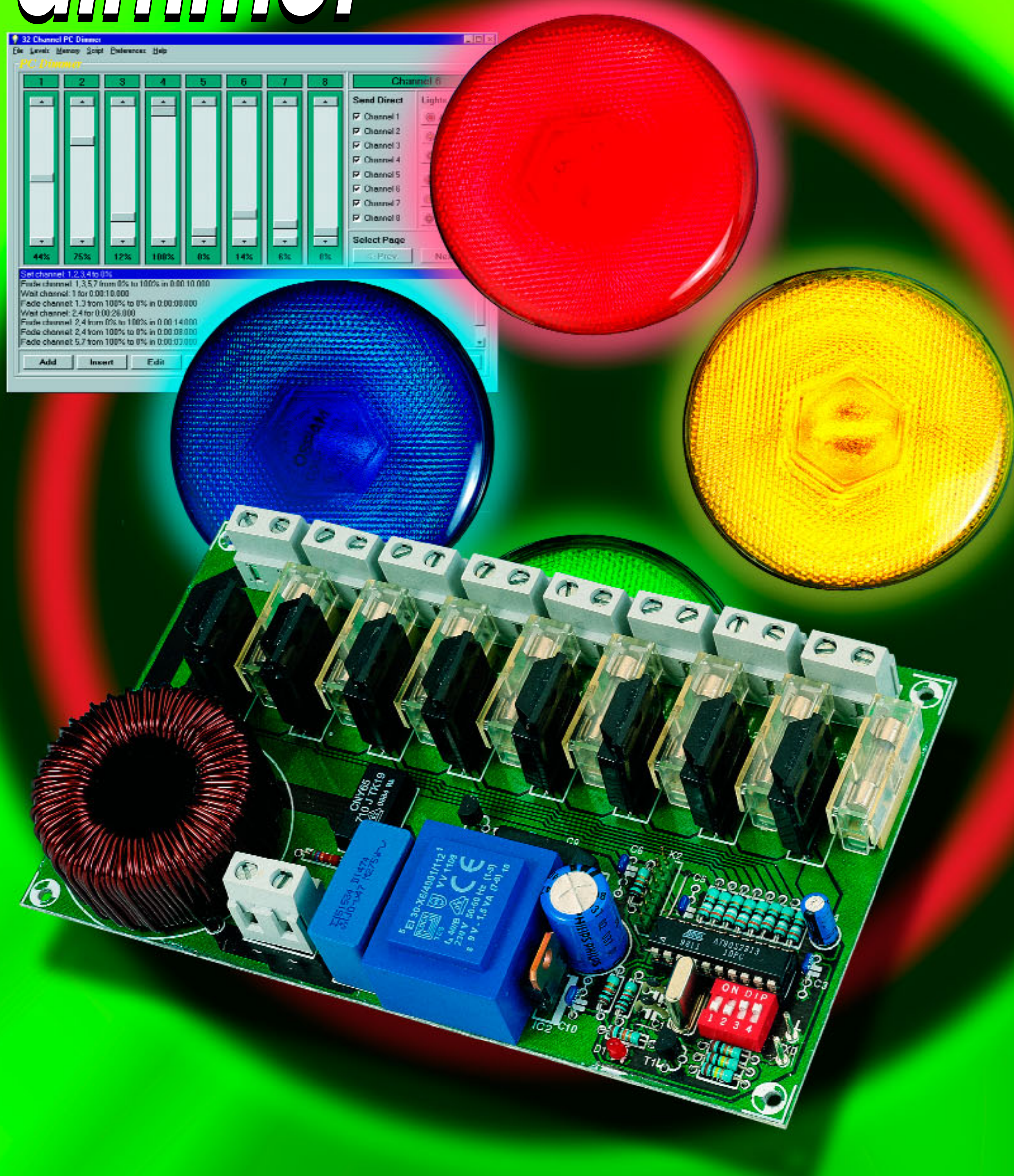

digital terrestrial television (DTT)

anemometer

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MISCELLANEOUS INFORMATION

THIS MONTH IN PC TOPICS:

Visit our new Web site at
http://www.elektor-electronics.co.uk

We thank all our readers for their continued support during the past year and wish you all a happy, prosperous, and peaceful 1999.

1998 International PC-SOFTWARE CONTEST

The Winners

Squarely aimed at anyone with an active interest in the subjects Measurement, Development and Communication, an International Software Design Competition was launched in our July/August 1998 magazine. Apparently, these subjects were much harder to master than last year, because 'only' 80 Competition entries landed on the desks of the Jury members. Despite the smaller number of entries, however, the Competition did produce some very nice projects.

International First Prize

Last year, as the Competition drew to a close and the points were totted up, the winning design quickly emerged because of the very high grades the Jury members gave to his project. This year, the Jury witnessed a photo finish, with fierce competition till the last moment between four extremely well-designed projects. In the end, the International First Prize, an **ST1000 EMC master Spectrum Analyzer/Tracking Generator** with a value of more than £3,000 (sponsored by EMC Master International), was awarded to the brothers **Jack and Mark Nowinski** from Ontario, Canada, for their superbly designed, novel and extremely useful **Electrocardiograph** project. The Jury is not only extremely pleased and honoured to be able to award the prize to these Canadian readers of Elektor Electronics, but also deeply impressed by the great software and hardware as well as the excellent multimedia presentation supplied by Messrs Nowinski.

National Prize Winners

Winners of the national prizes — mostly sponsored by advertisers in the UK version of Elektor — are listed in the table below. Congratulations! All prizewinners have been individually advised of their good fortune by our Editorial Secretariat. Unfortunately, because quite a few Competition entries we received did not meet the Competition Rules, a number of prizes could not be awarded. These prizes, it has been agreed with the relevant sponsors, will remain in store for next year's Competition.

Presentation of the designs

In the PC Topics Supplement of the January 1999 magazine we will endeavour to publish a selection of prize-winning projects. This will be an international choice, presenting Competition entries we received from many countries in which *Elektor Electronics* (or its sister magazines) is read, in other words, from all over the world! As a matter of course, the project that won the International First prize will also be included (in condensed form).
A number of prize-winning Competition entries will be copied integrally onto a CD-ROM which we hope to publish by early February 1999. Furthermore, a number of projects will be discussed in greater detail in forthcoming (1999) issues of *Elektor Electronics*.

(990010-1)

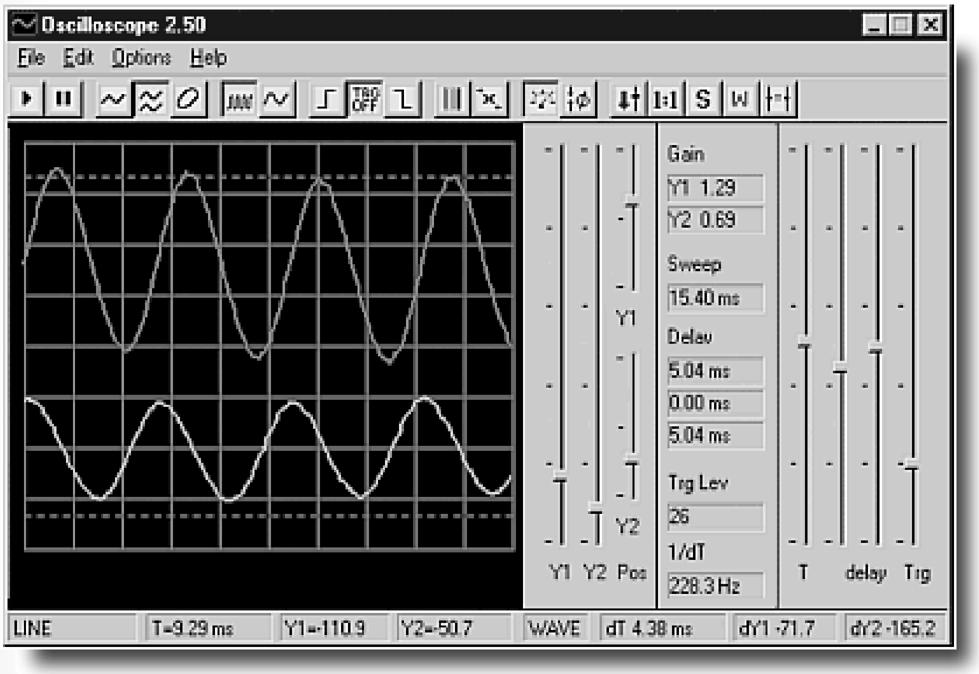
Prize no.	Description	Sponsor	Awarded to	Project
1	Ultiboard Challenger Unlimited; value £1833	Ultimate Technology	J. T. Kokkoris	Temperature Recorder; a wonderfully simple temperature logger based on the DS1620 transducer.
2	Proteus IV package; value £1625	Labcenter Electronics	H. Vasquez Matute	535 Simulator; use this intuitive program to simulate this CPU and others in the 80x51 family.
3	ANSI C Compiler for the 8051 + Source Level Simulator; value £1040	Crossware Products	D. D. Aggelos	UPIO; use the PC to control up to 8 relays.
4	EMC Master Info CD, Extended Version; value £775	EMC Master International	O. Kvindesland	Log2ps; a QSL card machine for the active radio amateur.
5	ADC-200 Multifunction Test Instrument; value £499	Pico Technology	S. Jonsson	Hell/CW Tx/Rx program; your PC turned into a HellSchreiber (for radio amateurs).
6 (1)	EMC Master Info CD; value £285	EMC Master International	M. A. Wisintainer et al.	Remote Experiments with PIC17C43; a remote, didactic laboratory wholly controlled by students via an Internet link.
6 (2)	EMC Master Info CD; value £285	EMC Master International	M. Hewitt	Frequency Counter on Printer port; for TTL pulses, max. frequency about 100 kHz on a P133 PC.

electronics on-line

oscilloscope software

soundcard becomes A-D converter

The oscilloscope is generally recognized as an indispensable test instrument when it comes to measuring electrical quantities. However, as an 'occasional' user, you may object to forking out £300 or more for a 'scope. If you restrict yourself to relatively low frequencies only, the PC oscilloscope is a much cheaper alternative.



There are several ways of implementing an oscilloscope function on a PC. To mimic a 'real' oscilloscope, you use a special type of insertion card or an external box containing the requisite amplifier stages and fast A-D converters. For many applications, however, it is sufficient to be able to just view the occasional low-frequency signal. In that case, consider exploiting the soundcard installed in your computer. After all, it will typically have an input amplifier and a stereo A-D converter with a maximum sampling rate of 44.1 kHz or even 48 kHz, and these are just what you need to realize an oscilloscope on your PC. Other ingredients include an oscilloscope program for proper processing and displaying of the data supplied by the soundcard. Besides, this software should provide the user controls normally found on an oscilloscope. With this in mind we set out to find PC-oscilloscope freeware, shareware and affordable software on the Internet.

If you want to keep things as simple as possible, and you are satisfied with viewing just one signal, you will find that Yukinon Yamasaki's shareware program **Graphic Level Meter** goes a long way. Besides a (pretty Spartan) two-channel oscilloscope, this Windows program offers three types of VU meter. Registration is just $10, and the download addresses may be found at:
http://www.hitsquad.com/smm/programs/Graphic_Level_Meter/download.shtml

A more sophisticated 'scope function is offered by **WinSpec32**. This program offers more than just an oscilloscope on your PC screen: being a spectrum analyser it also sports an FFT function. The front panel controls that appear on the screen allow you to set the sampling rate and trigger level, to mention but a few. The spectrum analyser is even capable of producing 3D plots. Registration costs $20 and the download site is at:
http://www.c3sys.demon.co.uk/winspec.htm

The website run by PAS has a number of measurement programs in store, including a spectrum analyser, a frequency generator and a signal analysis centre. Most programs also offer an oscilloscope function to view signal shapes. Although the programs are not as cheap as the two mentioned above, they do offer a lot, and the graphic presentation is impressive. 14-day trial versions of all programs may be downloaded from:
http://www.audio-software.com

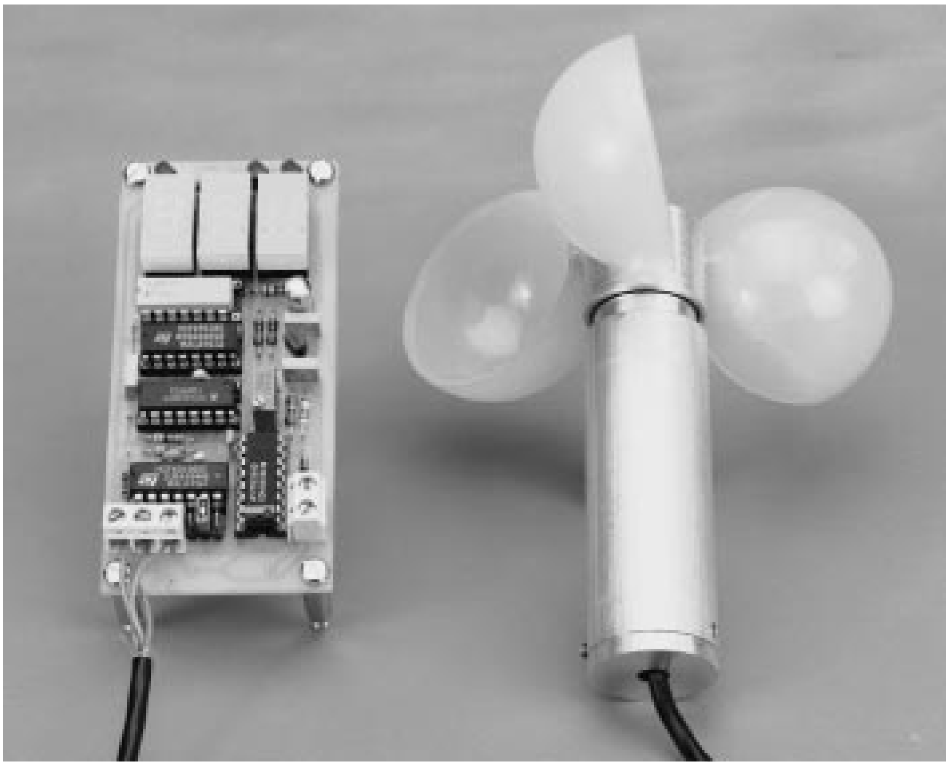
Oscilloscope for Windows is a remarkable program by almost any standard: firstly, it is free of charge; secondly, its size is less than 100 kB. It offers all normal 'scope functions: two channels, an XY function, a spectrum analyser, trigger controls, delay, memory function, and lots more. The program was written by Konstantin Zeldovich and may be downloaded from the Moscow server of the Russian State University:
http://polly.phys.msu.su/~ zeld/oscill.html

(985089-1)

anemometer

based on Hall-effect sensor

An anemometer is not an instrument one finds in many homes. It is only those whose hobby is sailing, surfing or meteorology who find an anemometer a desirable instrument. Since a good anemometer is fairly expensive, it is well worth while to consider building one yourself. The design presented here has been tested in all kinds of weather during which it stood up well and proved its worth.



INTRODUCTION

An anemometer is, in general, an instrument for measuring the rate of flow of a gas. More particularly, in meteorology, it is an instrument for measuring the speed of the wind. A common type consists of four hemispherical cups carried at the ends of four radial arms pivoted so as to be capable of rotation in a horizontal plane, the speed of rotation being indicated on a dial calibrated to read wind speed directly. An anemograph records the speed and sometimes the direction of the wind.

A different type is the hot-wire anemometer which is based on an electrically heated wire that is cooled by the fluid or gas passing around it. The faster the flow, the lower the temperature of the wire and the lower its resistance. So, the rate of flow can be calculated by measuring the resistance of the wire.

In yet another, simple, version, the force of the wind is used to depress a plate suspended from a spring. Any change in the length of the spring forms a directly readable measure of the wind speed. Such an instrument is, however, not terribly accurate.

Most practical meteorological anemometers consist of two distinct parts: a sensor and an indicator or dial. The sensor normally consists of an assembly of three or four hemispherical cups mentioned earlier.

The rotary movement of the sensor can be translated into an electrical signal in various ways. For instance, a light barrier or reflection sensor may be used to scan the light/dark pattern produced by a crenellated disk on a spindle. Another method is the use of a small magnet mounted on a spindle to generate pulses in a pickup coil or reed relay as seen in certain speedometers for bicycles. Yet another, and a very reliable, method is used in the present design. In this, a rotating small magnet is combined with a Hall-effect sensor.

FREQUENCY COUNTER

The basic setup of the anemometer is shown in the block diagram in **Figure 1.** This shows that the electronic part is very simple and, indeed, in this type of design the mechanical work is invariably the most tedious. In essence, the electronic part is nothing more than a frequency counter. The pulses generated by the rotating magnet in the Hall-effect sensor are counted in relation to time, then decoded, and finally applied to a light-emitting diode (LED) display. The display is calibrated in m s^{-1}.

The simplicity of the electronic cir-

Design by R. Veltkamp

1

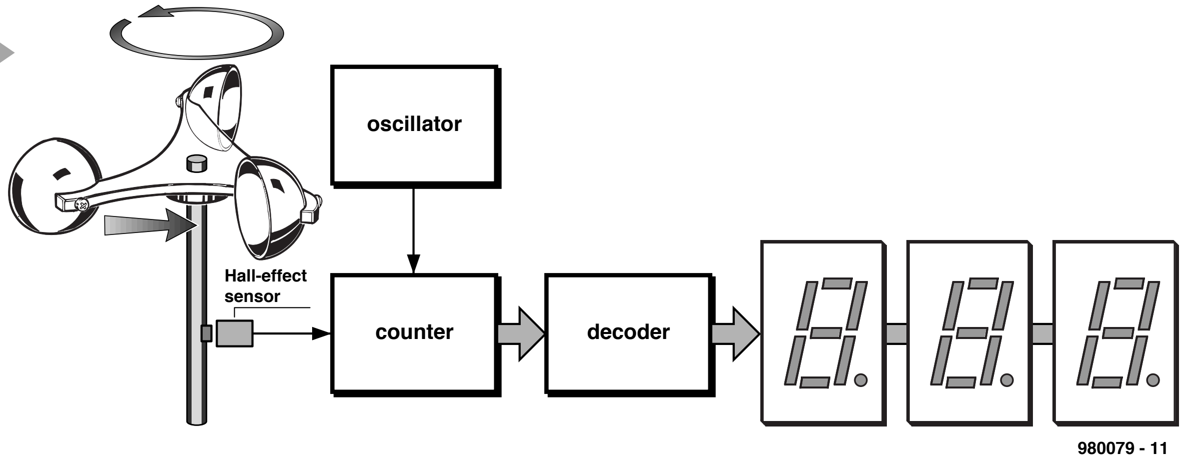


Figure 1. A simple frequency counter is used to determine the number of pulses induced by a rotating magnet in a Hall-effect sensor. The magnet is rotated by the wind via a hemispherical cup assembly.

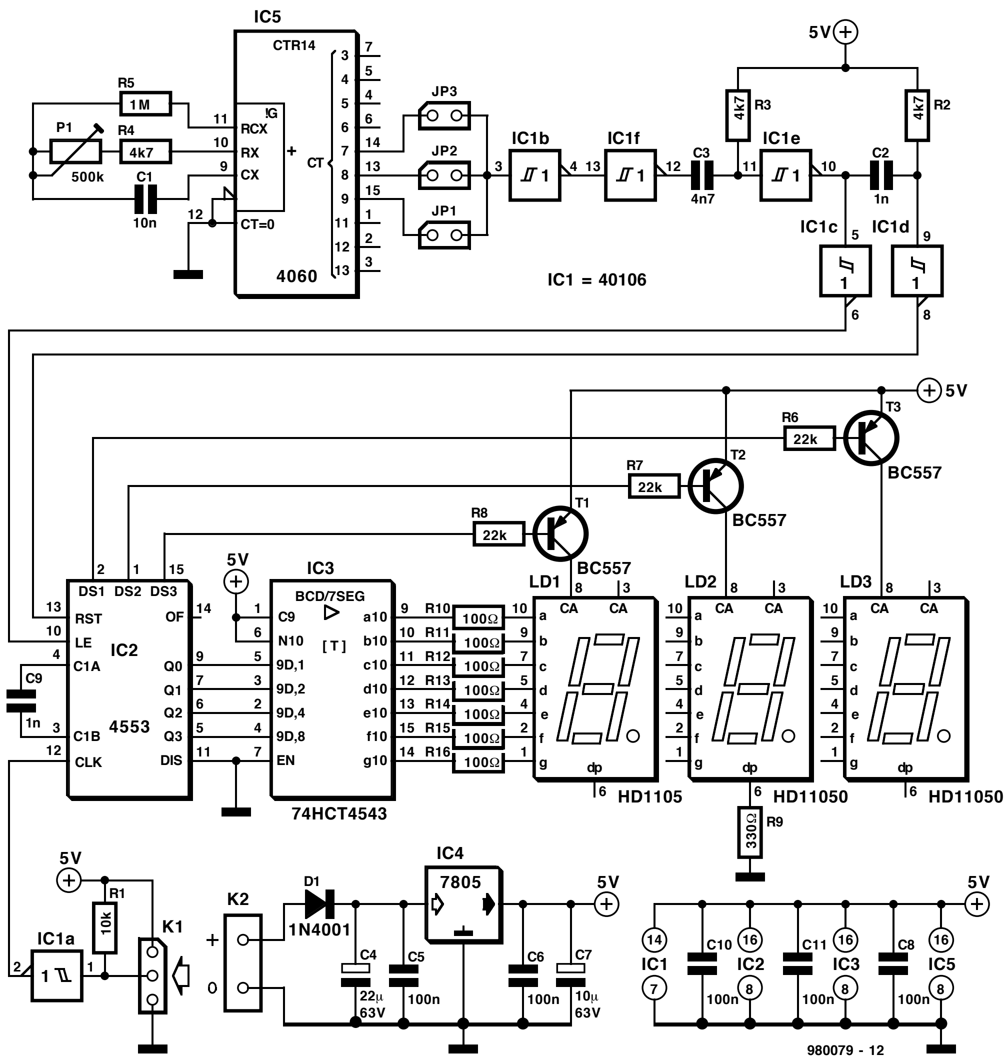
cuits indicated in the block diagram is reflected in the circuit diagram in **Figure 2.**

The circuit is based on a Type MC14553B counter, IC_2, to which the pulses from the Hall-effect sensor are applied via pin header K_1 and Schmitt trigger IC_{1a}. Circuit IC_2 consists of three binary-coded decimal (BCD) counters that are triggered by the trailing edges of the incoming pulses. The counters are cascaded in synchrony. A four-fold latch at the output of each counter enables any selected counting result to be stored. This information is then multiplexed into a single BCD coded output.

The counter state is stored in the latches when the input to the relevant latch is high. The data stored in the latches are read after the counters have been reset, provided that the latch enable input (pin 10) remains high during the whole reset cycle.

Figure 2. The circuit of the electronic part of the anemometer is based on counter IC_2.

2



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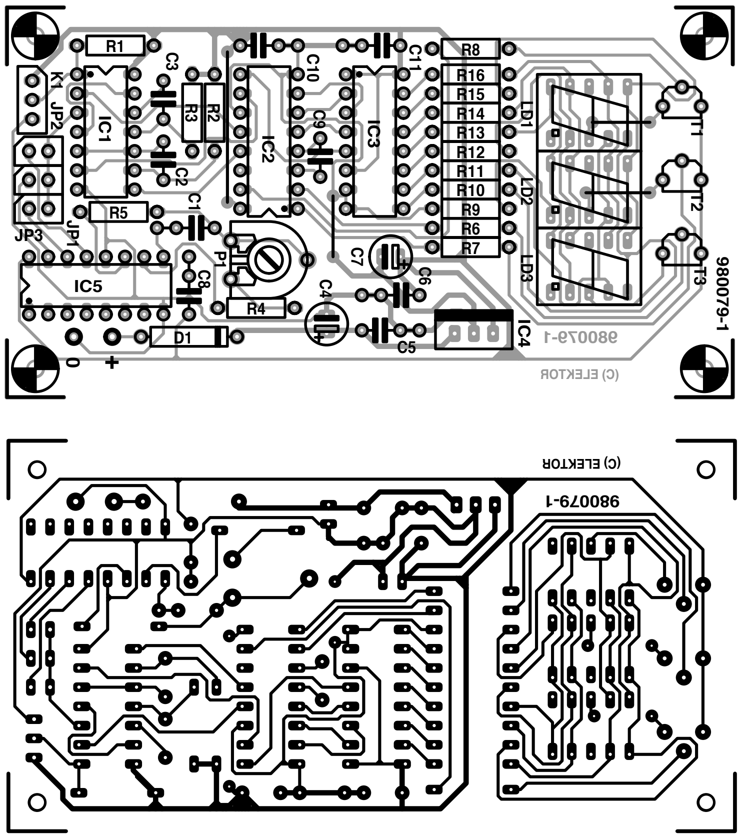
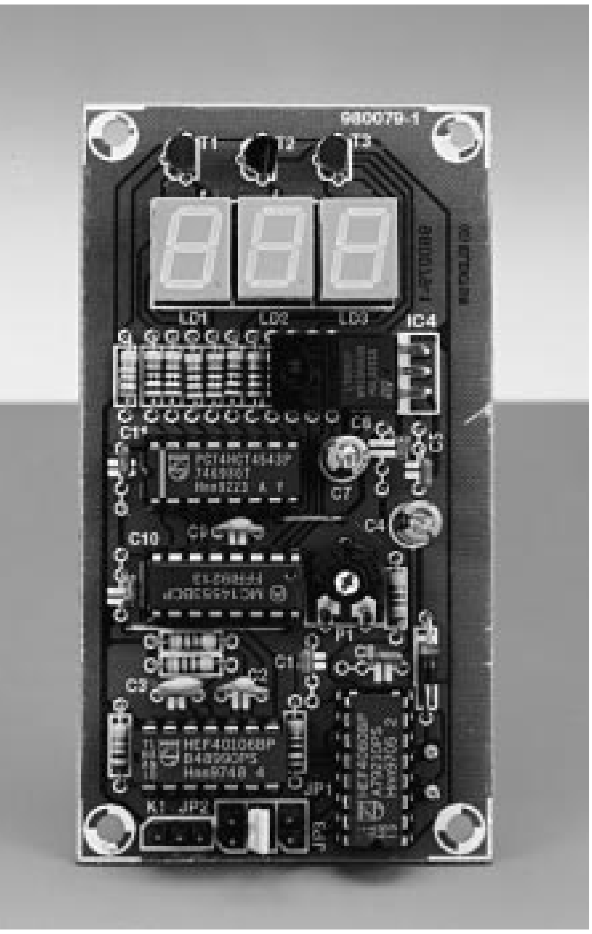


Figure 3. Populating the printed-circuit board for the electronic part of the anemometer is absolutely straightforward. The board is, however, not available ready-made.



It will be clear that the measurement interval is equal to the time lapse between resetting and latching. This time lapse is determined by oscillator IC_5, which gives the requisite switching pulses to IC_2 inputs LE and RST. Since IC_5 also contains a 14-bit binary counter, it is possible to use a much higher oscillator frequency than needed for the requisite measurement time, which improves the overall stability. The pin jumpers, JP_1–JP_3, at the relevant outputs of IC_5 enable setting an oscillator frequency that is $\times 2^8$, $\times 2^9$, or $\times 2^{10}$ higher than needed for the measurement time.

The required positions of the pin jumpers and the frequency set with P_1 depend entirely on the construction of the hemispherical cup assembly. Assuming a maximum readout of 99.9 m s^{-1}, and two sensor pulses per revolution, the basic requisite frequency in the prototype worked out at 82.6 Hz, which is multiplied by $\times 2^{10}$ (JP_3 closed – other two jumpers open).

In some cases it may be necessary to alter the value of one or more components in the oscillator circuit. The oscillator frequency, f_o, is given by:

$$f_o = 1/2.3C_1(R_4+P_1) \qquad \text{[Hz]}.$$

Parts list

Resistors:
R_1 = 10 kΩ
R_2–R_4 = 4.7 kΩ
R_5 = 1 MΩ
R_6–R_8 = 22 kΩ
R_9 = 330 Ω
R_{10}–R_{16} = 100 Ω
P_1 = 500 (470) kΩ preset potentiometer

Capacitors:
C_1 = 0.01 μF, ceramic
C_2, C_9 = 0.001 μF, pitch 5 mm
C_3 = 0.0047 μF, pitch 5 mm
C_4 = 22 μF, 40 V, radial
C_5, C_6, C_8, C_{10}, C_{11} = 0.1 μF, ceramic
C_7 = 10 μF, 63 V, radial

Semiconductors:
D_1 = 1N4001
T_1–T_3 = BC557B

Integrated circuits:
IC_1 = 40106
IC_2 = 4553
IC_3 = 74HCT4543
IC_4 = 7805
IC_5 = 4060

Miscellaneous:
Hall-effect sensor = OHN3040U (TRW)
LD_1–LD_3 = HD11050, Class >L
K_1 = 3-way pin header
K_2 = 2-way terminal block for PCB mounting
Mains adaptor: >9 V, 300 mA
JP_1–JP_3 = 2.54 mm pin strip with pin jumper
Roller bearing
Magnets 7 mm long, 15 mm dia.
Set ring
Brass spindle, 4 mm dia.

Note that, strictly speaking, the value of R_5 needs to be 2–10 times that of R_4+P_1. In practice, this is not very critical, however.

The BCD-to-7-segment conversion is carried out by IC_3, while the display is formed by LD_1–LD_3.

Resistors R_{10}–R_{16} prevent the current through each of the display diodes exceeding 10 mA.

POWER SUPPLY

Power for the anemometer is provided by a mains adaptor with an output of not less than 9 V. This voltage is regulated by IC_4 at 5 V.

CIRCUIT BOARD

The electronic circuits are best assembled on the printed-circuit board in **Figure 3**. As this board is not available ready-made, it needs to be produced privately.

Note that the board just visible in the introductory photograph is that of the first prototype and deviates in some important points from that shown in Figure 3.

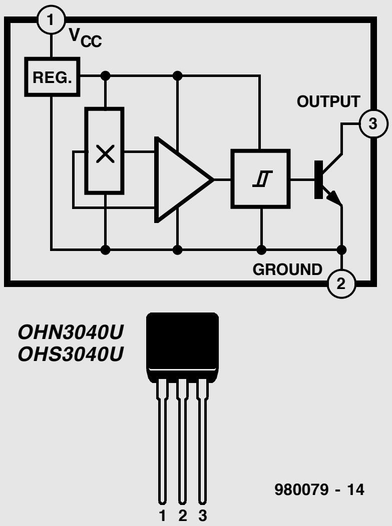
Hall-effect sensor

When a current-carrying conductor or semiconductor is placed in a magnetic field, a potential difference is generated between the two opposed edges of the conductor or semiconductor in a direction mutually perpendicular to both the field and the conductor or semiconductor. This effect is called the Hall effect, and the generated voltage, the Hall voltage. The voltage level depends on the kind and thickness of the conductor or semiconductor material, the magnitude of the current, and the strength of the magnetic field.

A Hall element is very suitable for use as a contact-less switch and as test probe for measuring direct currents. Since, in contrast to inductors, the voltage generated by the magnetic influence does not depend on the speed with which the field changes, a Hall-effect sensor is eminently suitable for measuring relatively slowly changing or even stable magnetic fields.

The Hall-effect sensor used in the anemometer is a Type OHN3040U from TRW, which is a monolithic IC that incorporates a Hall element, a linear amplifier, a Schmitt trigger, and a bandgap voltage regulator to allow operation over a wide range of supply voltages.

The device has a logic level output and provides sink currents of up to 15 mA. This allows the driving of more than five TTL loads, or any standard logic family using power supplies ranging from 4.5 V to 24 V. The output amplitude is constant at switching frequencies from d.c. to over 200 kHz.



MECHANICAL CONSTRUCTION

In its simplest form, the hemispherical cup assembly may be made from a rotor with three equidistant (120°) arms to each of which half a table tennis ball is glued (see Figures 1 and 4). A small magnet should be glued to the central axis, whereupon the Hall-effect sensor is mounted so that the distance between it and the small magnet is an absolute minimum.

For good and accurate performance, it is, of course, essential that the spindle moves as frictionless as possible. In the prototype, use is made of a roller (ball) bearing, combined with a magnetic trunnion to remove any axial pressure. If the magnets are not strong enough to bear the weight of the cup assembly, two or three on top of each other may be used.

In the prototype, hemispherical cups with a diameter of 47 mm are used. These are simply bolted to a special flange that serves as cover. The assembly is shown in diagrammatic form in **Figure 4**. The most important aspects are that the bearing is of good quality and that the Hall-effect sensor is positioned correctly.

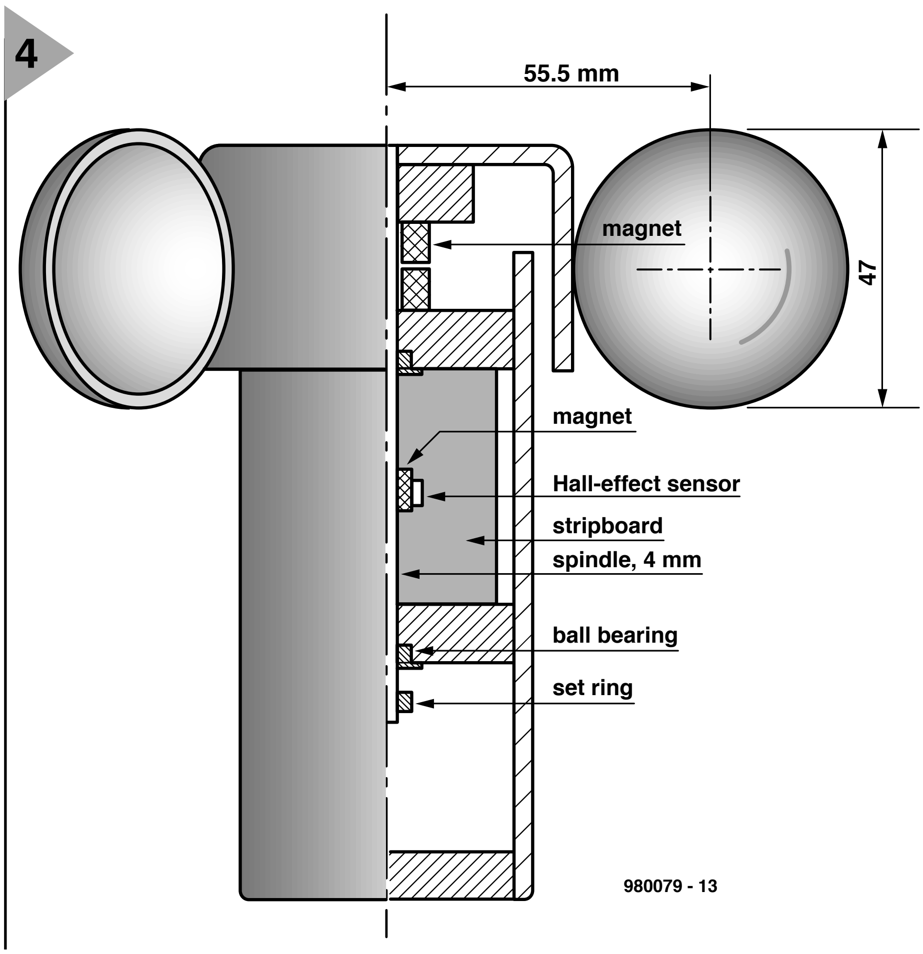
CALIBRATION AND INSTALLATION

The instrument can be set up correctly only with reference to a calibrated anemometer. Start by placing pin jumper JP_2 (leaving the other two open) and check that the correct value can be obtained by adjusting P_1. If the measurement is too large, remove JP_2 and place JP_1 or JP_3. Again adjust P_1 until the correct value is obtained. In the rare instance that the instrument can not be calibrated in this manner, it may be necessary to change the values of R_4 and C_1 to some extent.

Positioning the instrument correctly is, of course, of paramount importance. High buildings in the immediate vicinity lead undoubtedly to poor and unpredictable results. It is advisable to place the hemispherical cup assembly as high as possible on a sturdy mast. Strictly speaking, an anemometer should be placed at a height of 10 metres (33 ft) in an open area.

[980079]

Figure 4. A roller bearing is used in the prototype to ensure frictionless transfer of the windpower into rotational speed of the magnet.



RF signal generator

part 2 (final): construction, operation and adjustment

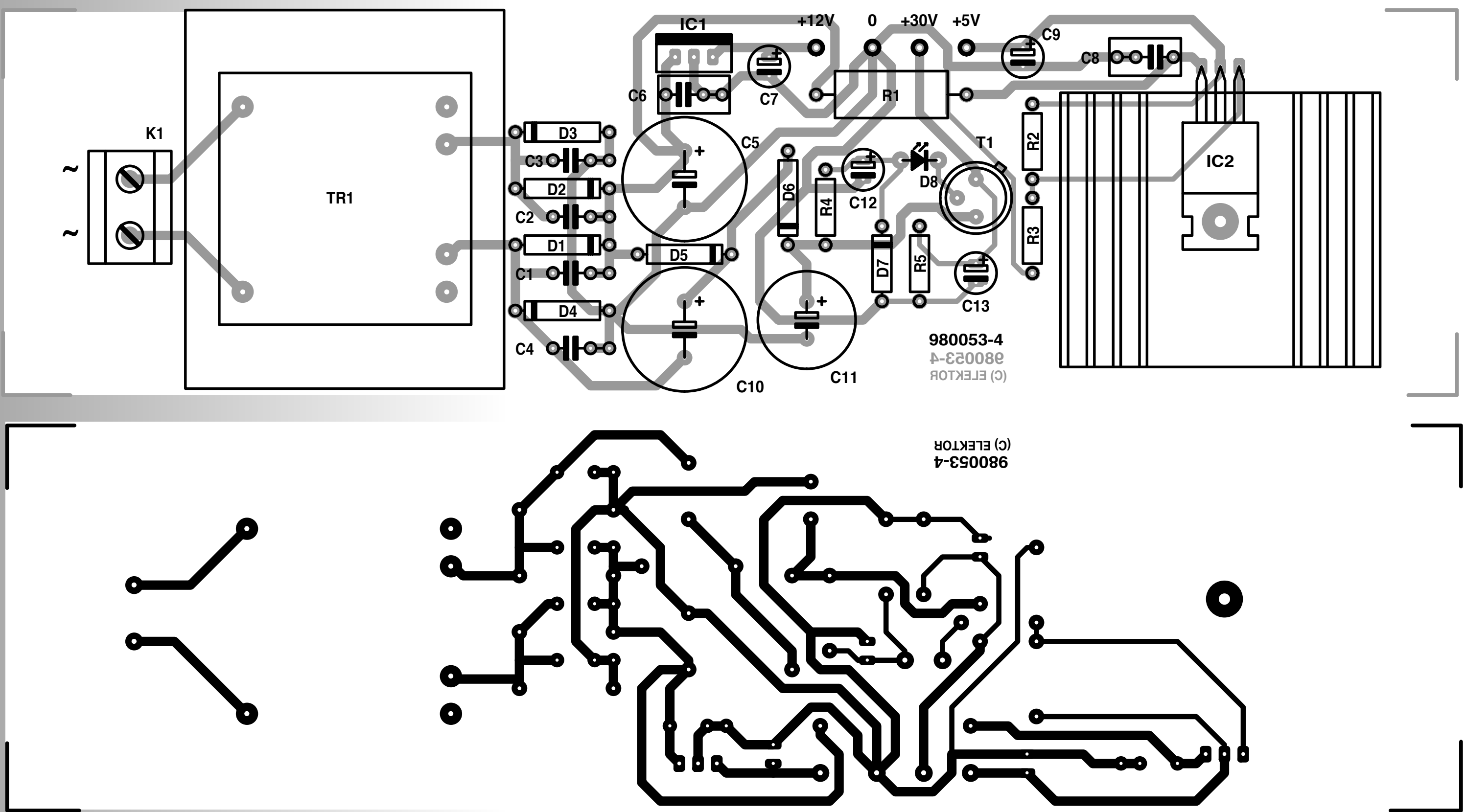


Figure 6. Copper track layout and component overlay of the power supply board.

Although the main subject of this month's second and final instalment is 'all matters constructional', there's also information on adjusting the instrument and, of course, on how to use it!

Design by G. Brunner

The RF signal generator is a quite complex instrument, and we should really advise beginners not to attempt to build this project without the help or guidance of someone with considerable experience in building RF and microcontroller circuits.

There are no fewer than four boards to build up, and each of these boards contains a fair number of components. Add to that the mounting of the four boards in a case and the inter-board

COMPONENTS LIST

POWER SUPPLY BOARD

Resistors:
R1 =22Ω 5W
R2 =270Ω
R3 =820Ω
R4 =1kΩ
R5 =10kΩ

Capacitors:
C1-C4 =47nF
C5 =1000µF 35V radial
C6,C8 =220nF MKT
C7,C9 =2µF2 16V radial
C10 = 470µF 63V radial
C11 = 220µF 63V radial
C12 = 1µF 63V radial
C13 = 10µF 63V radial

Semiconductors:
D1-D6 = 1N4001
D7 = 33V 400mW zener diode
D8 = LED, red, high efficiency
T1 = BC141
IC1 = 7812
IC2 = LM317T

Miscellaneous:
TR1 = mains transformer, 15V 8VA, Monacor/Monarch type VTR8115
K1 = PCB terminal block, 2-way, raster 7.5mm
K2 = mains socket, integral switch and fuseholder, with fuse 63mAT
Heatsink type SK59 37.5mm (Fischer, Dau Components)
PCB, order code 980053-4 (see Readers Services page)

7

Figure 7. Finished PSU board (prototype).

wiring, and you are looking at a project which should take even advanced hobbyists several hours, winter evenings or rainy Sunday afternoons to complete.

The four boards are built up one by one in the order indicated by the text to follow. As usual, great care should be taken to fit each and every part in the right position on the board. The component overlays and associated parts list should guide you through the process of assembling the boards. Particularly with the 1% resistors in the attenuator section, you should (1) ascertain the value and (2) look up the position on the board, before (3) fitting *any* resistor.

POWER SUPPLY BOARD

This board is the simplest to build. Populating it should be straightforward, using the relevant Components List and the component overlay shown in **Figure 6**. Resistor R1 may run fairly hot and should not touch the circuit board. The LM317T voltage regulator may be mounted directly on to the heatsink — an insulating washer is not required. The 'power on' LED is not fitted directly on the board — instead, it is connected up via a pair of thin wires with an length of about 20 cm.

This board is simple to test by provisionally connecting it to the mains and using a voltmeter to check the indicated output voltages: +5 V, +30 V and +12V. The finished PSU board is shown in **Figure 7**. Check your work against this photograph!

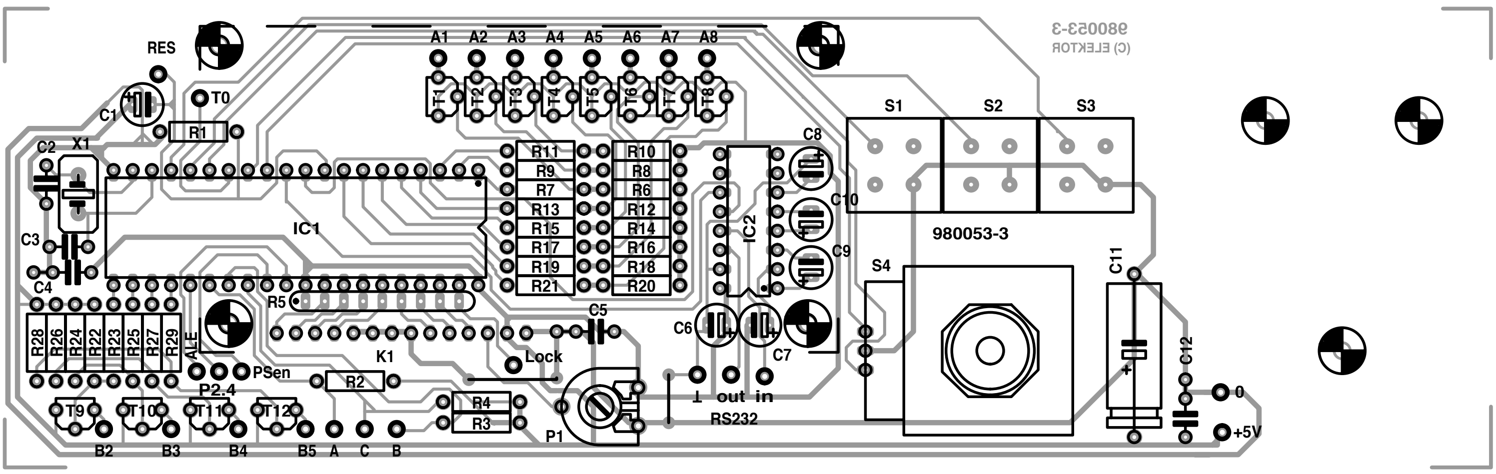
CONTROLLER BOARD

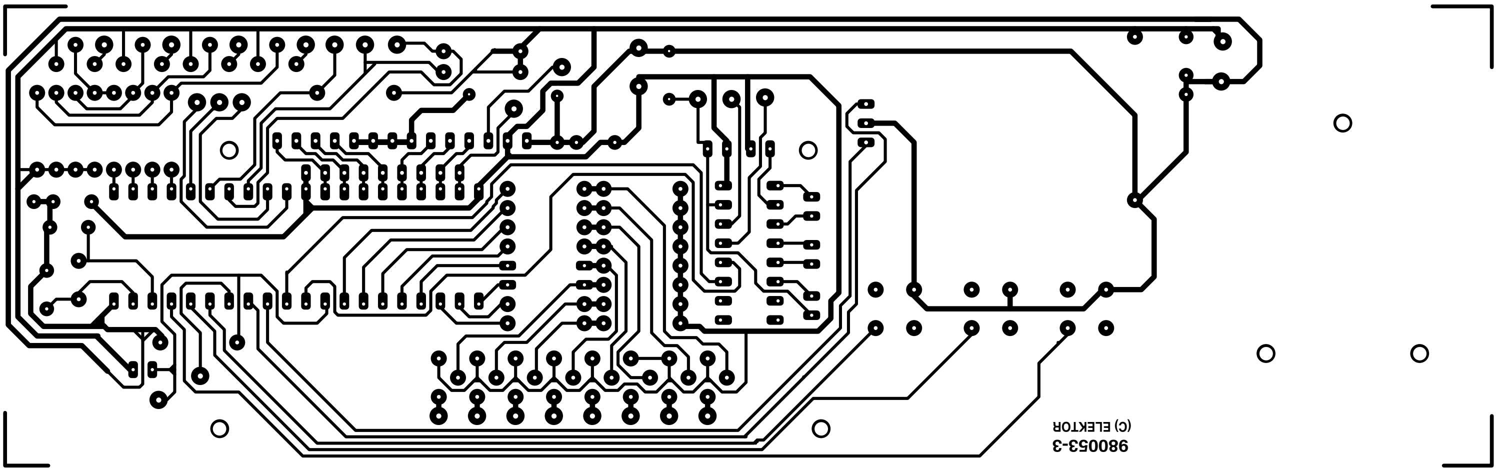
The controller board shown in **Figure 8** is far more densely populated than the PSU board. Hence, great care and precision is required when it comes to soldering the parts in place.

Start with the two wire links on the board — you'll find them near preset P1. Next, fit the components, the best order is probably from low-profile parts (resistors, IC sockets) to upright mounted parts (crystal, transistors, radial electrolytic capacitors).

The three push-buttons, S1, S2 and S3, are not mounted directly on to the board. Their pins are inserted in socket strips or stacked IC sockets so that their height can be adjusted a little. Alternatively, their pins are 'lengthened' using pieces of stiff wire. This is necessary to enable the cap tops to protrude a little through the front panel. The same mounting method is used for LCD. As with the push-buttons, the height of the LCD

Figure 8. Controller board artwork.





COMPONENTS LIST

CONTROLLER BOARD

Resistors:
R1 = 22kΩ
R2,R3,R4 = 4kΩ7
R5 = 10kΩ 8-way SIL array
R6,R8,R10,R12,R14,R16,R18,R20,R22,R24,R26,R28 = 1kΩ
R7,R9,R11,R13,R15,R17,R19,R21,R23,R25,R27,R29 = 3kΩ3
P1 =10kΩ preset, H

Capacitors:
C1 =1μF 16V radial
C2,C3 =33pF
C4,C5,C12 =100nF ceramic
C6-C10 =10μF 63V radial
C11 = 220μF 16V

Semiconductors:
T1-T12 = BC557B
IC1 =AT89C51-20PC or SC87C51CCN40 (order code 986515-1)
IC2 = MAX232

Miscellaneous:
X1 = 11.059MHz crystal
S1,S2,S3 = pushbutton, 1 make contact, ITT type D6-R-RD; cap type D6Q-RD-CAP (Eurodis)
K1 =14 way SIL pinheader
K2 =9-way sub-D socket (female)
S4 = rotary encoder, Bourns type ECW1J-B24-AC0024 (Eurodis)
LCD, 2x16 characters, Sharp type LM 16A211 (Eurodis)
PCB, order code 980053-3 (see Readers Services page)

above the controller board may need to be adjusted later, so do not mount it securely as yet. The rotary switch encoder, S4, is mounted directly on to the board, but its spindle is not yet cut off. Later, rectangular clearances are cut in the front panel to allow the LCD to be viewed, and the push-buttons to be pressed.

It is recommended to use sockets for IC1 and IC2. All holes in the PCB with a label printed near it (like A1, T0, Psen, Lock, etc.) are for inter-board wires. Solder pins are not strictly necessary — direct wire connections to the board are also fine. As with the PSU board, check your work against our fully working prototype. This time, refer to the photograph in **Figure 9**. The board is fitted vertically behind the metal front plate (which has to be purchased separately). It is held in position by a pair of slots moulded on the bottom plate of the case. Several slots are available, and the pair you actually choose to use should ensure that the metal frame around the face of the LCD is pressed firmly against the inside of the front panel. The three type 'D6' push-buttons should then protrude a little from the front panel.

The holes marked 'In', 'Out' and 'ground' to the right of preset P1 are for an *optional* 3-wire RS232 link to a PC. If you do not require PC control, the MAX232 may be omitted. The practical use of the RS232 interface will be reverted to further on.

VFO/PLL BOARD

As you can see from the PCB artwork in **Figure 10**, this is the board with the highest component density of all four. Care and precision are essential if you want to avoid a tedious faultfinding session. Identify and check each part before fitting it, and double-check its value and position using the Components List and the component overlay.

As usual, start with the wire links (there are three), so they are not forgotten or overlooked. Then follow the low-profile parts and, finally, the vertically mounted parts. IC sockets should *not* be used for the NE592 and the SAA1057 on this board.

The value of the inductors is usually printed on these parts in the form of colour bands (like resistors) or dots.

The PLL/VFO board is fitted in a tinplate enclosure from Teko. After the solder work, inspect the board, and compare yours with our prototype shown in **Figure 11**.

Figure 9. Finished controller board (prototype).

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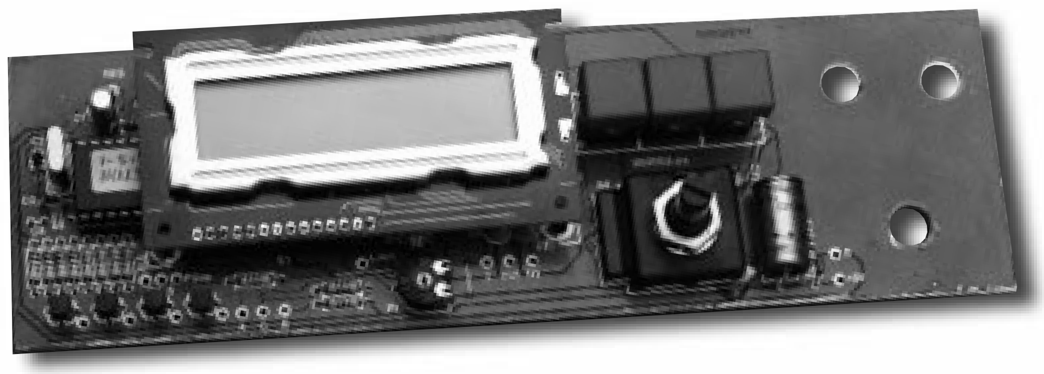

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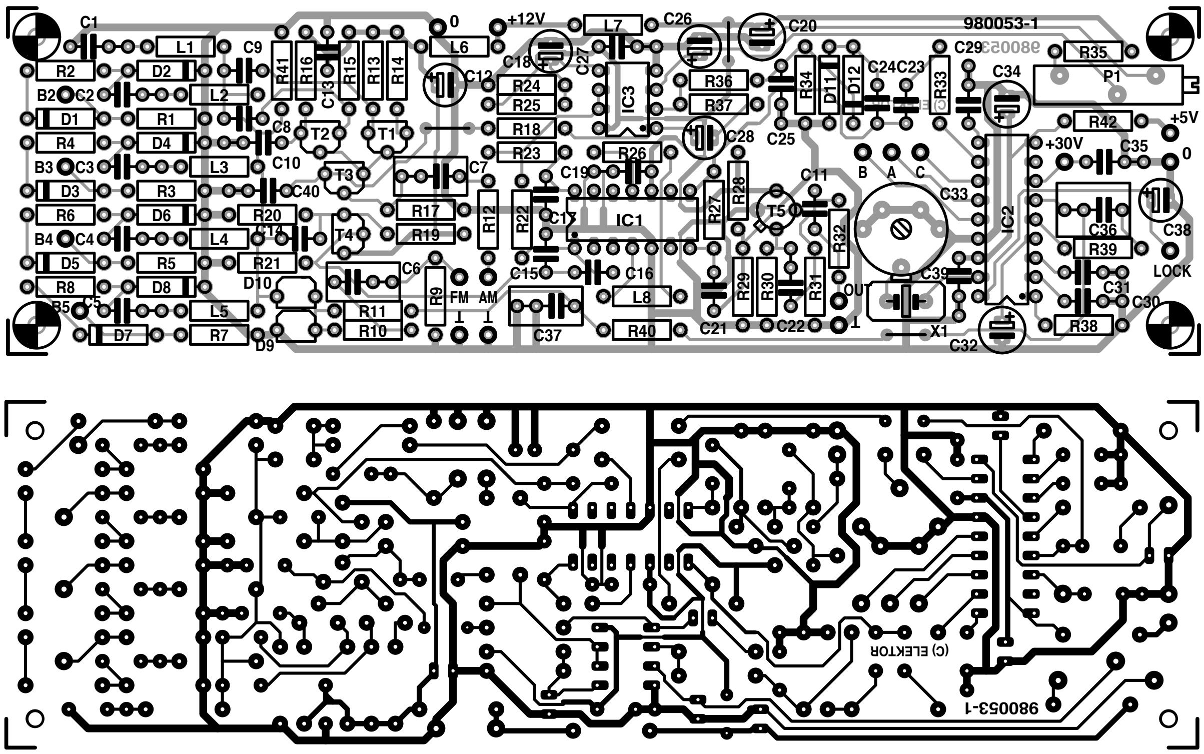

Figure 10. VFO/PLL PCB design.

Attenuator board

The main point to mind about assembling the attenuator board (**Figure 12**) is that each close-tolerance (1%) resistor goes to the right position on the board. One error in this respect may cause wrong attenuation levels later, with possibly difficult to explain behaviour of some of the radio equipment you may be aligning! Our advice is, therefore: read the Components List carefully, check the colour code, use a DMM to measure the value of each resistor, and then check its position on the board.

The attenuator board has relatively large copper areas to assist in screening and preventing unwanted signals from being generated and picked up by the circuit. The attenuator board is shown in **Figure 11**, together with the VFO/PLL board. For RF screening purposes, both boards are fitted in Teko tinplate cases.

Adjustment

The boards may be wired up experimentally for an initial test and a few adjustments.

To begin with, set the two presets and the trimmer to the centre of their travel. It is assumed that the power supply board has been tested already (with good results, of course).

After applying power, the first thing to do is set the LCD contrast with preset P1. Next, use an oscilloscope to check that the VFO/PLL board supplies an RF signal to the attenuator board.

The output frequency supplied by the generator may be checked with a calibrated frequency meter, a frequency standard (off-air Rugby MSF or similar) or a calibrated SW receiver (zero-beat). The relevant adjustment is trimmer capacitor C33.

COMPONENTS LIST

VFO/PLL BOARD

Resistors:
R1,R3,R5,R7,R12,R22,R23,R31,R34,R36,R37 = 10kΩ
R2,R4,R6,R8 = 390Ω
R9,R14,R15,R21,R27,R33,R40 = 1kΩ
R10,R41 = 330kΩ
R11,R13,R16,R18 = 100kΩ
R17,R26 = 100Ω
R19 = 2MΩ2
R20 = 1MΩ
R24,R25,R35 = 22kΩ
R28,R29 = 3kΩ3
R30 = 560Ω
R32 = 47Ω
R38 = 180Ω
R39 = 18kΩ
R42 = 10Ω
P1 = 2kΩ multiturn preset, H

Capacitors:
C1-C5,C10,C22 = 33nF ceramic
C6,C25,C30 = 2nF2 ceramic
C7 = 220nF MKT
C8,C9,C16,C23,C24 = 330pF ceramic
C11 = 68pF ceramic
C12,C18,C38 = 10µF 63V radial
C13 =100nF ceramic 5mm
C19,C27,C35,C39,C40 =100nF ceramic
C14 = 180 p ceramic
C15 = 27 p ceramic
C17 = 33n ceramic 5mm
C20,C32 = 47µF 16V radial
C21 = 4n7 ceramic
C26,C28 = 2µF2 16V radial
C29,C31 = 10nF ceramic
C33 = 40pF trimmer
C34 = 100µF 10V radial
C36 = 1µF MKT
C37 = 330nF MKT

Inductors:
L1 = 330µH
L2 = 100µH
L3 = 22µH
L4 = 3µH9
L5 = 0µH56
L6,L7 = 39µH
L8 = 3µH3

Semiconductors:
D1,D3,D5,D7 = 1N4148
D2,D4,D6,D8 = BA243
D9,D10 = BB130
D11,D12 = AA113
T1,T2,T3 = BF494
T4 = BF256B
T5 = 2N5179
IC1 =NE592N (N14)
IC2 = SAA1057 (Philips)
IC3 = LM358P

Miscellaneous:
X1 = 4MHz
Tinplate case, Teko, size 160x25x49mm
Case, Bopla type Ultramas UM52011 (size 224x72x199mm)
Front panel type FP50011 or FPK50011
PCB, order code 980053-1 (see Readers Services page)

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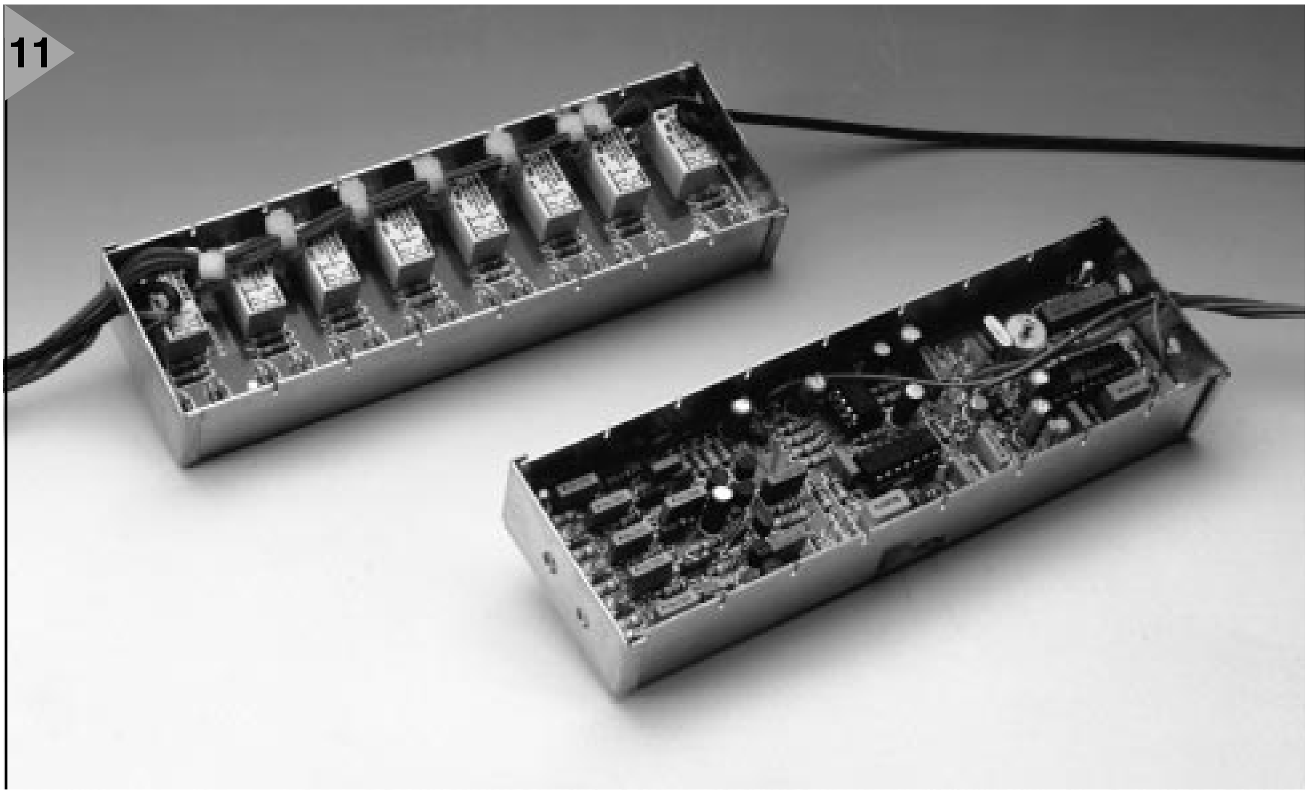


Figure 11. Finished PLL/VFO board (below) and attenuator board (above), both fitted in 'Teko' ready-made tinplate cases.

Adjustment of the RF signal level is only possible if you have an accurate and calibrated RF voltmeter. With the attenuation set to 0 dB, preset P1 may be adjusted for an output level of 630 mV_{pp} into 50 Ω at the generator output. Failing the necessary test equipment, you may leave the multi-turn preset at mid-travel.

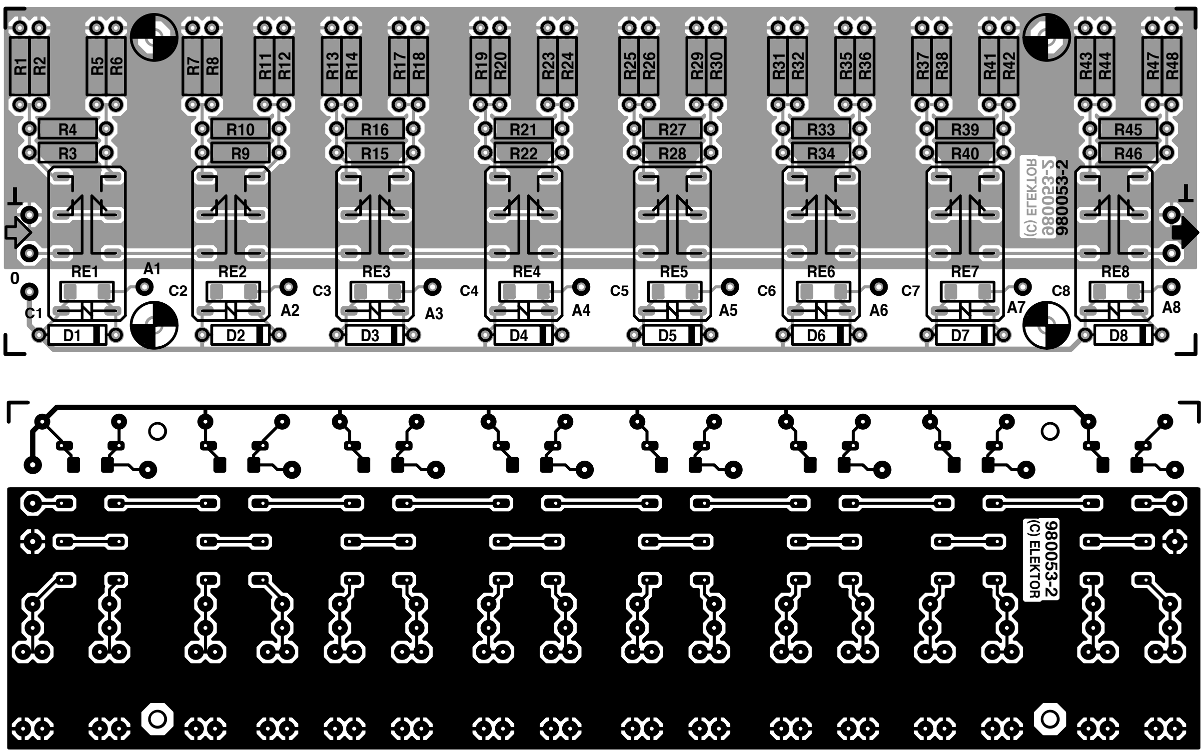
Wiring and mechanical work

Although there are quite a few wire connections between the boards, there are no special precautions in this respect. The RF signal connection between the PLL/VFO board and the attenuator board must, of course, be made in coax cable. The same goes for the connections between the AM and FM inputs on the PLL/VFO board and the associated BNC sockets on the front panel. If you can get hold of it, use the 3-mm dia. type RG174/U, else, the much thicker RG50/U or /CU is a good alternative.

All other inter-board connections are made in light-duty flexible wire or flatcable, although slightly thicker wire should be used for the 0-V, 5-V and 12-

Figure 12. Attenuator PCB artwork.

12



COMPONENTS LIST

ATTENUATOR BOARD

Resistors (all 1%):
R1,R5,R21 = 909Ω
R2,R6 = 20kΩ
R3 = 6Ω81
R4 =39Ω2
R7,R11 = 475Ω
R8,R12 = 6kΩ19
R9 = 368Ω
R10 = 12Ω1
R13,R17 = 243Ω
R14,R18 = 2kΩ74
R15,R20,R24 = 3kΩ65
R16 = 24Ω3
R19,R23 = 121Ω
R22 = 56Ω2
R25,R29,R31,R35,R37,R41,R43,R47 = 75Ω
R26,R30,R32,R36,R38,R42,R44,R48 = 825Ω
R27,R33,R39,R45 = 3kΩ92
R28,R34,R40,R46 = 162Ω

Capacitors:
C1-C8 = 100nF SMD

Semiconductors:
D1-D8 = 1N4148

Miscellaneous:
RE1-RE8 = relay, 2 x change-over, type V23042-A1001-B101 or V23042-A2001-B101 Siemens (Eurodis, ElectroValue)
PCB, order code 980053-2 (see Read-

V supply wiring. Do not make any of the wires longer than necessary to prevent digital noise being picked up from the controller board.

The wires and the coax cables to and from the PLL/VFO board and the attenuator board should pass through holes drilled in the short side panels of the Teko tinplate cases. Once these boards are fully operational, the top covers are fitted for optimum RF screening.

Guidance for mounting the four boards into the Bopla enclosure may be obtained from the photographs in this article, and in particular, **Figure 13**. Note that the solder side of the power supply board is protected by a perspex cover plate cut to roughly the same size as the board. The VFO/PLL and attenuator boards are screened by tinplate boxes, and mounted horizontally on to the bottom plate of the enclosure. As already mentioned, the PSU board is fitted vertically, using a pair of the moulded PCB slots towards the back panel. The three holes at the 'empty' right-hand side of the controller board are drilled to a diameter of about 8 mm to allow the coax cables to the three front-panel mounted BNC sockets to pass.

The mains voltage is switched on and off by a double-pole switch integrated into a mains socket fitted onto the plastic rear panel of the enclosure. The wires between the mains socket/switch combination and the PCB terminal block on the PSU board should be mains-rated and properly isolated. At the PCB side in particular, the 'live' and 'neutral' wires should not be stripped longer than strictly necessary, and they should be inserted into the clamps right up to the insulation. Finally, once the wires are connected, the terminals on the mains socket/switch combination must be insulated using heat-shrink sleeving.

The metal front panel is cut, drilled and lettered using the template shown in **Figure 13**. This front panel foil is not available ready-made.

In the (ABS plastic) back panel, you have to cut rectangular clearances for the mains socket/switch combination and, optionally, for the RS232 connector (a 9-pin sub-D type).

Figure 13. A look inside our prototype of the RF Signal Generator. The covers of the tinplate cases of the VFO/PLL board and the attenuator board were removed for this photograph.

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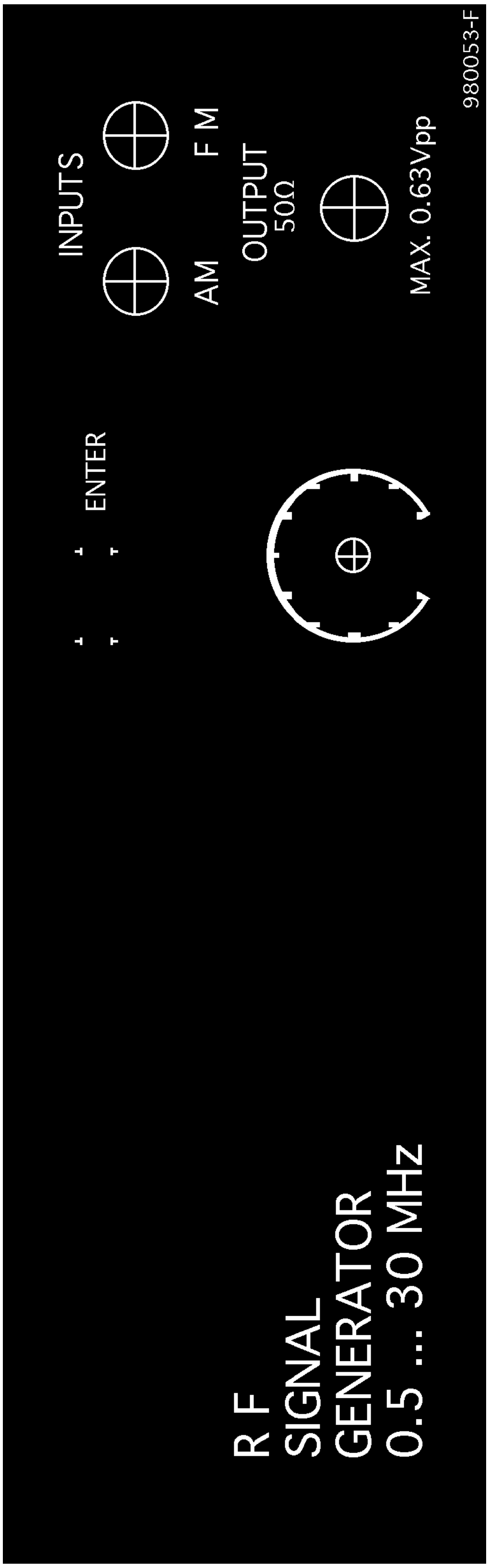


Figure 14. Suggested front-panel layout. Use it as a template to drill the metal front panel of the instrument, and apply the lettering/symbols.

Operation

The instrument is controlled by means of three pushbuttons and a rotary encoder, all accessible on the front panel. The instrument communicates with you via an LCD with two lines of 16 characters.

The functions of the 'left' and 'right' pushbuttons are self-evident, we reckon, because they move the cursor on the LC display in the direction indicated by the arrows on the front panel

From the starting position (cursor on 'MHz'), the cursor may be moved to the left on to any of the post-decimal positions of the frequency. The number at which the cursor arrives may then be changed by turning the rotary encoder. The frequency set in this way is however not actually generated until you press the 'Enter' pushbutton (asynchronous operation, this is indicated in the upper right-hand corner of the display). After any frequency change, the PLL status is indicated by 'lock' in the left-hand bottom corner of the readout.

From the initial position to the right, the cursor jumps to 'M0' (memory 0). This indicates two memories, M0 and M1, in which frequency and attenuator settings may be stored. You press the Enter key to change between these memories. In this way, you can quickly change between two previously stored settings, which may be useful, for example, for aligning a filter. Alternatively, you may use the same frequency twice, but with two different attenuator settings. This facility is useful for adjusting, say, a receiver AGC (automatic gain control).

Moving further to the right, the cursor jumps on 'asy'. Here you can switch to asynchronous operation by pressing 'Enter'. In synchronous mode, any frequency change requested by way of the rotary encoder is immediately passed on to the VFO/PLL unit. In this mode, the RF output frequency is continuously adjustable, but only within the selected range (one of five). If you turn the encoder to a frequency outside a certain range, the PLL will drop out of lock, and the 'lock' indication will disappear from the LCD. By pressing any key, the PLL is returned to asynchronous mode, and the last selected frequency is automatically restored. If you then move the cursor to a decimal digit of the frequency readout, and press the Enter key, the generator changes to the relevant frequency range, allowing you to change to synchronous mode again and continue 'tuning' again using continuous frequency variation.

One more position to the right, the cursor reaches the 'dB' position, indicating the currently valid attenuation. The desired attenuation may be set with the aid of the rotary encoder. As with the frequency setting, the desired attenuation becomes effective only when you press the Enter key. This is done to reduce wear and tear on the relays.

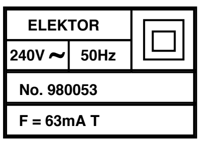
Optional RS232 interface

The RS232 interface on the controller board is an optional extension whose function has not been fully developed out by the author/designer. Basically, it was designed into the circuit to enable the generator frequency and output signal attenuation to be controlled by a PC.

The communication parameters are as follows: 9600 bits/s, 8 data bits, 1 stop bit. The communication works with character strings, and is easily tested with the aid of a terminal program. To set the frequency you have to send an 'F' (for 'frequency'), then five numbers for the frequency in kilohertz, and, finally, a carriage-return (CHR$(13)). An additional Line-Feed (CHR$(10)) will be ignored. If everything is correct (first character is 'F', a total of 6 characters and the frequency in the right range), the controller returns a 'D' (for 'done'), followed by a CR-LF sequence, otherwise, an 'E' (for 'error') and a CR-LF.

The attenuation is set by sending an 'A', two numbers and CR. Again the controller answers as described. The main purpose of the serial interface was to create a basis for using the generator in an environment like LabView™.

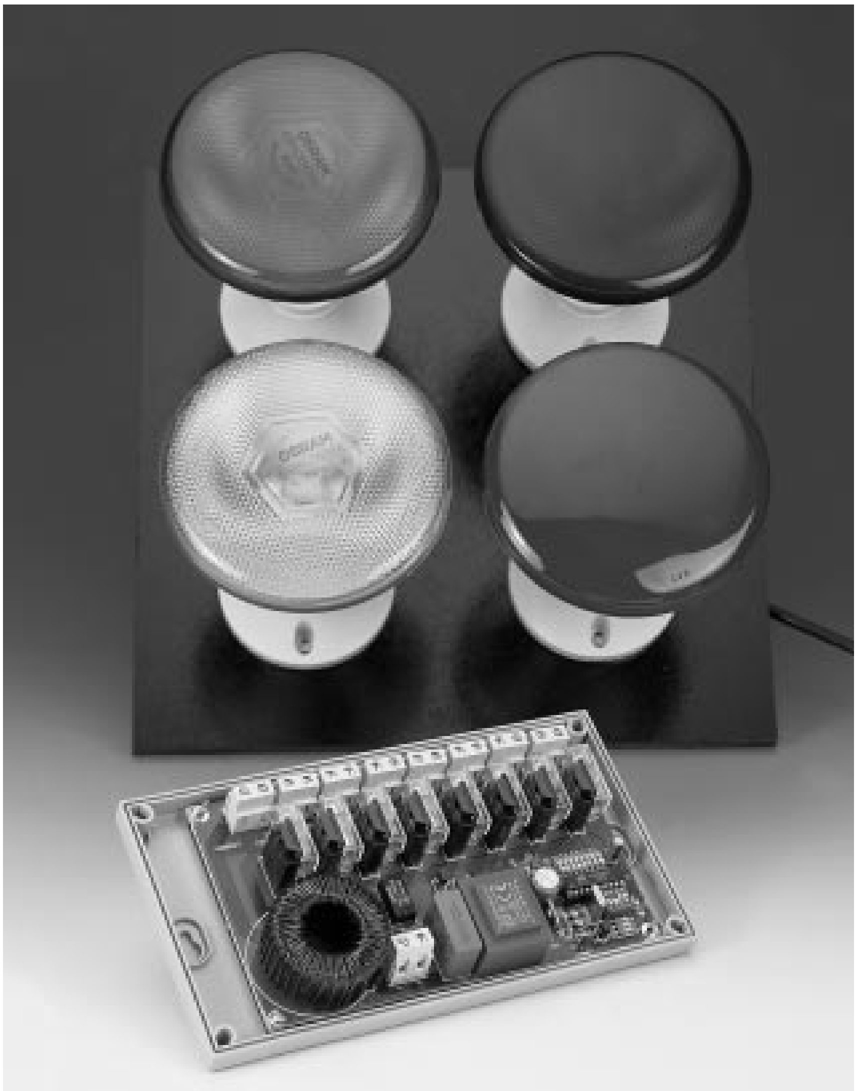
(980053-2)



32-channel PC-controlled light dimmer

de luxe controller for ohmic and inductive loads

Over the years this magazine has published a variety of light dimmers. Invariably, these were designs based on discrete components. This time, the design is a more advanced one that may be controled with the aid of a computer. Each channel is capable of controlling ohmic as well as inductive loads rated at up to 300 watts.



Hardware design by H. Bonekamp
Software design by G. Janssen

INTRODUCTION

To most people a light dimmer is a gadget with a rotary control that takes the place of the usual light switch. The dimmer described in this article is rather different. The usual standard components are replaced by a microcontroller driven by a computer program. The design may consist of 8–32 channels, each of which can be individually controlled with Windows software.

1

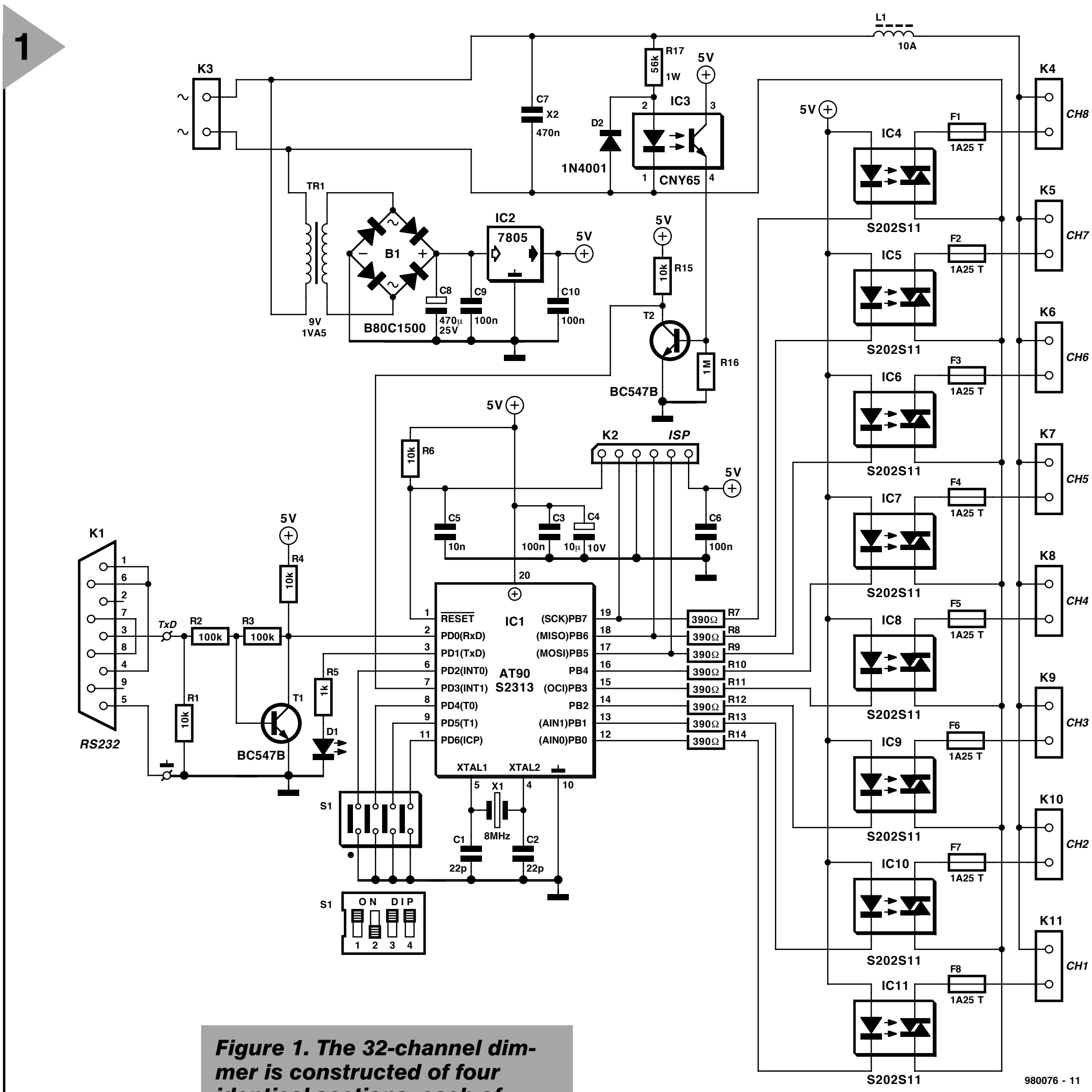


Figure 1. The 32-channel dimmer is constructed of four identical sections, each of which supports eight channels. One section is shown here.

The dimmer may be used for stage lighting in a small or home theatre, for controlling domestic lights, or for illuminating an aviary or aquarium. Another use is the controlled but random switching on and off of domestic lighting during the residents's absence as a burglar deterrent.

Since the design can handle inductive loads, the dimmer may also be used for controlling motors, halogen lamps and transformers.

The unit can control up to 32 individual loads for which it contains four modules, each with eight controllers. It may, of course, be used with just one, two or three modules.

DESIGN SPECIFICATION

The original requirements laid down that the design must be of the open type, that is, it must be possible for individual users to develop a program or alter an existing one for controlling the dimmer. There is a standard electronic mixer for 32 (light) channels available that works under Windows 95.

Communication between the computer and dimmer takes place via the usual RS232 link on the basis of a compact and clear protocol. This will be reverted to later.

The instructions sent by the computer to the dimmer include the final value of the setting and fade time per step. Dimming occurs of course in synchrony with the mains voltage. The controller stores, of each and every

Features:

- *controls eight loads per module*
- *up to four modules per serial gate*
- *suitable for ohmic and inductive loads*
- *switches in synchrony with mains*
- *operates with simple protocol*
- *four integral control curves*
- *simple to program*
- *batch operation possible*
- *up to 300 watt loading per channel*
- *uses Windows 95 software*
- *all source codes available*

1

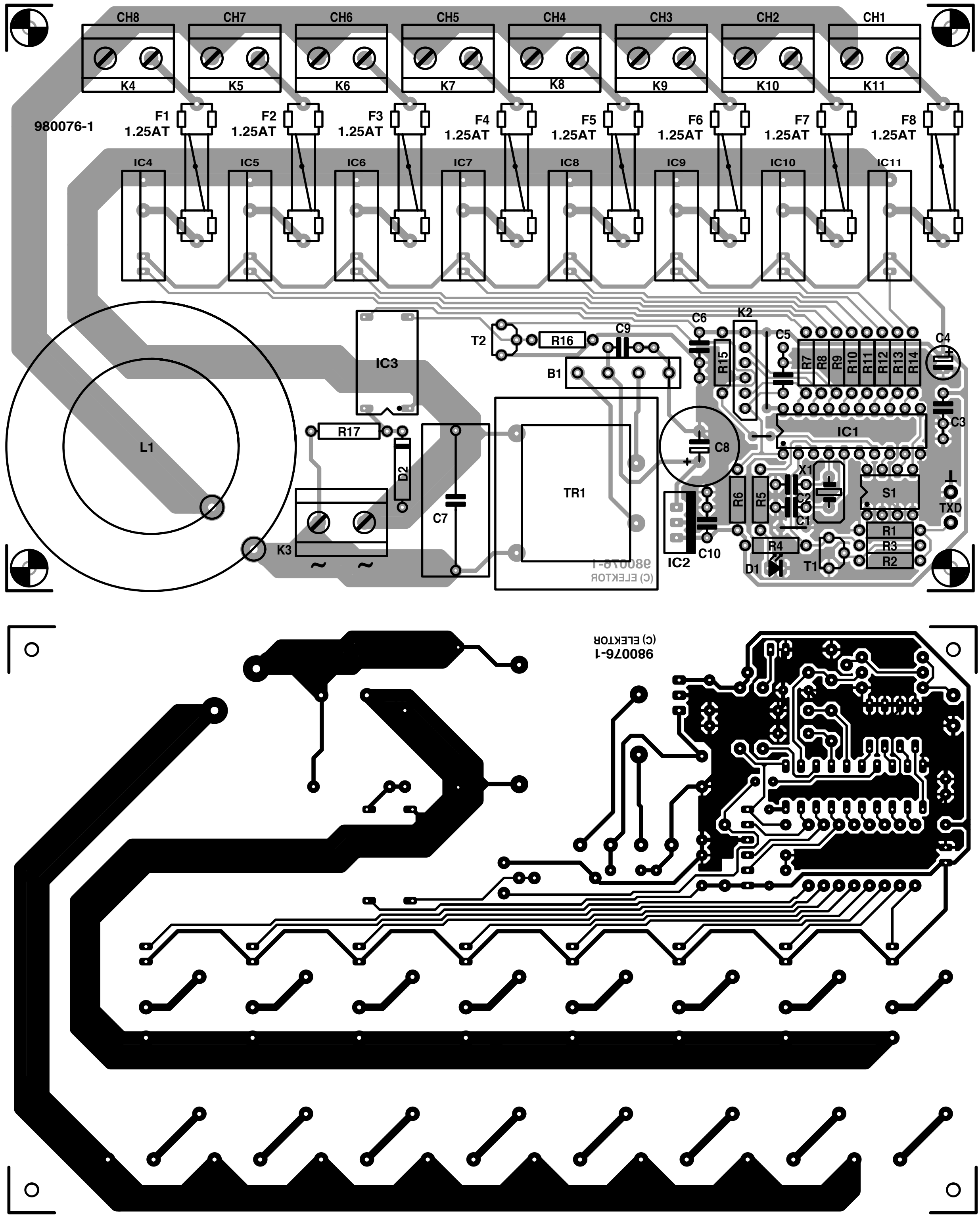


Figure 2. The printed-circuit board for a single section of the 32-channel dimmer. To be able to use all 32 channels, four of these boards are needed.

channel, the set maximum value, the fade time and the actual value. Every time there is an alteration, the computer sends a new set of parameters. This arrangement ensures that there is relatively little need for the exchange of information between the computer and the controller.

CIRCUIT DESCRIPTION

It is clear from the circuit diagram of one of the four modules in **Figure 1** that the key component of the design is a Type AT90S2313 microcontroller (from Atmel). The controller, which is linked to the computer via a TxD line, drives optotriacs IC_4–IC_{11} via pins PB_0–PB_7. One set of terminals of the optotriacs is connected to the mains voltage, whereas the set of terminals at the other side is switched to earth via the I/O lines of the controller and a 390 Ω resistor.

The optotriacs are protected at the mains voltage terminals by a fuse rated at 1.25 A. At a mains voltage of 240 V, this means that loads of up to about 300 watts can be controlled.

Since the optotriacs incorporate a snubber network, the dimmer can be used for controlling inductive loads.

Optoisolator IC_3 serves to detect the mains zero crossing. The zero crossing is used for synchronizing the controller. The optoisolator is linked directly to the mains supply since the transformer causes a small phase shift that may cause errors in

Parts list

Resistors:
R_1, R_4, R_6, R_{15} = 10 kΩ
R_2, R_3 = 100 kΩ
R_5 = 1 kΩ
R_7–R_{14} = 390 Ω
R_{16} = 1 MΩ
R_{17} = 56 kΩ, 1 W, 400 V

Capacitors:
C_1, C_{22} = 22 pF, ceramic
C_3, C_6, C_9, C_{10} = 0.1 μF, ceramic
C_4 = 10 μF. 10 V, radial
C_5 = 0.01 μF, ceramic
C_7 = 0.47 μF, 250 V a.c., Class X2
C_8 = 470 μF, 25 V, radial

Inductors:
L_1 = suppressor coil, 10 A, Type T60405M6108X2 (Siemens)

Semiconductors:
B_1 = rectangular rectifier bridge Type B80C1500
D_1 = LED, red, high efficiency
D_2 = 1N4001
T_1, T_2 = BC547B

Integrated circuits:
IC_1 = AT90S2313 (Order no. 986524-1 – see Readers Services towards the end of this issue)
IC_2 = 7805
IC_3 = CNY65
IC_4–IC_{11} = S202S11 (Sharp)

Miscellaneous:
X_1 = crystal, 8 MHz
S_1 = 4-section DIP switch
K_1 = 9-pole female sub-D connector
K_2 = 6-pole SIL header
K_3–K_{11} = two-way terminal block, pitch 7.5 mm
Tr_1 = mains transformer, secondary 9 V/1.5 A
F_1–F_8 = fuse holder with 1.25 slow fuse
Enclosure Bopla EG2050L (available from Phoenix Tel. 01296 398 355)
PCB 980076–1*
Windows 95 software, incl. source code, EPS 986025*
Source code for programming Atmel controller: EPS 986033–1*
Programmed controller EPS 986524–1*

* (see Readers Services towards the end of this issue)

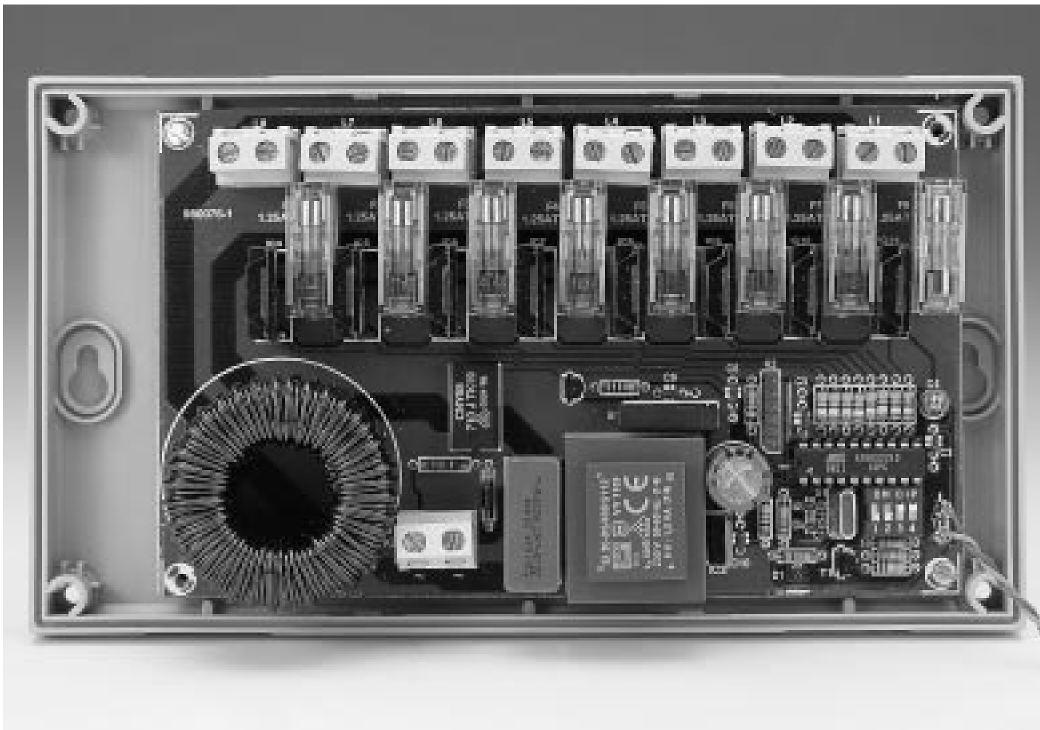


Figure 3. The completed prototype of a single section.

the measurements.

The synchronizing pulse is buffered by transistor T_2 and applied to input PD_3 (INT1).

To ensure correct operation, the dimmer should be connected to a frequency-stable mains supply only.

DIY PROGRAMMING

As the controller is in principle InCircuit Programmable, the printed-circuit board incorporates a special ICP interface. This interface is linked to pins RESET, SCK, GND, MIS0, MOS1, and V_{cc}, and to six-pole header K_2. The interface enables the user to program the controller.

Some pins of the controller are linked to DIP switches, which enable the user to select a number of options. This will be reverted to later.

The serial interface has been kept simple. The controller uses only the TxD signal of the computer. The RS232 voltage level is converted to TTL level with the aid of a transistor and some resistors and then inverted. No other RS232 lines need to be linked in the connector, provided only the programme supplied is used. If a self-written program is used, some other lines may well need to be linked.

The clock generator, C_1-C_2-X_1 provides a clock of 8 MHz.

The circuit contains all the electronics for controlling up to eight loads. The available Windows 95 software is arranged so that four of these circuits can be driven simultaneously.

CONSTRUCTION

The circuit is best built on the printed-circuit board shown in **Figure 2**. It will be seen that in spite of the versatility of the unit, the board has been kept compact.

Most of the work is straightforward as long as the specified components are used. This is particularly true of the suppressor circuit, L_1-C_7. The capacitor must be a Class X2 component.

The inductor consists of two parallel wound windings that are twisted at the ends. Untwist them about 1 cm, remove the enamel from the ends, push them through the relevant holes in the board and bend them. It is important that they are soldered over a larger length than usual to ensure good electrical contact. Note that at maximum load, a current of up to 10 A flows through the coil: a good reason for ensuring good contacts.

The finished prototype board is shown in **Figure 3**.

Readers who have obtained the ready programmed controller via the Readers Services (towards the end of this issue) can start work immediately.

Insert the controller into its socket and assemble the board in the specified plastic case.

Since several tracks carry the full mains voltage, extreme care is required in the assembly. Always unplug the unit from the mains before doing any work or checking something after assembly.

Finally, make the serial link with the computer. After the Windows software (only 95 or 98) has been installed, the dimmer is ready for use. The intuitive user interface (see the screendump in **Figure 4**) readily points the user into the right direction.

Readers who wish to extend the program can do so right away because the program is supplied with the source code (in Delphi format).

HANDYMAN PROGRAMMER

As mentioned earlier, it is possible to programme the controller to one's own wishes and requirements. This is possible because the diskette containing the Windows software also contains the source code and the ROM file of

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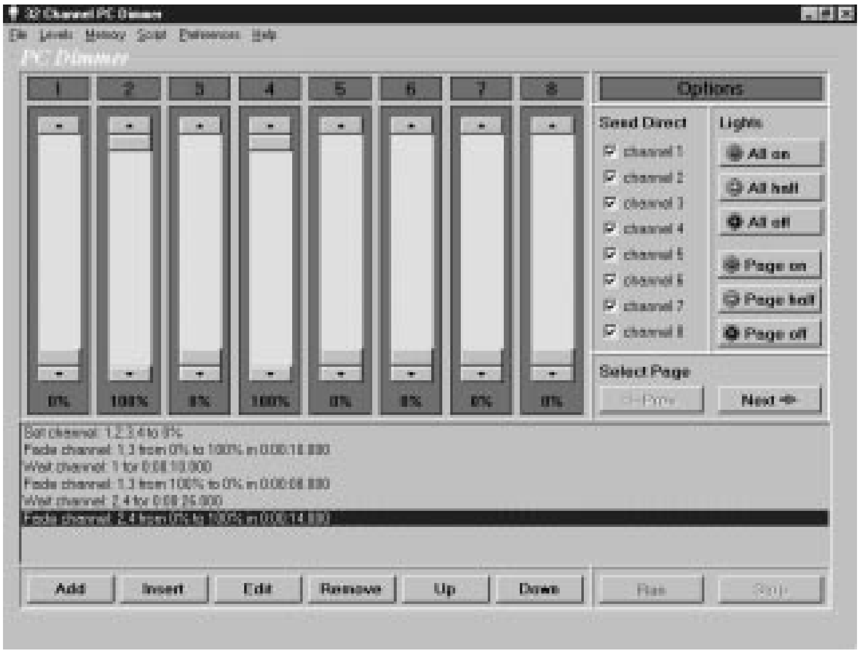


Figure 4. The software available through the Readers Services enables the easy control of up to 32 channels via Windows 95 or Windows 98.

the controller program. Moreover, the board has a special ICP connection.

The Handyman Programmer featured in the December 1997 issue of this magazine may be used for this purpose once an adaptor lead is made to link the 10-way box header to the single-row header on the board.

When the associated program is started, add the switch /8515 to the instruction line, which then looks as follows:

`C:/...handyman.exe../AT90S8515`.

This option appreciably enlarges the reserved memory range of the controller. The capacity of the 2313 used is about twice that of the 1200. The 8515 processor is the largest in the Atmel catalogue so that the memory range can now be set to maximum.

[980076]

ELEKTOR	
240V~	50Hz
No. 980076	
F = 8x 1.25A T	

the software revisited

In essence, the program consists of two interrupt routines. The 'sync' routine synchronizes the phase modulator with the zero crossing of the mains supply voltage. The phase modulator counter runs from 0 to 240; there is also a half-period counter which is 0 or 0xff. The first counter scans the half-period, while the second indicates which part of the period.

At each call, the timer interrupt routine increments the counter of the phase modulator. Each time the count cycle is finished, the half-period counter is inverted.

The synchronization interrupt routine checks whether the phase modulator counter is in synchrony and in phase with the mains voltage. If they are not in phase, the routine increments or reduces the counter content by one unit, depending on the nature and extent of the detected error. In this way, the phase difference will be eradicated within a very short time.

Space is reserved in the RAM for each and every channel for storing all relevant data:

fade time:	*3 bytes*
fade counter:	*3 bytes*
end value:	*1 byte*
step size:	*1 byte*
phase modulator:	*1 byte*

Five bytes are also reserved to enable an RS232 frame to be stored in its entirety.

Fade time is expressed in units of 10 milliseconds (ms). The end value is 0–127, which is multiplied ×2 by the software.

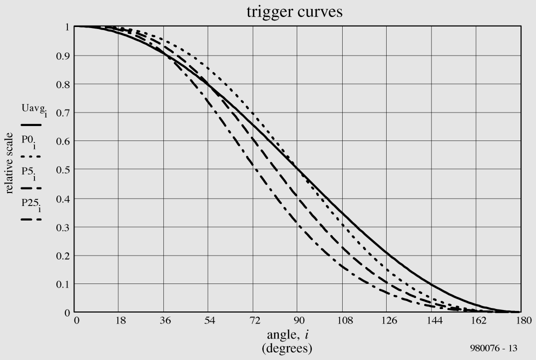
There are four tables, each with 256 reference values, in the program memory. The reference values are 0–249 and are composed so that they represent a specific control operation (i.e., characteristic curve).

New data are stored at the appropriate memory addresses via the RS232 port.

The instantaneous content of the phase modulator is incremented or reduced until the final value is reached. The speed at which this happens depends on the set fade time. The fade time is entered into the fade counter, whereupon the content of this counter is reduced every 10 ms until it is 1. Subsequently, the setting of the phase modulator is modified by 1, whereupon the fade value is again entered into the fade counter. This process is repeated until the end value has been reached. When the fade value is 0, the setting of the phase modulator is made equal to the end value. All these routines are initiated via a timer interrupt.

dip switch positions

Sections 1 and 2 of DIP switch S_1 enable setting a curve along which the controller sets the ignition angle of the optotriacs. Four curves are provided which enable the most propitious angle for a number of applications to be



selected (see illustration below).

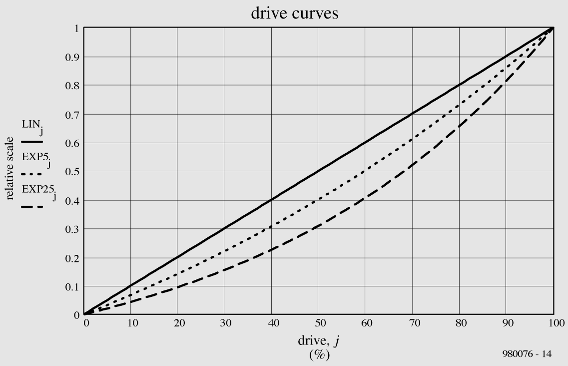
$S_1(1)$	$S_1(2)$	Curve
on	on	1
on	off	2
off	on	3
off	off	4

Curve 1 represents control based on the average voltage.
Curve 2 represents control based on power consumption with $\beta = 0$.
Curve 3 represents semi-logarithmic control based on power consumption with $\beta = 5$.
Curve 4 represents logarithmic control based on power consumption with $\beta = 25$.

Sections 3 and 4 are for setting the address of the controller board.

$S_1(3)$	$S_1(4)$	
on	on	channels 1–8
off	on	channels 9–16
on	off	channels 17–24
off	off	channels 25–32

For those readers who may not be very interested in these curves, but rather know what the effect of shifting the controls is, the curves in the illustration below give the answers. The two linear curves (power consumption with $\beta = 0$ and average voltage) are identical and form a straight line. The other two (exponential)curves are those pertaining to the situation where $\beta = 5$ and $\beta = 25$.



protocol

The protocol used in the dimmer makes control straightforward. The serial format is:

9600 baud
1 start bit
1 stop bit
no parity bit

1st byte: $b_7 = 1$ (marking the start of a frame). With all other bytes, $b_7 = 0$.

Frame construction
A frame consists of five words and is composed as follows:

byte 1
1 A4 A3 A2 A1 A0 S1 S0

byte 2
0 T20 T19 T18 T17 T16 T15 T14

byte 3
0 T13 T12 T11 T10 T9 T8 T7

byte 4
0 T6 T5 T4 T3 T2 T1 T0

byte 5
0 D6 D5 D4 D3 D2 D1 D0

where
A = address of channel
S = size of step
T = duration of step
D = end value

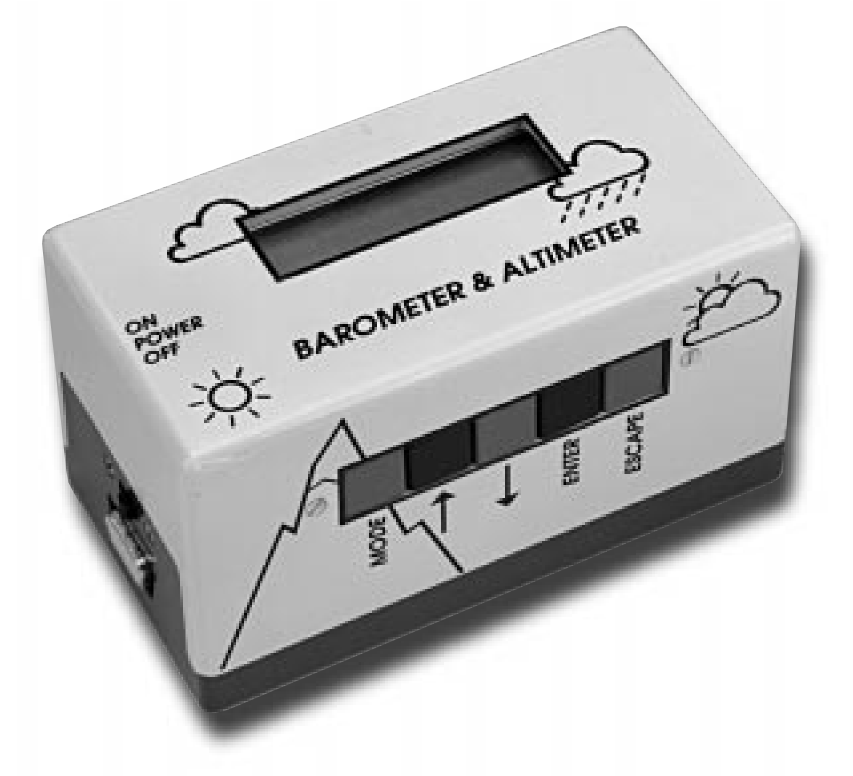
If, on reception, there is an error within a frame, the entire frame will be ignored and the system will wait for the start of a new frame ($b_7 = 1$).

When the step resolution is 1, a control time of 255 × 10 ms is needed. When a short fade time has been set, an error of not more than 1.27 s may ensue. For this reason, the software will alter the size of step to 2, 3 or 4 when the fade time is shorter than 2.55 seconds. This ensures that the fade time is altered to ensure that the error never exceeds 15 per cent.

barometer/altimeter Part 2

programming, calibration, software and operation

When construction of the barometer/altimeter, which is based on a precision air-pressure sensor with integral signal processor from Motorola, is completed, the unit has to be programmed and calibrated in accordance with the correct reference. Evaluation, storing, display and interchange of data are effected by a microcontroller system.



Design by H. Bonekamp

CHOICE OF PROGRAM

Since the 12 kbyte flash-ROM in the Atmel microcontroller is relatively small for the present application, two versions of the program are available:

normal.hex
which arranges for the reference pressure to be derived from the indicated height;

vsl.hex (virtual sea level)
which allows the pressure at sea level as given by the nearest airport or weather station to be input.

PROGRAMMING

Programming of the microcontroller can be effected only with software diskette Type 986031-1 available through our Readers' Services (see towards the end of this issue). The diskette should be copied to a hard disk or other write medium that is not write-protected.

1. Start program SISP (Serial in System Programmable) in the real DOS mode (not in the DOS window of Windows).

1

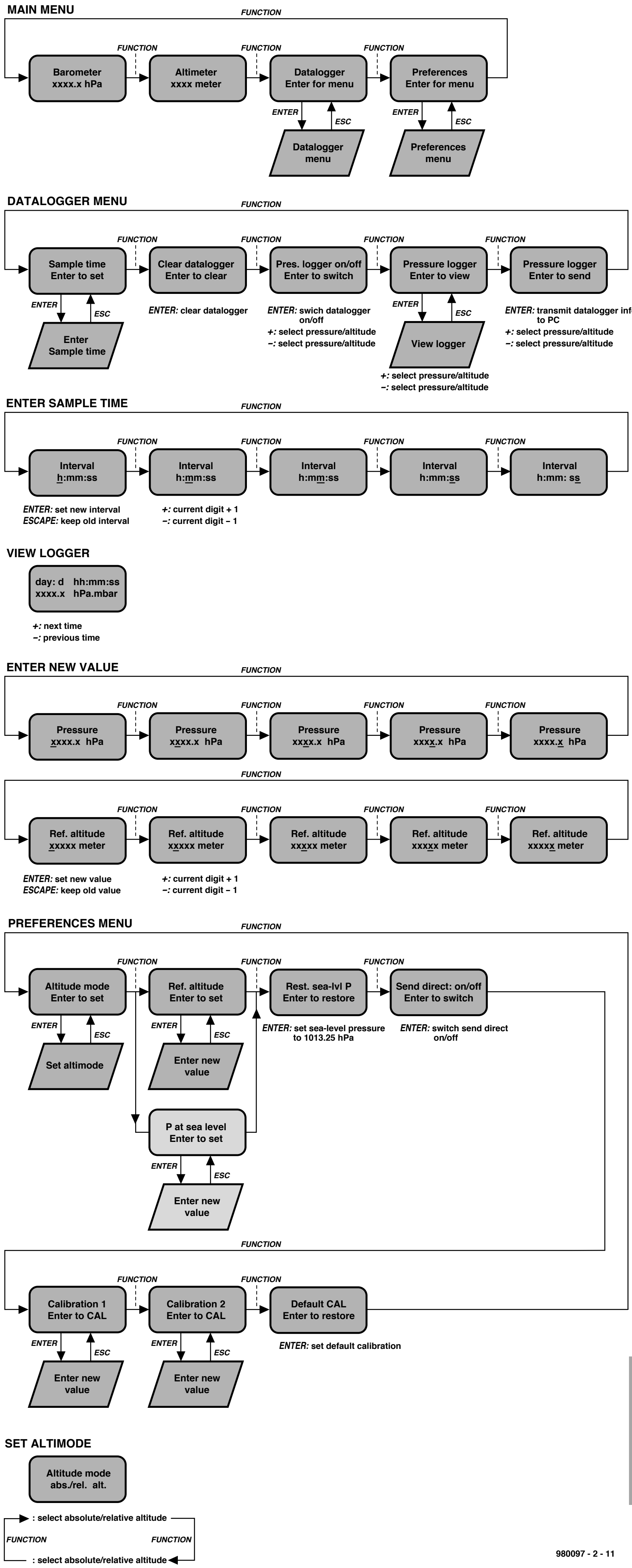


Figure 1. The main menu contains four sub-menus, two of which, data logger and preferences, are themselves divided into sub-sub-menus.

2. Select input file normal.hex or vsl.hex at option 5.

3. Enter the wanted serial interface (com-port) at option 5.

4. Link the barometer to the computer via an RS232 cable (1:1). Do not yet switch on the unit.

5. Set pin jumper JP2 to position I (ISP).

6. Switch on the barometer unit.

7. Insert the pin jumper into JP1.

8. Select option 1 of SISP (write code memory) and wait until the programming has been completed

9. In case of an error message, remove the pin jumper from JP1, switch off the barometer, and repeat the foregoing procedure from point 6.

10. Insert the pin jumper of JP2 into position R (RS232) and remove the pin jumper from JP1.

CALIBRATION

Without correct calibration, the barometer/altimeter would not have the requisite accuracy. Only correct calibration ensures that the program uses the standard transconductance and offset values (0.01509 and 0.1518 respectively) of the sensor specified by the manufacturers to compute the air pressure. Apart from this, there is no other compensation of the tolerances of the voltage divider. There are two methods of calibrating the unit: the single-point and the two-point. In single-point calibration only the offset is corrected, which shifts the pressure vs output voltage characteristic upwards.

Single-point calibration requires a reliable, actual value of the atmospheric pressure, which can normally be obtained from the nearest airport or weather station. This value must be input into calibration 1 in the preferences menu. This value is then compared with the actually measured value and a new offset value computed. The new offset value is stored in the eeprom.

In the two-point method, not only the offset of the sensor is compensated, but also its transconductance. The procedure is as with method 1, but in this case a second value is entered in calibration 2 in the menu. This value must differ by at least 5 hPa (5 millibar) from the first one. This margin is needed by the software to reduce the effects of rounding-off errors and any interference. The greater the difference between the two values, the more accurate the transconductance is computed.

The second value is obtained after the atmospheric pressure has risen or fallen sufficiently with respect to the first value. Check this second value with the nearest weather station and input it at CALIBRATION 2.

OPERATION

Before commencing any measurements, leave the barometer/alitmeter for about two minutes to enable it to 'warm up'. The unit is operated by five touch keys on the front panel:

MODE (S2), which enables any of the functions on each of the menus to be accessed and utilized.

↑ (UP) (S3) with which the value of the selected digits is increased.

↓ (DOWN) (S4) with which the value of the selected digits is decreased.

ENTER (S5) with which a selection is confirmed or altered (stored in the eeprom).

ESCAPE (S6) to return to the next higher menu without this being stored in the eeprom.

These keys take the operator through all the menus as shown in Figure 1.

The Main Menu consists of four sub-menus: barometer, altimeter, data logger, and preferences.

The data logger sub-menu itself consists of a number of sub-sub-menus as shown.

When Input Sample Time has been selected, the sample interval may be set between 10 seconds and 8 hours with a resolution of one second.

The measurement results stored in the data logger can be viewed by selecting View Logger.

The elaborate Preferences menu contains the functions 'Ref. Altitude' (normal.hex) and 'P at Sea Level' (vsl.hex). In both, as during calibration, enter the value for altitude or atmospheric pressure into 'Input New Value' digit by digit.

'Set Altimode' determines whether the relative or absolute altitude is displayed. The absolute altitude is the height above sea level, and the relative altitude is the height above a set reference. In both cases, the actual atmospheric pressure at sea level must be entered. The reference atmospheric pressure at sea level, that is, 1013.25 millibar (1013.25 hPa), is reset with 'Rest. sea-|v|P'. 'Default cal' provides a similar function: it erases any calibration entries and replaces them with standard values.

Not much more can be said about the display than that the atmospheric pressure is shown in hPa (hecto-Pascal, which is identical to millibars) and the altitude in meters. Two arrows indicate whether the measurement is moving up or down. When the data logger is active, the display shows an asterisk.

DATA TO COMPUTER

Communication between the barometer/altimeter and the computer is in 8-bit format, 9600 baud, no parity bit, and no handshake. Basically, any PC using DOS or Windows is suitable. There are two ways of transferring data to the computer: individual, by which a measurement value is transferred to the computer as soon as it is available, or stream, by which the content of the data logger is transferred in one operation.

[980097]

home-made software

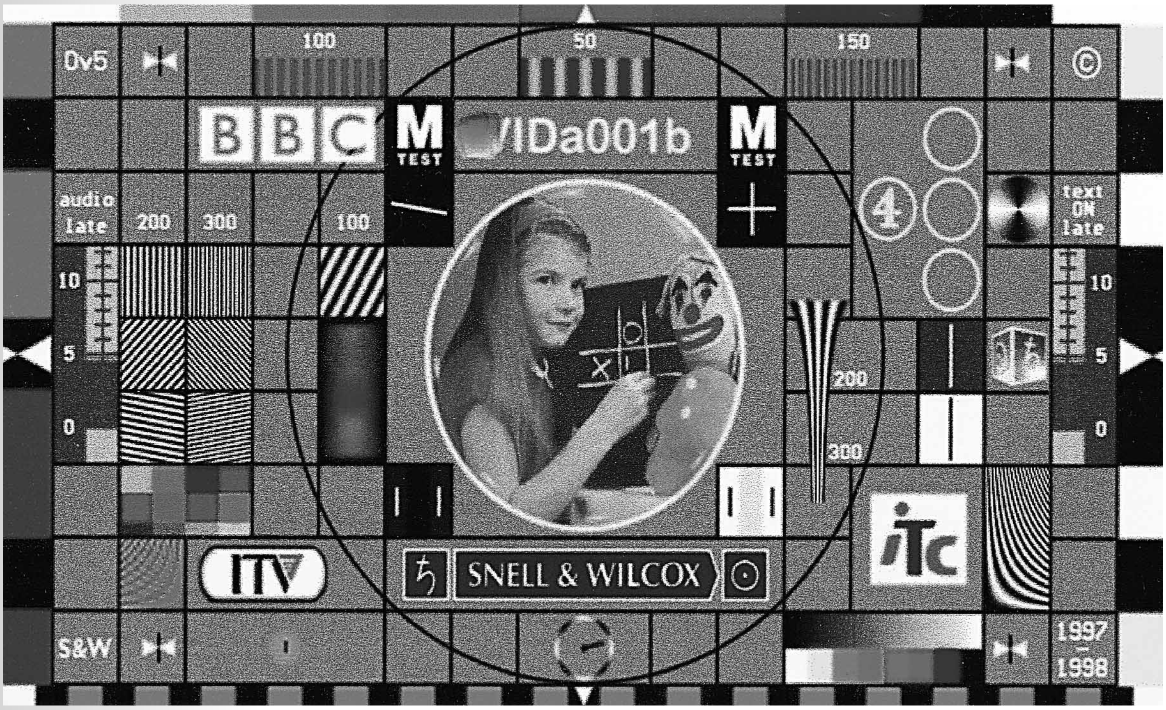
Since the diskette contains the source code of the software, competent readers are able to incorporate their own requirements and other special functions, provided they have a Tasking C development tool available. A demonstration of the C compiler is available as freeware on the Internet: www.tasking.com

Communication with the display is made possible by _IOWRITE in the basic function. This enables the standard print commands in C to be used. This means that conversions must not be written. Printer formats are supported by FPRINTF().

In the same way, communication via the RS232 interface takes place with _IOREAD() and _IOWRITE(). The total package of I/O routines available in C must be entered.

digital terrestrial television (DTT)

An inevitable development



The future of television is largely in digital terrestrial broadcasting. Britain is forging the way in this new field, but other countries in Europe and North America, as well as Japan, are bound to follow soon. Most of the European development was carried out under the DVB (Digital Video Broadcasting) Project launched in 1993 and approved by almost 200 signatories from 25 countries in 1995. The launching group consisted of representatives from industry, public and private broadcasters,tele-communicationss companies, research institutes and the European Commission. Because of this wide-ranging participation, the DVB Project has taken over the leading role in the introduction of digital television in Europe.

Based on a report by Bill Higgins

INTRODUCTION

Digital terrestrial television provides substantial benefits to viewers: more quality channels, better sound, better pictures, and new services. In the near future, many of these services will be interactive, enabling the viewer to shop, bank, send e-mail messages, and others. Many countries in Europe will undoubtedly discontinue the analogue transmissions and switch to digital broadcasts over the next 10–15 years' time (in Britain, suggestions have already been made for this to happen as early as 2008).

Viewers will need a new digital television receiver or a digital terrestrial set-top box to receive the new broadcasts. Viewers who wish to make use of the interactive services need a telephone socket near their TV set. There will be no 'ghosting' or 'snow' with digital terrestrial TV. Viewers with widescreen television receivers will be able to take advantage of the higher proportion of widescreen material included in digital broadcasts.

HOW DOES DIGITAL TELEVISION WORK?

An analogue signal can be sampled and digitized as shown in **Figure 1**, and then represented by a stream of logic 1s and 0s . The planning of digital terrestrial television started in the USA in the late 1980s and soon thereafter, in the early 1990s, separate projects and pilot developments came about in Europe also.

Essential to the concept of digital video coding is data compression technology which makes possible much narrower bandwidths than with analogue TV signals. The compression is applied to quantized images provided by the television camera. The picture area is sampled pixel by pixel and a value is allocated in each case to the luminance value *Y* and the chrominance values *R-Y* and *B-Y* of each pixel. This pulse-code modulation is carried out initially with binary numbers. The resultant bit stream has a data rate of 166 Mbit/s.

In the case of High Definition Tele-Vision (HDTV), a new aspect ratio of 16:9 is used (current standard is 4:3). The scanning rate of the HDTV luminance signal is increased to 72 MHz (currently, 13.5 MHz). This results in a total data rate of 1.52 Gbit/s for the *Y-U-V* video source signal.

A giant step forward in coding technology was achieved by motion-compensating coding. In this, to start the data compression, only the pixels of a new frame that have moved owing to the movement of the object are transmitted. The unchanged pixels are redundant and are reproduced from the frame buffer in the receiver. A further reduction in the number of pixels to be transmitted is obtained by specifying the frame-to-frame displacement of a moving object by a displacement vector. This vector is used in the receiver to take the pixel for the moving object from the buffer. The background data must, of course, also be transmitted in this method, which is called interframe DPCM (Differential Pulse Code Modulation) with motion compensation.

In the decomposition, use is made of the Discrete Cosine Transform (DCT), which is a variant of the Discrete Fourier Transform (DFT). There are two kinds of DCT: the hybrid and the intraframe, both of which have advantages and disadvantages. Therefore, full frames are normally intraframe coded at fixed intervals in the hybrid method. This is the basis of the ISO MPEG standard.

The ISO (International Standardization Organization), which is a consultant to the United Nations, was decisively involved in the

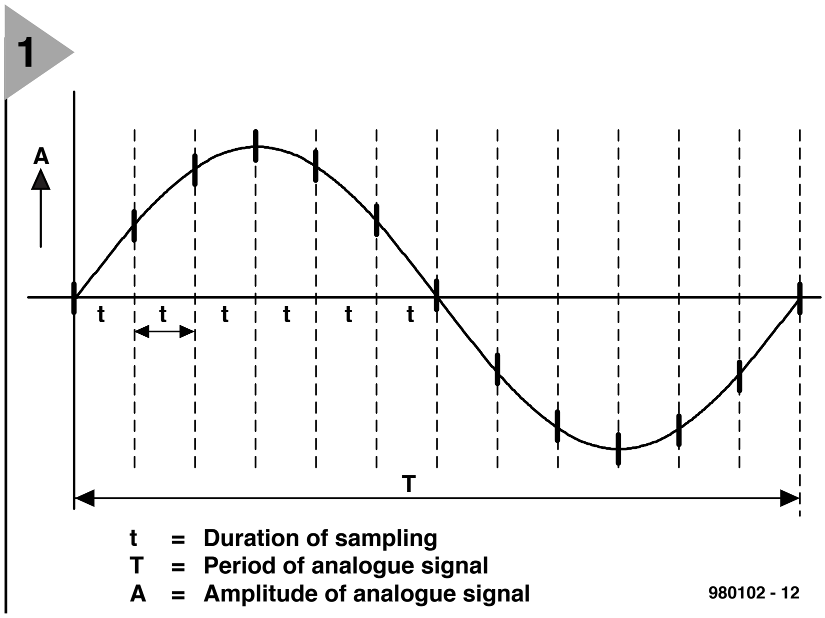


Figure 1. Principle of sampling and quantizing an analogue signal.

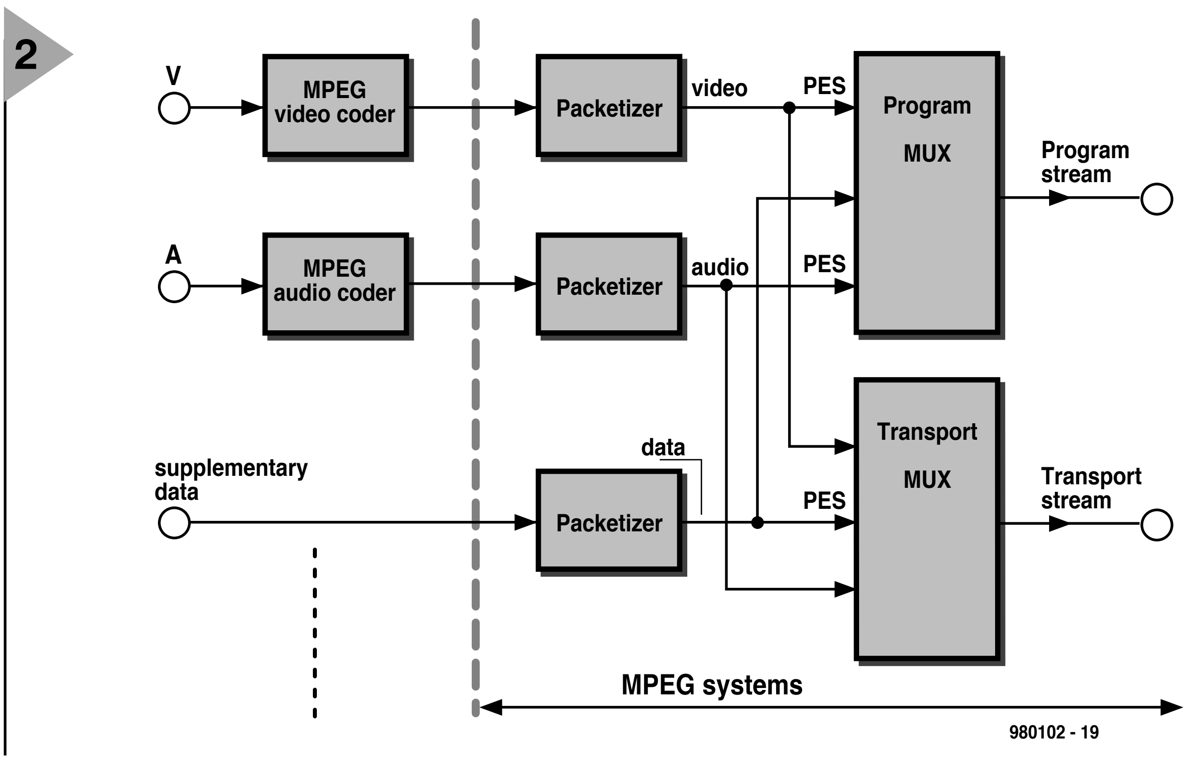


Figure 2. The program and transport stream in the MPEG-2 standard.

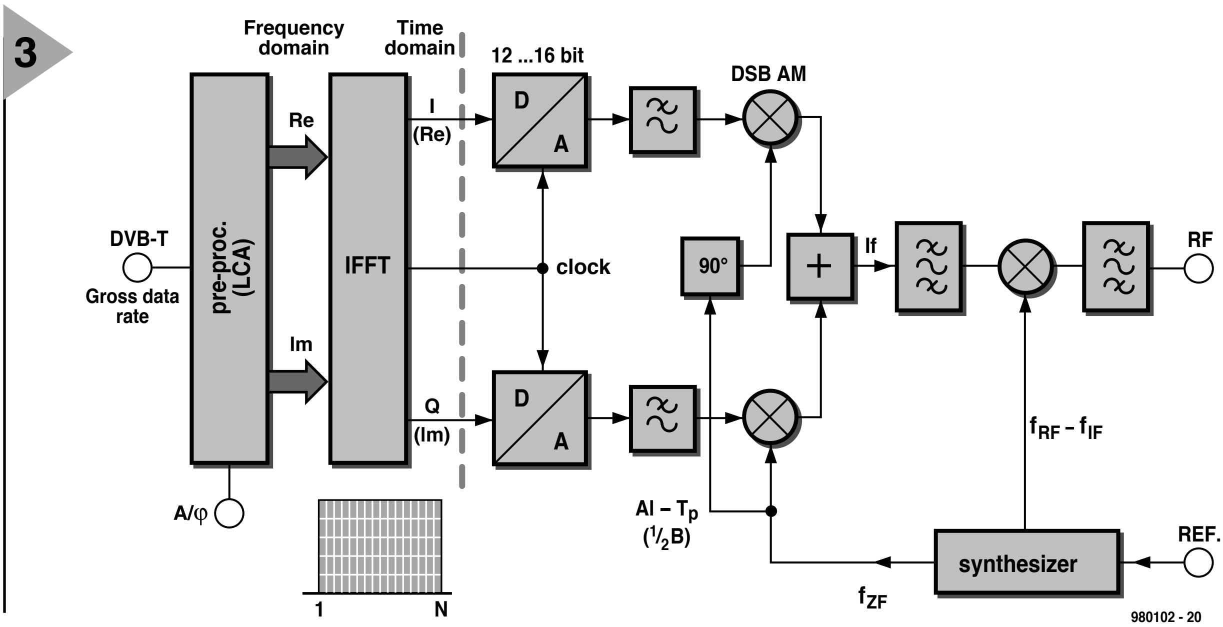


Figure 3. Typical OFDM (Orthogonal Frequency Division and Multiplexing) modulator used in DTT (Digitial Terrestrial Television).

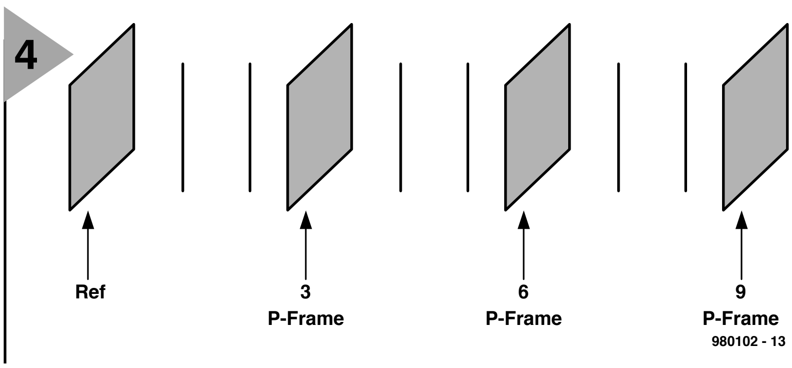
advances made in video coding. Since the early 1990s, the ISO and IEC (International Electrotechnical Commission) have been coordinated by the JCT1 (Joint Technical Committee 1) in the telecommunications field.

A subgroup (called the Motion Pictures Expert Group – MPEG) of the JCT1 was set up to define a standard for full-video communication. The standard specifies storage, for instance, in multi-media workstations, and can also be applied to transmissions on the established media.

The MPEG-1 standard is suitable for the coding of small-format images with low data rates (up to 1.5 Mbit/s). The second project phase, MPEG-2, is a specification for a method that is compatible with MPEG-1, but which allows the coding of enhanced PAL (Phase Alternate Line) quality. It also includes HDTV. The standard specifies sampling rates of 2–15 Mbit/s. The correlation between the sampling rate, the duration of the sample, and the number of samples per television line is shown in **Table 1**.

MPEG-2 has become a world standard for video, applying both to the transmitting side and to the receiver at the output of the demodulator: OFDM (Orthogonal Frequency Division and Multiplexing) for terrestrial reception, QSPK (Quadrature Phase Shift Keying) for satellite reception, and 64QAM (Quadrature Amplitude Modulation) for reception via a cable network.

Figure 4. P(redicted)-frames at 3, 6, and 9 frames from the reference frame.



The MPEG-2 Video Coding Standard was conceived as a generic, that is, application-independent solution. In other words, the syntax of its algorithm makes it suitable for many different applications and their relevant data rates. In addition to this flexibility in source formats, MPEG-2 allows different ‘profiles’. A profile offers a collection of compression tools that together make up the coding system.

The specification for DVB-T (Digital Video Broadcasting Terrestrial) was finalized in late 1995. It lays down that the DVB-T transmission system contains the following fundamentally new elements: baseband coding for video and audio, MPEG-2 transport stream (see **Figure 2**), terrestrial channel coding, OFDM modulation, and coverage using single-frequency network technology. The OFDM modulation method (see **Figure 3**) and the single-frequency network technology lead to a number of system engineering consequences.

One of the consequences of applying data compression at the signal source is that conventional sinewave test methods with swept-frequency signals in the frequency domain and with reference to test line signals in the time domain are not usable for the digital transmission channel.

The consequence of transporting signals in time division multiplex is

Sampling rate (Mbit/s)	Duration of sample (μs)	Number of samples per TV line
2	0.5	128
6.75	0.148	432
13.5	0.074	865
15	0.066	969

Table 1. Correlation of MPEG-2 sampling rates, duration of sample, and number of samples per television line.

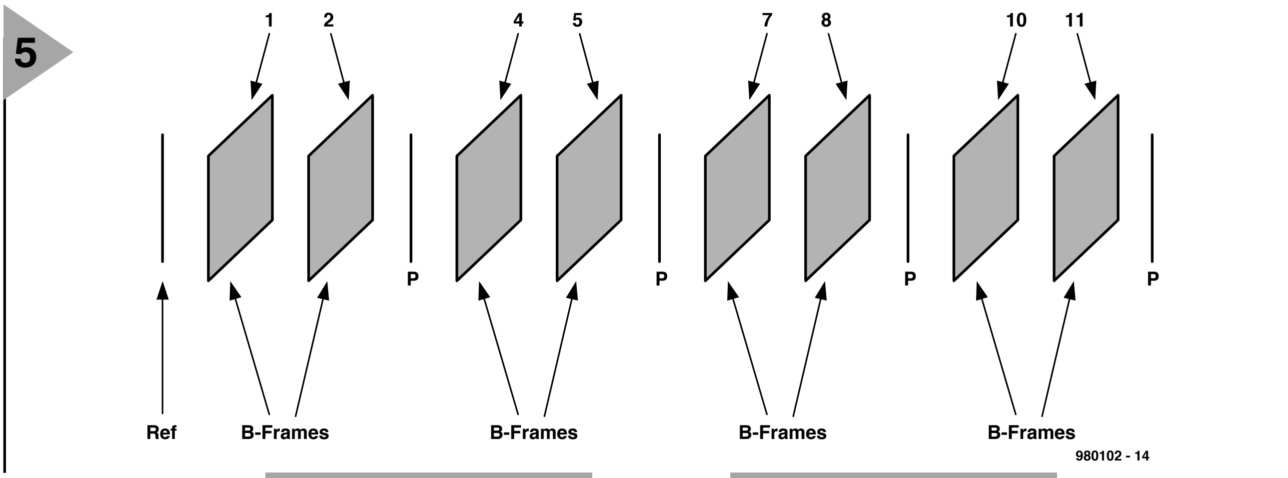
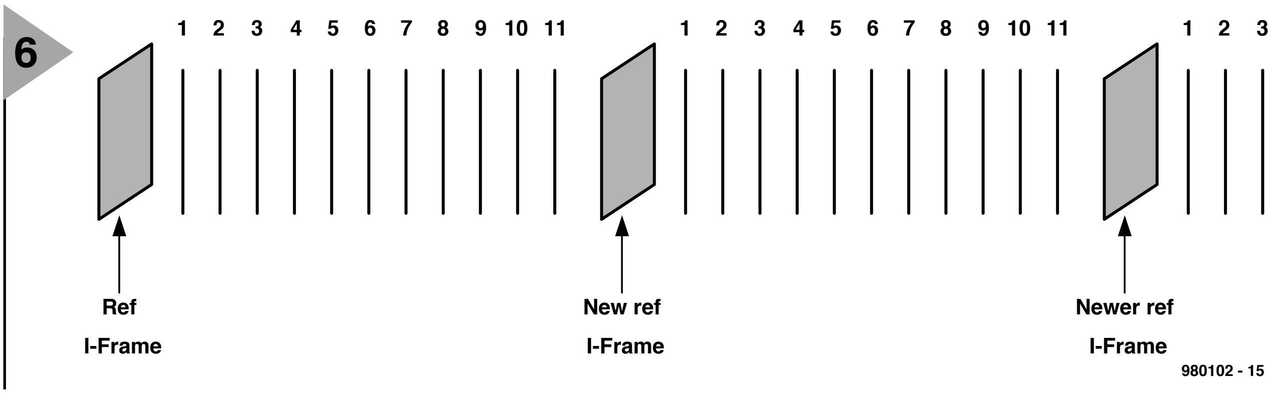


Figure 5.Location of B(i-directional) frames.

Figure 6. Location of I(ntraframe) signals.



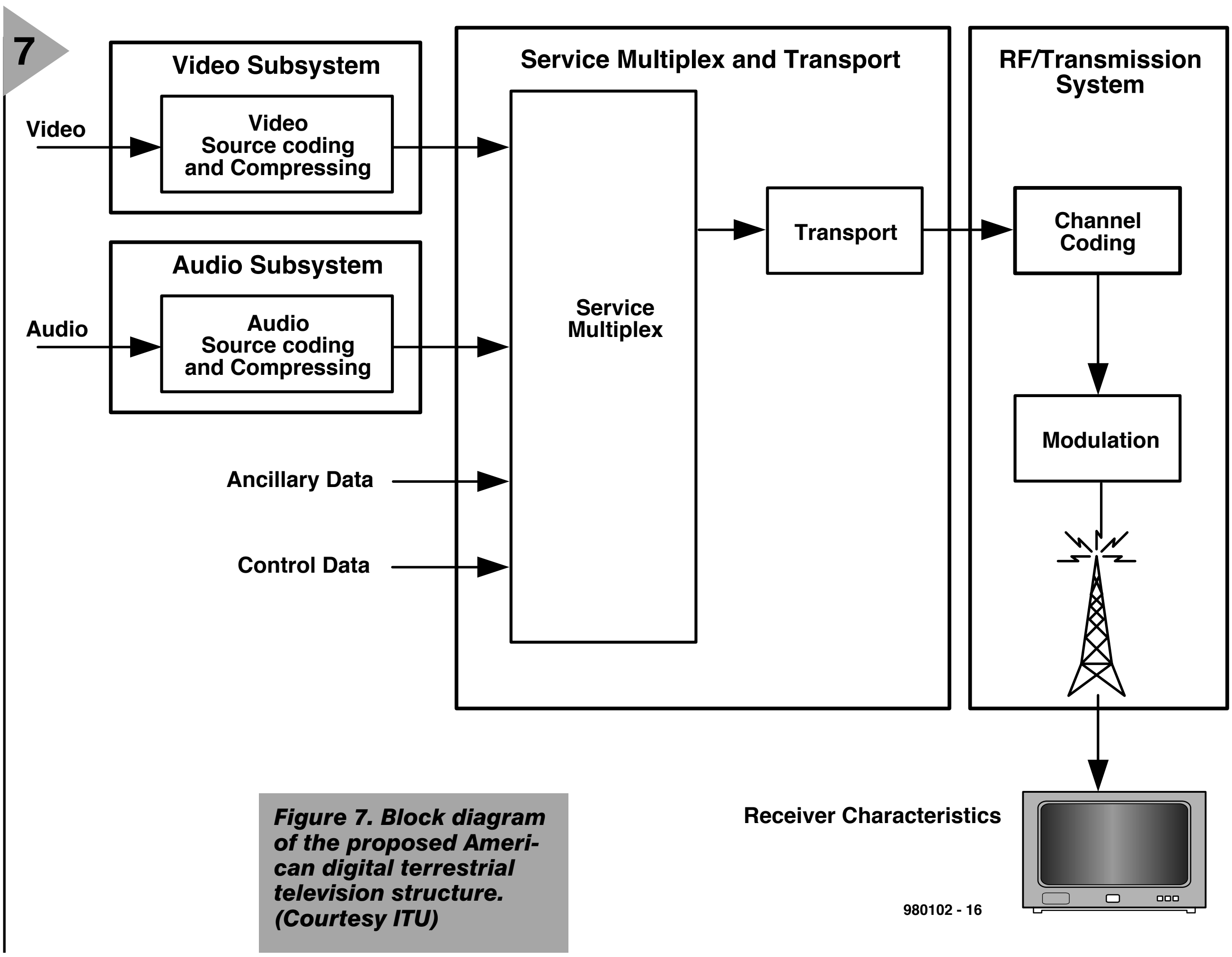


Figure 7. Block diagram of the proposed American digital terrestrial television structure. (Courtesy ITU)

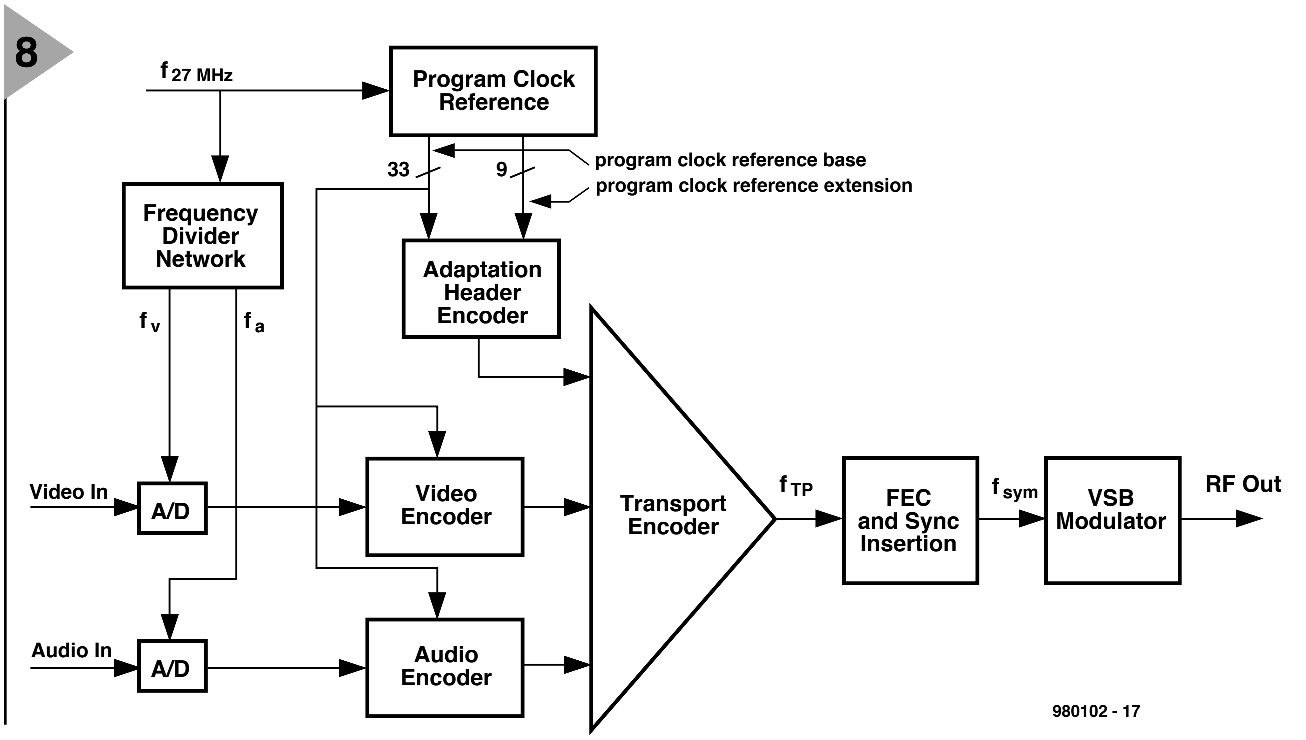


Figure 8. High-level encoding equipment for the proposed American standard of DTT. (Courtesy ITU)

that digital terrestrial broadcasting service technology need not remain confined to the transmission of television signals and associated data, but that video, audio and data signals can be freely assembled and transmitted transparently for multi-layer services.

Programme distribution can be effected via standard copper lines, optical fibre cables or microwave links.

Digital terrestrial transmitter engineering requires novel transmitter measuring techniques. This involves parameters such as BER (Bit Error Rate), pattern analysis, spectrum analysis, OFDM power measurement and the measurement of the operating characteristics of multi-carrier power amplifiers.

Operating DVB-T transmitters in single-frequency mode presupposes frequency- and bit-synchronous operation by the transmitters. This requires new approaches to frequency and time synchronization providing operational reliability on a regional and national level.

Terrestrial transmission over radio paths with single transmitters and single-frequency transmitters requires novel coverage measuring techniques in which, in addition to the traditional field-strength measurement, parameters such as channel impulse response, raw bit error rate, intersymbol interference, and selective C/I (Carrier over Interferer) are important factors.

MULTIPLEXING

Multiplexing is the process of transmitting two or more signals over the same path without interaction. This can be achieved by separating the signals in time or frequency.

Frequency division multiplexing (FDM) is an analogue technique which is still used on satellite and microwave links, although many of these now use digital techniques.

Time division multiplexing (TDM) is a method of interleaving digital signals from a number of channels on to one circuit. For instance, six 600 bit/s channels may be multiplexed on to one 3600 bit/s circuit. Both ends of the circuit must be synchronized to ensure that the data on one channel input reaches the correct channel output at the far end.

DATA COMPRESSION

In general, data compression is a method to reduce the amount of transmitted data by applying an algorithm to the basic data at the point of transmission. A decompression algorithm expands the data back at the receiver into its original format. There are two major methods in use: Interframe and intraframe.

Interframe

The interframe method is based on a difference signal generated by the the frames before and after the present frame. These difference signals are termed P(redicted) frames and B(i-directional) frames. P-frames are predicted from the preceding reference frame and are normally 3, 6 or 9 frames from the reference as shown in **Figure 4.**

B-frames are generated by interpolation from P-frames and the reference frame and are therefore called bi-directional. As shown in **Figure 5**, they slot in between the reference and the P-frames, at 1, 2, 4, 5, 7, 8, 10, and 11, frames from the reference.

Intraframe

In the interframe system, the reference frame occurs every 12 frames. This is effectively the intraframe signal or I-frame. A new I-frame occurs after every eleven interframe difference signals throughout the transmission (see **Figure 6**).

AMERICAN STANDARD

In the USA, the Advisory Committee on Advanced Television Service (ACATS), which was set up by the FCC (Federal Communications Commission), and the Advanced Television Test Center (ATTC), a collaboration between broadcast service operators and the television receiver industry, have devised a different standard for digital television.

The specification is basically the Digital Spectrum Compatible HDTV (DSC-HDTV) proposal by Zenith and AT&T. Basically, it splits the digital TV system into:

- source coding and compression
- service multiplex and transport
- RF transmission.

A block diagram of the system for digital terrestrial television is shown in **Figure 7.** The coding is based on the MPEG-2 standard, but uses 27 MHz sampling and special digital extensions to allow for any new formats in the future, picture extensions, and indica-

Although digital satellite signals can be received all over the UK, digital terrestrial signals will, at least for the time being, not cover the whole country. Where possible, the new digital transmission will be on frequencies close to those used for the current analogue TV broadcasts, which means that viewers can continue to use their existing antennas.

The BBC has already started transmitting DTT signals: from 23 September last, viewers with suitable equipment have been able to watch the first regular digital terrestrial channel in the world, together with wide-screen versions of BBC1 and BBC2. The inset table correlates some UHF channels with analogue and digital TV programmes (based on Crystal Palace transmitter).

Channel number (UHF)	DTT programme	Analogue programme
22	DMx2	
23		ITV (London)
25	DMx1	
26		BBC1 (London)
28	DMx4	
29	DMx6	
30		Channel 4 (London)
32	DMx3	
33		BBC2 (London)
34	DMx5	
37		Channel 5 (London)
DMx = Digital Multiplex		

The existing terrestrial channels will be free-to-air on digital terrestrial TV, as will the new digital channels: BBC News 24, BBC Choice, and existing free channels on analogue TV: Sky News, CNN, and Eurosport.

Open standard integrated digital television sets are now available in the shops, enabling viewers to receive all the digital services from the BBC, ITV, Channel 4, Channel 5, and ONdigital without the need for set-top boxes, satellite dishes or cable connections. Six manufacturers will be producing set-top boxes needed to receive DTT on analogue TV sets: Grundig, Nokia, Pace, Philips, Sony and Toshiba.

Video cassette recorders (VCRs) can be used with digital terrestrial set-top boxes. Unfortunately, there is the same serious and annoying flaw as with satellite receivers: only the digital programme being watched can be recorded. This is an area where analogue TV will retain its current popularity for some time to come.

ONdigital, formerly called British Digital Broadcasting (BDB), has been granted 24-year licenses by the Independent Television Commission to be the terrestrial pay-TV platform in the United Kingdom. ONdigital is a partnership between Carlton Communications and the Granada Group.

A smart card, which looks like an ordinary credit or banker's card, but which has an integrated microprocessor and memory instead of a magnetic recording strip, will be used by ONdigital for subscription services and pay-per-view events.

Smart cards are already used for mobile telephones, loyalty cards and by BSkyB to control subscription viewing to its satellite service. Owing to the extended reach of Astra 2, Sky smart cards will be needed for the free-to-air channels to prevent viewers in other European countries tuning in to programmes that have been copyright-cleared for the UK only.

Electronic Programme Guide (EPG) is a technique used in digital TV that allows for the transmission of data about programmes that are being broadcast. One of the features of this programme is that it allows viewers to order TV channels for special events or films on the appropriate pay-TV channel. EPG also has the possibility of providing interactivity with the home customer in offering a whole range of consumer services, from home banking to home shopping.

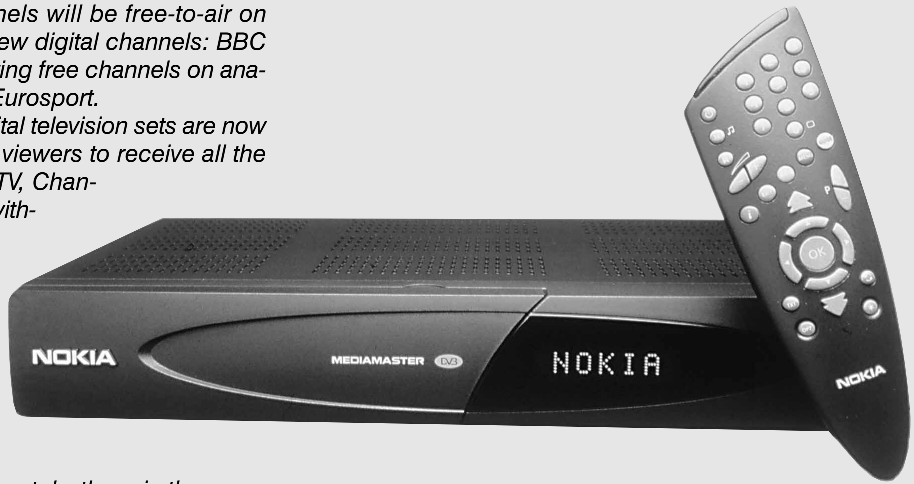
Useful telephone numbers:

BBC: 0990 118 833 for a digital information pack; 0870 010 0123 for other queries; web site: www.bbc.co.uk/digital/

ONdigital: 0171 819 8000; web site www.ondigital.co.uk

BSkyB: 0870 242 4200

The photograph (A) shows the Mediamaster 9850T from Nokia which was introduced at the Cable & Satellite Show in London earlier this year. It is fully compliant with the ONdigital standard, ready for pay-per-view and other interactive services, and has an integrated modem. It comes with a remote controller for ease of operation.



MOTOROLA TO REVOLUTIONIZE MULTIMEDIA IN THE HOME

In a move to set the standard in the home entertainment market, Motorola has launched the first system to bring together digital TV, audio, Internet, 3D computer games and other multimedia applications in one box.

At a time when digital TV-based services establish themselves in Europe, as defined by the DVB-T project, the open system will liberate viewers by allowing them to receive the services of all competing digital services, whether terrestrial, cable or satellite.

The system, named *Blackbird*, has been offered to manufacturers of set-top boxes and other manufacturers.

Consumers can expect to see set-top box products, based on *Blackbird*, within the next six months.
Motorola 0049 172 678 9545 (David Jones), or 0044 802 365 956 (Una Kent)

tions of use of the system in the appropriate signal.

A high-level view of the encoding equipment is shown in **Figure 8**. In this drawing, f_{TP} is is the transmission frequency of the transport stream, while f_{sym} is the frequency of the vestigial sideband (VSB). These frequencies, which must be locked, are related by:

$$f_{TP} = (188/208)(312/313)f_{sym}.$$

As stated earlier, the MPEG-2 Video Coding Standard is used by all proposed DTT systems to achieve an adequate throughput of the vast amounts of data required by the American HDTV and DVB's multi-channel Standard Definition TeleVision (SDTV). Although current DTT plans in Europe are geared towards SDTV, it should be noted that the choice of HDTV or SDTV has nothing to do with the transmission of the MPEG-2 bit stream. In other words, there are no obstacles to broadcasters using DVB-T to carry HDTV: all of the MPEG-2 compliant formats in the American HDTV standard can be delivered by DVB-T.

The difference between the American and DVB-T standards exists largely in their RF modulation technique. The American system uses the single-carrier, 8-VSB (Vestigial Side Band) modulation scheme, whereas DVB-T uses multiple-carrier COFDM (Coded OFDM). Even in the USA, there is considerable interest in CODFM because it provides the most rugged and flexible delivery mechanism for information available today.

[980102]

Sources:

Digital Terrestrial Television Broadcasting by Paul Dambacher, Springer Verlag, 1998. (Reviewed in the September 1998 issue of this magazine)

BBC, London, England
www.bbc.co.uk

BSkyB, Isleworth, England
www.sky.co.uk

DigiTag: www.digitag.org

Digital TV Group, Hants, England
dtg.org.uk

Echostar, the Netherlands

FCC: www.fcc.gov

General Instruments: www.gi.com

Grundig: www.grundig.com

ITC: www.itc.org.uk

ITU, Geneva, Switzerland
www.itu.int/newsroom

MPEG: www.mpeg.org

NEC Benelux, the Netherlands

Newman Carter Hill, London, England

Nokia Multimedia, London, England
www.nokia.com

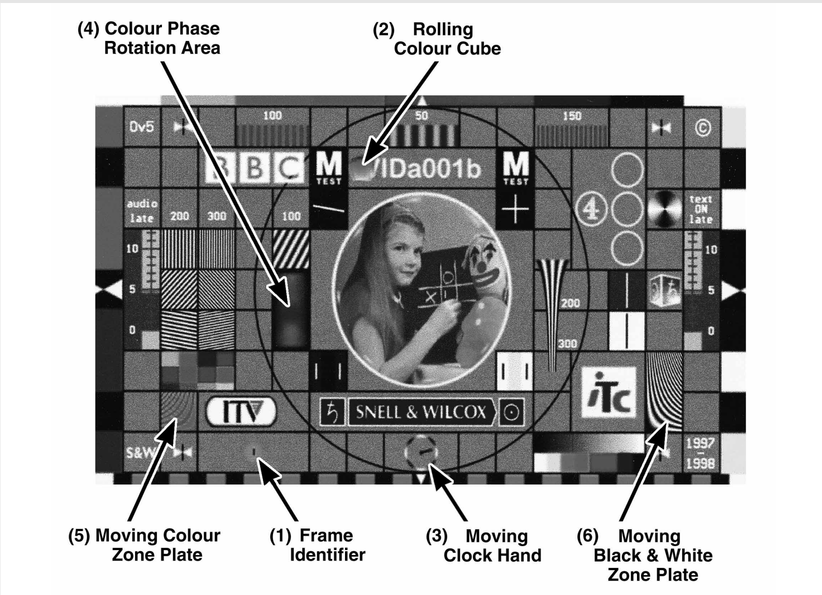
Pace Micro Technology, Shipley, England; www.pacemicro.com

Panasonic: www.panasonic.co.uk
Snell & Wilcox, Peterfield, England

TV/COM International, Weybridge, England; www.tvcom.com

Test Card M

The Test Card for use in digital video broadcasts (DVB), called Test Card M as illustrated, is based on existing test cards, showing (in the UK) the familiar girl, blackboard and balloon, together with graticules, circles, and so on. Additional test areas for digital broadcasts are shown in the diagram for location.



1) *Frame identifier, which shows which frame is present, I, B or P, and gives it a number, for instance, 2nd B or 3rd P. This is considered the most useful parameter for fault diagnosis.*
2) *Rolling colour cube. Since difference signals are generated by the digital equipment, it is useful to have some movement in the test card and this is provided by the cube moving across the screen from left to right, weaving in front and behind the letters BBC, M test, VID001g, and so on, always on the same line of the card.*
3) *Moving clock hand, which moves on every second; useful for time/movement measurement analysis.*
4) *Colour phase rotation area. This is to show colour difference signal since the colour spectrum is changing continuously.*
5) *Moving colour zone plate to determine any image impairment on colour caused by cascading of multiplex stages.*
6) *Moving black&white zone plate to determine any impairment on B&W pixels owing to cascading multiplex stages.*

Test card M has been sponsored by the Department of Trade and Industry (DTI) under the 'Test Bed Programme'. Leaders of the project are Snell & Wilcox, while other members include the BBC, ITV, Channel 4, and ITC. It is financed by and geared to the European market

The structure of the test card enables rapid diagnosis of faults, system stress, and so on, without the need of specialized (expensive) equipment. A quick look at the test card should in many cases be sufficient to ascertain the nature of a fault.

The test card does not provide test sequences for the American system. It is expected that the American organizations will produce their own in due course.

OHN3040U, OHS3040U

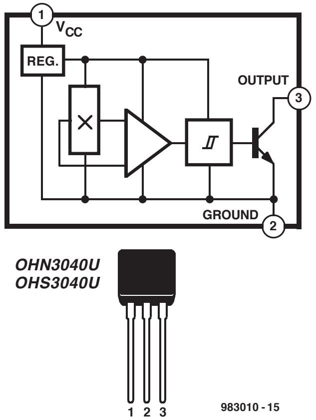
Sensors
Hall-Effect

ELEKTOR ELECTRONICS
DATASHEET 12/98

OHN3040U
OHS3040U
Hallogic™ Hall Effect Sensors

Manufacturer
TRW Communications, 1900 Richmond Road, Cleveland, Ohio 44124, USA.

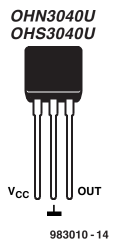
983010 - 13



Functional block diagram

Application Example
Anemometer, *Elektor Electronics* December 1998.

Features
- ➥ Operates over a broad range of supply voltages
- ➥ Excellent temperature stability to operate in harsh environments
- ➥ Drive capability up to 5 TTL loads
- ➥ Hall element, linear amplifier, and Schmitt trigger on a single Hallogic™ silicon chip



OHN/S3040U pinout

Absolute maximum ratings
(T_A = 25ºC unless otherwise noted)

Supply voltage V_{CC}	25 V
Storage temperature range, T_S	–65 C to +150 ºC
Operating temperature range, T_A	
- OHN3040U	–20 ºC to +85 ºC
- OHS3040U	–55 ºC to +125 ºC
Lead soldering temperature	
(3.2 mm from case for 5 sec. with soldering iron)(1)	260 ºC
Output ON current, I_{SINK}	25 mA
Output OFF voltage, V_{OUT}	V
Magnetic flux density, B	unlimited

(1) heat sink leads during hand soldering

XC9536

Integrated Circuits
Programmable Logic

ELEKTOR ELECTRONICS
DATASHEET 12/98

XC9536
In-System Programmable CPLD

Manufacturer
Xilinx Inc., 2100 Logic Drive, San Jose, CA 95124-3400, USA. Tel. (408) 559-7778, fax: 408-559-7114. Internet: www.xilinx.com.

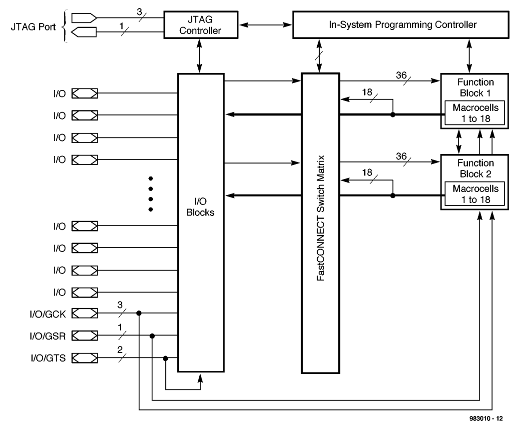
983010 - 11

UK: Xilinx Ltd., Benchmark House, 203 Brooklands Rd., Weybridge, Surrey KT13 ORH. Tel. (01932) 349401.

Features
- ➥ 5 ns pin-to-pin logic delays on all pins
- ➥ f_{CNT} to 100 MHz
- ➥ 36 macrocells with 800 usable gates
- ➥ Up to 34 user I/O pins
- ➥ 5 V in-system programmable (ISP)
- ➥ Endurance of 10,000 program/erase cycles
- ➥ Program/erase over full commercial voltage and temperature range
- ➥ Enhanced pin-locking architecture
- ➥ Flexible 36V18 Function Block
- ➥ 90 product terms drive any or all of 18 macrocells within Function Block
- ➥ Global and product term clocks, output enables, set and reset signals
- ➥ Extensive IEEE Std 1149.1 boundary-scan (JTAG) support
- ➥ Programmable power reduction mode in each macrocell
- ➥ Slew rate control on individual outputs
- ➥ User programmable ground pin
- ➥ Extended pattern security features for design protection
- ➥ High-drive 24 mA outputs
- ➥ 3.3 V or 5 V I/O capability
- ➥ PCI compliant (-5, -6, -7, -10 speed grades)
- ➥ Advanced CMOS 5V FastFLASH technology
- ➥ Supports parallel programming of more than one XC9500 concurrently
- ➥ Available in 44-pin PLCC, 44-pin VQFP, and 48-pin CSP packages

Description
The XC9536 is a high-performance CPLD providing advanced in-system programming and test capabilities for general-purpose logic integration. It is comprised of two 36V18 Function Blocks, providing 800 usable gates with propagation delays of 5 ns.

Application example
Compact Multiburst Generator, *Elektor Electronics* January 1999.



XC9536

Integrated Circuits
Programmable Logic

ELEKTOR ELECTRONICS

DATASHEET 11/98

XC9536 I/O pins (PC44 case only)

Function Block	Macrocell	Pin (PC44 case)	Bscan Order	Notes
1	1	2	105	
1	2	3	102	
1	3	5	99	[1]
1	4	4	96	
1	5	6	93	[1]
1	6	8	90	
1	7	7	87	[1]
1	8	9	84	
1	9	11	81	
1	10	12	78	
1	11	13	75	
1	12	14	72	
1	13	18	69	
1	14	19	66	
1	15	20	63	
1	16	22	60	
1	17	24	57	
1	18	—	54	
2	1	1	51	
2	2	44	48	
2	3	42	45	[1]
2	4	43	42	
2	5	40	39	[1]
2	6	39	36	[1]
2	7	38	33	
2	8	37	30	
2	9	36	27	
2	10	35	24	
2	11	34	21	
2	12	33	18	
2	13	29	15	
2	14	28	12	
2	15	27	9	
2	16	26	6	
2	17	25	3	
2	18	—	0	

Note: [1] Global control pin

Power Management

Power dissipation can be reduced in the XC9536 by configuring macrocells to standard or low-power modes of operation. Unused macrocells are turned off to minimize power dissipation. Operating current for each design can be approximated for specific operating conditions using the following equation:

$$I_{CC}\text{ (mA)} = MC_{HP}\ (1.7) + MC_{LP}\ (0.9) + MC\ (0.006\text{ mA/MHz})\ f$$

Where:

MC_{HP} = Macrocells in high-performance mode
MC_{LP} = Macrocells in low-power mode
MC = Total number of macrocells used
f = Clock frequency (MHz)

XC9536 Global, JTAG and Power Pins (PC44 case only)

Pin type	Pin (PC44 case only)
I/O/GCK1	5
I/O/GCK2	6
I/O/GCK3	7
I/O/GTS1	42
I/O/GTS2	40
I/O/GSR	39
TCK	17
TDI	15
TDO	30
TMS	16
VCCINT 5 V	21, 41
VCCIO 3.3 V/5 V	32
GND	23, 10, 31
No Connects	

OHN3040U, OHS3040U

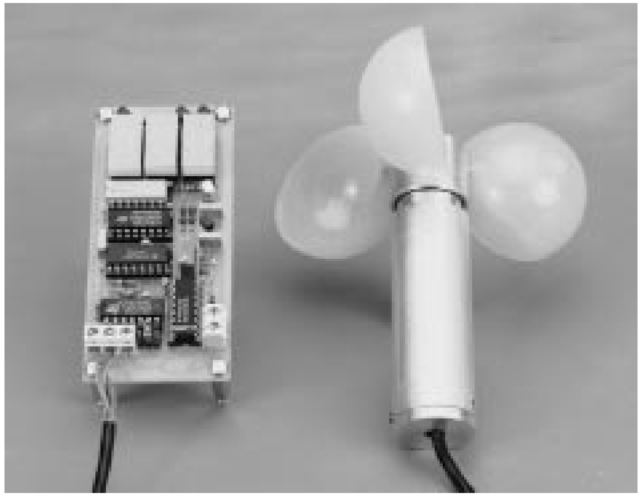
Sensors
Beaufort Scale

ELEKTOR ELECTRONICS

DATASHEET 11/98

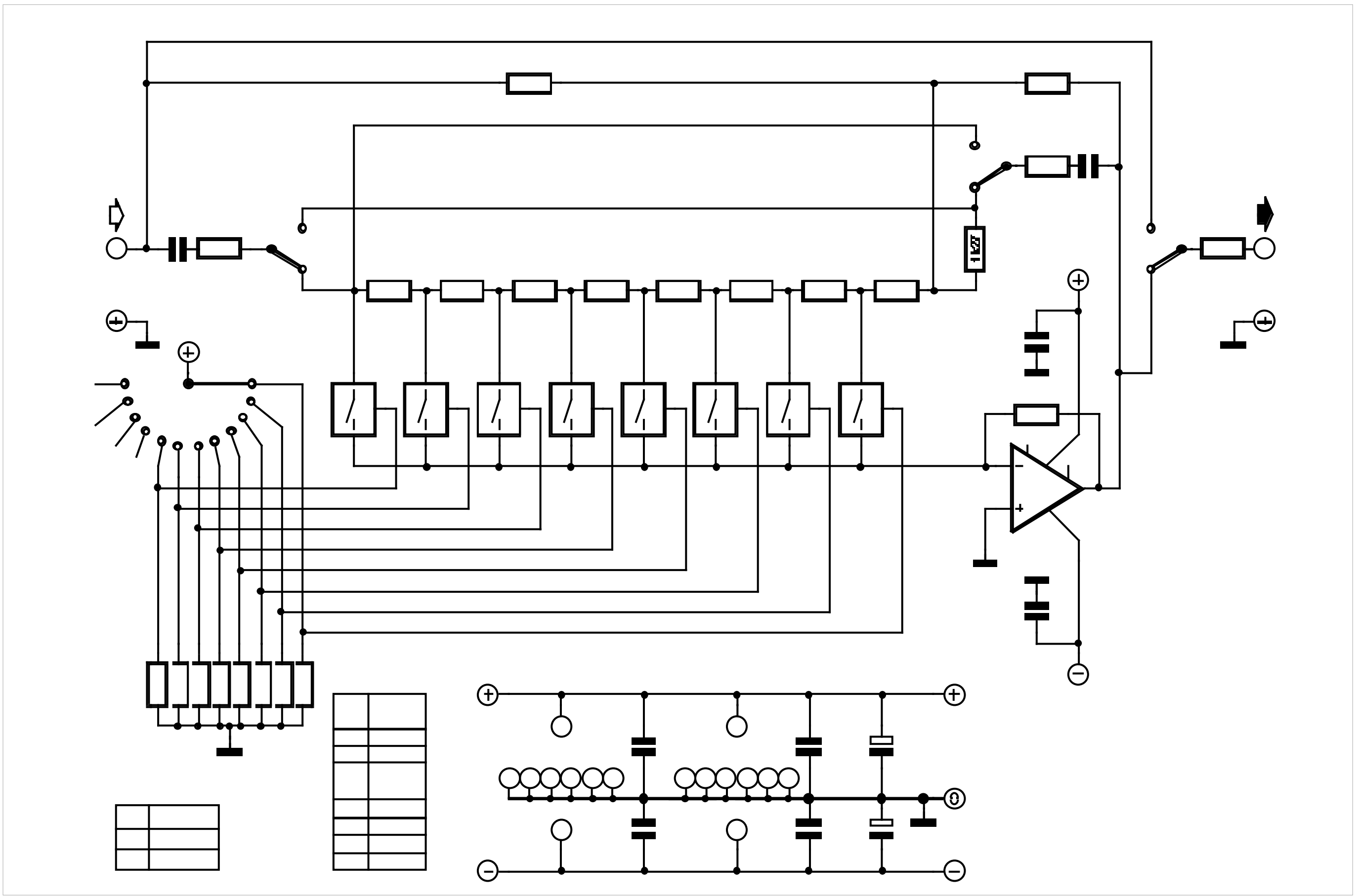
Beaufort scale and correlated wind speeds

Description	Wind speed		
	m/s	mph	knots
calm	0 – 0.2	0 - 1	0 – 1
light air	0.3 – 1.5	1-3	1 – 3
light breeze	1.6 – 3.3	4 – 7	4 – 6
gentle breeze	3.4 – 5.4	8 – 12	7 – 10
moderate breeze	5.5 – 7.9	13 – 18	11 – 16
fresh breeze	8.0 – 10.7	19 – 24	17 – 21
strong breeze	10.8 – 10.7	25 – 31	22 – 27
moderate gale	13.9 – 17.1	32 – 38	28 – 33
fresh gale	17.2 – 20.7	39 – 46	34 – 40
strong gale	20.8 – 24.4	47 – 54	41 – 47
whole gale	24.5 – 28.4	55 – 63	48 – 55
storm	28.5 – 32.6	64 – 75	56 – 65
hurricane	> 32.6	> 75	>65



001

treble tone control



Design: T. Giesberts

The treble control works in a similar manner as the bass control elsewhere in this issue, but contains several modifications, of course. One of these is the series network C_1-C_2–R_1–R_{11}.

The d.c. operating point of IC_3 is set with resistors R_{12} and R_{13}. To ensure that these resistors do not (adversely) affect the control characteristics, they are coupled to the junction of R_9 and R_{10}. In this way they only affect the low-frequency noise and the load of the op amp. Their value of 10 kΩ is a reasonable compromise.

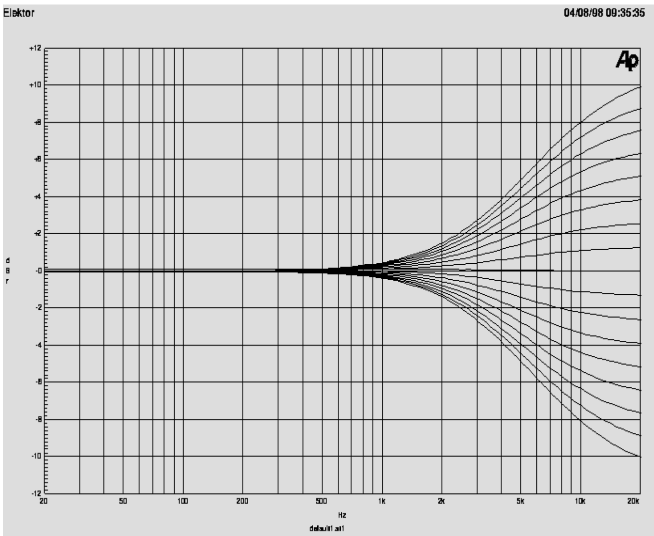
The functions of switches S_1–S_3 are identical to those of their counterparts in the bass tone control; their influence is seen clearly in the characteristics.

Good symmetry between the left-hand and right-hand channels is obtained by the use of 1% versions of R_1–R_{13} and C_1, C_2.

The value of resistors R_2–R_{10} is purposely different from that of their counterparts in the bass tone control. In the present circuit, the control range starts above 20 kHz. To make sure that a control range of 10 dB is available at 20 kHz, the nominal amplification is ×3.5 (11 dB).

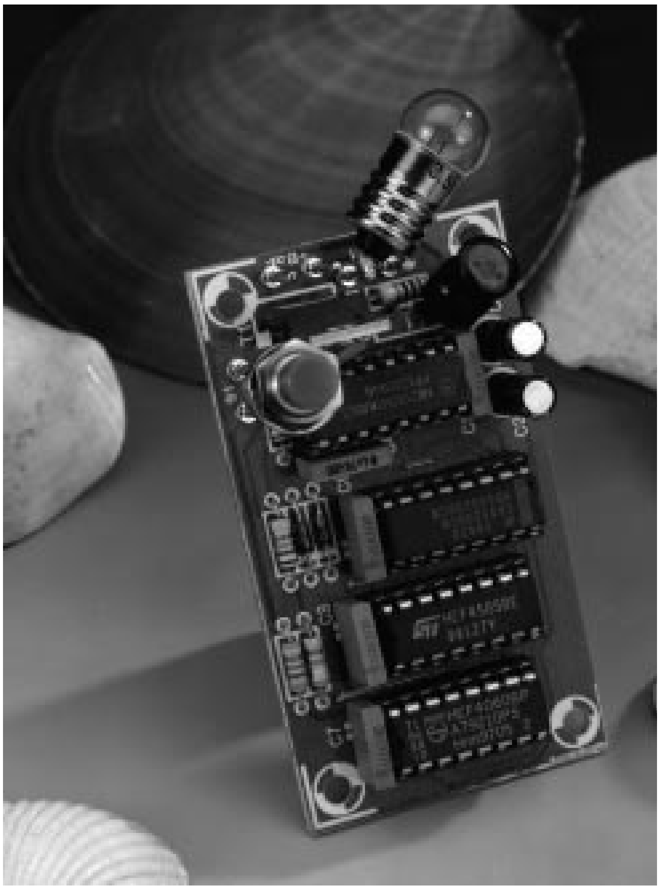
The control circuit draws a current of about ± 10 mA.

[984115]



002

torchlight dimmer

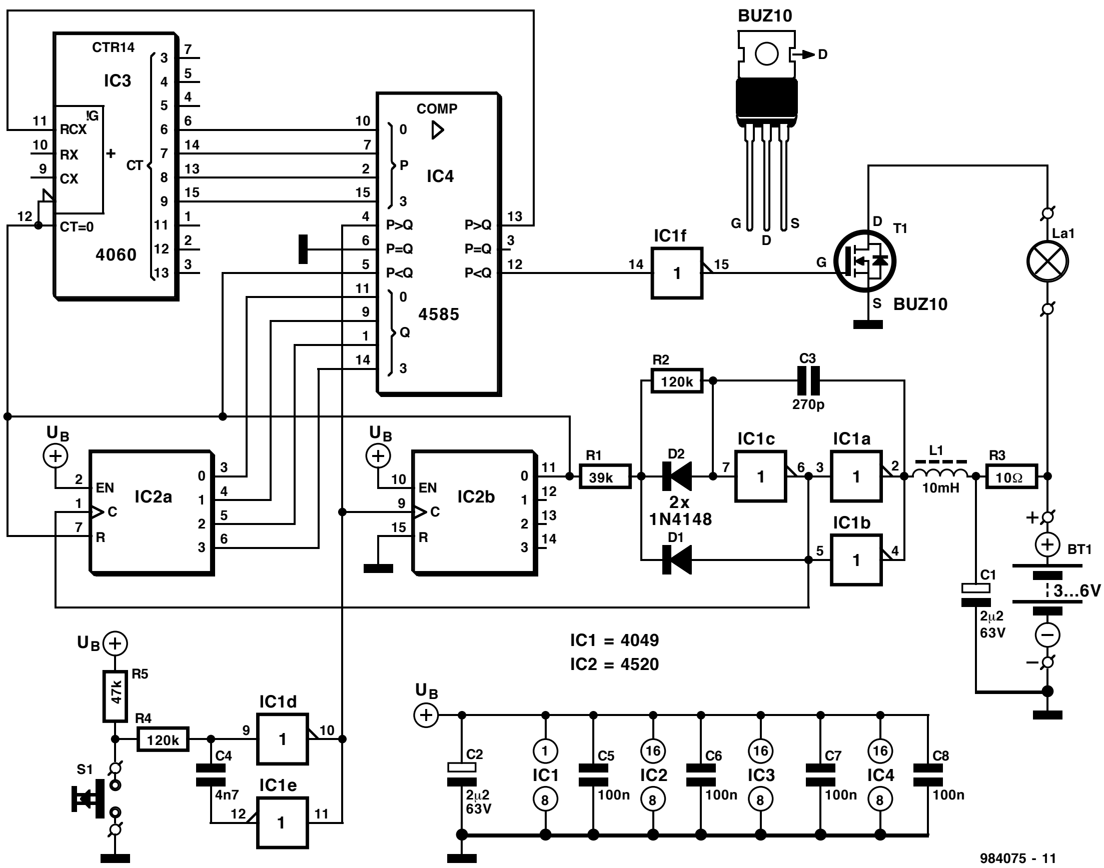


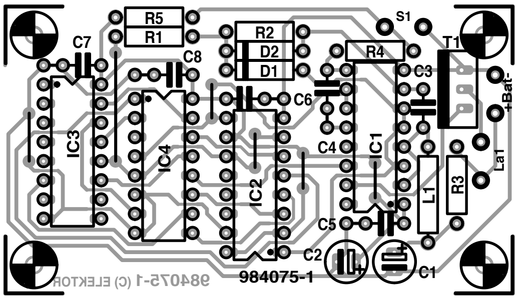
Design: F. Rimatzki

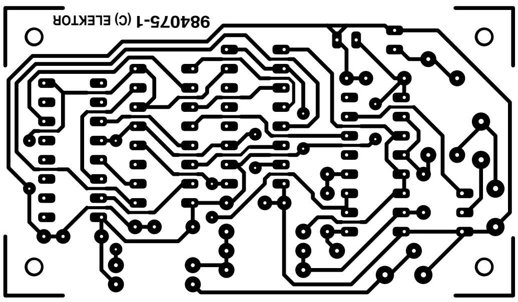
This circuit was originally designed to control the brightness of an electric torchlight, but may find many other applications because of its high efficiency, ease of operation and ability to control (lamp) loads drawing several amps. The dimmer offers brightness control from nil to maximum in 16 steps by means of a small push-button. When the push-button is released, the selected brightness is retained. One of the most remarkable things about this circuit is that it hardly adds to the battery load, its own current consumption amounting to no more than about 4 mA (at a battery voltage of 3.5 V).

The 16 discrete brightness values are obtained by comparing two counter states. One of these actually determines the lamp brightness, while the other performs a cyclic count from 0 to 15. The lamp current is then only switched on if the second value is smaller than or equal to the first. To make sure the switching losses remain as small as possible, a power MOSFET with a very low on-resistance is used. The BUZ10 used here does, however, call for a drive voltage of at least 6 V, so that an additional voltage step-up converter is required.

Counter IC2b only acts as a bistable to allow the circuit to be switched on by means of the lamp brightness push-button, S1. The circuit is switched off (current consumption: less than 5 μA) if output Q0 of IC2b (pin 11) supplies a logic high level. The 27-kHz (approx.) oscillator built from gates IC1a, IC1b and IC1c is then disabled, so that the outputs of IC1a and IC1b are logic high. The ICs in the circuit are then powered via choke L1 and the output transistors of IC1a and IC1b. This is unusual but possible because these transistors can also pass a voltage level at the IC outputs to the supply connection, instead of the other way around (which is far more usual). Because of the logic-high level at the reset inputs of counters IC3 and IC2a, comparator IC4 receives input







COMPONENTS LIST

Resistors:
R1 = 39kΩ
R2,R4 = 120kΩ
R3 = 10Ω
R5 = 47kΩ

Capacitors:
C1,C2 = 2µF2 63V radial
C3 = 270pF ceramic
C4 = 4nF7
C5-C8 = 100nF

Inductor:
L1 = 10mH choke

Semiconductors:
D1,D2 = 1N4148
T1 = BUZ10 (Siemens)
IC1 = 4049
IC2 = 4520
IC3 = 4060
IC4 = 4585

Miscellaneous:
S1 = push-button, 1 make contact
Bt1 = torchlight battery, 3-6V
La1 = torchlight lamp

data which causes it to pull its P<Q output (pin 12) logic high. The result is that inverter IC1f pinches off the BUZ10 MOSFET, and the lamp remains off. When the push-button is actuated for the first time, the bistable in IC2b receives a clock pulse from switch debouncing circuit IC1c-IC1d. Next, the counters and the oscillator are enabled. The duty factor of the oscillator signal is determined by resistors R1 and R2. The oscillator output signal is filtered by R3 and C1. Although the step-up converter is only capable of supplying a few mA, that is sufficient for the CMOS ICs and the BUZ10 MOSFET. For battery voltages between 3 and 6 V, the indicated values of R1 and R2 enable a voltage of 8.5 V to about 16 V to be created for powering the ICs and driving the BUZ10.

As long as the push-button is held depressed, the level at the cascading input of IC4, pin 4, causes the P>Q output, pin 13, to be enabled, so that IC3 is clocked. The counter slowly increases the value at the 'P' inputs of the comparator, thereby controlling the duty factor (mark/space ratio) of the signal at the comparator's P<Q output, pin 12. As soon as IC3 reaches its maximum counter state, the signal at pin 13 of IC4 no longer changes, so that the counter is not started at 0 again. The P<Q output then also remains at 0, so that T1 is driven hard and the lamp lights at maximum brightness. If the push-button is released before the maximum brightness is reached, counter IC3 no longer receives clock pulses and 'freezes' at the current state, causing the lamp to light at the selected brightness. The next action on S1 resets the entire circuit and switches the lamp off If so desired, the brightness control rate may be reduced by doubling or trebling the value of C3. To compensate the resultant drop in the IC supply voltage, the value of choke L1 then has to be increased proportionally. The IC supply voltage should always be between 8 V and 16 V (maximum value of 4000 series CMOS ICs).

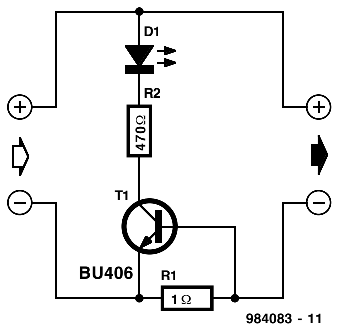
The circuit is best built on a printed circuit board of which the templates are shown here. Unfortunately this board is not available ready-made through the Publishers.

(984075-1, Gb)

003 battery-charging indicator for mains adaptor

Design: J. Gonzalez

Although you may well be the proud owner of the very latest NiCd battery charger, you may still come across the odd 'incompatible' battery, for example, one having a rare voltage or requiring a much higher charging current than can be supplied by your off-the-shelf charger. In these cases, many of you will resort to an adjustable mains adaptor (say, a 500-mA type) because that is probably the cheapest way of providing the direct voltage required to charge the battery. Not fast and not very efficient, this 'rustic' charging system works, although subject to the following restrictions:

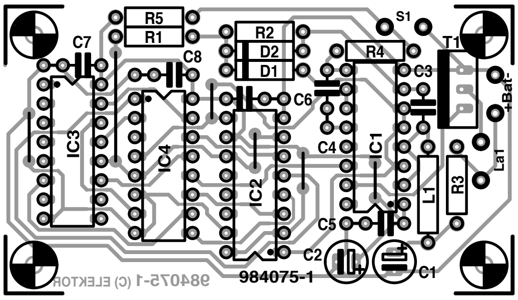


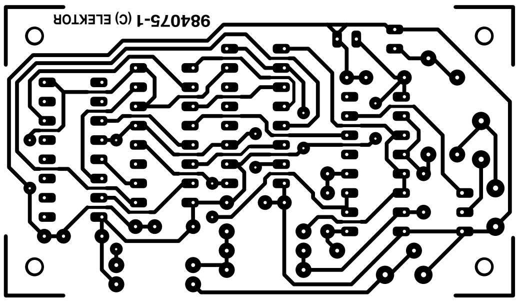
1. You should have some idea of the charging current. In case you use an adaptor which is adjustable but of the unregulated, low output current type, you can adjust the current by adjusting the output voltage.
2. You have to know if the current actually flows through the battery. A current-detecting indicator is therefore much to be preferred over a voltage indicator.
3. To prevent you from forgetting all about the charging cycle, the indicator should be visible from wherever you pass by frequently.

Using the circuit shown here, the LED lights when the base-emitter potential of the transistor exceeds about 0.2 V. Using a resistor of 1 Ω as suggested this happens at a current of about 200 mA, or about 40 mA if R_1 is changed to 4.7 Ω.

The voltage drop caused by this indicator can never exceed the base-emitter voltage (U_{BE}) of the transistor, or about 0.7 V. Even if the current through R_1 continues to increase beyond the level at which U_{BE} = 0.7 V, the base of the transistor will 'absorb' the excess current. The TO-220 style BU406 transistor suggested here is capable of accepting base currents up to 4 A.

Using this charging indicator you have overcome the restrictions 2 and 3 mentioned above.





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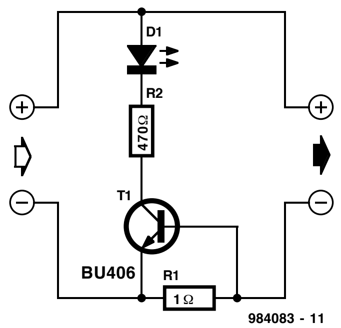

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Using this charging indicator you have overcome the restrictions 2 and 3 mentioned above.

What remains is the problem of knowing the required current. As long as U_{BE} remains below 0.6 V or so, the voltage across R_1 is a faithful indication of the charging current. Alternatively, you may insert an ammeter to find out about the charging currents produced at different output voltage settings on the adaptor. Next, you choose between reasonably fast charging, say, in about 5 hours using C(Ah)/5, or slower, say, 10 hours at C(Ah)/10. C(Ah) is the battery capacity in (milli-) ampere-hours, which is usually printed on the battery. In general, the lower the charging current, the smaller the risk of damage to the battery if you forget to switch off the charger.

In some cases it will be possible to incorporate the circuit into the mains adaptor. That may be dangerous, however, because of the presence of the mains voltage in the adaptor housing. A safer alternative is to install the circuit in a remote control box.

The circuit is not protected against reversal of the battery polarity. If such protection is required, a fuse or circuit breaker should be added.

[984083]

004

playback amplifier for cassette deck

Design: T. Giesberts

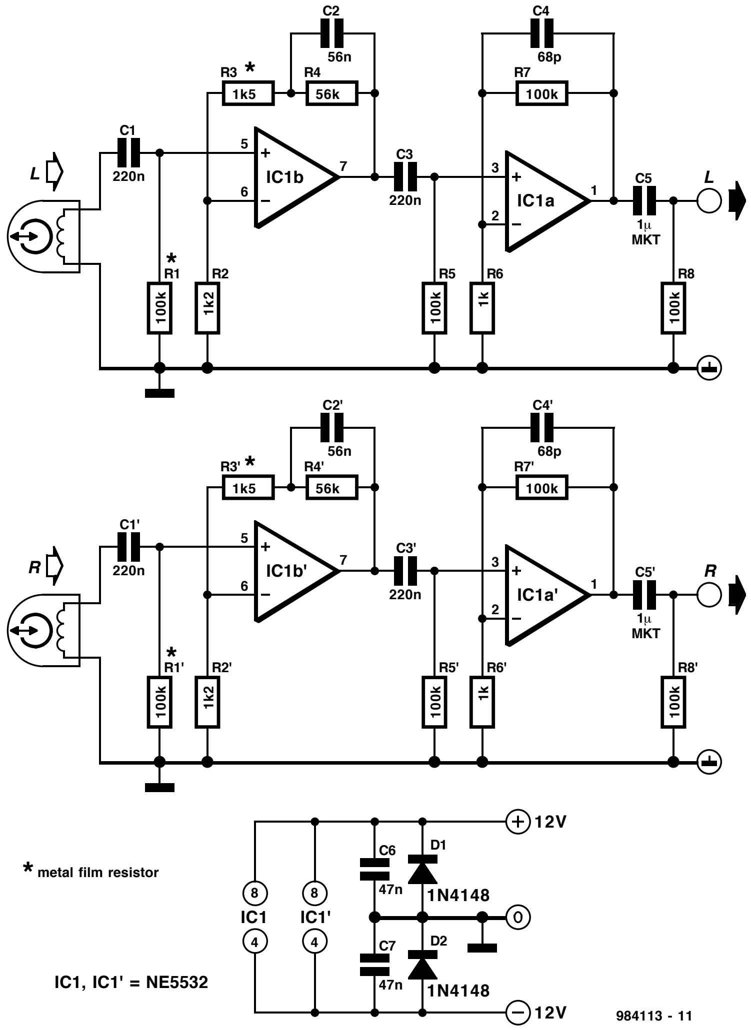
For some time now, there have been a number of tape cassette decks available at low prices from mail order businesses and electronics retailers. Such decks do not contain any electronics, of course. It is not easy to build a recording amplifier and the fairly complex magnetic biasing circuits, but a playback amplifier is not too difficult as the present one shows.

The stereo circuits in the diagram, in conjunction with a suitable deck, form a good-quality cassette player. The distortion and frequency range (up to 23 kHz) are up to good standards. Moreover, the circuit can be built on a small board for incorporation with the deck in a suitable enclosure.

Both terminals of coupling capacitor C_1 are at ground potential when the amplifier is switched on. Because of the symmetrical ±12 V supply lines, the capacitor will not be charged. If a single supply is used, the initial surge when the capacitor is being charged causes a loud click in the loudspeaker and, worse, magnetizes the tape.

The playback head provides an audio signal at a level of 200–500 mV. The two amplifiers raise this to line level, not linearly, but in accordance with the RIAA equalization characteristic for tape recorders. Broadly speaking, this characteristic divides the frequency range into three bands:

- Up to 50 Hz, corresponding to a time constant of 3.18 ms, the signal is highly and linearly amplified.
- Between 50 Hz and 1.326 kHz, corresponding to a time constant of 120 μs, for normal tape, or 2.274 kHz, corresponding to a time constant of 70 μs, for chromium dioxide tape, the signal is amplified at a steadily decreasing rate.
- Above 1.326 kHz or 2.274 kHz, as the case may be, the signal is slightly and linearly amplified.



This characteristic is determined entirely by A_1 (A_1'). To make the amplifier suitable for use with chromium dioxide tape, add a double-pole switch (for stereo) to connect a 2.2 kΩ resistor in parallel with R_3 (R_3').

The output of A_1 (A_1') is applied to a passive high-pass rumble filter, C_3-R_5 (C_3'-R_5') with a very low cut-off frequency of 7 Hz. The components of this filter have exactly the same value as the input filter, C_1-R_1 (C_1'-R_1').

The second stage, A_2 (A_2') amplifies the signal ×100, that is, to line level (1 V r.m.s.).

What remains is the problem of knowing the required current. As long as U_{BE} remains below 0.6 V or so, the voltage across R_1 is a faithful indication of the charging current. Alternatively, you may insert an ammeter to find out about the charging currents produced at different output voltage settings on the adaptor. Next, you choose between reasonably fast charging, say, in about 5 hours using C(Ah)/5, or slower, say, 10 hours at C(Ah)/10. C(Ah) is the battery capacity in (milli-) ampere-hours, which is usually printed on the battery. In general, the lower the charging current, the smaller the risk of damage to the battery if you forget to switch off the charger.

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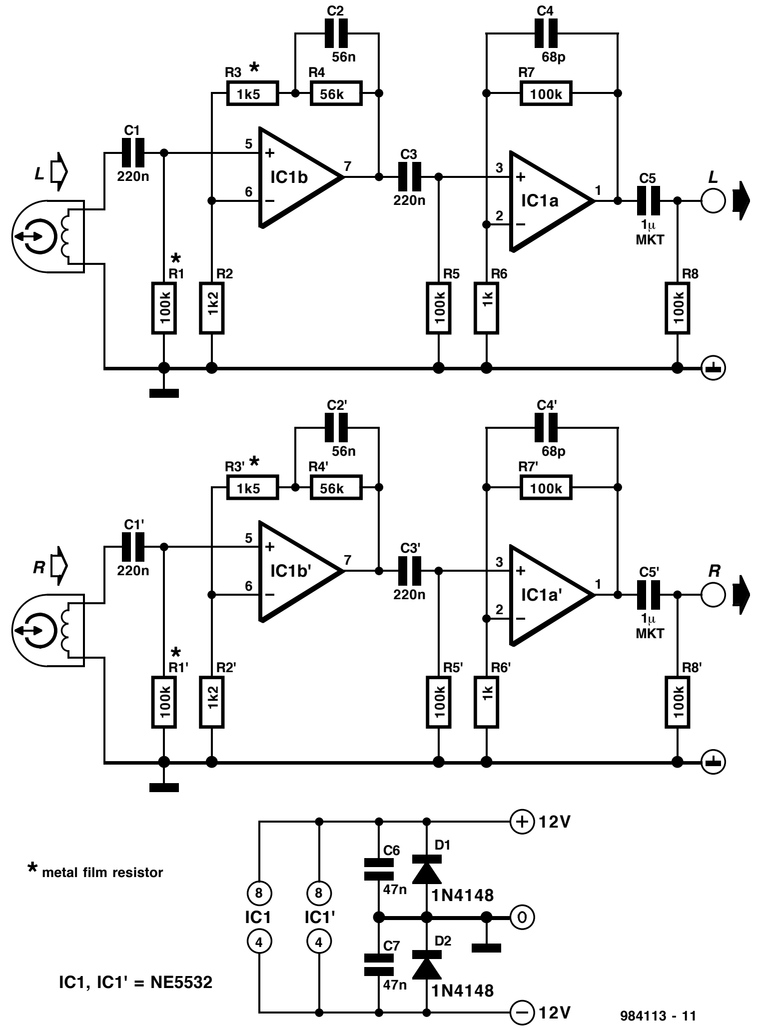
Both terminals of coupling capacitor C_1 are at ground potential when the amplifier is switched on. Because of the symmetrical ±12 V supply lines, the capacitor will not be charged. If a single supply is used, the initial surge when the capacitor is being charged causes a loud click in the loudspeaker and, worse, magnetizes the tape.



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- Above 1.326 kHz or 2.274 kHz, as the case may be, the signal is slightly and linearly amplified.

This characteristic is determined entirely by A_1 (A_1'). To make the amplifier suitable for use with chromium dioxide tape, add a double-pole switch (for stereo) to connect a 2.2 kΩ resistor in parallel with R_3 (R_3').

The output of A_1 (A_1') is applied to a passive high-pass rumble filter, C_3-R_5 (C_3'-R_5') with a very low cut-off frequency of 7 Hz. The components of this filter have exactly the same value as the input filter, C_1-R_1 (C_1'-R_1').

The second stage, A_2 (A_2') amplifies the signal ×100, that is, to line level (1 V r.m.s.).

Capacitor C_4 limits the upper frequency range to avoid r.f. interference and any tendency of the amplifier to oscillate.

The amplifier needs a symmetrical ±12 V power supply that can provide a current of up to 0.5 A. The greater part of this current is drawn by the motor of the deck; the electronic circuits draw only 15 mA.

[984113]

005 memory change-over tip

Design: L. Lemmens

When the contents of two existing memory address have to be interchanged for one reason or an other, there is usually a need for an additional address or variable:

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MOV dummy,var1
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```

This dummy variable is not always necessary:

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XOR var1,.var2
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This tip may well be of use when the memory space is limited. It may also be used with higher programming languages to save having to declare an additional variable.

[984080]

006 improved power-down for the 8051

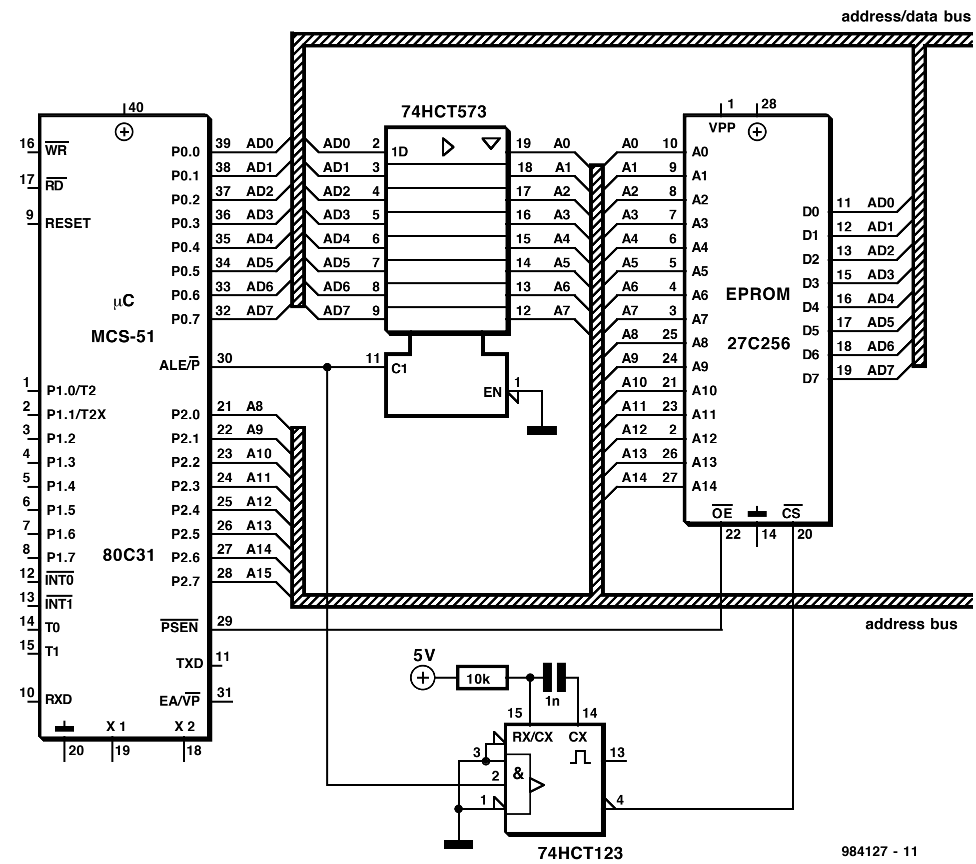
Design: G. Kleine

Members of the 8051 family of microcontrollers (MCS51) are well-known and widely used. The controllers have a power-down mode in which the program processing is suspended by the clock oscillator and ended with a power-down instruction. To reduce the current drain, the supply voltage is reduced to a minimum of 2 V after the powered-down mode has been selected. This mode can only be disabled by a reset, for which the supply voltage needs to be returned to 5 V.

In simple applications of the 8051, the EPROM containing the program to be executed is enabled by making $\overline{\text{PSEN}}$ (program storage enable) active via its $\overline{\text{OE}}$ (output enable) terminal. There are also circuits where $\overline{\text{PSEN}}$ acts on the $\overline{\text{CS}}$ (chip select) terminal of the EPROM.

Use of the power-down mode has a drawback: line ALE (address latch enable), like $\overline{\text{PSEN}}$, remains low during the power-down mode and so holds the EPROM active. It occupies the address/data bus with the accidentally same addressed byte.

This drawback can be removed by the circuit in the diagram. A retriggerable monostable evaluates the low and high edges of the ALE signal, which after a power-down and before a reset has a clock pulse. The output of the monostable sets a high on the $\overline{\text{CS}}$ input of the EPROM when the power-down mode is selected (and when, consequently, the disabled quartz oscillator can no longer generate an ALE pulse).This arrangement ensures that the EPROM can also be switched to the power-down mode.



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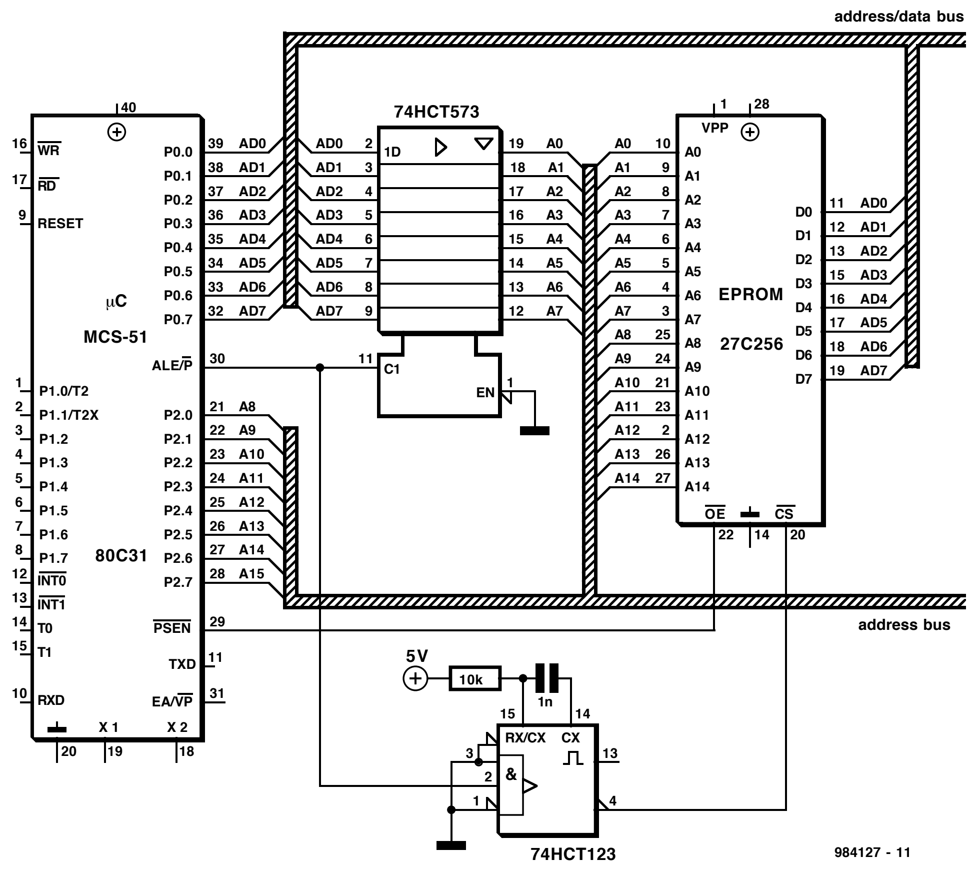

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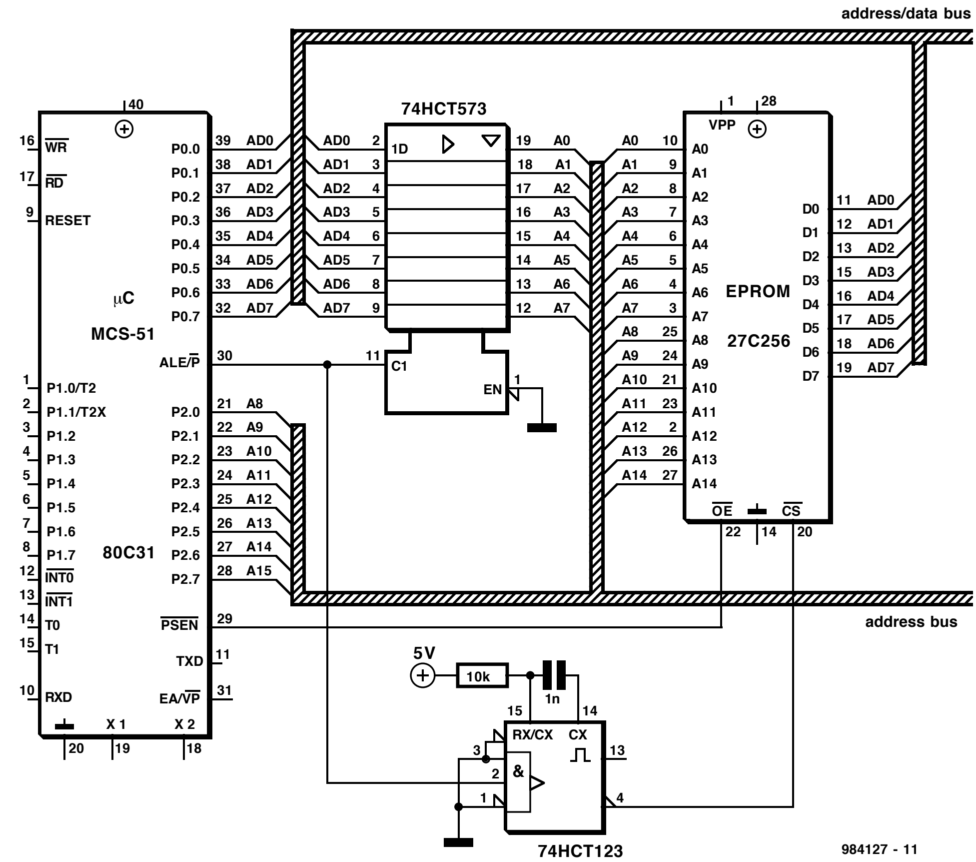

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output may also be set on the address decoder of the system, or the combined $\overline{CS}$ lines of other peripheral equipment, so that these are also in the power-down mode.

Note that with component values as specified, the monostable has a time constant of about 4.5 μs.

[984127]

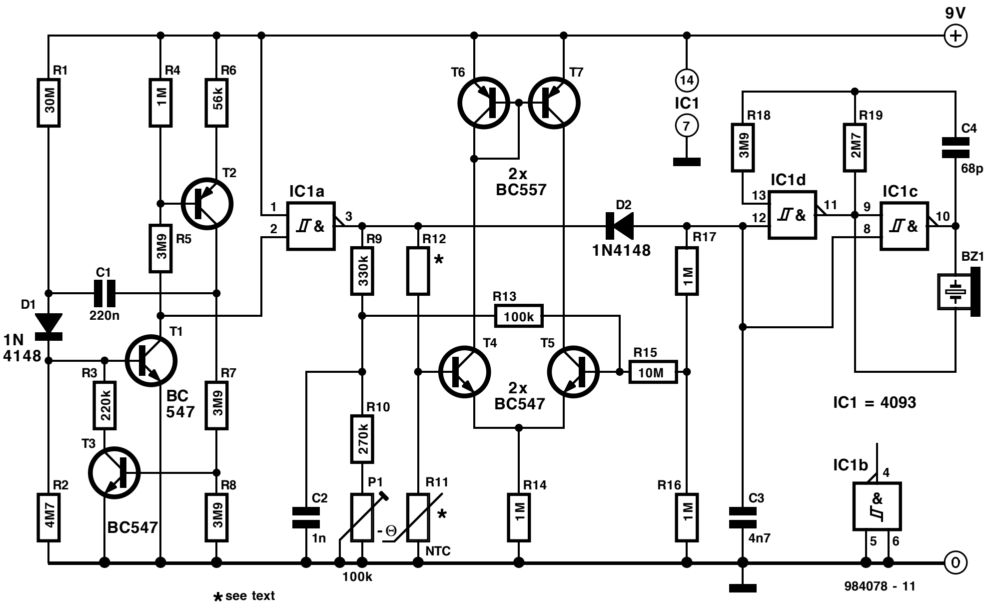
Further reading:

Elektor Electronics, March 1998, '80C32-BASIC control computer'.

Elektor Electronics, September 1997, 'Data acquisition system'

Elektor Electronics, June 1997, 80C537 microcontroller board'

007 general-purpose alarm



Design: K. Syttkus

The alarm may be used for a variety of applications, such as frost monitor, room temperature monitor, and so on.

In the quiescent state, the circuit draws a current of only a few microamperes, so that, in theory at least, a 9 V dry battery (PP3, 6AM6, MN1604, 6LR61) should last for up to ten years. Such a tiny current is not possible when ICs are used, and the circuit is therefore a discrete design.

Every four seconds a measuring bridge, which actuates a Schmitt trigger, is switched on for 150 ms by a clock generator. In that period of 150 ms, the resistance of an NTC thermistor, R_{11}, is compared with that of a fixed resistor. If the former is less than the latter, the alarm is set off.

When the circuit is switched on, capacitor C_1 is not charged and transistors T_1–T_3 are off. After switch-on, C_1 is charged gradually via R_1, R_7, and R_8, until the base voltage of T_1 exceeds the threshold bias. Transistor T_1 then comes on and causes T_2 and T_3 to conduct also. Thereupon, C_1 is charged via current source T_1-T_2-D_1, until the current from the source becomes smaller than that flowing through R_3 and T_3 (about 3 μA). This results in T_1 switching off, so that, owing to the coupling with C_1, the entire circuit is disabled.

Capacitor C_1 is (almost) fully charged, so that the anode potential of D_1 drops well below 0 V. Only when C_1 is charged again can a new cycle begin.

It is obvious that the larger part of the current is used for charging C_1.

Gate IC_{1a} functions as impedance inverter and feedback stage, and regularly switches on measurement bridge R_9–R_{12}-C_2-P_1 briefly. The bridge is terminated in a differential amplifier, which, in spite of the tiny current (and the consequent small transconductance of the transistors) provides a large amplification and, therefore, a high sensitivity.

Resistors R_{13} and R_{15} provide through a kind of hysteresis a Schmitt trigger input for the differential amplifier, which results in unambiguous and fast measurement results.

Capacitor C_2 compensates for the capacitive effect of long cables between sensor and circuit and so prevents false alarms.

If the sensor (R_{11}) is built in the same enclosure as the remainder of the circuit (as, for instance, in a room temperature monitor), C_2 and R_{13} may be omitted. In that case,C_3 will absorb any interference signals and so prevent false alarms.

To prevent any residual charge in C_3 causing a false alarm when the bridge is in equilibrium, the capacitor is discharged rapidly via D_2 when this happens.

Gates IC_{1c} and IC_{1d} form an oscillator to drive the buzzer (an a.c. type).

Owing to the very high impedance of the clock, an epoxy resin (not pertinax) board must be used for building the alarm. For the same reason, C_1 should be a type with very low leakage current.

If operation of the alarm is required when the resistance of R_{11} is *higher* than that of the fixed resistor, reverse the connections of the elements of the bridge and thus effectively the inverting and non-inverting inputs of the differential amplifier.

An NTC thermistor such as R_{11} has a resistance at –18 °C that is about ten times as high as that at room temperature. It is, therefore, advisable, if not a must, when precise operation is required, to consult the data sheet of the device or take a number of test readings.

For the present circuit, the resistance at –18 °C must be 300–400 kΩ. The value of R_{12} should be the same. Preset P_1 provides fine adjustment of the response threshold.

Note that although the prototype uses an NTC thermistor, a different kind of sensor may also be used, provided its electrical specification is known and suits the present circuit. [984078]

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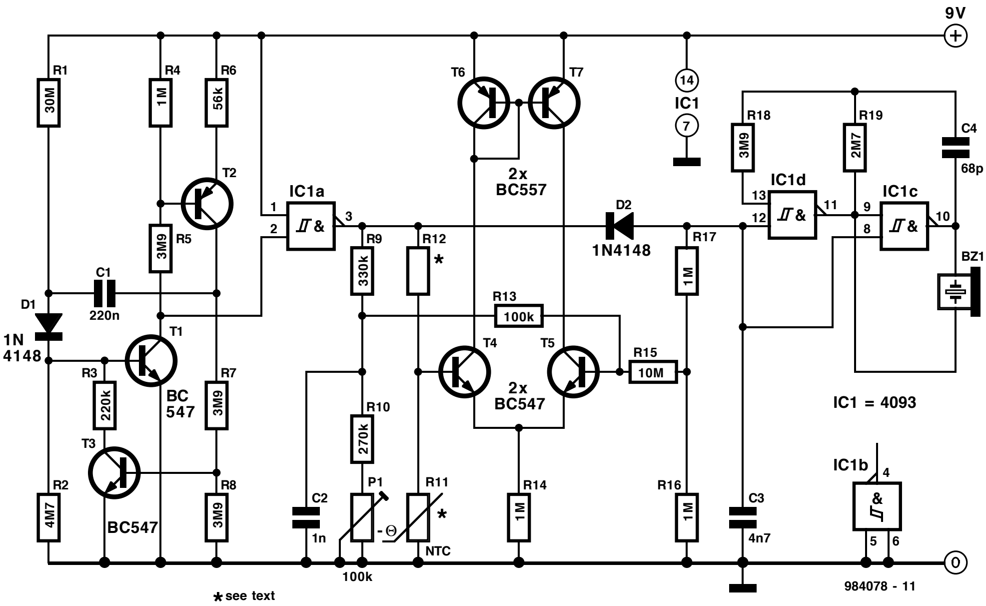

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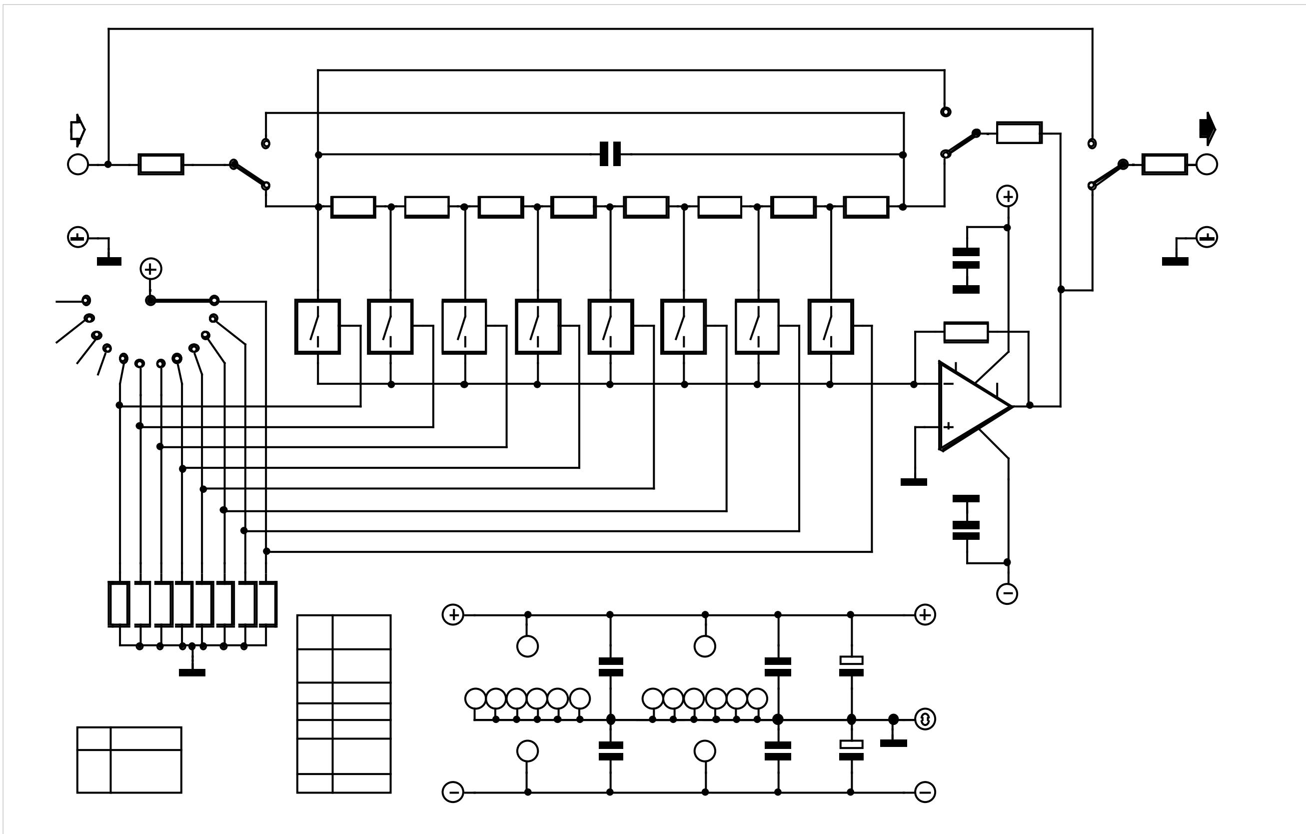
008

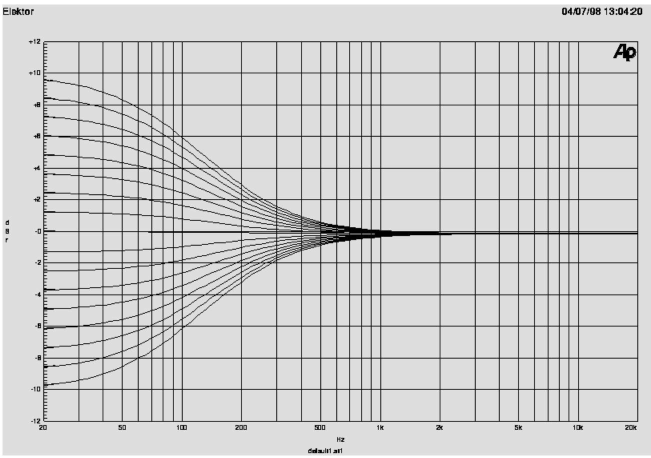
accurate bass tone control



Design: T. Giesberts

A difficult problem in the design of conventional stereo tone controls is obtaining synchronous travel of the potentiometers. Even a slight error in synchrony can cause phase and amplitude differences between the two channels. Moreover, linear potentiometers are often used in such controls, and these give rise to unequal performance by human hearing. Special potentiometers that counter these difficulties are normally hard to obtain in retail shops.

A good alternative is a control based on a rotary switch and a discrete potential divider. The problem with this that for good tone control more than six steps are needed, and switches for this are also not readily available. Fortunately, electronic circuits can remove these difficulties.



The analogue selectors used may be driven by mechanical switches, standard logic circuit or a microcontroller. The selectors used in the present circuit are Type SSM2404 versions from Analogue Devices, which switch noiselessly. Each IC contains four selectors, so that a total of eight are used. The step size is 1.25 dB at 20 Hz with a maximum of 10 dB.

The circuit can be mirrored with S_1, which means that a selection may be made of amplification or attenuation of bass frequencies. The user can choose between attenuation only and extending the range by dividing R_9. The control can be bridged by switch S_2.

To prevent the output impedance of the circuit having too much effect on the operation of the circuit, the output impedance must be ≤ 10 Ω. Resistor R_{12} protects the circuit against too small a load.

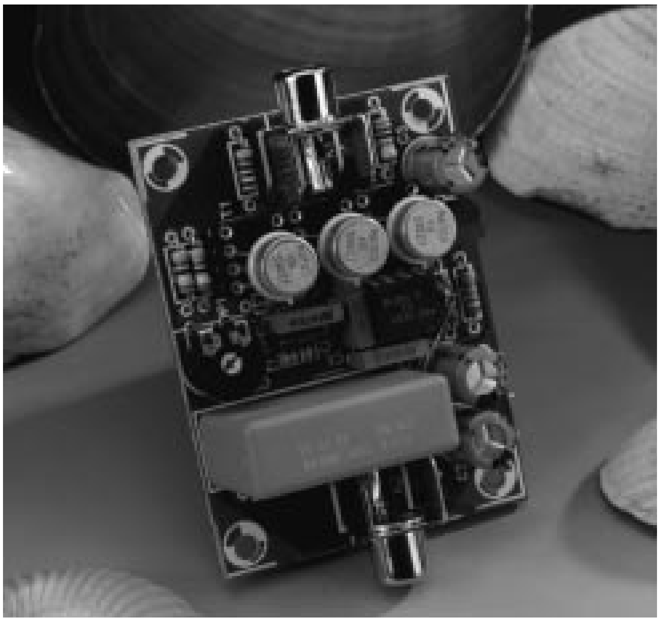
At maximum bass amplification at U_{in} = 1 V r.m.s., the THD+ N < 0.001% for a frequency range of 20 Hz to 20 kHz and and a bandwidth of 80 kHz.

The circuit draws a current of about 10 mA.

[984117]

009

ultra low-noise MC preamplifier



COMPONENTS LIST

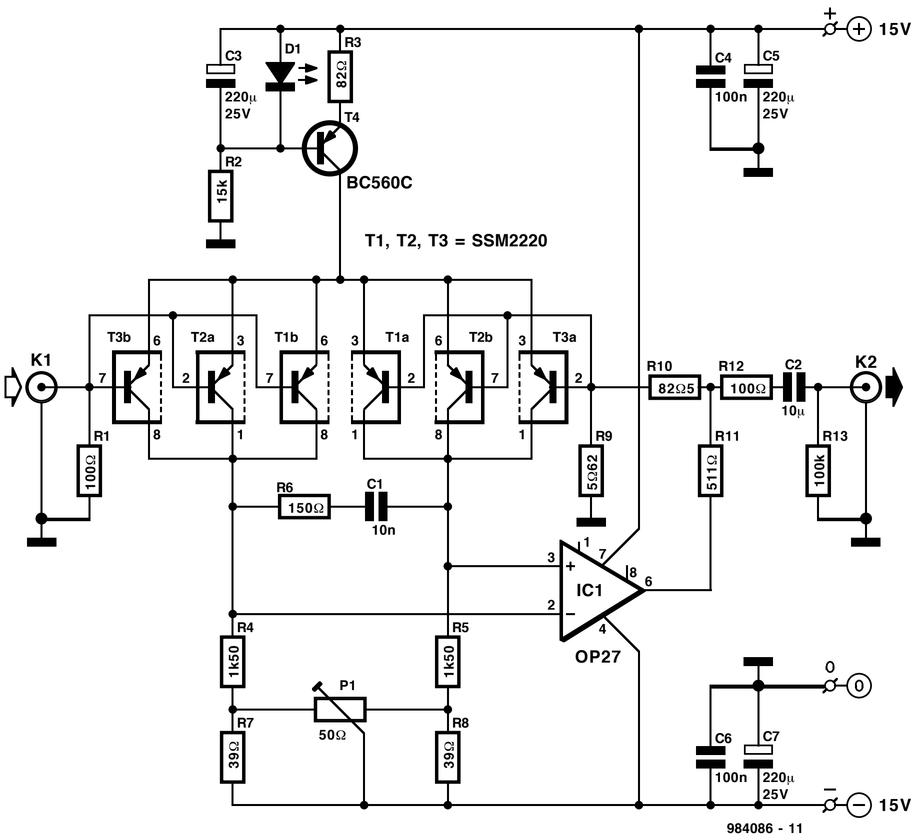
Resistors:
R1,R12 = 100Ω
R2 = 15kΩ
R3 = 82Ω
R4,R5 = 1kΩ50
R6 = 150Ω
R7,R8 = 39Ω
R9 = 5Ω62
R10 = 82Ω5
R11 = 511Ω
R13 = 100kΩ
P1 = 50Ω preset H

Capacitors:
C1 = 10nF
C2 = 10µF MKT (Siemens) raster 22.5mm or 27.5mm
C3,C5,C7 = 220µF 25V radial
C4,C6 = 100nF

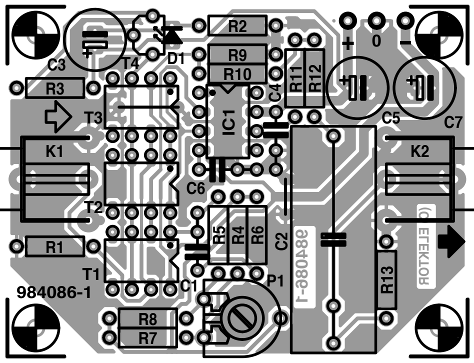
Semiconductors:
D1 = red LED, flat
T1,T2,T3 = SSM2220 or MAT03 (Analog Devices)
T4 = BC560C
IC1 = OP27GP (Analog Devices)

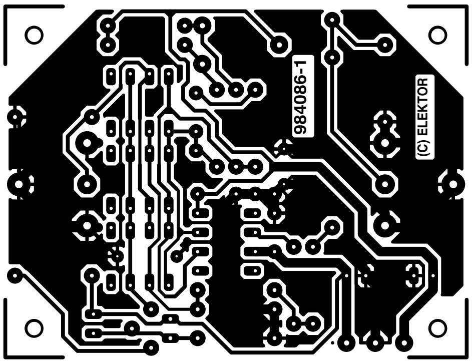
Miscellaneous:
K1,K2 = phono (line) socket, PCB mount, gold-plated, e.g. T-709G from Monacor/Monarch (available from C-I Electronics or Stippler Electronics)

This preamplifier was designed for low-impedance signal sources like MC (moving-coil) pick-up cartridges used in high-end record players (yes, they still exist). The actual input impedance of the preamplifier is 100 Ω. To keep the input noise as low as possible, three dual transistors type SSM2220 or MAT03 transistors are connected in parallel to form a discrete difference amplifier. By connecting this amplifier ahead of an opamp (OP27), the input noise of the opamp becomes immaterial. The base connections of the discrete amplifier then function as the inputs of a super-opamp with a very low input noise level. An advantage of the p-n-p transistors used here over their n-p-n counterparts is their much lower low-frequency noise level. On the down side, a fairly large bias current of about 5.5 µA is created at the input. This is the result of the 2-mA setting for each transistor in combination with the relatively low gain of the p-n-p devices.



Preset P1 and resistors R7/R8 enable you to iron out any tolerances on R4 and R5 in the difference amplifier output. Transistor T4 and LED D1 ensure a stable current setting for the difference amplifier. D1 should be a flat, red, LED which is fitted face-to-face against T4 for thermal coupling. Because the input noise level amounts to 0.4 nV/√Hz (theoretical value for a 10-Ω resistor), it is essential that the feedback adds as little as possible to the overall noise figure. Consequently, the impedance of the feedback circuit must be much lower than 10 Ω. Furthermore, the OP27 demands a certain minimum load impedance, so that the feedback impedance may not be less than 600 Ω. To ensure that a low value can be used for R9, a compromise had to be found between maximum gain (here, approx. 24 dB or 15.7 times) on the one hand, and the value of R9. By fitting an additional resistor, R11, ahead of the actual feedback, the opamp is not excessively loaded, while R9 adds 'only' 0.3 nV/√Hz to the input noise level, which, based on measurement data, amounts to 0.52 nV/√Hz. If more gain is needed, a noise figure of about 0.4 nV/√Hz may be achieved at a lower value of R9. The obvious disadvantage of adding R11 is a higher internal gain, causing a smaller bandwidth and a lower drive margin. Fortunately, these factors are of little consequence in the case of moving-coil elements.
There are two ways to adjust P1. The first is to adjust the output





voltage to nil (measure at IC1 pin 6). The second option is to measure the input offset, for example, 0.55 mV across 100 Ω. Assuming that the offset caused by T1, T2 and T3 is negligible, then the output voltage should be 15.68 x 0.55 mV for perfect symmetry, in other words, junction R10-R11-R12 should be at 8.62 mV with respect to ground.

Those of you who like to experiment may want to try the effects of reducing the number of input transistors from three to just one. You may want to do this, for example, to reduce the input bias current. Resistor R3 then has to be changed into 249 Ω. Do remember, however, that the input noise level then rises by 2.5 dB!

The output has a large, solid 10 μF MKT (metal theraphtelate, ask your local Siemens distributor) capacitor to prevent a large offset voltage being applied to the input of an MD amplifier.

The preamplifier is powered by a symmetrical, regulated 15-V supply, and draws about 16 mA on each rail. Finally, here are a few key figures measured on our prototypes:

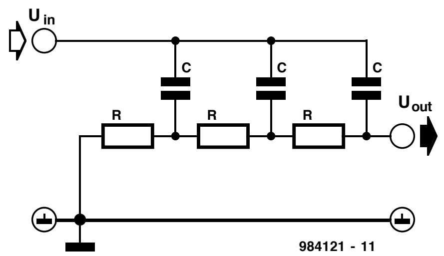
Configuration: 3 x SSM2220/MAT03		
	signal: 0.5 mV/25 Ω	input short-circuited
S/N (BW = 22 kHz)	71.2 dB	74 dB
	74 dBA	76.2 dBA
Configuration: 1 x MAT03 (R3 = 249 Ω)		
S/N (BW = 22 kHz)	69.5 dB	71 dB
	72.3 dBA	73.7 dBA

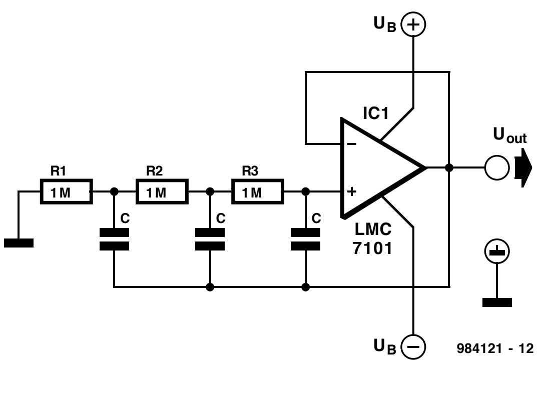
The preamplifier is best built on the printed circuit board whose artwork is shown here. Construction is uncritical, but do not forget the wire links under transistor T3 and next to capacitor C2. The PCB is unfortunately not available ready-made from the Publishers.

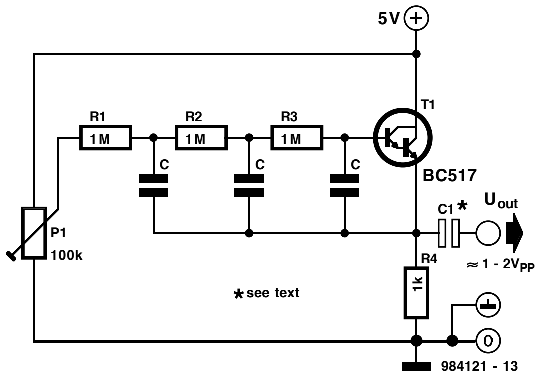
(984086-1, Gb)

010

Philbrick oscillator





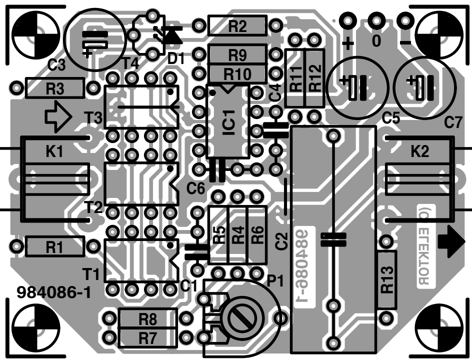


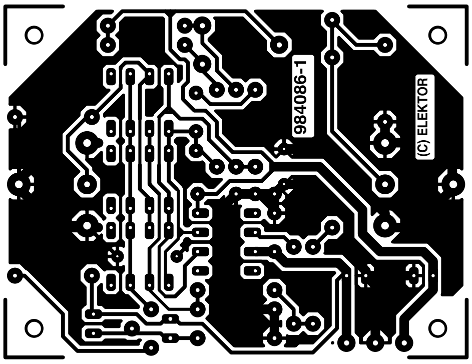
Design: G. Kleine

The Philbrick oscillator is a little known design, patented by the American scientist George A Philbrick in 1956. It generates signals at low amplitude and uses fairly standard components. The circuit, consisting of three resistors and three capacitors (see **Figure 1**), was originally used for d.c. decoupling at the input of oscilloscopes.

Since the step (transient) response of the *RC* network is greater than 1000, it may be used to build an oscillator by feeding back the output signal to the input via a high-resistance voltage follower. The resulting oscillator can generate even very low audio frequencies.

The diagrams in **Figures 2** and **3** show two versions of the oscillator. The one with the op amp has the disadvantage that it needs a symmetrical





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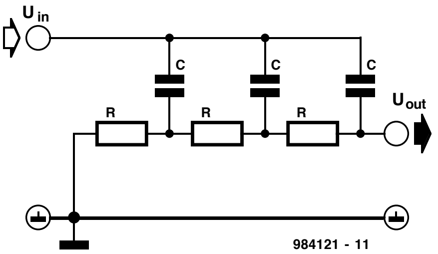
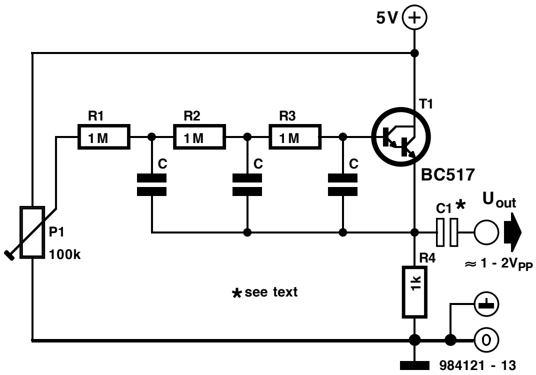
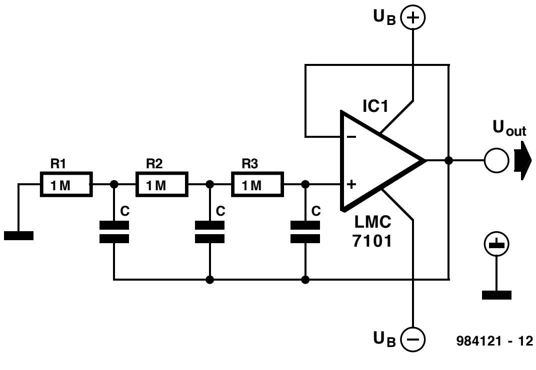
The preamplifier is best built on the printed circuit board whose artwork is shown here. Construction is uncritical, but do not forget the wire links under transistor T3 and next to capacitor C2. The PCB is unfortunately not available ready-made from the Publishers.

(984086-1, Gb)

010 Philbrick oscillator







Design: G. Kleine

The Philbrick oscillator is a little known design, patented by the American scientist George A Philbrick in 1956. It generates signals at low amplitude and uses fairly standard components. The circuit, consisting of three resistors and three capacitors (see **Figure 1**), was originally used for d.c. decoupling at the input of oscilloscopes.

Since the step (transient) response of the *RC* network is greater than 1000, it may be used to build an oscillator by feeding back the output signal to the input via a high-resistance voltage follower. The resulting oscillator can generate even very low audio frequencies.

The diagrams in **Figures 2** and **3** show two versions of the oscillator. The one with the op amp has the disadvantage that it needs a symmetrical

power supply of ±1.5–±7.5 V. If that is a problem, the circuit based on a transistor can be used. This operates from a power supply of +5 V.

The operating point of the emitter follower circuit in Figure 3 is set with P_1 so that oscillations and maximum output voltage are guaranteed.

When the output of the transistor version contains very low near-sinusoidal frequencies, it should be applied to the following stage via an electrolytic capacitor. This capacitor may have to be polarized, depending on whether there is any direct voltage at the input of the following stage.

In the transistor oscillator with resistor values as specified, the following frequencies were measured with the stated capacitor values.

C (nF)	*f* (Hz)
100	5
10	50
1	500

[984121]

011 presence filter

From an idea by R. Sontheimer

To make a certain musical instrument in a group stand out, a so-called presence filter is normally used. Unfortunately, the types usually found in amplifiers and mixers can only raise the level of the instrument output, but not attenuate it.

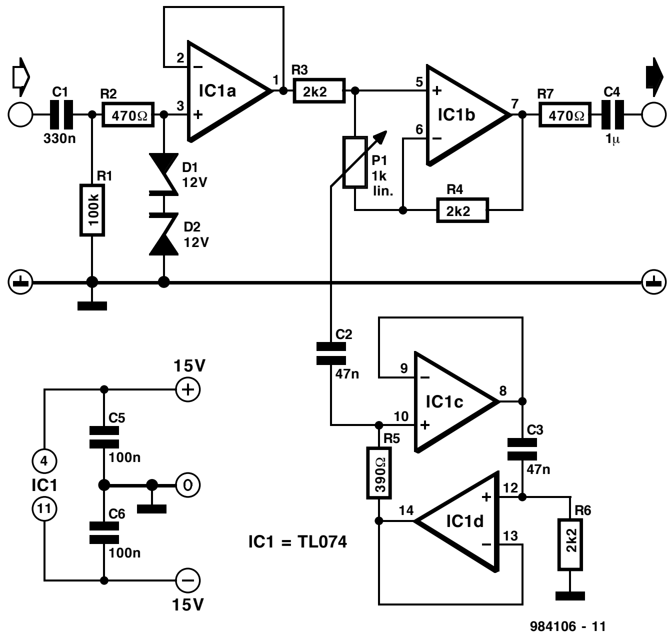
The filter in the diagram provides amplification (15 dB) as well as attenuation (15 dB) over the presence range (see **Figure 1**). When potentiometer P_1 is at its centre position, the signal is unaltered.

The input signal (see **Figure 2**) is applied to impedance converter A_1. Capacitor C_1 blocks any d.c. on the signal. Resistor R_1 sets the input resistance of the circuit. Diodes D_1 and D_2 protect the input against high voltages. Resistor R_2 limits the current to the input of the impedance converter.

The actual filter process is carried out by op amps A_2 and A_3 and associated components. The filter behaves as a frequency-dependent resistance whose value is a minimum at about 3.5 kHz. At very high and very low frequencies, the resistance of the filter is high. Depending on the setting of P_1, the filter forms a potential divider with R_3 or part of the feedback loop with R_4. When P_1 is at its centre position, the filter attenuates the signal to the same degree as it is amplified by A_2.

[984106]

1



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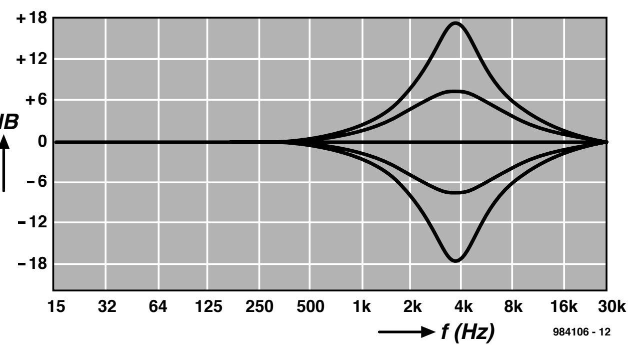

012 polarity reverser

Design K. Lorenz

There are systems in which it is imperative that the supply voltage of, say, a motor, always has the correct polarity. It is, of course, possible to use a bridge rectifier for this, but if large currents are involved, this is not always possible. This may be because large voltage drops across diodes result in appreciable heat dissipation, or that the peak current exceeds the current rating of a diode. Fortunately, a good, inexpensive mechanical rectifier may be

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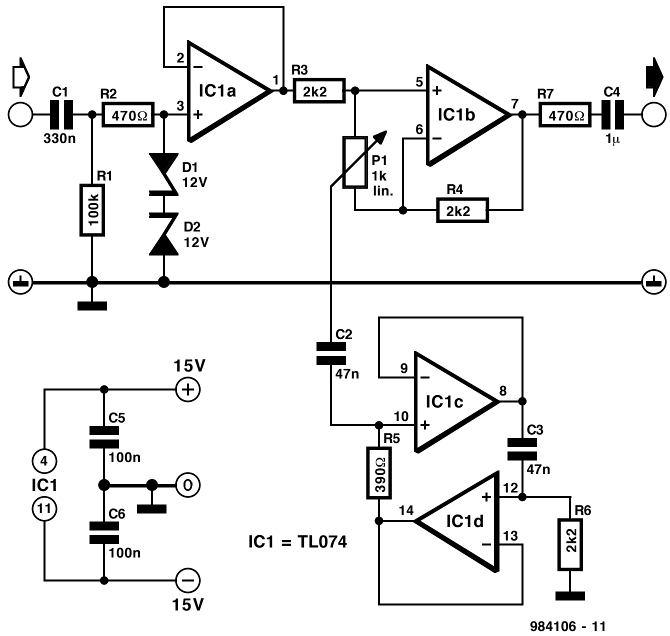

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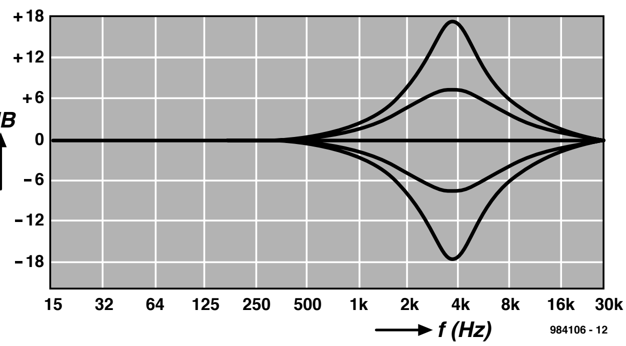

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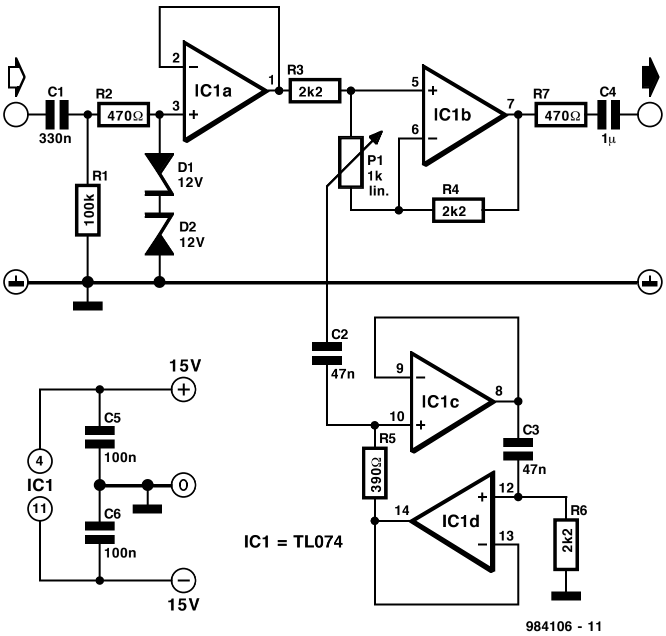

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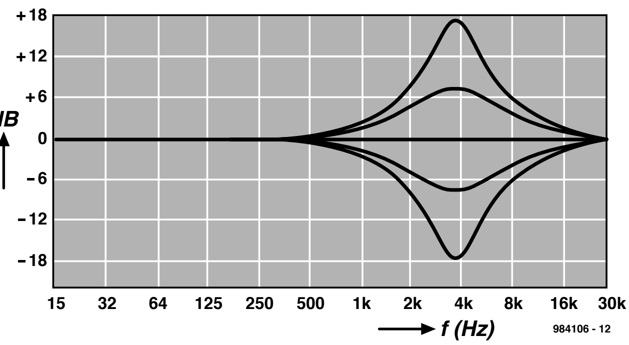

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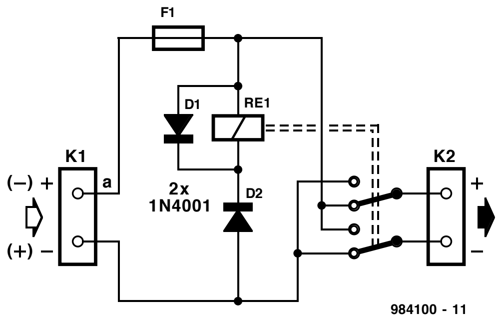
constructed with the aid of a relay.

In the diagram, the supply voltage is applied to K_1, while the motor that needs a supply with correct polarity is linked to K_2. Provided fuse F_1 is intact, a positive potential at terminal a of K_1 will be applied to the positive terminal of K_2. Diode D_2 prevents the relay being energized. When the polarity at K_1 is reversed, the relay will be energized via D_2. The relay contacts then interchange the connections to the terminals of K_2



to ensure that the previous polarity of the supply to the load is retained.

Diode D_1 is a freewheeling diode for the relay coil.

The type of relay to be used depends on the requisite operating voltage and the current through its contacts. Other parts of the circuit are not critical.

It stands to reason that the circuit is not suitable for use with a small battery, since the relay coil draws a fairly large current.

[984100]

013 mains phase indicator

Design M. Hahn

The three voltages of a three-phase supply, L_1, L_2 and L_3 (or R, G and B) are 120° out of phase with one another—see **Figure 1a**. When, for instance, the positive half-wave of L_1 (pin 1) begins, the instantaneous value of L_2 (pin 2) is still negative. The positive half-wave of L_2 starts 120° later and cuts the waveform of L_1 at a level of about half the peak voltage at 150°. At 180° L_1 becomes negative; at 270°, L_2.

When two connections are interchanged as in **Figure 1b**, the positive half-wave of L_1 appears first at pin 2 and then that of L_2 at pin 1. This always happens when connections are interchanged. It is, therefore, necessary only to establish in what order the half-waves arrive at two given terminals to determine the phase. The third connection is not needed.

This requirement is met by the circuit in **Figure 2**. It uses a pair of thyristors, which are arranged so that the first one to be triggered cuts off the other. It should be noted that the circuit is completely symmetrical. Diodes D_1 and D_2 ensure that only positive half-waves are taken into account. The current is limited by R_1 and R_2. The two phases are combined by D_3 and D_4. The higher of the two positive voltages is always at A, and its phase is between 0° and 270°. The potential at A rises until the breakdown voltage (39 V) of

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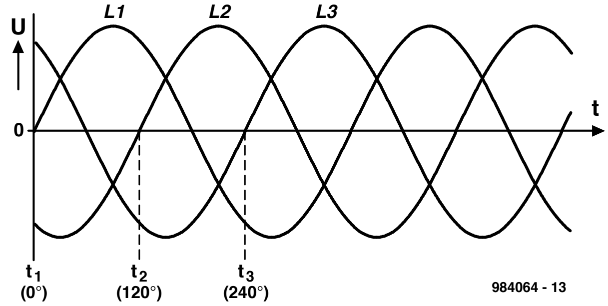

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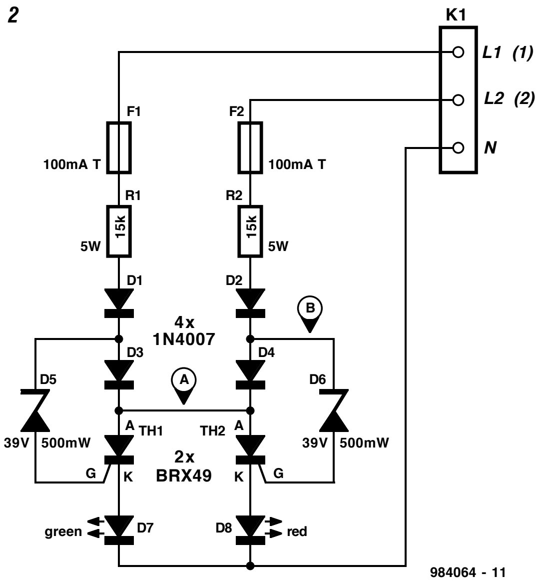

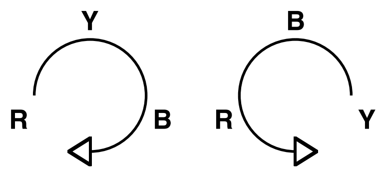

984064 - 12

zener diode D_5 is reached, whereupon thyristor Th_1 comes on and D_7 (green) lights. The potential at A then drops to a level equal to the sum of the breakdown voltage of Th_1 and the drop across D_7.

When a positive half-wave appears at pin 2, the potential at B can be higher only by the diode voltage of D_4 than that at A and this cannot be as high as the zener voltage of D_6. Instead, diode D_7 draws current from terminal 2 in the time interval between 150° and 270°. Thyristor Th_1 is cut off at 270° when L_2 drops below zero and the hold current of the thyristor ceases.

When, however, both terminals are interchanged, Th_2 is triggered first and draws current from terminal 1 at 150°, so that only the red LED (D_8) can light. The 20 ms interval between 270° and 360° cannot be discerned by the human eye.

Since the circuit operates with and from the mains supply, appropriate safety measures must be observed during the construction. It is imperative that the enclosure is strapped to the mains earth. Plugs and sockets used must, of course, be of the appropriate standard, and cable inlets must be provided with a strain relief. Do not use inferior materials!

[984064]

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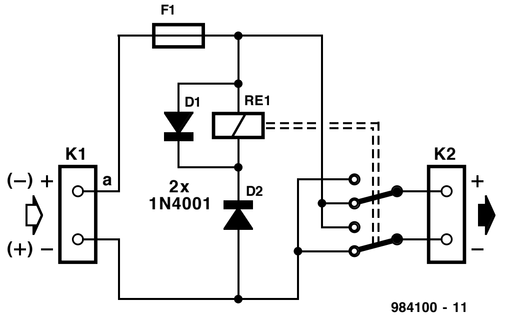

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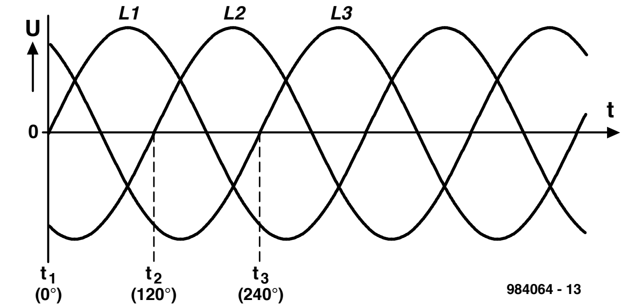

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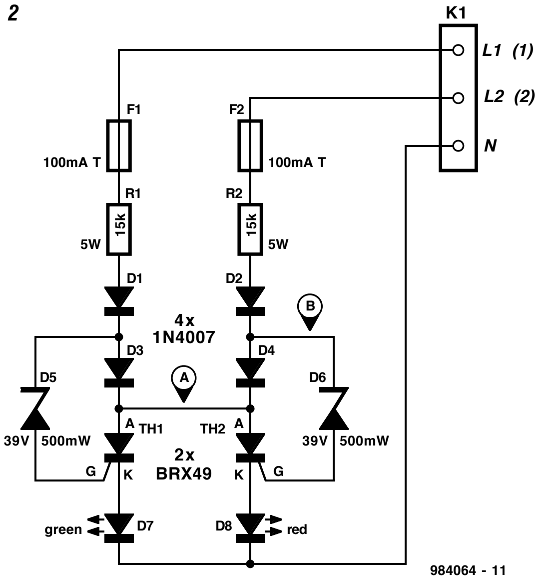

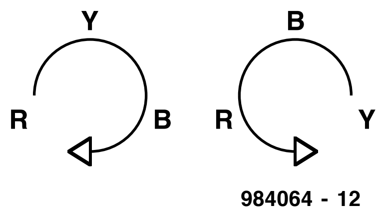

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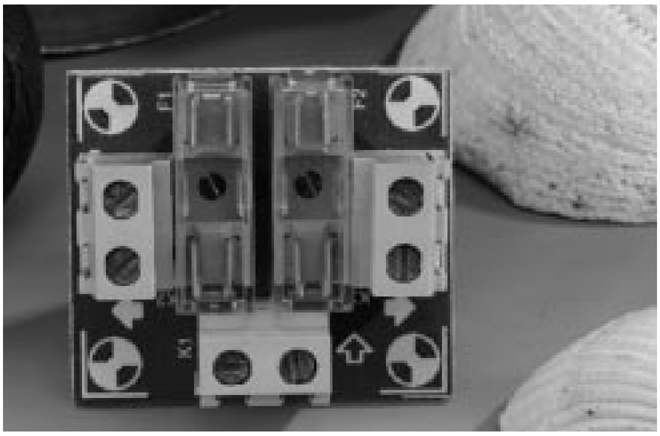
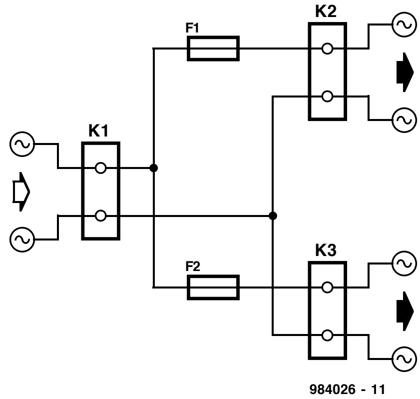
[984064]

014

mains splitter for AF power amps

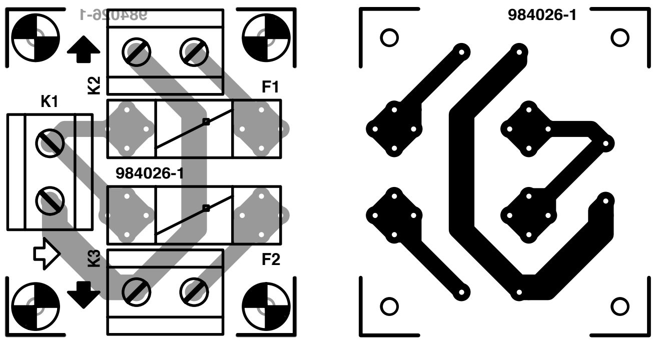




In many home-brew AF power amplifiers, including quite a few built according to the noble art of 'high-end audio amplifier construction', the primaries of the mains transformers are simply connected in parallel and protected by a single, large, fuse. There may be one, hefty, transformer inside the case, or two, each powering a monoblock, or even three, where a smaller one is used to power an ancillary circuit like a protection circuit. Using a single fuse to protect the lot is undesirable because this fuse has to be rated for the rush-in current of the large transformers. Moreover, when the fuse burns out you never know which monoblock, or indeed which other part of the amplifier, is the culprit (although that may be easy to find out by sniffing around or looking for smoke signals…).

The small circuit board shown here allows the mains input voltage to be distributed in a safe manner to two loads, each with its own (properly rated) fuse. Because the 'circuit' does not include an earth line, it may not be used as an external unit, that is, outside an earthed enclosure. For essential notes on electrical safety with mains-operated circuits like this one, please review the Safety Guidelines page which appears occasionally in *Elektor Electronics*. A copy of this page may be obtained from the Publishers.

(984026-1, Gb)



COMPONENTS LIST

K1,K2,K3 = 2-way PCB terminal block, pin distance 7.5mm
F1,F2 = fuseholder, PCB mount, with cap
Two fuses, ratings as required by application
PCB (not available ready-made)

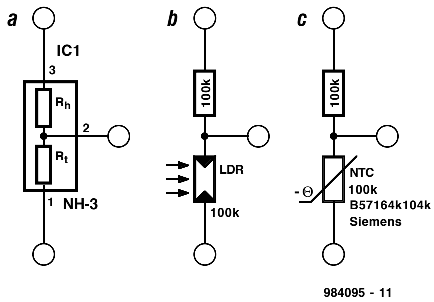
015

modified humidity control

Design: H. Bonekamp

The Type NH-3 humidity sensor used in the 'automatic air humidifier' (July/August 1998) may be replaced by a light-dependent resistor, LDR, or a resistor with negative temperature coefficient, NTC—see diagram. There are other devices as well: the main requirement is that they can be driven by an alternating voltage, that is, that they are non-polarized.

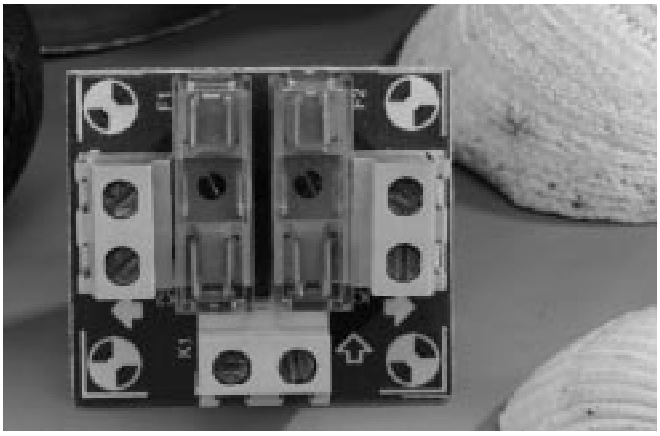
An LDR is usually connected in series with a fixed resistor whose resistance should be equal to that of the LDR when it is not exposed to light.

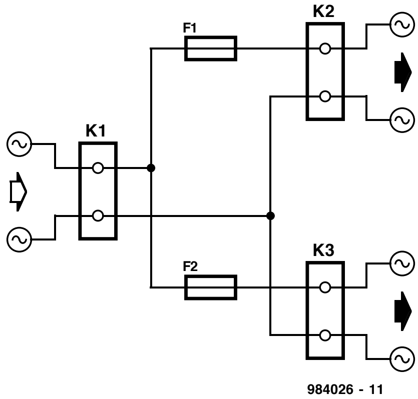


In the diagram (b), the network is connected as a twilight switch, that is, the circuit switches on the mains when the LDR is in darkness. When the two components are interchanged, the circuit switches the mains on when the LDR is exposed to light.

In network (c), the fixed resistor should also have the same value as that with an NTC (at 20° C). This network is suitable for use as thermostat in a greenhouse. When the temperature in there drops below a value set with P_1, the network switches on a heater. [984095]

014 mains splitter for AF power amps





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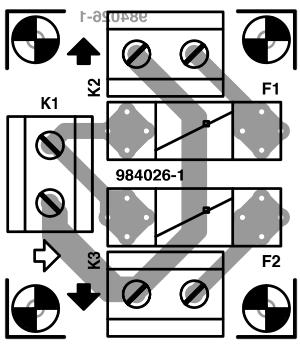

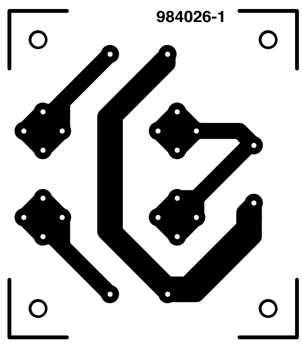

COMPONENTS LIST

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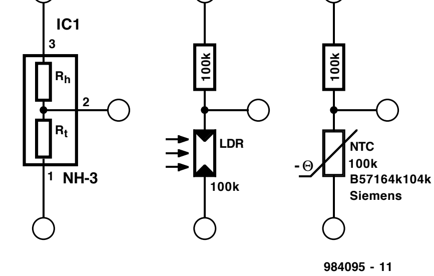

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016

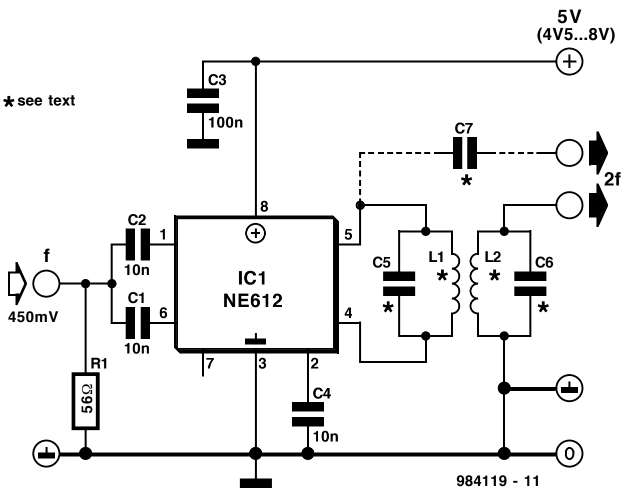
circuit ideas for the NE612

Designs: G. Kleine

The NE612 is an active mixer/oscillator that is used in numerous r.f. circuits. Three unusual applications of this versatile building block are described in this article.

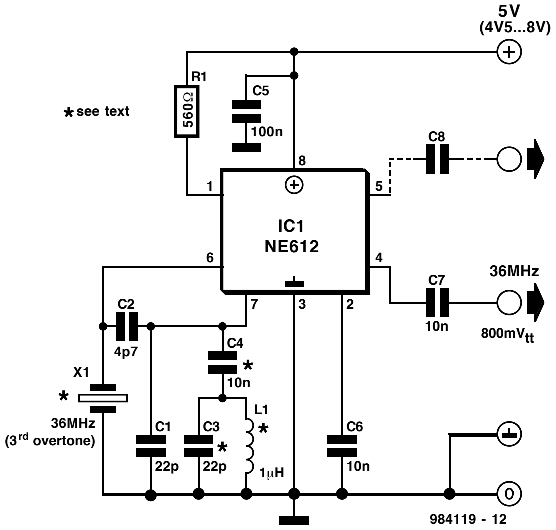
The IC can be arranged as a frequency doubler (**Figure 1**). In this application, pin 6, which is normally linked to the tuned oscillator circuit, is connected to the input. The internal oscillator transistor (base at pin 6; emitter at pin 7) then functions as a linear amplifier. The frequency of the output signal is twice that of the input signal.

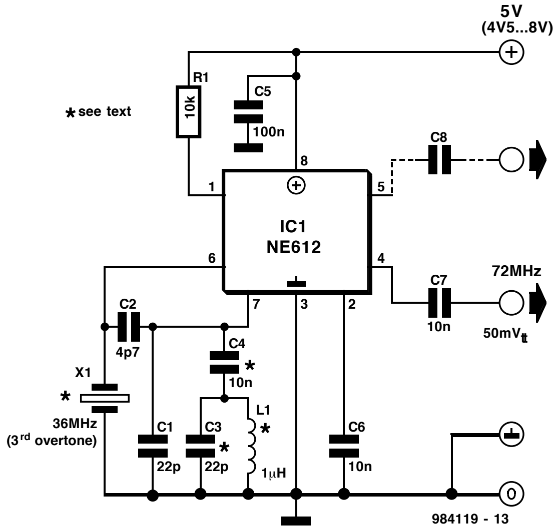
The fundamental frequency, *f*, and harmonics 3*f*, 4*f*, and so on, are only 10 dB away from the output frequency, 2*f* if the output is taken from C_7. It is, therefore, advisable to use a bandpass filter at the output if the circuit works permanently with a fixed input frequency.



The optional bandpass filter consists of two inductively coupled tuned circuits, L_1-C_5 and L_2-C_6. If losses at higher harmonics are acceptable, these circuits may be tuned for use with such harmonics.

The NE612 can be configured as an overtone oscillator (**Figure 2**). The internal oscillator is normally not accessible and mixes or multiplies the input signal with the oscillator signal, so yielding an output $f_{in}+f_{lo}$. If, however, the r.f. input of the multiplier, pin 1, is linked to pin 8 via resistor R_1, the mixer produces a high-level output at the oscillator frequency.





The maximum output level is a function of the value of R_1 and the supply voltage. It has been found by trial and error that a resistor value of 560 Ω is optimal. The desired output level is optimal at a supply voltage of +5 V.

Although the first harmonic (72 MHz) is only 10 dB away from the fundamental (36 MHz), the higher harmonics are more than 25 dB lower. Greater distances may be obtained by the use of a bandpass filter at the output.

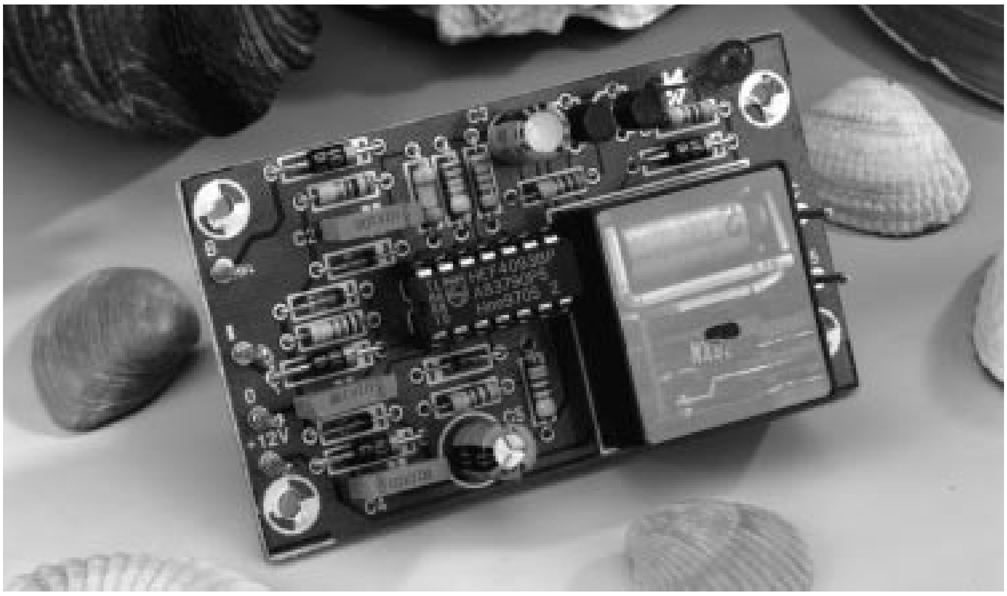
If a fundamental-frequency crystal is used, circuit L_1-C_3, which is tuned to the first harmonics, as well as coupling capacitor C_4, may be omitted.

The two applications just discussed may be combined into a third: an overtone oscillator with frequency doubler. In this, the mixer input (pin 1) is linked to the emitter of the oscillator transistor (pin 7) via resistor R_1. In this application, a value of 10 kΩ for this resistor proved optimal.

It should be noted that the output voltage possible with an optimal output range is lower than in the previous application: about 50 mV peak-to-peak. On the other hand, the harmonics in the output range are 20 dB away from the output frequency of 72 MHz. This means that in most cases a bandpass filter at the output will not be required.

[984119]

017 car immobilizer



Design M. Lawton

A starter motor immobilizer is an effective (but not certain) means of protecting your car against theft. It has the drawback that a would-be thief will try to render it inoperative and in the process damages your car. The present circuit is a simple version of car immobilizer and tends to confuse the thief. This is because the car appears to function normally, but it does not start. Has it broken down or is there some sort of protection circuit active?

The circuit does not need additional controls, indicators, switches or keypads to be fitted in the car. The setup is 'invisible'. The only external sensor is

Parts list

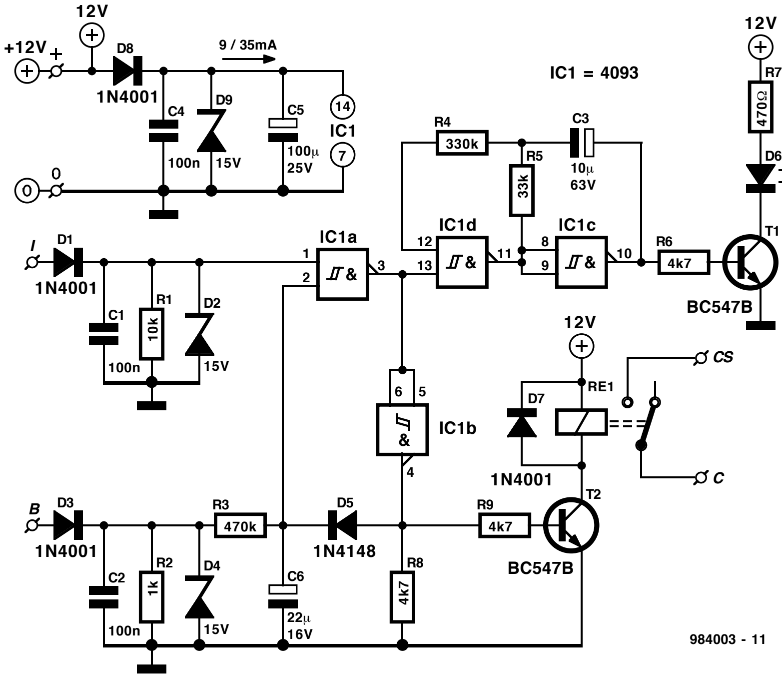
Resistors:
R_1 = 10 kΩ
R_2 = 1 kΩ
R_3 = 470 kΩ
R_4 = 330 kΩ
R_5 = 33 kΩ
R_6, R_8, R_9 = 4.7 kΩ
R_7 = 470 Ω

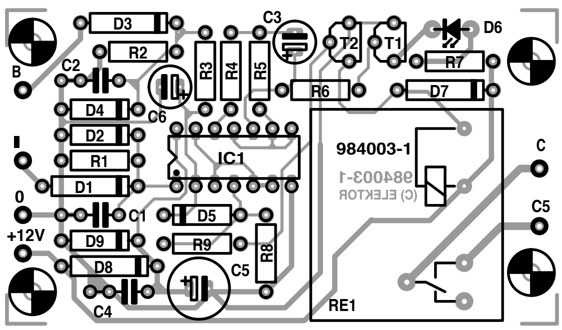
Capacitors:
C_1, C_2, C_4 = 0.1 μF
C_3 = 10 μF, 63 V, radial
C_5 = 100 μF, 25 V
C_6 = 22 μF, 16 V, radial

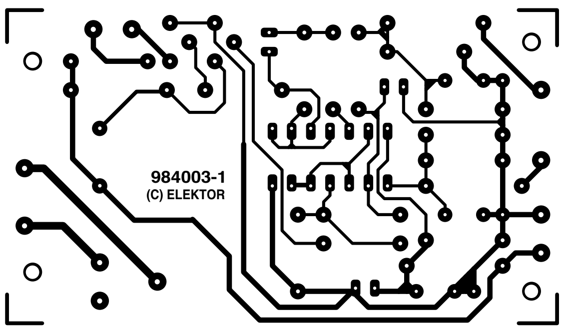
Semiconductors:
D_1, D_3, D_7, D_8 = 1N4001
D_2, D_4, D_9 = zener diode, 15 V, 400 mW
D_5 = 1N4148
D_6 = LED
T_1, T_2 = BC547B

Integrated circuits:
IC_1 = 4093

Miscellaneous:
PC_1–PC_6 = PCB terminal (pin)
Re_1 = 12 V car-type relay, 1 change-over contact







the brake pedal. After the ignition has been switched on, the brake pedal has to be pressed for at least five seconds before voltage is applied to the coil. Since the thief does not know this, he/she will try everything to get the car started. Since it is only the coil to which voltage is not applied, all other electrical functions will work normally, but it is just impossible to get the car started.

When the ignition is switched on, the circuit is powered by the voltage at terminal PC2. Until the brake pedal is pressed, the potential at terminal PC1 remains low, so that the relay remains unenergized. When the brake pedal is pressed, capacitor C6 is charged via resistor R3. The time, t, it takes for the capacitor to become fully charged is determined by network R8–C6. When this time has elapsed, the output of IC1a goes low, whereupon voltage is applied to the base of T2 via IC1b. When T2 is on, the relay is energized , whereupon its contact changes over and voltage is applied to the coil. After the pressure on the brake pedal is released, diode D5 ensures that the voltage remains applied.

Gates IC1c and IC1d form an oscillator, which causes diode D6 to flash when the starter is immobilized. This has the disadvantage, of course, that it discloses the protection circuit.

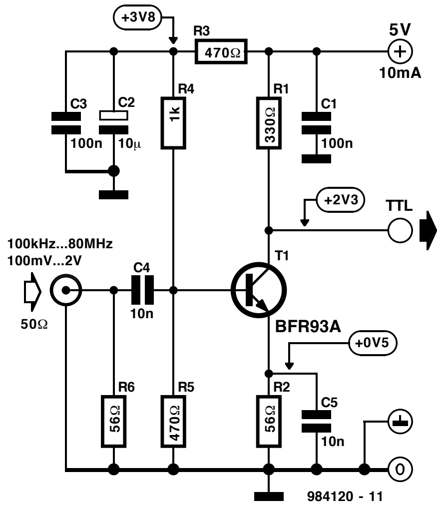
Finding the right points to which to connect the circuit should not be a problem in most cars. The ignition voltage is normally available at the radio/cassette terminals, while the potential coupled to the brake pedal is usually available at the brake lights. [984003]

018 sine wave to TTL converter

Design: G. Kleine

As the title implies, the present circuit is intended to convert sinusoidal input signals to TTL output signals. It can handle inputs of more than 100 mV and is suitable for use at frequencies up to about 80 MHz.

Transistor T_1, configured in a common-emitter circuit, is biased by voltage divider R_3–R_5 such that the potential across output resistor R_1 is about half the supply voltage. When the circuit is driven by a signal whose amplitude is between 100 mV and TTL level (about 2 V r.m.s.), the circuit generates rectangular signals. The lowest frequencies that could be processed by the prototype were around 100 kHz at an input level of 100 mV, and about 10 kHz when the input signals were TTL level.



Resistor R_6 holds the input resistance at about 50 Ω, which is the normal value in measurement techniques. It ensures that the effects of long coaxial cables on the signal are negligible.

If the converter is used in a circuit with ample limits, R_6 may be omitted, whereupon the input resistance rises to 300 Ω. [984120]

019 input impedance booster

Design: H. Bonekamp

The input impedance of a.c.-coupled op amp circuits depends almost entirely on the resistance that sets the d.c. operating point. If CMOS op amps are used, the input is high, in current op amps up to 10 MΩ.

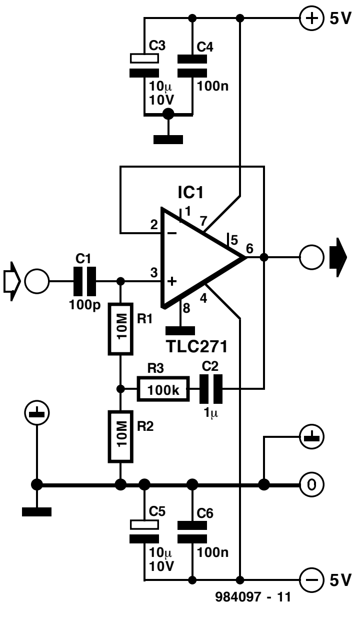
If a higher value is needed, a bootstrap may be used, which enables the input impedance to be boosted artificially to a very high value.

In the diagram, resistors R_1 plus R_2 form the resistance that sets the d.c. operating point for op amp IC_1. If no other actions were taken, the input impedance would be about 20 MΩ. However, part of the input signal is fed back in phase, so that the alternating current through R_1 is smaller. The input impedance, Z_{in}, is then:

$$Z_{in}=(R_2+R_3)/R_3)(R_1+R_2).$$

With component values as specified, Z_{in} has a value of about 1 GΩ.

The circuit draws a current of about 3 mA. [984097]



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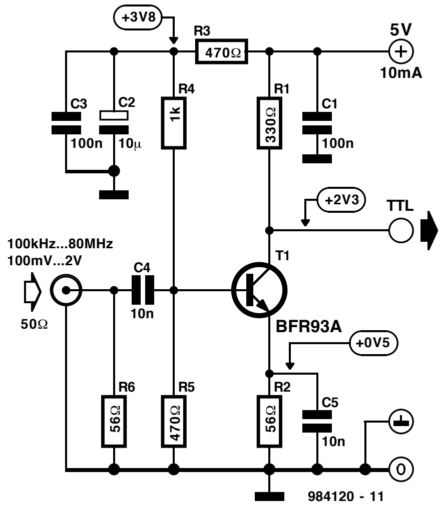

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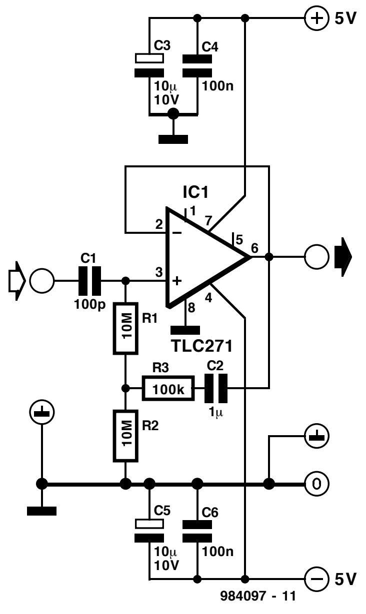

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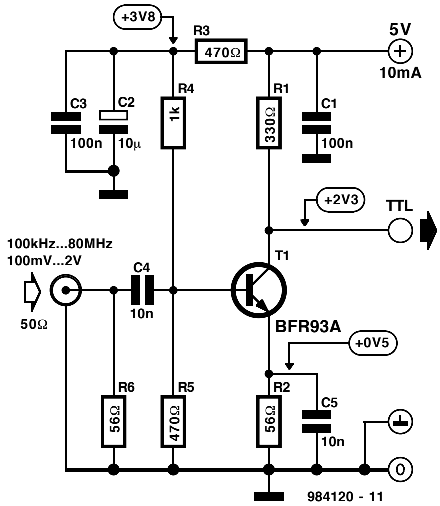

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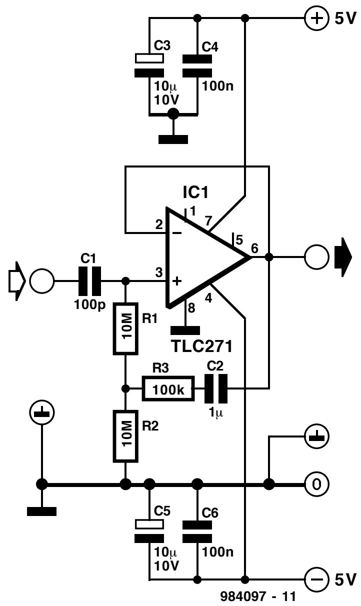


020 1-watt BTL audio amplifier

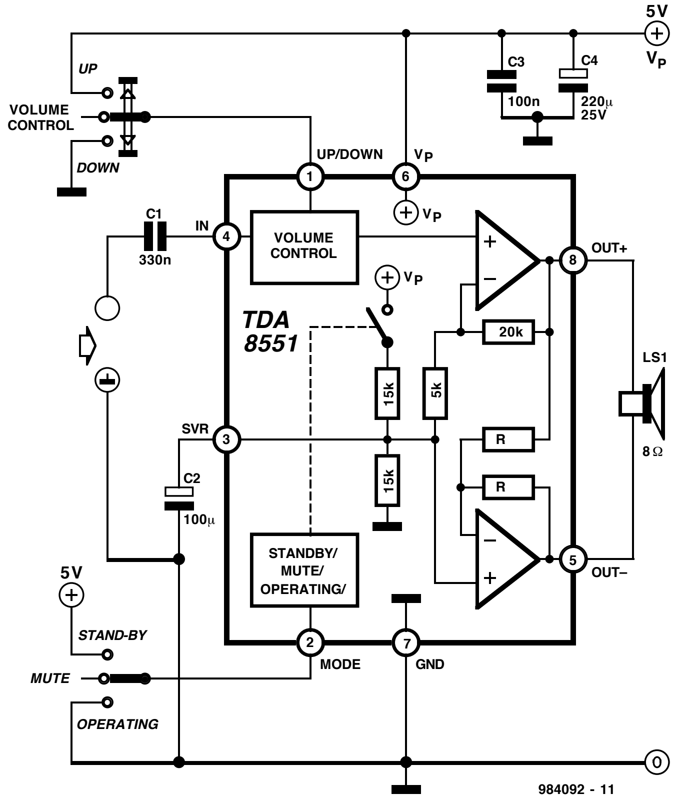
Source: Philips Semiconductors Preliminary Specification

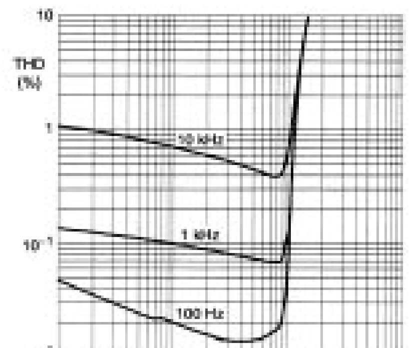
The TDA8581(T) from Philips Semiconductors is a 1-watt Bridge Tied Load (BTL) audio power amplifier capable of delivering 1 watt output power into an 8-Ω load at THD (total harmonic distortion) of 10% and using a 5-V power supply. The schematic shown here combines the functional diagram of the TDA8551 with its typical application circuit. The gain of the amplifier can be set by the digital volume control input. At the highest volume setting, the gain is 20 dB. Using the MODE pin the device can be switched to one of three modes: standby (MODE level between V_p and V_p–0.5 V), muted (MODE level between 1 V and V_p–1.4 V) or normal (MODE level less than 0.5 V). The TDA8551 is protected by an internal thermal shutdown protection mechanism.

The total voltage loss for both MOS transistors in the complementary output stage is less than 1 V. Using a 5-V supply and an 8-Ω loudspeaker, an output power of 1 watt can be delivered.

The volume control has an attenuation range of between 0 dB and 80 dB in 64 steps set by the 3-state level at the UP/DOWN pin: *floating*: volume remains unchanged; *negative pulses*: decrease volume; *positive pulses*: increase volume Each pulse at he Up/DOWN pin causes a change in gain of 80/64 = 1.25 dB (typical value). When the supply voltage is first connected, the attenuator is set to 40 dB (low volume), so the gain of the total amplifier is then –20 dB. Some positive pulses have to be applied to the UP/DOWN pin to achieve listening volume. The graph shows the THD as a function of output power. The maximum quiescent current consumption of the amplifier is specified at 10 mA, to which should be added the current resulting from the output offset voltage divided by the load impedance.

(984092-1, Gb)





021 simple electrification unit

From an idea by P. Lay

The circuit is intended for carrying out harmless experiments with high-voltage pulses and functions in a similar way as an electrified fence generator. The p.r.f. (pulse repetition frequency) is determined by the time constant of network R_1-C_3 in the feedback loop of op amp IC_{1a}: with values as specified, it is about 0.5 Hz.

The stage following the op amp, IC_{1b}, converts the rectangular signal into narrow pulses. Differentiating network R_2-C_4, in conjunction with the switching threshold of the Schmitt trigger inputs of IC_{1b}, determines the pulse period, which here is about 1.5 ms.

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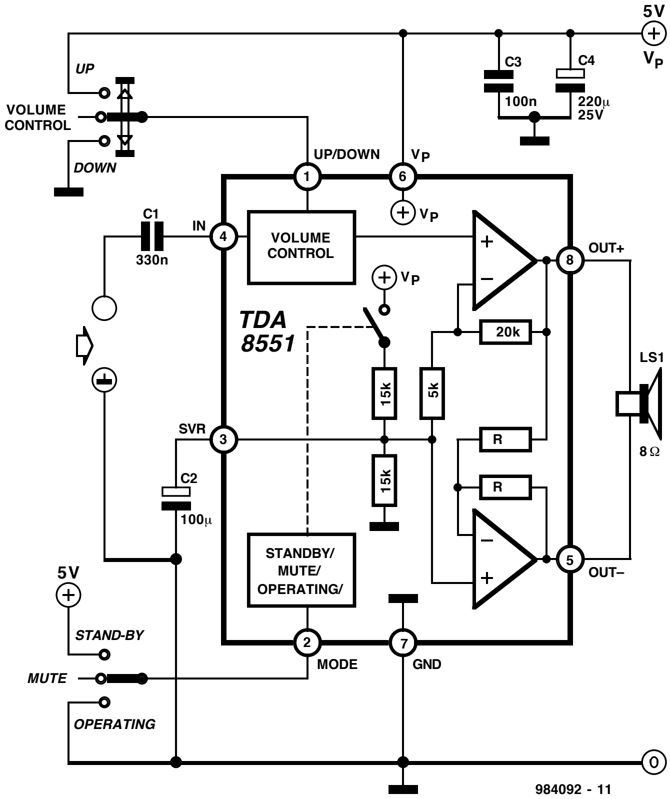

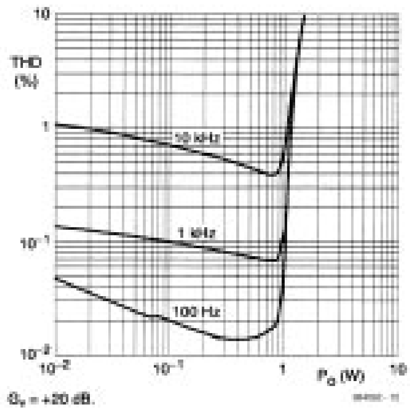

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The output of IC_{1b} is linked directly to the gate of thyristor THR_1, so that this device is triggered by the pulses.

The requisite high voltage is generated with the aid of a small mains transformer, whose secondary winding is here used as the primary. This winding, in conjunction with C_2, forms a resonant circuit.

Capacitor C_3 is charged to the supply voltage (12 V) via R_3. When a pulse output by IC_{1b} triggers the thyristor, the capacitor is discharged via the secondary winding. The energy stored in the capacitor is, however, not lost, but is stored in the magnetic field produced by the transformer when current flows through it.

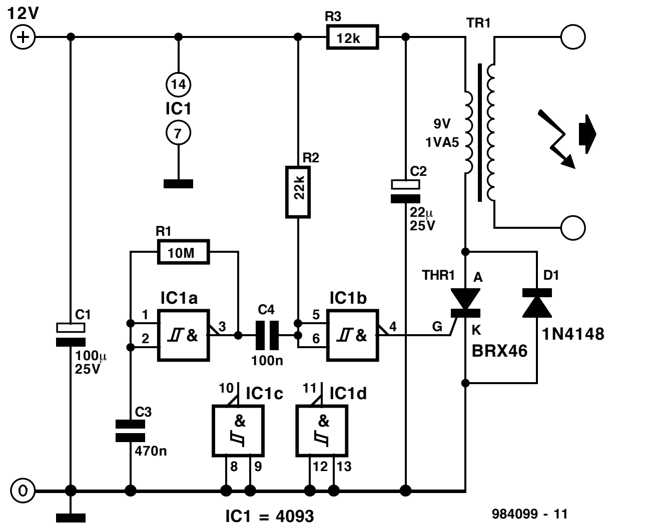
When the capacitor is discharged, the current ceases, whereupon the magnetic field collapses. This induces a counter e.m.f. in the transformer winding which opposes the voltage earlier applied to the transformer. This means that the direction of the current remains the same. However, capacitor C_2 is now charged in the opposite sense, so that the potential across it is negative.

When the magnetic field of the transformer has returned the stored energy to the capacitor, the direction of the current reverses, and the negatively charged capacitor is discharged via D_1 and the secondary winding of the transformer. As soon as the capacitor begins to be discharged, there is no current through the thyristor, which therefore switches off. When C_2 is discharged further, diode D_1 is reverse-biased, so that the current loop to the transformer is broken, whereupon the capacitor is charged to 12 V again via R_3.

At the next pulse from IC_{1b}, this process repeats itself.

Since the transformer after each discharge of the capacitor at its primary induces not only a primary, but also a secondary voltage, each triggering of the thyristor causes two closely spaced voltage pulses of opposite polarity. These induced voltages at the secondary, that is, the 230 V, winding, of the transformer are, owing to the higher turns ratio, much higher than those at the primary side and may reach several hundred volts. However, since the energy stored in capacitor C_2 is relatively small (the current drain is only about 2 mA), the output voltage cannot harm man or animal. It is sufficient, however, to cause a clearly discernible muscle convulsion.

[984099]



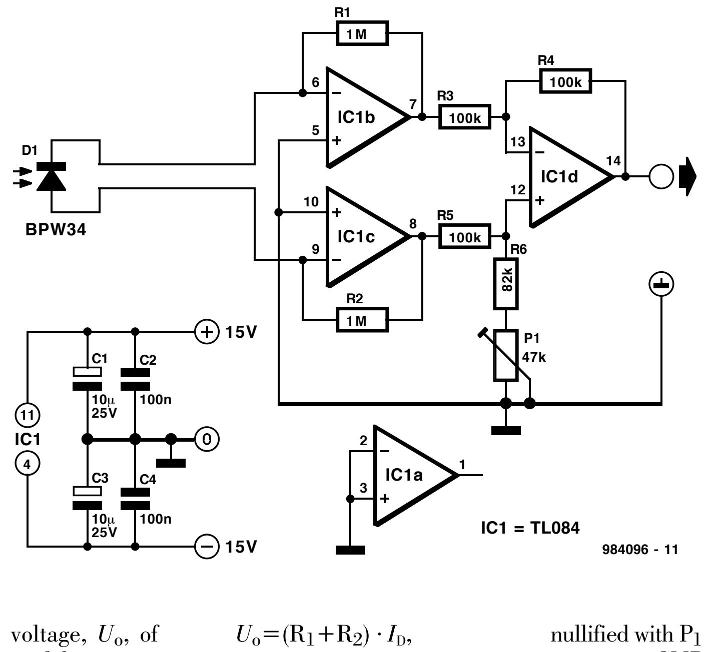
022 balanced amplifier for photo-diode

Design: H. Bonekamp

A photo-diode is a p-n diode whose reverse current depends on the amount of light falling on its junction. The reverse current is greatly dependent on the temperature since heat can liberate more covalent bonds. As light can also do this, the diode can be housed in a transparent case.

When a photo-diode is located at some distance from the associated electronic circuits, noise may be picked up in the connecting cable, even when this is screened. Such noise can, fortunately, be suppressed easily, provided it is common mode, that is, when the diode is not connected to earth ('floats').

A differential amplifier enables a feedback signal to be amplified, but does not respond to common-mode signals. In the diagram, the differential amplifier consists of two op amps, IC_{1b} and IC_{1c}, which convert the diode current into a voltage. The current-to-voltage conversion depends on R_1 and R_2, so that gain setting in amplifier IC_{1d} is not necessary.



The output voltage, U_o, of the differential amplifier is

$$U_o = (U_{in1} - U_{in2}) \cdot R_4/R_3.$$

When $R_3 = R_5 = R_4 = R_6 + P_1$, the amplification is unity. In that case,

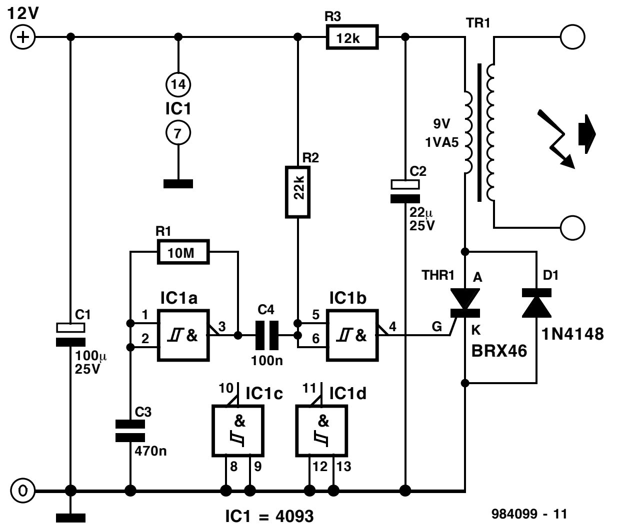
$$U_o = (R_1 + R_2) \cdot I_D,$$

where I_D is the diode current.

The Common Mode Rejection, CMR, depends on the equality of the resistors as stipulated earlier. Their tolerances, and those of R_1 and R_2, can be nullified with P_1 so as to achieve optimum CMR. A Common Mode Rejection Ratio, CMRR, of >60 dB is obtained when the specified op amps are linked to the photo-diode by a twisted pair.

The circuit draws a current of about 10 mA.

[984096]



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Capacitor C_3 is charged to the supply voltage (12 V) via R_3. When a pulse output by IC_{1b} triggers the thyristor, the capacitor is discharged via the secondary winding. The energy stored in the capacitor is, however, not lost, but is stored in the magnetic field produced by the transformer when current flows through it.

When the capacitor is discharged, the current ceases, whereupon the magnetic field collapses. This induces a counter e.m.f. in the transformer winding which opposes the voltage earlier applied to the transformer. This means that the direction of the current remains the same. However, capacitor C_2 is now charged in the opposite sense, so that the potential across it is negative.

When the magnetic field of the transformer has returned the stored energy to the capacitor, the direction of the current reverses, and the negatively charged capacitor is discharged via D_1 and the secondary winding of the transformer. As soon as the capacitor begins to be discharged, there is no current through the thyristor, which therefore switches off. When C_2 is discharged further, diode D_1 is reverse-biased, so that the current loop to the transformer is broken, whereupon the capacitor is charged to 12 V again via R_3.

At the next pulse from IC_{1b}, this process repeats itself.

Since the transformer after each discharge of the capacitor at its primary induces not only a primary, but also a secondary voltage, each triggering of the thyristor causes two closely spaced voltage pulses of opposite polarity. These induced voltages at the secondary, that is, the 230 V, winding, of the transformer are, owing to the higher turns ratio, much higher than those at the primary side and may reach several hundred volts. However, since the energy stored in capacitor C_2 is relatively small (the current drain is only about 2 mA), the output voltage cannot harm man or animal. It is sufficient, however, to cause a clearly discernible muscle convulsion.

[984099]

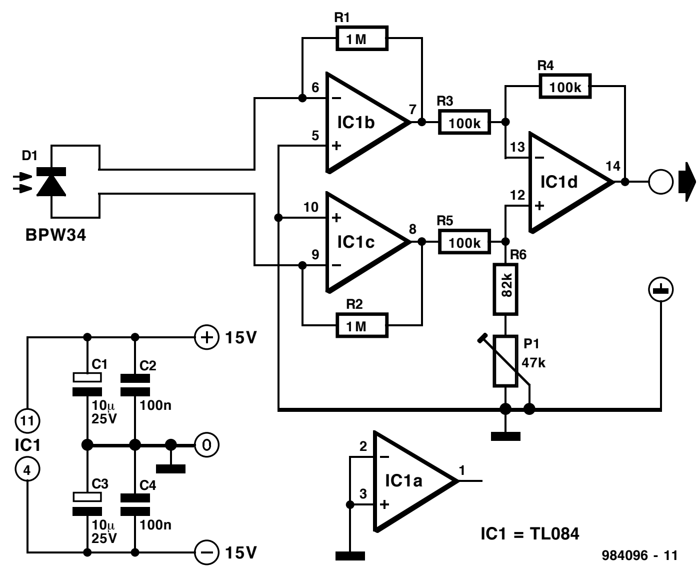
022 balanced amplifier for photo-diode

Design: H. Bonekamp

A photo-diode is a p-n diode whose reverse current depends on the amount of light falling on its junction. The reverse current is greatly dependent on the temperature since heat can liberate more covalent bonds. As light can also do this, the diode can be housed in a transparent case.

When a photo-diode is located at some distance from the associated electronic circuits, noise may be picked up in the connecting cable, even when this is screened. Such noise can, fortunately, be suppressed easily, provided it is common mode, that is, when the diode is not connected to earth ('floats').

A differential amplifier enables a feedback signal to be amplified, but does not respond to common-mode signals. In the diagram, the differential amplifier consists of two op amps, IC_{1b} and IC_{1c}, which convert the diode current into a voltage. The current-to-voltage conversion depends on R_1 and R_2, so that gain setting in amplifier IC_{1d} is not necessary.



The output voltage, U_o, of the differential amplifier is

$$U_o=(U_{in1}-U_{in2})\cdot R_4/R_3.$$

When $R_3=R_5=R_4=R_6+P_1$, the amplification is unity. In that case,

$$U_o=(R_1+R_2)\cdot I_D,$$

where I_D is the diode current.

The Common Mode Rejection, CMR, depends on the equality of the resistors as stipulated earlier. Their tolerances, and those of R_1 and R_2, can be nullified with P_1 so as to achieve optimum CMR. A Common Mode Rejection Ratio, CMRR, of >60 dB is obtained when the specified op amps are linked to the photo-diode by a twisted pair.

The circuit draws a current of about 10 mA.

[984096]

023

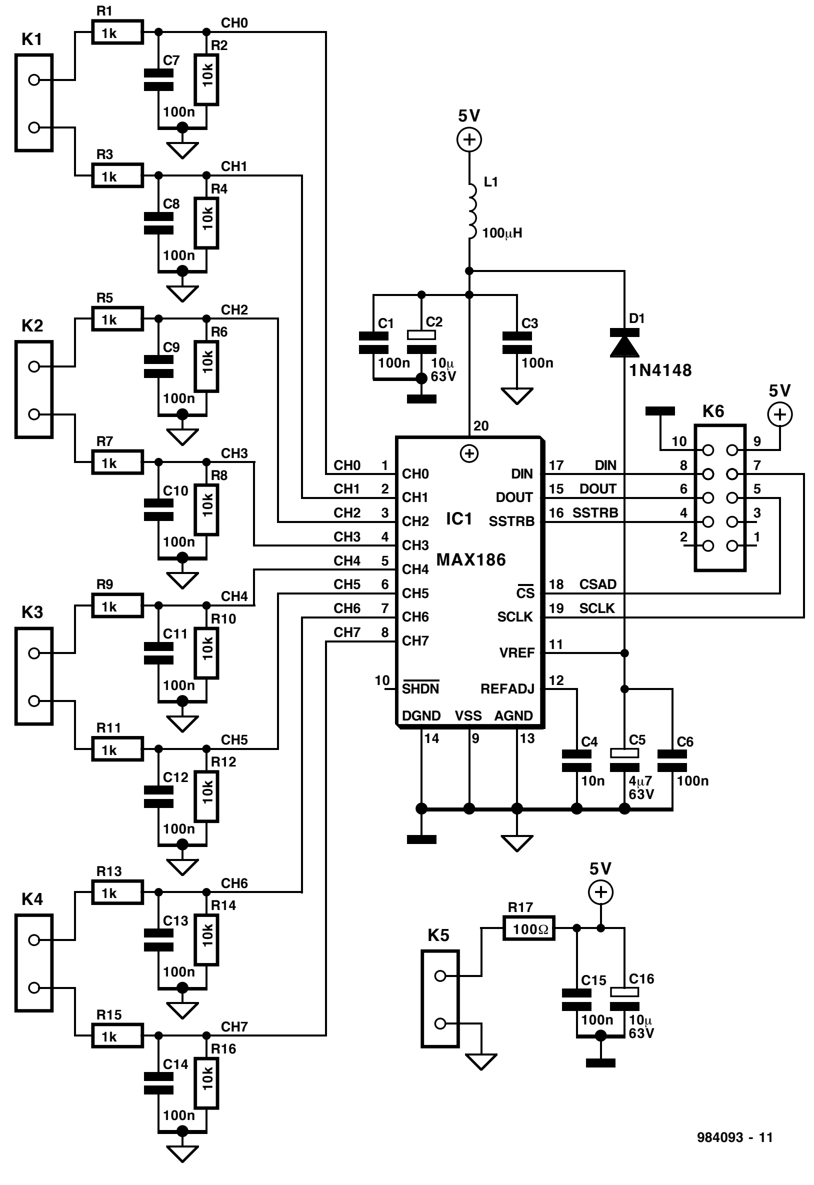
A-D converter for MatchBox BASIC computer

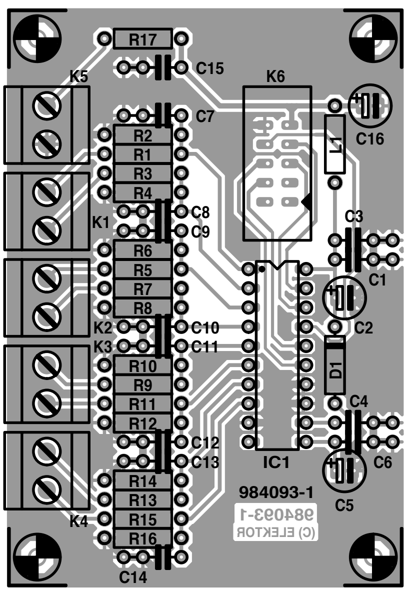
Design: K. Walraven

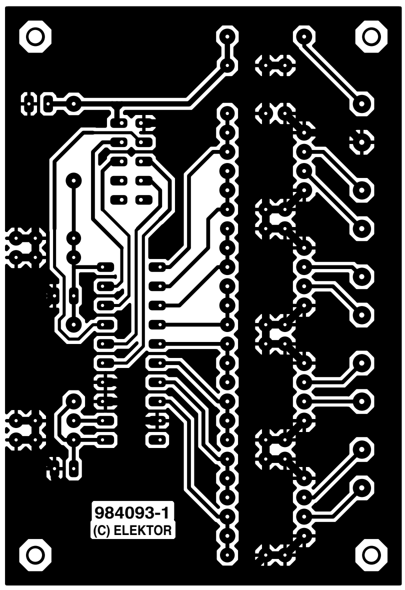
In their book *MatchBox BASIC Computer* [1] the authors describe a way of connecting a 12-bit analogue-to-digital converter (ADC) Type MAX187 to the small computer board originally described in *Elektor Electronics* magazine.

For the present article the MAX186 is employed, which is a similar converter with eight analogue inputs instead of just one. The connection with the computer board is made via a length of 10-way flatcable. Although K_4 would appear to be the right connector for this link, K_1 was eventually chosen because bit operations are not possible on port P_2. A disadvantage of using K_1 is, however, that the 1-way cable has to be connected to a 20-way pinheader. Note that the converter may, in principle, be connected to any port as long as the supply voltage is at the right pins.

The inputs of the present circuit are fitted with overvoltage protection resistors (R_1, R_3, etc.) as well as pull-up resistors (R_2, R_4 etc.). Consequently, inputs which are left 'open' are still held at a defined level, while additional ESD (electrostatic discharge) protection is







COMPONENTS LIST

Resistors:
R_1, R_3, R_5, R_7, R_9, R_{11}, R_{13}, R_{15} = 1kΩ
R_2, R_4, R_6, R_8, R_{10}, R_{12}, R_{14}, R_{16} = 10kΩ
R_{17} = 100 Ω

Capacitors:
C_1, C_3, C_6–C_{15} = 0.1 µF
C_2, C_{16} = 10 µF, 63 V, radial
C_4 = 0.01 µF
C_5 = 4 µF, 63 V, radial

Inductor:
L_1 = 100 µH

Semiconductors:
D_1 = 1N4148

Integrated circuits:
IC_1 = MAX186DCPP or MAX186BEPP

Miscellaneous:
K_1–K_5 = 2-way PCB terminal block
K_6 = 10-way box header

```
READ_AD:

; This subroutine reads the MAX186 12-bit A-D converter.
; Before calling this routine the code for the desired
; channel has to loaded into integer variabele TEMP,
; as follows:
; TEMP:=1XYZ1110B where XYZ indicates the desired channel.
; The conversion result is then returned in TEMP.
;
; MAX187 connections:
; Data to 187     P1.0
; Serial clock    P1.1
; Data from 187   P1.2
; Chip select     P1.3
; Strobe          P1.4
;

INIT_AD:
  P1.0:=0                  ;data in
  P1.1:=0                  ;clock
  P1.3:=0                  ;CS active

WRITE_AD:
  CNTR:=8
  WHILE CNTR>0 DO          ;send 8 bits of A-D command
    P1.0:=TEMP.7           ;msb bit to A-D
    TEMP:=TEMP SHL(1)      ;hold next bit ready
    P1.1:=1                ;clock-in data on pos. edge
    P1.1:=0
    CNTR:=CNTR-1
  WHEND

READ_AD:
  TEMP:=0                  ;store result in this variable
  CNTR:=12
  P1.0:=0                  ;read zeroes (else conversion starts)
  WHILE CNTR>0 DO          ;fetch 12 bit data
    P1.1:=1                ;supply clock pulse
    P1.1:=0                ;data valid after neg. edge
    TEMP:=(TEMP SHL 1)+P1.2 ;read data
    CNTR:=CNTR-1
  WHEND
  P1.3:=1                  ;CS, turn off A-D
RETURN
```

provided. Note that the resistors cause a certain amount of attenuation if an input voltage is applied. The resistor values have been selected such that an external temperature sensor with an output of 1 μA °C^{-1} (for instance, AD590 or LM334) provides the desired voltage gradient of 10 mV °C^{-1}.

An example of a control program for the converter is listed here. The converter receives eight databits; bit 7 is the startbit, while bits 4, 5 and 6 indicate which input is being selected. Bit 3 is used to signal that the measurement is to take place between ground and V_{REF}, while bit 2 tells the converter to perform a single-ended (i.e., non-differential) measurement. Bits 1 and 0, finally, initiate an A-D conversion based on the internal clock. Next, the 12-bit result can be read back.

A final remark: the channel selection bits are mixed up: bit 6 is the LSB, bit 5, the MSB, and bit 4, the middle bit.

The circuit draws a current not greater than 2 mA. The printed circuit board shown here is not available ready-made. For more information on the MAX186, visit Maxim's Internet site at www.maxim-ic.com.

Reference:
[1]. *MatchBox BASIC Computer*, by K.H. Dietsche and M. Ohsmann, Elektor Electronics (Publishing), ISBN 0-905705-53-X.

[984093]

024 ten-band equalizer

Design: P. Staugaard

The equalizer presented in this article is suitable for use with hi-fi installations, public-address systems. mixers and electronic musical instruments.

The relay contacts at the inputs and outputs, in conjunction with S_2, enable the desired channel to be selected. The input may be linked directly to the output, if wanted. The input impedance and amplification of the equalizer are set with S_1 and S_3. The audio frequency spectrum of 31 Hz to 16 kHz is divided into ten bands.

Ten bands require ten filters, of which nine are passive and one active. The passive filters are identical in design and differ only in the value of the relevant inductors and capacitors. The requisite characteristics of the filters are achieved by series and parallel networks. The filter for the lowest frequency band is an active one to avoid a very large value of inductance. It is based in a traditional manner on op amp A_1.

The inductors used in the passive filters are readily available small chokes. The filter based on L_1 and L_2 operates at about the lowest frequency (62 Hz) that can be achieved with standard, passive components.

The Q(uality) factor of the filters can, in principle, be raised slightly by increasing the value of R_{19} and R_{23}, as well as that of P_1–P_{10}, but that would be at the expense of the noise level of op amp IC_1.

With component values as specified, the control range is about ±11 dB, which in most case will be fine. A much larger range is not attainable without major redesign.

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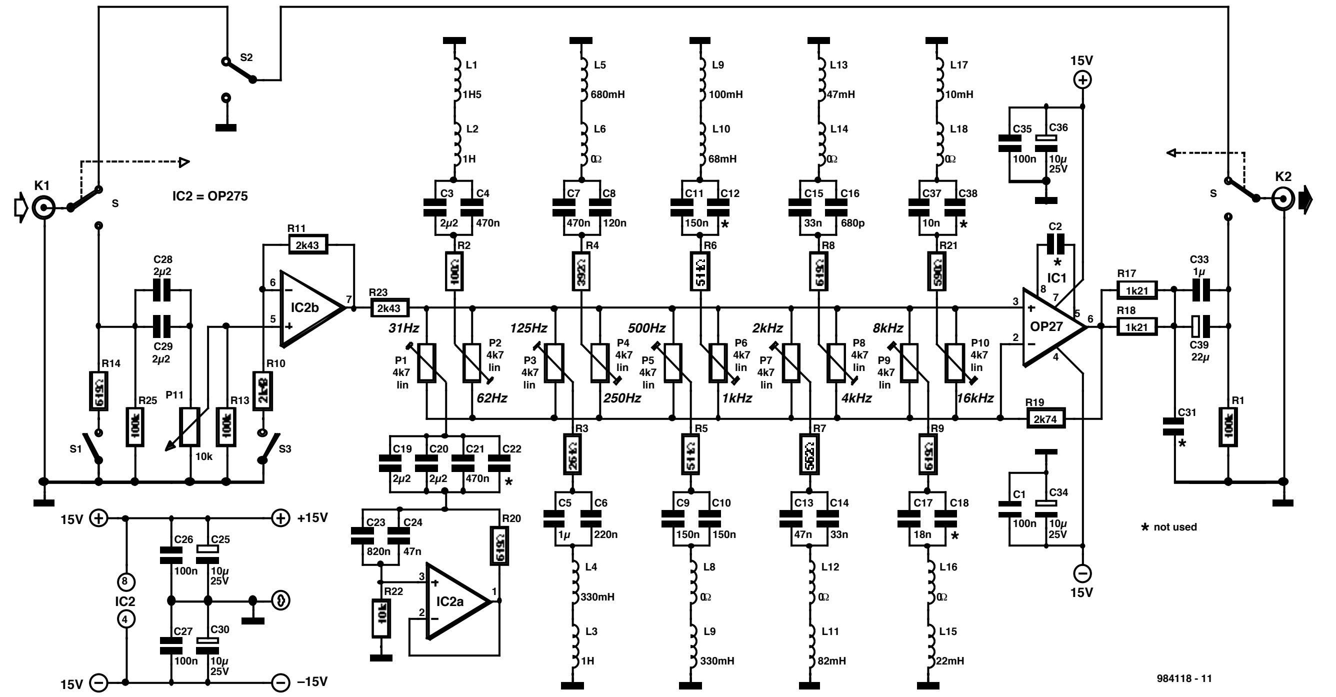


The input level can be adjusted with P_1, which may be necessary for adjusting the balance between the channels or when a loudness control is used in the output amplifiers.

Several types of op amp can be used: in the prototype, IC_1 is an LT1007, and IC_2, an OP275. Other suitable types for IC_1 are OP27 or NE5534; and for IC_2, AD712, LM833 and NE5532. If an NE5534 is used for IC_1, C_2 is needed; in all other cases, not.

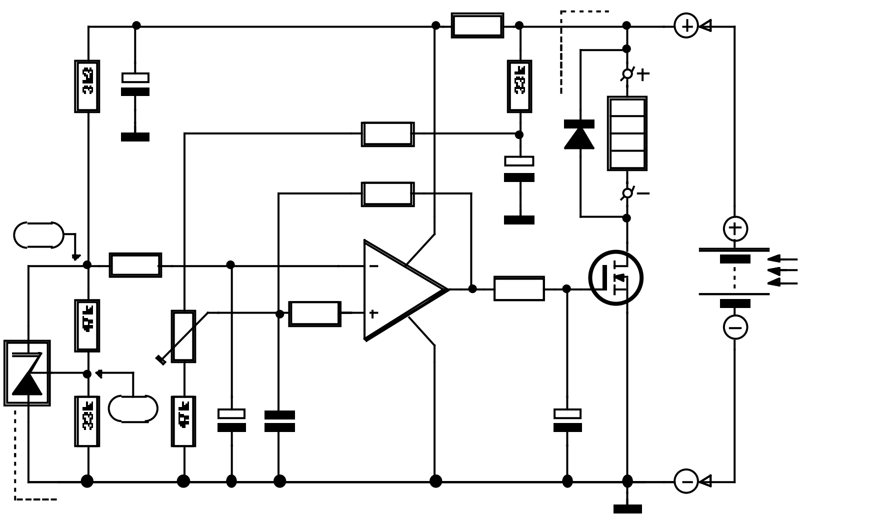
The circuit needs to be powered by a regulated, symmetrical 15 V supply. It draws a current of not more than about 10mA.

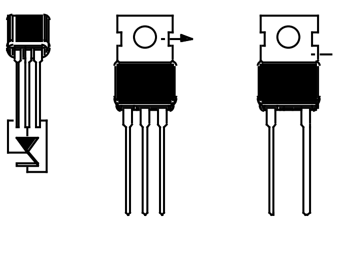
[984118]

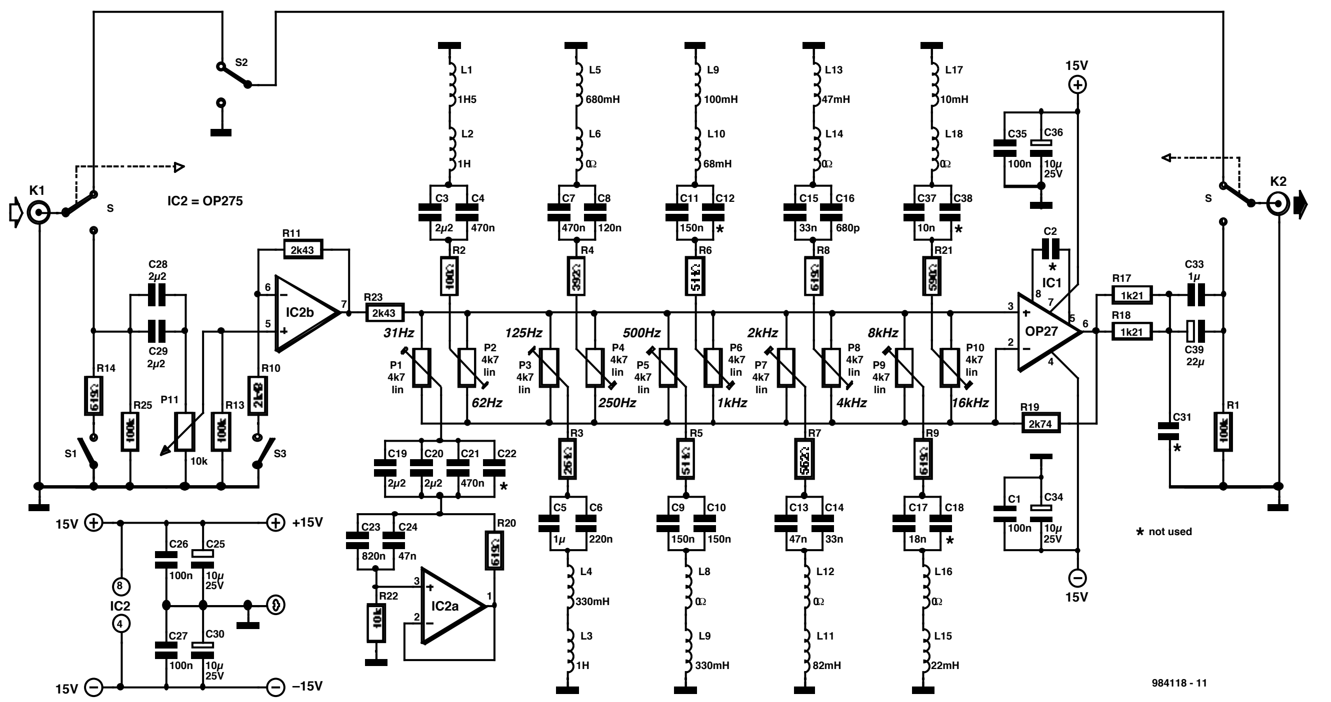
025 lead-acid-battery regulator for solar panel systems

The design of solar panel systems with a (lead-acid) buffer battery is normally such that the battery is charged even when there is not much sunshine. This means, however, that when there is plenty of sunshine, a regulator is needed to prevent the battery from being overcharged. Such controls usually arrange for the superfluous energy to be dissipated in a shunt resistance or simply for the solar panels to be short-circuited. It is, of course, an unsatisfactory situation when the energy derived from a very expensive system can, after all, not be used to the full.

The circuit presented diverts the energy from the solar panel when the battery is fully charged to another user, for instance, a 12 V ice box with Peltier elements, a pump for drawing water from a rain butt, or a 12 V ventilator. It is, of course, also possible to arrange for a second battery to be charged by the superfluous energy. In this case,







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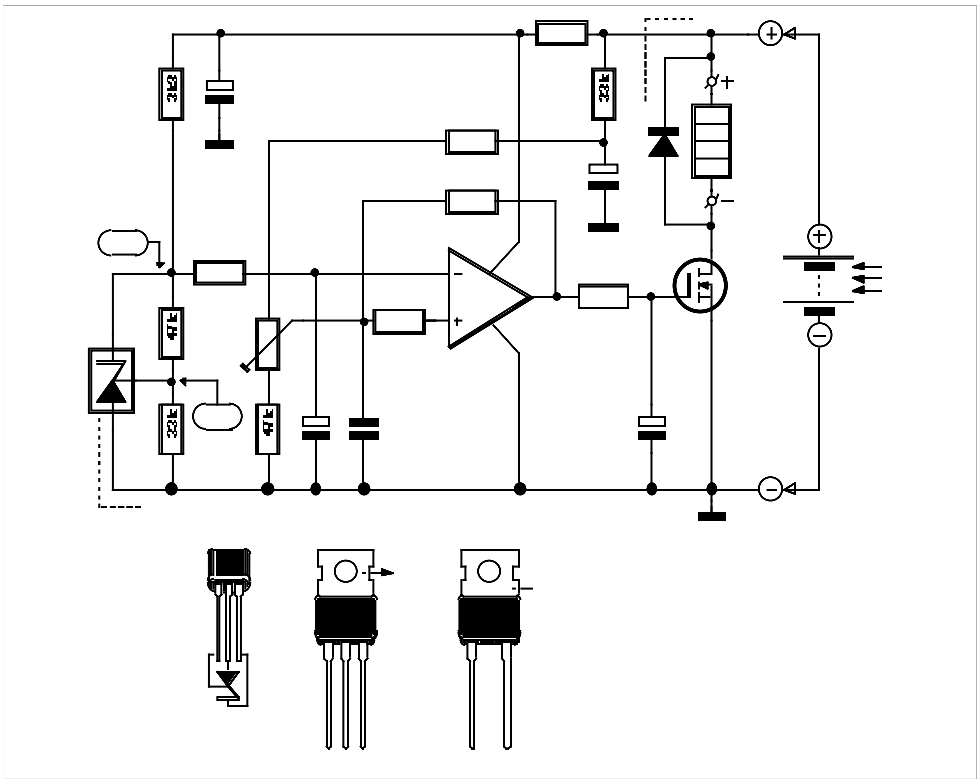


however, care must be taken to ensure that when the second battery is also fully charged, there is also a control to divert the superfluous energy.

The shunt resistance needed to dissipate the superfluous energy must be capable of absorbing the total power of the panel, that is, in case of a 100 W panel, its rating must be also 100 W. This means a current of some 6–8 A when the operating voltage is 12 V. When the voltage drops below the maximum charging voltage of 14.4 V owing to reduced sunshine, the shunt resistance is disconnected by an n-channel power field effect transistor (FET), T_1. The disconnect point is not affected by large temperature fluctuations because of a reference voltage provided by IC_1. The necessary comparator is IC_2, which owing to R_9 has a small hysteresis voltage of 0.5 V. Capacitor C_5 ensures a relatively slow switching process, although the FET is already reacting slowly owing to C_4. The gradual switching prevents spurious radiation caused by steep edges of the switched voltage and also limits the starting current of a motor (of a possible ventilator). Finally, it prevents switching losses in the FET that might reach 25 W, which would make a heat sink unavoidable.

Setting up of the circuit is fairly simple. Start by turning P_1 so that its wiper is connected to R_5. When the battery reaches the voltage at which it will be switched off, that is, 13.8–14.4 V, adjust P_1 slowly until the output of comparator IC_2 changes from low to high, which causes the load across T_1 to be switched in.

Potentiometer P_1 is best a 10-turn model. When the control is switched on for the first time, it takes about 2 seconds for the electrolytic capacitors to be charged. During this time, the output of the comparator is high, so that the load across T_1 is briefly switched in.

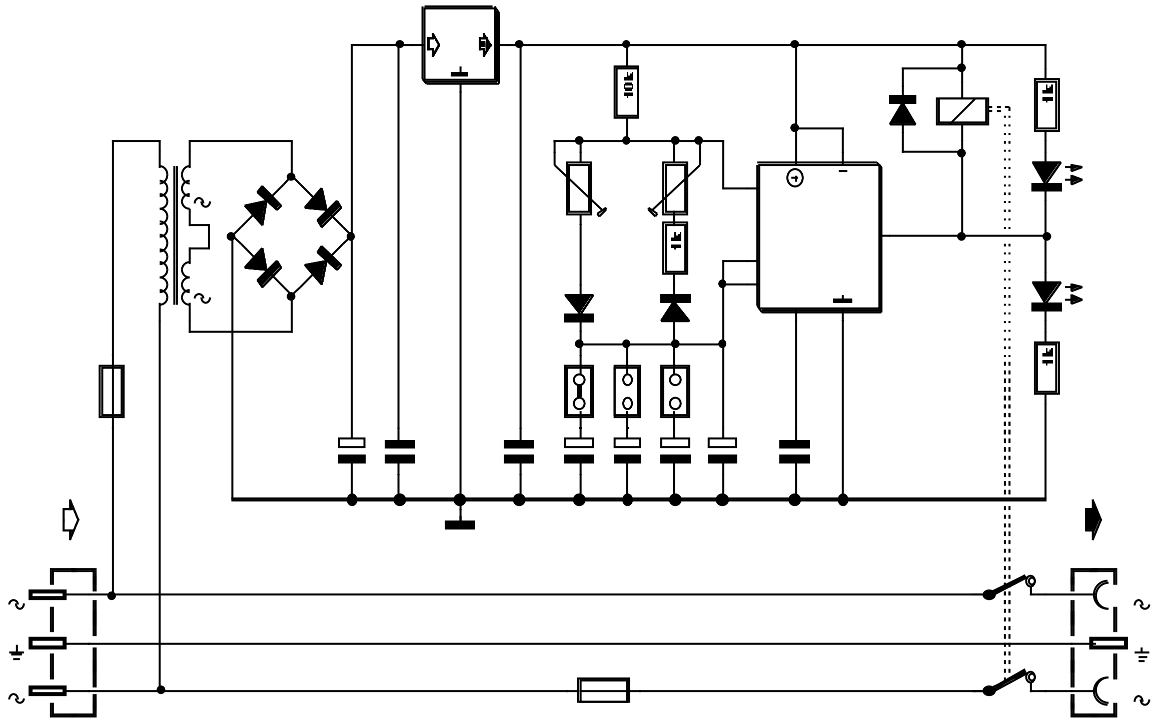
In case T_1 has to switch in low-resistance loads, the BUZ11 may be replaced by an IRF44, which can handle twice as much power (150 W) and has an on-resistance of only 24 mΩ.

Because of the very high currents if the battery were short-circuited, it is advisable to insert a suitable fuse in the line to the regulator.

The circuit draws a current of only 2 mA in the quiescent state and not more than 10 mA when T_1 is on.

[Zeiller – 984072]

026 mains pulser



Design: G. Kleine

The pulser is intended to switch the mains voltage on and off at intervals between just under a second and up to 10 minutes. This is useful, for instance, when a mains-operated equipment is to be tested for long periods, or for periodic switching of machinery.

Transformer Tr_1, the bridge rectifier, and regulator IC_1 provide a stable 12 V supply rail for IC_2 and the relay. The timer is arranged so that the period-determining capacitor can be charged and discharged independently. Four time ranges can be selected by selecting capacitors with the aid of jumpers. Short-circuiting positions 1 and 2 gives the longest time, and short-circuiting none the shortest. In the latter case, the 10 μF capacitor at pins 2 and 6 of the timer IC determines the time with the relevant resistors. The value of this capacitor may be chosen slightly lower.

The two preset potentiometers enable the on and off periods to be set. The 1 kΩ resistor in series with one of the presets determines the minimum discharge time.

The timer IC switches a relay whose double-pole contacts switch the mains voltage.

The LEDs indicate whether the mains voltage is switched through (red) or not (green).

The 100 mA slow fuse protects the mains transformer and low-voltage circuit. The 4 A medium slow fuse protects the relay against overload.

[984122]

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[Zeiller – 984072]

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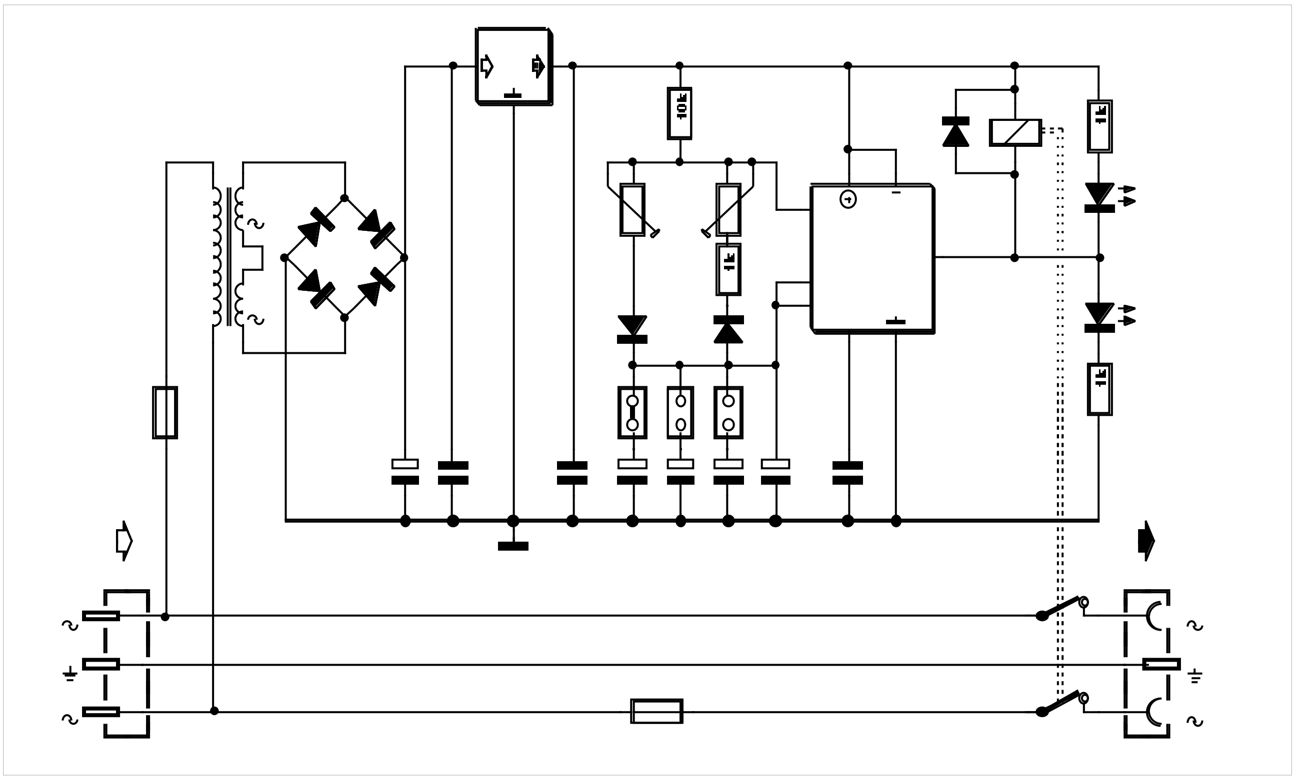

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The timer IC switches a relay whose double-pole contacts switch the mains voltage.

The LEDs indicate whether the mains voltage is switched through (red) or not (green).

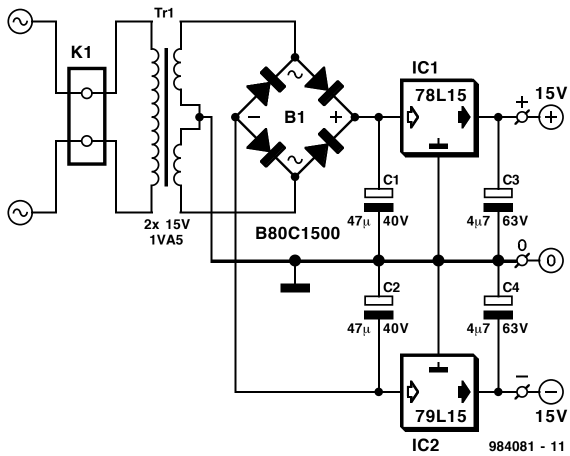
The 100 mA slow fuse protects the mains transformer and low-voltage circuit. The 4 A medium slow fuse protects the relay against overload.

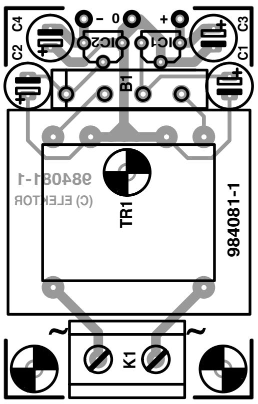
[984122]

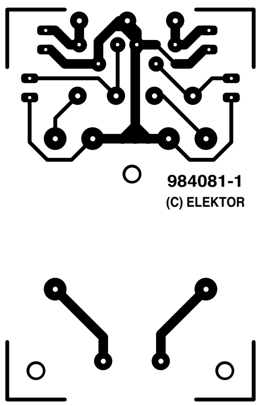
027 XS symmetrical supply

This extra-small (XS) symmetrical supply is useful in those cases where a symmetrical power supply is required with an output capacity of just a few milliamps. The example circuit shows a ±15-V supply capable of delivering a continuous output current of about 25 mA, or 100 mA peak. By using other transformers and/or voltage regulators, the supply can be dimensioned for output voltages of ±5 V, ±9 V, ±12 V, ±15 V, ±18V and ±24 V. For the latter two voltages, however, the negative-voltage regulator may be hard to obtain. Thanks to its modest size, the XS symmetrical supply is easily incorporated into existing equipment.

A disadvantage of small (low-VA) mains transformers as used for this supply is that they often supply relatively high no-load secondary voltages. Under no-load conditions, the indicated Monacor/Monarch transformer, for example, supplies no less than 32 V to the regulator inputs (measured at a mains voltage of 230 V). In some cases, the no-load secondary voltage may exceed the maximum permissible input voltage of the low-power voltage regulator. Typically this will be 30 V for 5-V regulators, 35 V for 12-V and 15-V types, and 40 V for 18-V and 24-V types. When the no-load voltage can be expected to approach the absolute maximum level specified for the voltage regulator, you should connect shunt resistors (bleeders) across the transformer secondaries. Keep the value of these resistors as high as possible to avoid unnecessary dissipation. In most cases, a bleeder current of a few mA is already sufficient to drop the regulator input-voltage to a safe level.







Although the Hahn transformers suggested in the parts list have the same footprint as the Monacor/Monarch types, the 3.2-VA type is taller. If this particular transformer is used, the continuous output current capacity of the supply rises to about 55 mA, provided C1 and C2 are increased to, say, 100 µF/25V. Note, however, that you may have to reduce the no-load secondary voltage as described above. The printed-circuit board shown here is, unfortunately, not available ready-made from the Publishers.

(984081-1, Gb)

COMPONENTS LIST

Capacitors:
C1,C2 = 47µF 40V radial
C3,C4 = 4µF7 63V radial

Semiconductors:
IC1 = 78L15 (see text)
IC2 = 79L15 (see text)
B1 = B80C1500, straight case (80V piv, 1.5A cont.)

Miscellaneous:
K1 = 2-way PCB terminal block, raster 7.5mm
Tr1 = mains transformer, see text.
Examples :
2x15V 1.5VA: type VTR1215 (Monacor/Monarch) or type BV EI 302 2028 (Hahn)
2x15V 3.2VA: type BV EI 306 2078 (Hahn)
Note: Monacor/Monarch and Hahn transformers are supplied by C-I Electronics and Stippler Electronics

028 speech eroder

Design: T. Giesberts

Nowadays, the speech quality on our telephone systems is generally very good, irrespective of distance. However, there are occasions, for instance, in an amateur stage production, or just for fun, when it is desired to reproduce the speech quality of yesteryear.

The eroder circuit accepts an acoustic (via an electret micro-

027 XS symmetrical supply

This extra-small (XS) symmetrical supply is useful in those cases where a symmetrical power supply is required with an output capacity of just a few milliamps. The example circuit shows a ±15-V supply capable of delivering a continuous output current of about 25 mA, or 100 mA peak. By using other transformers and/or voltage regulators, the supply can be dimensioned for output voltages of ±5 V, ±9 V, ±12 V, ±15 V, ±18V and ±24 V. For the latter two voltages, however, the negative-voltage regulator may be hard to obtain. Thanks to its modest size, the XS symmetrical supply is easily incorporated into existing equipment.

A disadvantage of small (low-VA) mains transformers as used for this supply is that they often supply relatively high no-load secondary voltages. Under no-load conditions, the indicated Monacor/Monarch transformer, for example, supplies no less than 32 V to the regulator inputs (measured at a mains voltage of 230 V). In some cases, the no-load secondary voltage may exceed the maximum permissible input voltage of the low-power voltage regulator. Typically this will be 30 V for 5-V regulators, 35 V for 12-V and 15-V types, and 40 V for 18-V and 24-V types. When the no-load voltage can be expected to approach the absolute maximum level specified for the voltage regulator, you should connect shunt resistors (bleeders) across the transformer secondaries. Keep the value of these resistors as high as possible to avoid unnecessary dissipation. In most cases, a bleeder current of a few mA is already sufficient to drop the regulator input-voltage to a safe level.

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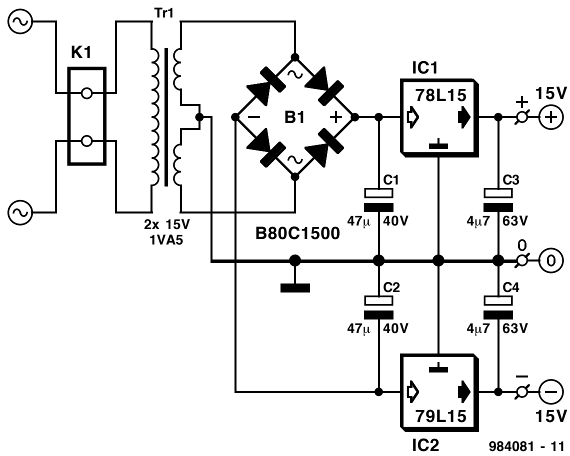

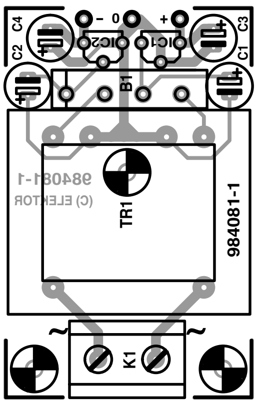

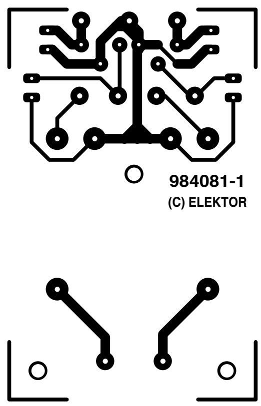

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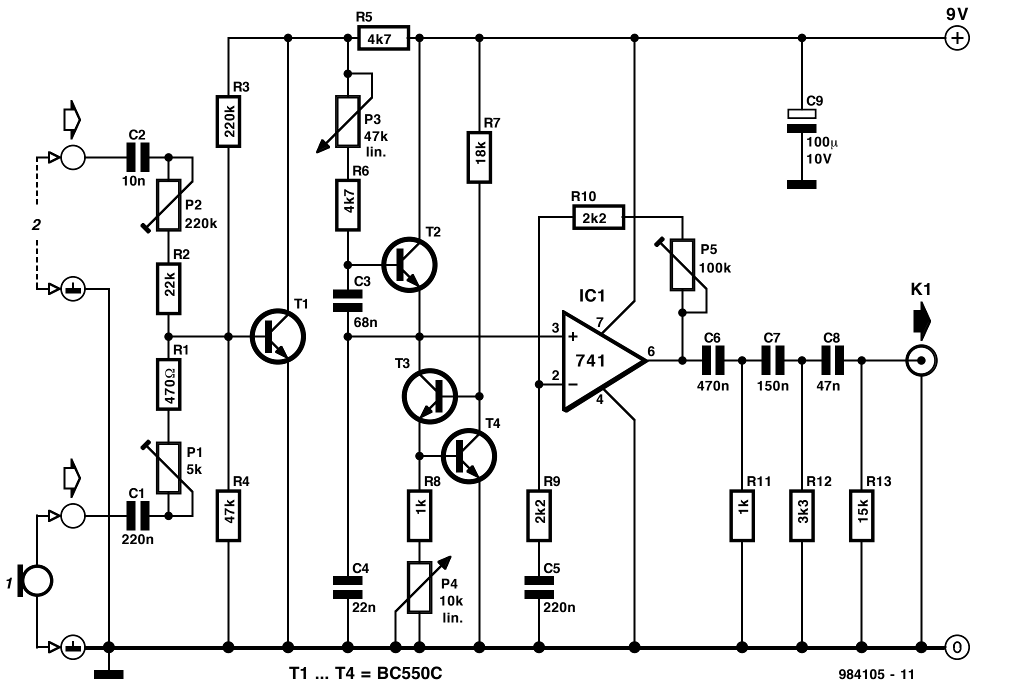


phone) or electrical signal. The signals are applied to the circuit inputs via C_1 and C_2, which block any direct voltage. The input cables should be screened.

The signals are brought to (about) the same level by variable potential dividers P_1-R_1-R_4 and P_2-R_2-R_3, and then applied to the base of transistor T_1. The level of the combined signals is raised by this preamplifier.

The preamplifier is followed by an active low-pass filter consisting of $T_{2-T}4$, C_3, C_4, R_6–R_8, and P_4.

Although, strictly speaking, P_3 serves merely to adjust the volume of the signal, its setting does affect the filter characteristic. Note, by the way, that the filter is a rarely encountered current-driven one in which C_3 and C_4 are the frequency-determining elements. It has a certain similarity with a Wien bridge.

Transistors T_3 and T_4, and resistors R_8 and P_4 form a variable current sink.

The position of P_4 determines the slope of the filter characteristic and the degree of overshoot at the cut-off frequency.

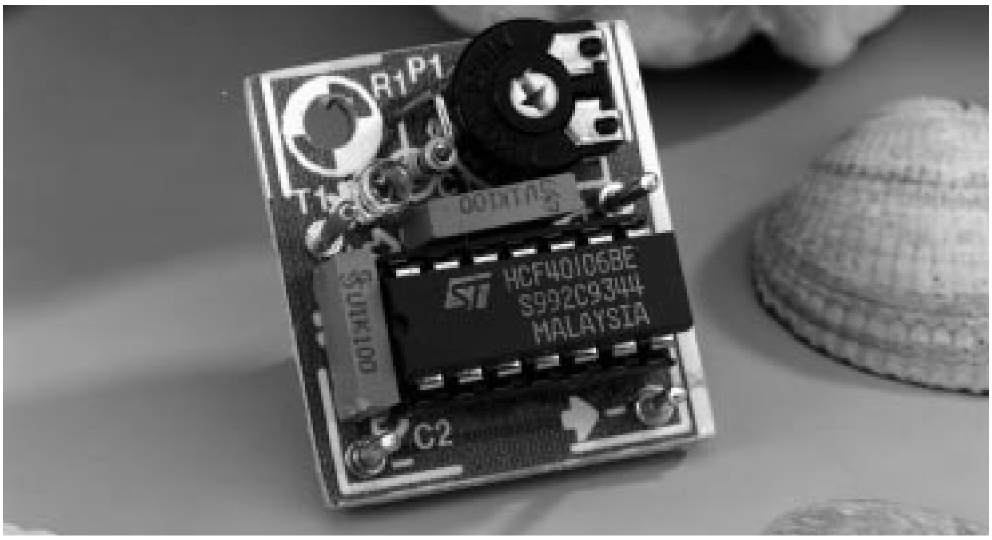
The low-pass filter is followed by an integrated amplifier, IC_1, whose amplification is matched to the input of the electronic circuits connected to the eroder with P_5.

The final passive, third-order high-pass filter is designed to remove frequencies above about 300 Hz.

The resulting output is of a typical nasal character, just as in telephones of the past.

[984105]

029 thrifty light-controlled switch



This circuit shows that only a handful of parts is needed to make a light-controlled switch with a digital power buffer output capable of switching up to 25 milliamps. The circuit is intended mainly for use in low-power battery-powered equipment. The SFH309-4 is a phototransistor from Siemens, its pin connection is included in the circuit diagram. In this application, the SFH309 draws only a few tens of µA. At a certain ambient light intensity level, the voltage at the input of gate IC1f drops below the switching threshold of the Schmitt trigger, and the output consequently toggles to logic high. This level is again inverted by the five remaining gates in the '106 which are connected in parallel to boost their output drive capacity. The effect of stray light picked up from remote controls and other infrared transmitters is suppressed to some extent by R1-C1. If interference is still a problem, then C1 may be increased a little.

The ambient light intensity at which the output changes state is adjusted to individual requirements with preset P1. The supply voltage should be reasonably clean and not exceed

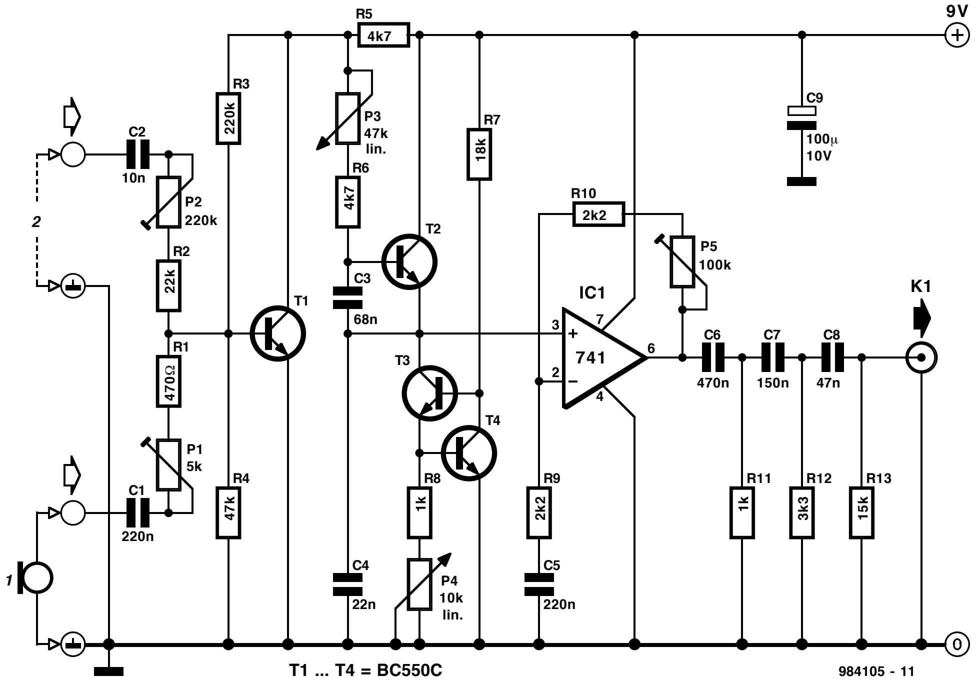
COMPONENTS LIST

Resistor:
R1 = 10MΩ
P1 = 1MΩ preset H

Capacitors:
C1,C2 = 100nF

Semiconductors:
T1 = SFH309-4 (Siemens, ElectroValue)
IC1 = 40106

Miscellaneous:
PCB, not available ready-made



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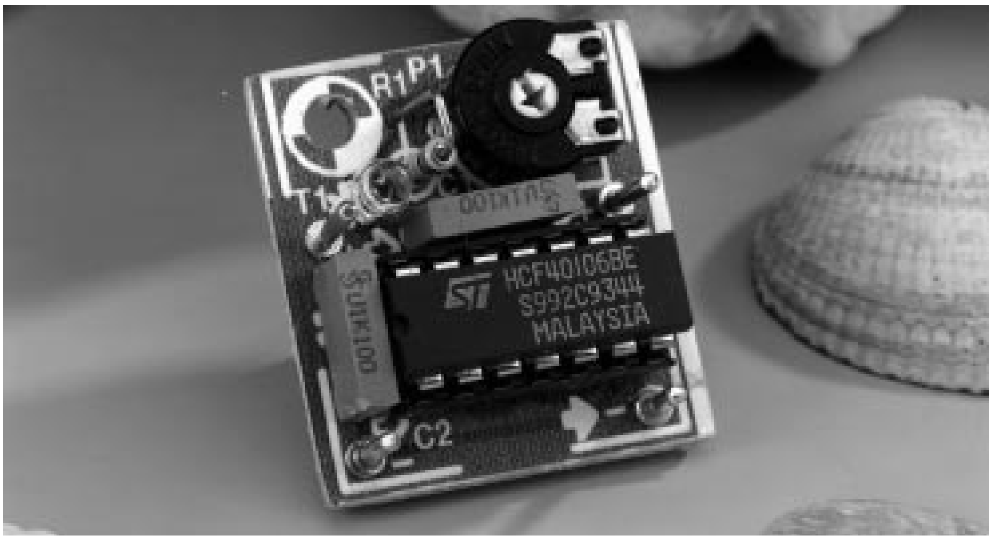

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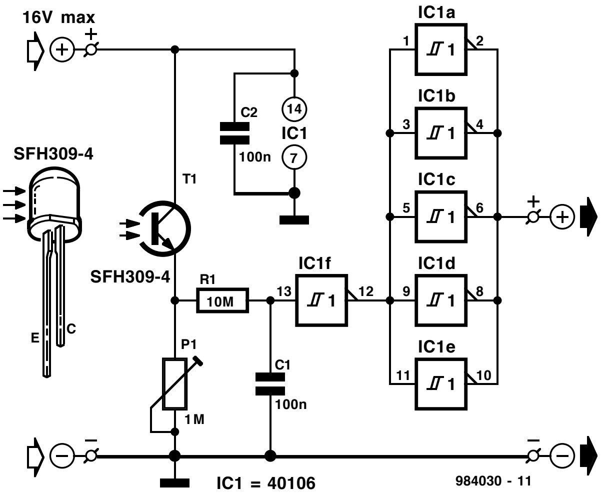
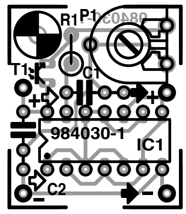
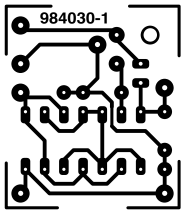






16 volts d.c. The circuit is best built on the miniature printed circuit board shown here. When fitting the phototransistor, make sure it is connected the right way around — the shorter pin is the collector.

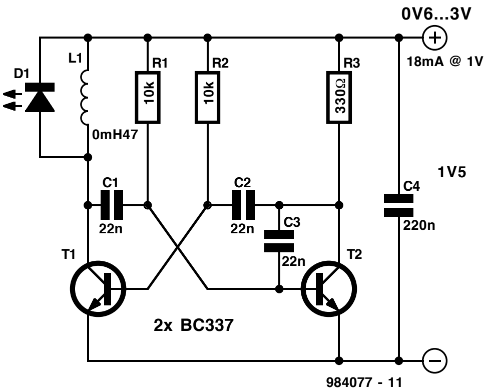
Current consumption of the circuit is 1 to 2μA in the dark, and about 20μA when light is detected (at a 9-V supply and with P1 set to mid-travel). Finally, the switching function of the circuit may be reversed by exchanging P1 and T1, and connecting R1 to the collector.

(984030-1, Gb)

030 light from flat batteries

Button or coin cells that appear to be flat in their normal function may yet be discharged further. This is because in many cases, for instance, a quartz watch stops to function correctly when the battery voltage drops to 1.2 V, although it can be discharged to 0.8 V.

Normally, however, not much can be done with a single cell. In the present circuit, a superbright LED is made to work from voltages between 1 V and 1.2 V. This may be used for map-reading lights, a keyhole light, or warning light when jogging in the dark. When a yellow, superbright LED is used with a fresh battery, it may be used as an emergency reading light or to read a front door nameplate in the dark or to find an non-illuminated doorbell.



Normally, LEDs light at voltages under 1.5 V (red) or 1.6–2.2 V (other colours) only dimly or not at all.

The present circuit uses a multivibrator of discrete design that oscillates at about 14 kHz. The collector resistor of one of the transistors has been replaced by a fixed inductor, which is shunted by the LED. Because of the self-inductance, the voltage across the LED is raised, so that the diode lights dimly at voltages as low as 0.6 V and becomes bright at voltages from about 0.8 V up.

The circuit requires a supply voltage of 0.6–3 V and draws a current of about 18 mA at 1 V.

[Zeiller – 984077]

031 infra-red burglar alarm

Design G. Pradeep

The alarm circuit uses infra-red light beams to bridge distances between 3 m and 5 m (10 ft to 16 ft), but if the transmit diode is given a reflector, larger distances are possible. When the beam is interrupted, a buzzer sounds.

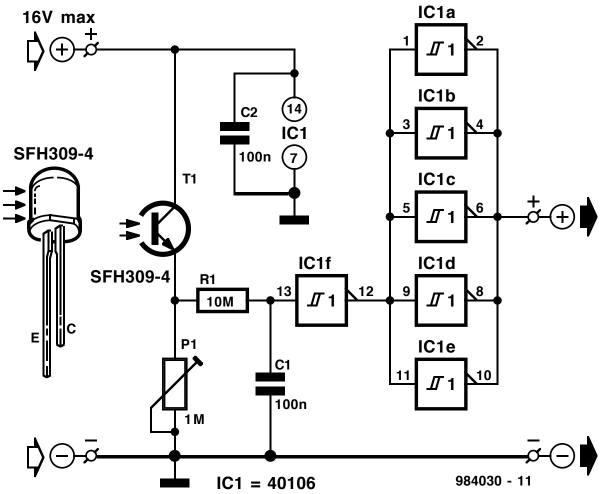
The transmitter is based on a Type 555 timer circuit, which generates 10 μs wide pulses at a rate of 20 kHz. During the pulse, a current of about 100 mA flows through the transmit diode. The average current drawn by the transmitter is about 12 mA, which will normally preclude a battery-operated supply.

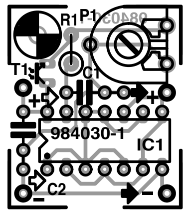
The receiver is rather more complex. The receive diode is normally cut off, but comes on when it is exposed to infra-red light. The more intense the infra-red light, the larger the photo current. This means that the received pulses cause an alternating voltage across resistor R_1.

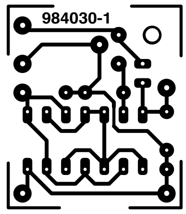
The a.c.-coupled amplifier based on transistors T_1–T_4 provides an amplification of ×200 at a frequency of 20 kHz.

The bandwidth of the receiver is purposely limited to enhance the stability of the circuit.

The pulses arriving from the







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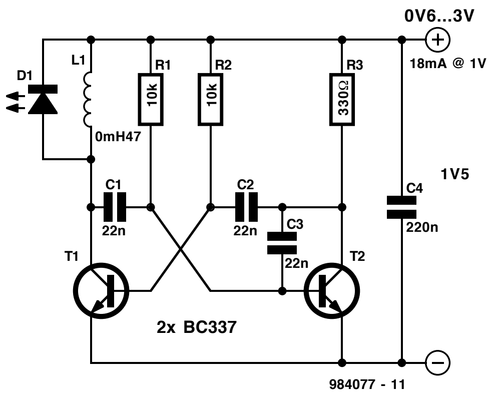

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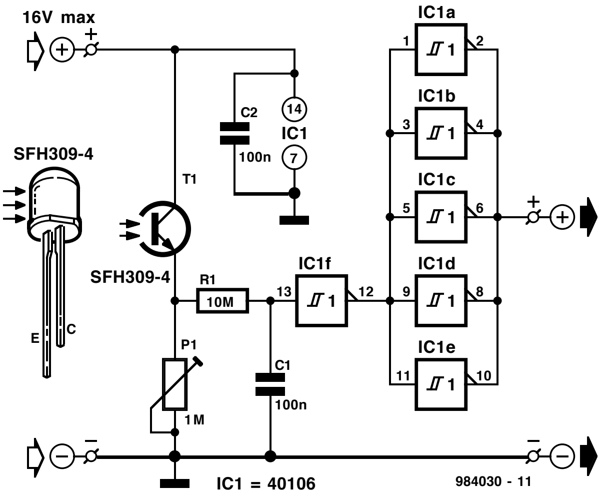

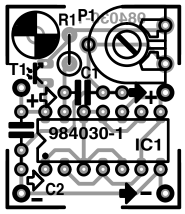

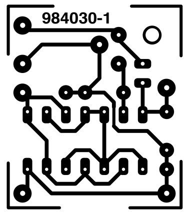

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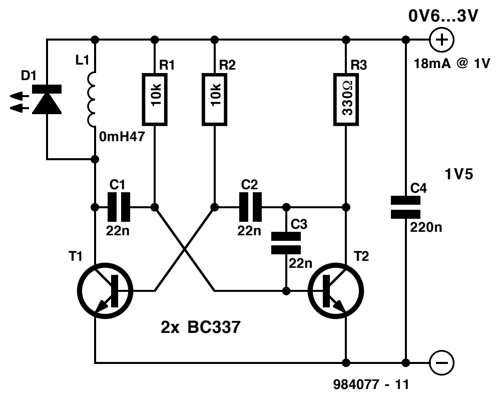

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The pulses arriving from the

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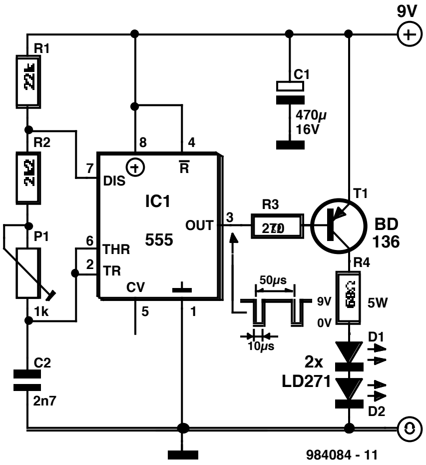
When the infra-red signal fails (because the beam is broken), the bistable is set, whereupon the buzzer is actuated. The sounding of the buzzer cannot be interrupted with switch S_1. When, however, the beam is restored, pressing S_1 causes the



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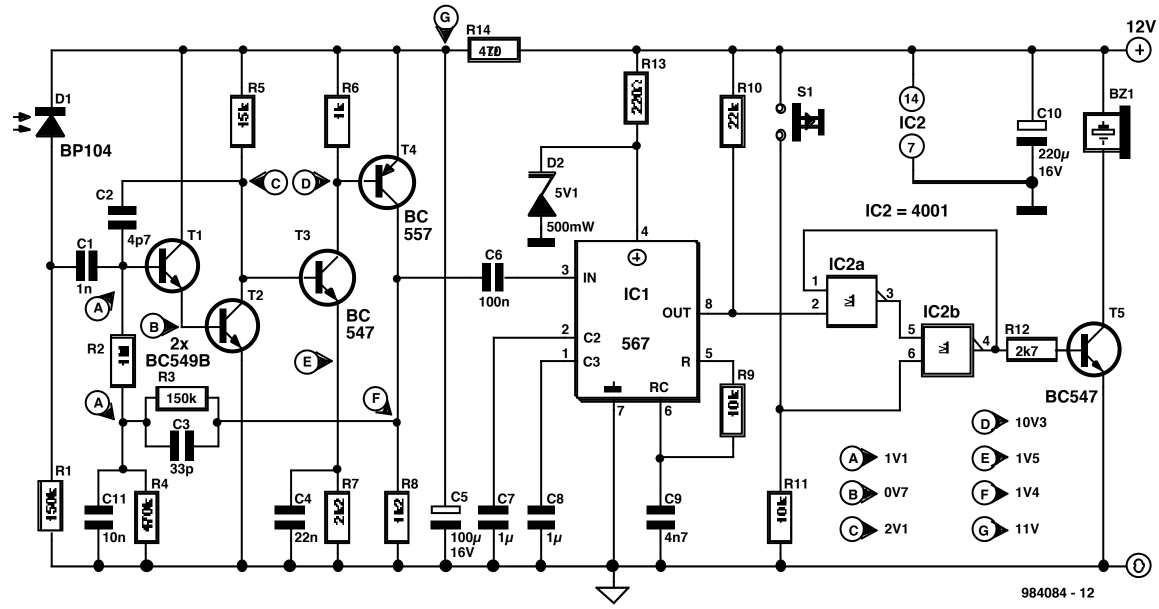
The receiver draws a current of about 30 mA in the quiescent state, which rises to about 50 mA when the buzzer sounds.

The relatively large currents make battery operation uneconomical; it is far better (and safer) to use an appropriate mains adaptor.

[984084]

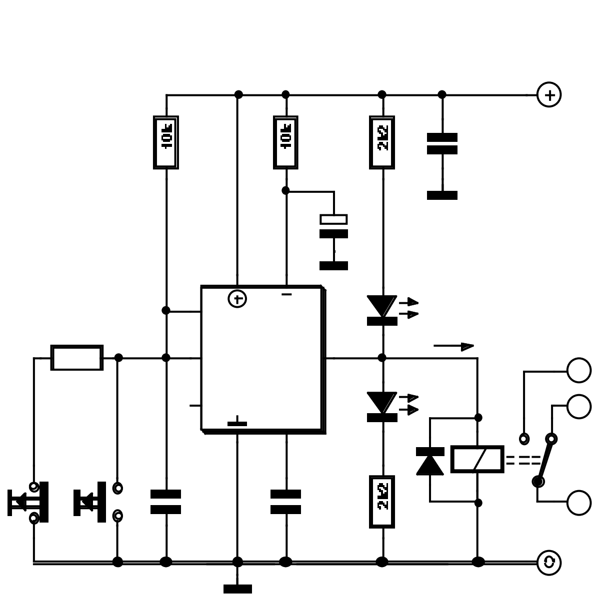


032 latch uses 555 in memory mode

By G. Kleine

The familiar Type 555 can be used to switch currents up to 200 mA. Less well-known is its use as a latch with control input. When the input pins 2 (trigger) and 6 (threshold) are linked and connected to half the supply voltage, the output can be switched as follows. When the potential at pins 2 and 6 is raised to full supply voltage level, the output is switched to ground. When pins 2 and 6 are linked to ground potential, the output assumes supply voltage level.

The circuit in the diagram uses this mode of operation of the 555 to realize a two-wire



on/off switch. The combination S_1 (closed), R_2 and R_1 provides half the supply voltage to the input (pins 2, 6) of IC_1. When S_2 is closed, the output, pin 3, goes high so that D_2 (on) lights. When S_1 is opened, the input at pins 2, 6 of IC_1 rises to above 2/3 of the supply voltage, whereupon IC_1 is disabled and the output goes low. Diode D_1 (off) then lights.

Network R_3-C_1 at the reset input, pin 4, forces the latch to come up in the off state when power is first applied.

(Source: Electronic Design, November 6, 1995)

[984126]

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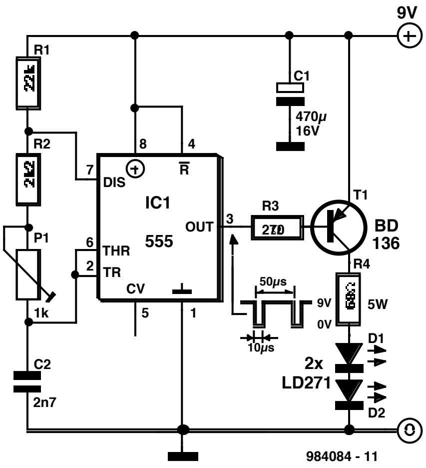
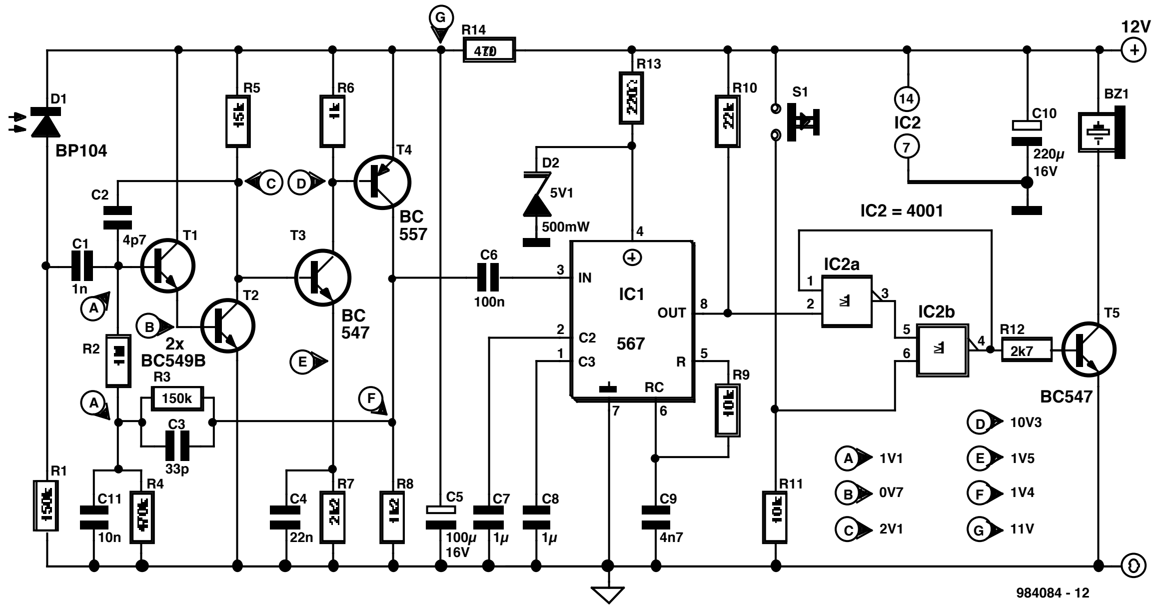




032 latch uses 555 in memory mode

By G. Kleine

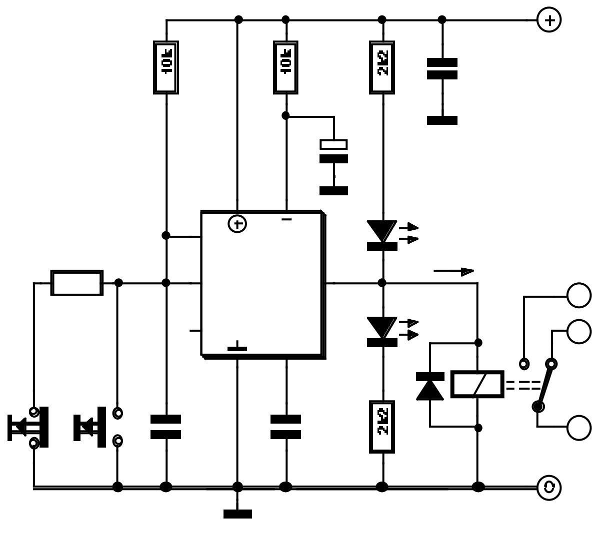
The familiar Type 555 can be used to switch currents up to 200 mA. Less well-known is its use as a latch with control input. When the input pins 2 (trigger) and 6 (threshold) are linked and connected to half the supply voltage, the output can be switched as follows. When the potential at pins 2 and 6 is raised to full supply voltage level, the output is switched to ground. When pins 2 and 6 are linked to ground potential, the output assumes supply voltage level.

The circuit in the diagram uses this mode of operation of the 555 to realize a two-wire on/off switch. The combination S_1 (closed), R_2 and R_1 provides half the supply voltage to the input (pins 2, 6) of IC_1. When S_2 is closed, the output, pin 3, goes high so that D_2 (on) lights. When S_1 is opened, the input at pins 2, 6 of IC_1 rises to above 2/3 of the supply voltage, whereupon IC_1 is disabled and the output goes low. Diode D_1 (off) then lights.

Network R_3-C_1 at the reset input, pin 4, forces the latch to come up in the off state when power is first applied.

(Source: Electronic Design, November 6, 1995)

[984126]



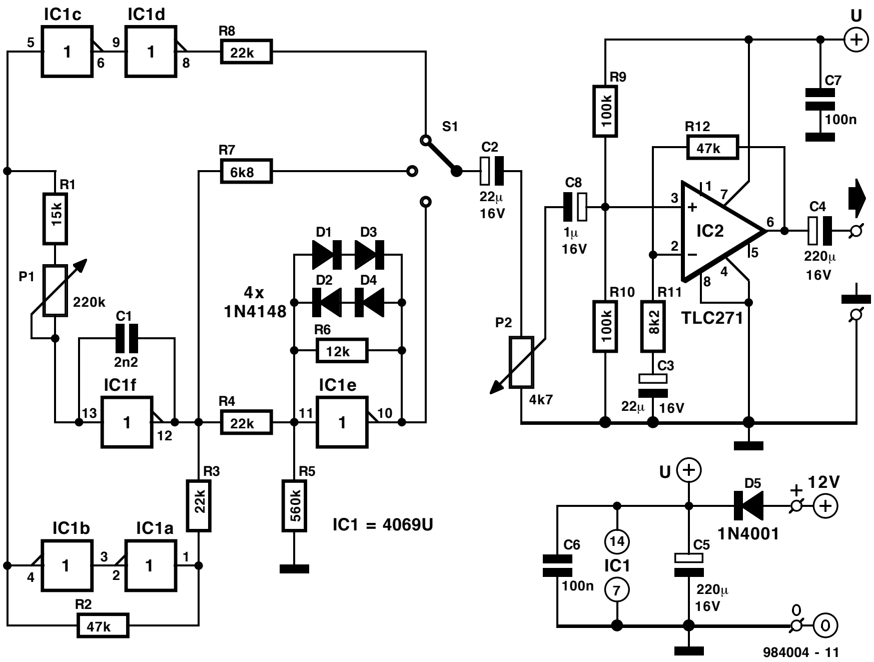
033

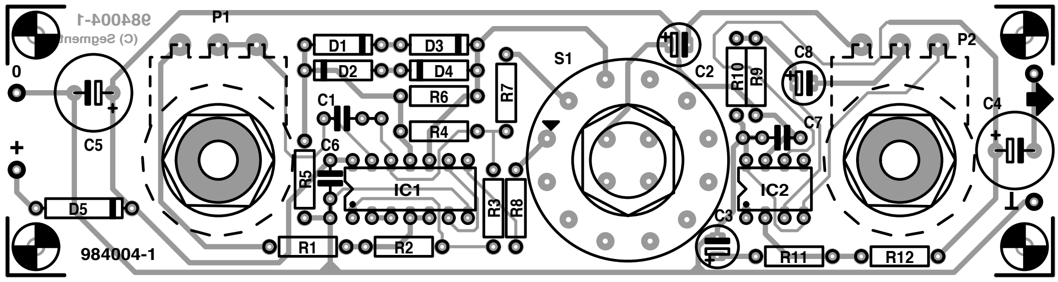
low-cost function generator

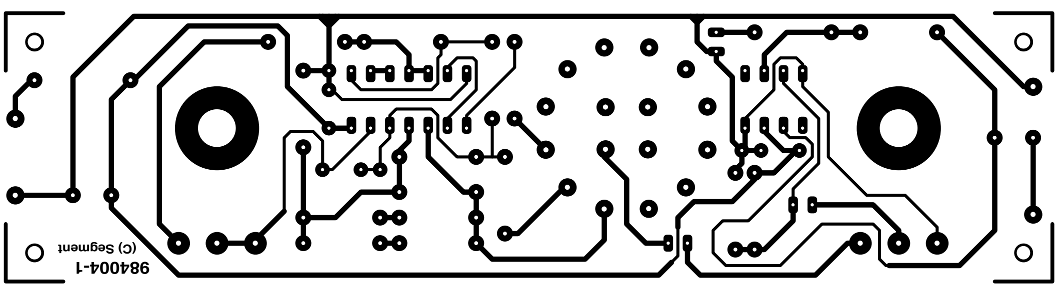
Design: G.Baars

Here's a function generator that won't break the bank yet offers perfectly acceptable output waveforms for many applications in your workshop. The generator supplies sinewave, rectangle and triangle waveforms within the frequency range 1 kHz to about 15 kHz. The output level is adjustable between 0 and about 10 V_{pp}. Though frugal, these specifications make the generator a useful piece of test equipment for audio design, experimentation and repair purposes. Since only common-or-garden components are used, the generator can be built at a modest outlay.

Here's how it works. Inverter gates IC1a and IC1b are connected to resistors R2 and R3 to form a buffer with some hysteresis. Another gate from the '4069U integrated circuit, IC1f, acts as an integrator together with R1, P1 and C1. The potentiometer, P1, defines the integrator's time-constant. A buffer acting as a comparator with hysteresis, together with the integrating effect provided by IC1f results in an oscillator whose output frequency is controlled by potentiometer P1. The buffer supplies a rectangular output signal, the integrator, a triangular one. The rectangular signal is further shaped and buffered by two more gates, IC1c and IC1d, before it is applied, via R8, to one of the contacts of the waveform selection switch, S1. The triangular signal is also applied to the switch, by way of R7. The triangular signal supplied by IC1f is fed to a sinewave shaper







COMPONENTS LIST

Resistors:
R1 = 15kΩ
R2,R12 = 47kΩ
R3,R4,R8 = 22kΩ
R5 = 560kΩ
R6 = 12kΩ
R7 = 6kΩ8
R9,R10 = 100kΩ
R11 = 8kΩ2
P1 = 220kΩ linear potentiometer
P2 = 4kΩ7 linear potentiometer

Capacitors:
C1 = 2nF2 MKT (Siemens)
C2,C3 = 22µF 16V radial
C4,C5 = 220µF 16 V radial
C6,C7 = 100nF Sibatit (miniature ceramic, Siemens)
C8 = 1µF 16 V radial

Semiconductors:
D1-D4 = 1N4148
D5 = 1N4001
IC1 = 4069U (U = unbuffered version!)
IC2 = TLC271CP

Miscellaneous:
S1 = 3-way rotary switch, 4 poles, PCB mount

consisting of IC1e, R4, R6, R5 and diodes D1 through D4. The output signal is applied directly to the waveform selection switch.

Because the three waveforms have different individual levels, the sinewave being the smallest, they have to be made roughly equal before they can be applied to the output amplifier, IC2. This levelling is achieved with the aid of the aforementioned resistors R7 and R8 for the triangle and rectangle wave respectively, in combination with the output level control pot, P2. The TLC271 opamp is wired for a gain of 6.7 times in order to achieve a maximum (no-load) output level of about 10 V_{pp}. The minimum load impedance to be observed is about 600 ohms.

The generator is powered from a regulated 12-volt source, and its current consumption will be of the order of 20 mA, depending, of course, on the load connected to the output.

The printed circuit board designed for the generator also contains all the controls, i.e., the frequency control pot, the waveform selection switch and the output level control pot, so that no tedious wiring is required. The project is conveniently boxed by drilling holes for the pot and switch shafts in the front panel, and then mounting the completed circuit board against the inside of the front panel. Unfortunately, the PCB for this project is not available ready-made through the Publishers.

(984004)

034 1.5 A step-down switching regulator

By G. Kleine

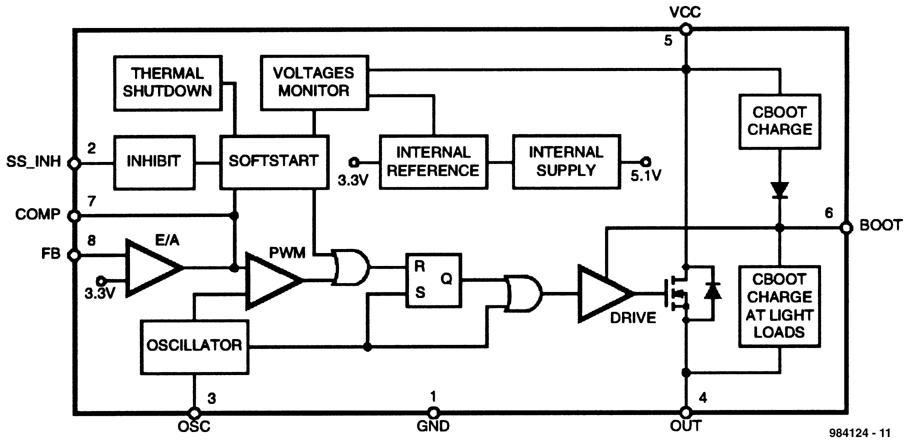
Step-down regulators are more and more often used to derive low supply voltage from a higher voltage without incurring losses. Owing to the steadily increasing switching rates, the suppression of interference and noise that are unavoidable by-products of the switching has become much simpler.

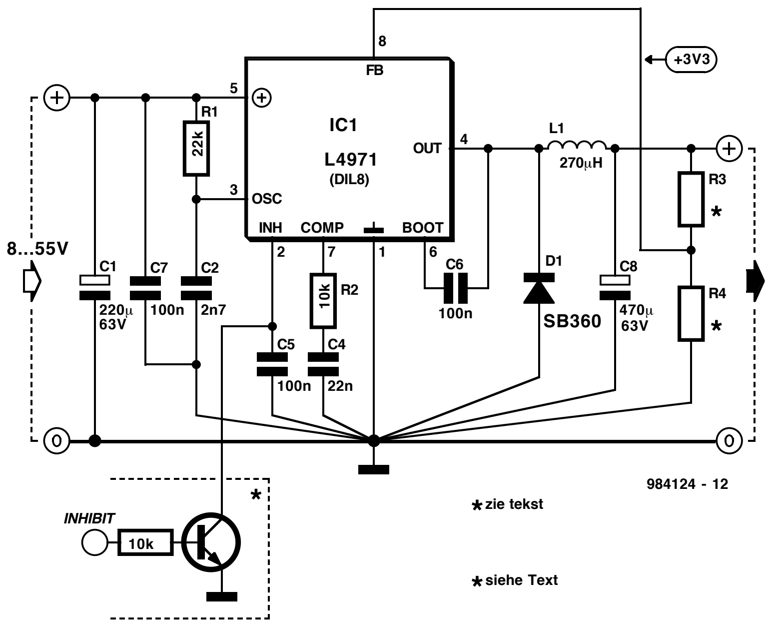
The L4971 from SGS-Thomson is a step-down monolithic power switching regulator delivering 1.5 A at a voltage between +3.3 V and +40 V (selected by a simple external divider). Realized in BCD mixed technology, the regulator is housed in a DIL8 or SO16-SMD enclosure.

The input voltage ranges from 8 V to +55 V, while the efficiency is 85%, rising to 95% when the input voltage is only a few volts higher than the output voltage.

The external divider to set the output voltage is formed by R_3-R_4. For an output of 3.3 V, V_{out} may be linked directly to FB (pin 8). Resistor values for some usual output voltages are given in the table.

The normal switching frequency of the L4971 is 100 kHz, but this may be increased to 500 kHz by making $R_1 = 12\ k\Omega$ and $C_2 = 0.001\ \mu F$.





V_{out} (V)	R_3 (kΩ)	R_4 (kΩ)
3.3	0	∞
5.1	2.7	4.4
9.0	8.2	4.7
12	12	4.7
15	16	4.7
18	20	4.7
24	30	4.7

The soft start/inhibit input (SS_INH pin 2) may be used with an open-collector output to inhibit the controller or capacitor C_5 for the soft start function.

A suitable core for L_1 is the Type T-94-26 from Amidon. The specified inductance is obtained with 65 turns close-wound 0.5 mm dia. enamelled copper wire.

Other features include pulse by pulse current limit, hiccup mode for short-circuit protection, voltage feed forward regulation, protection against feedback loop disconnection, inhibit for zero current drain and thermal shutdown.

[984124]

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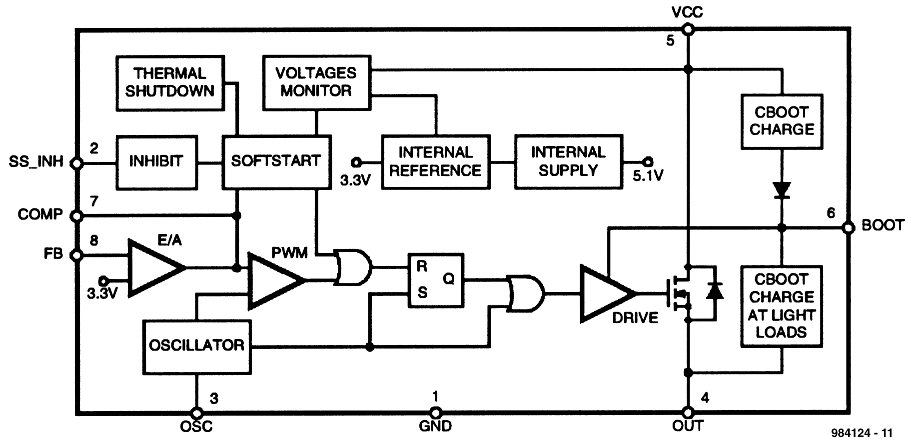

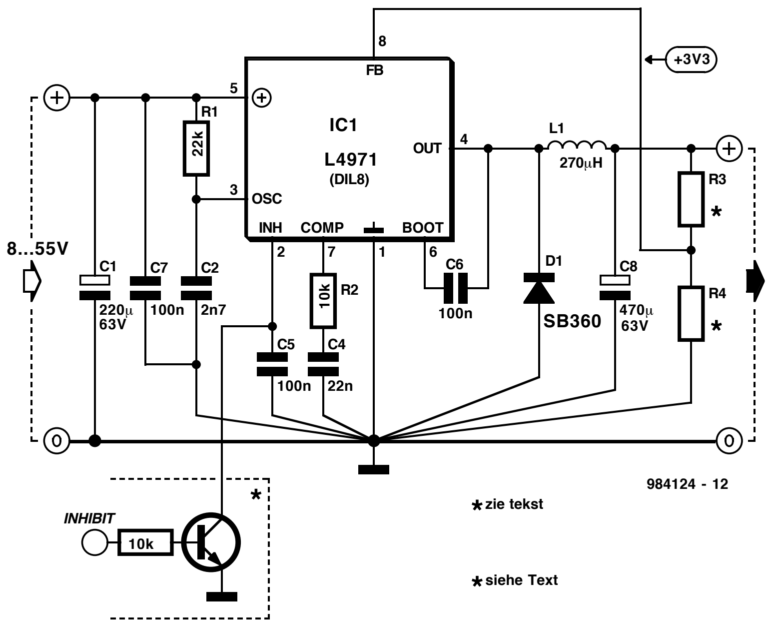

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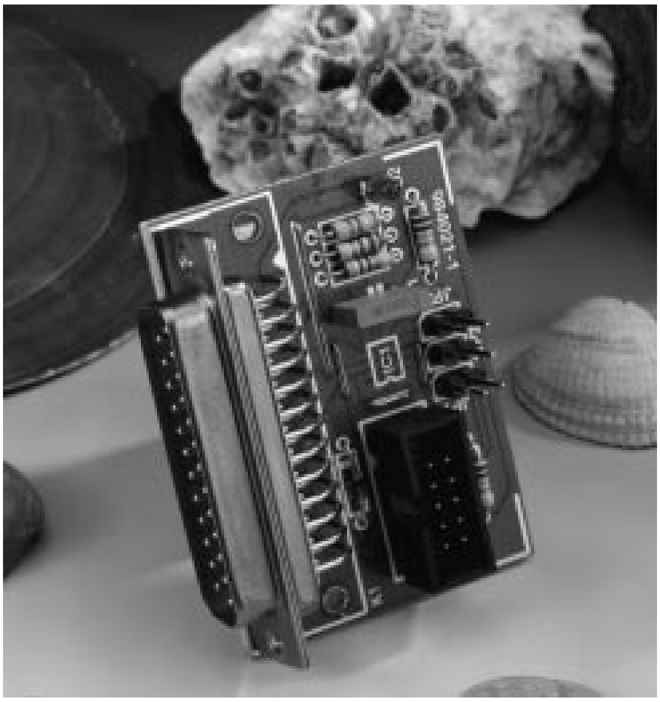
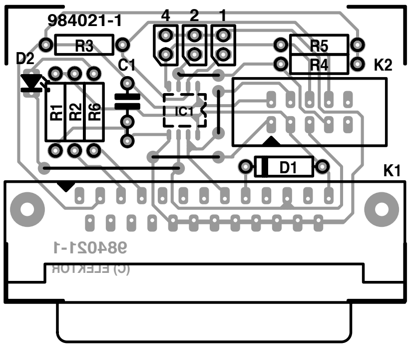
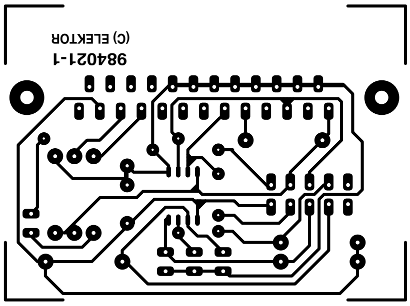
[984124]

035

I²C temperature sensor







Design K. Walraven

The LM75 from National Semiconductor is a temperature sensor, Delta-Sigma analogue-to-digital converter (ADC), and digital over-temperature detector with I²C™ interface. It is manufactured in surface-mount technology (SMT) for operation from 5 V or 3.3 V. The temperature may be read in half degrees in the range –55 °C to +125 °V. It provides a 9-bit output in twos complement (that is, 0FAH is +125 °C; 192H is –55 °C; 001H is +0.5 °C; 1FFH is –0.5 °C).

The LM75 can operate as a stand-alone temperature switch, for which purpose an upper and a lower switching level may be programmed in. The output of the device goes low when the set temperature is exceeded. This output may also be used as an interrupt for a computer or microcontroller. At power-up, the switching levels are fixed at 80 °C and 75 °C

The circuit shown is based on an LM75 and may be connected to the Centronics port of

Parts list

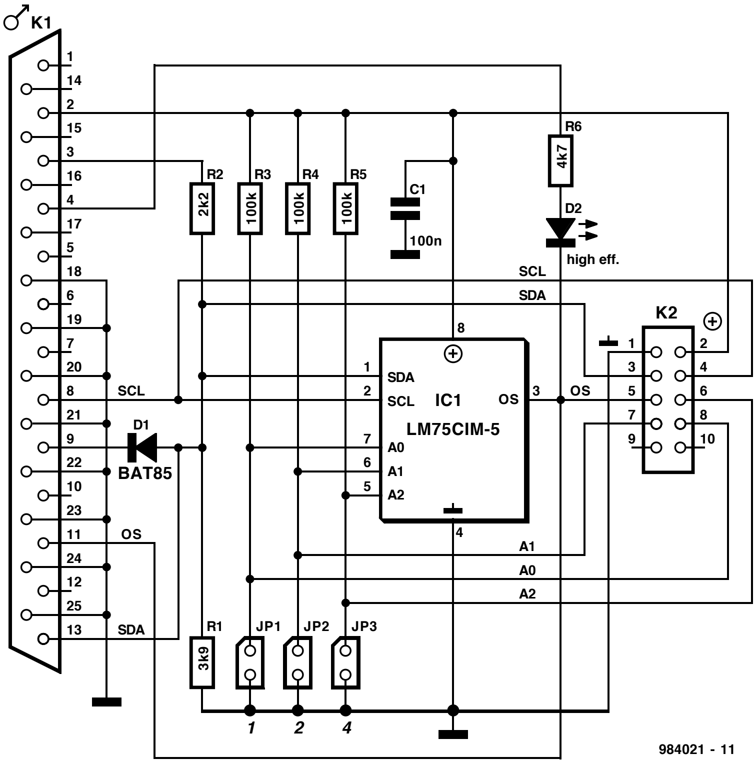
Resistors:
R_1 = 3.9 kΩ
R_2 = 2.2 kΩ
R_3–R_5 = 100 kΩ
R_6 = 4.7 kΩ

Capacitors:
C_1 = 0.1 μF

Semiconductors:
D_1 = BAT85
D_2 = LED, high efficiency

Integrated circuits:
IC_1 = LM75CIM-5

Miscellaneous:
K_1 = DB25 connector, male, right-angled, for board mounting
K_2 = 10-way box header for board mounting
JP_1–JP_3 = 2-way pin strip header with jumper link
PCB Order No. 984021 (see Readers Services towards the end of this issue)



a computer via a 25-way 1:1 cable. The port then functions as an I^{2}C interface. The necessary software, datasheet and application note may be downloaded from www.national.com/pf/LM/LM75.html.

Operation of the software is simple: at the top left is a button which when set to 'off' renders the Centronics port voltageless.Connect the board and select the relevant Centronics address and an I^{2}C address. This means that the highest address on the board (lowest in the list) must be selected without the use of jumpers. Set the button to 'on' and temperature monitoring starts.

Since the circuit draws current from the Centronics port, a dedicated power supply is not required. However, readers who worry about the additional load placed on their PC, may note that the LM75 draws a current not exceeding 250 μA.

[984021]

036 fast voltage-driven current source

Design: H. Bonekamp

The current source in the diagram, which react very fast to changes in the input signal, may be used, for instance, in certain measurements.

Differential amplifier IC$_1$ ensures that the potential across R$_2$ is equal to the input voltage:

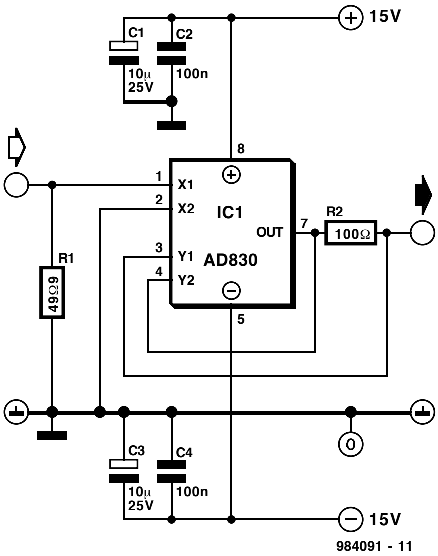
$$I_{out} = U_{in}/R_2.$$

The bandwidth, B, is given by

$$B = R_2 f / R_L,$$

where f=80 MHz, and the load impedance $R_L \geq R_2$ (both in ohms).

The input is terminated into



R$_1$ to give the usual 50 Ω impedance required by measuring instruments. At the same time, this resistor sets the d.c. operating point. If the link to the driving signal source is short and d.c. coupled, R$_1$ may be omitted.

The peak voltage between pins 1 and 2 of the IC is limited to 2.1 V to prevent too large a current at the output. Therefore, the peak output current is 2.1/100=21 mA.

[984091]

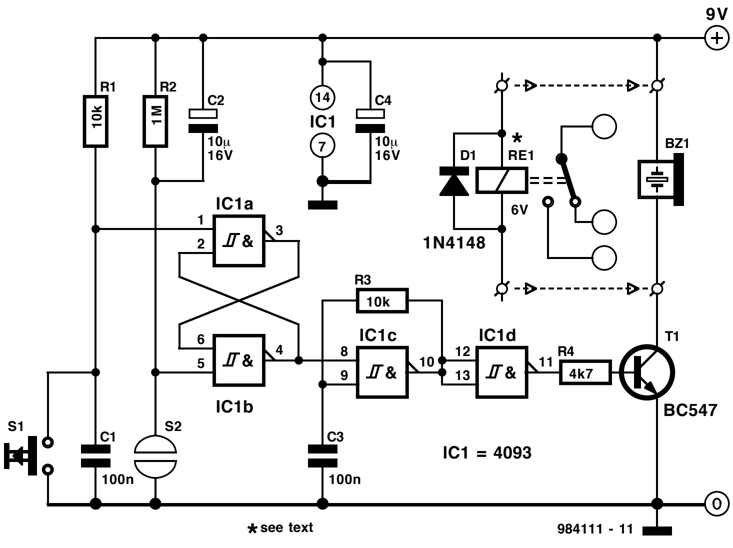
037 simple moisture detector

Design: Pradeep G.

The function of this circuit is to sound a buzzer, or, optionally, actuate a relay, when a certain moisture level is detected between a pair of probes.

The circuit has a 'memory' in the form of a flip-flop, IC1a-IC1b, which enables or disables a tone oscillator, IC1c. The flip-flop is reset either by C1 and C2 when the supply voltage appears, or by push-button S1. This may not reset the alarm, however, which will sound again until the probes are 'dry'.

The (passive) buzzer may be replaced by a relay actuating an externally connected sounder, lamp or other high-power signalling device. Because the duty



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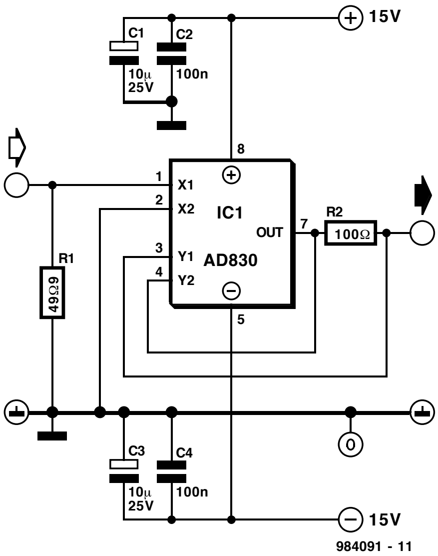

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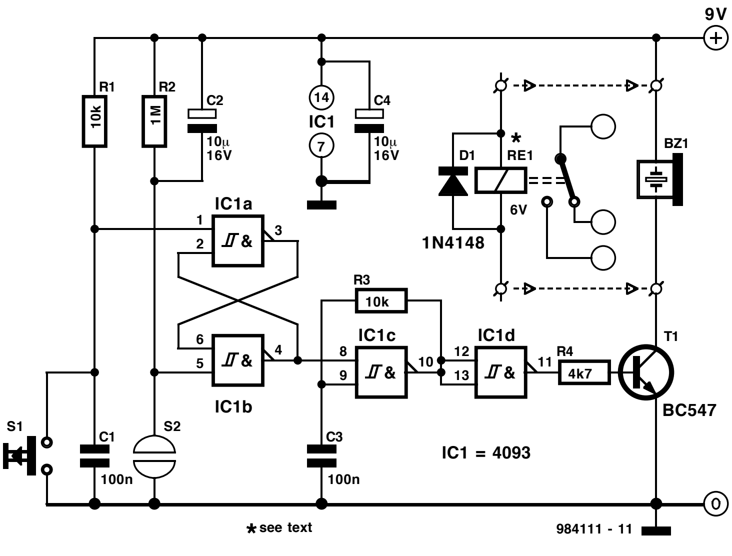

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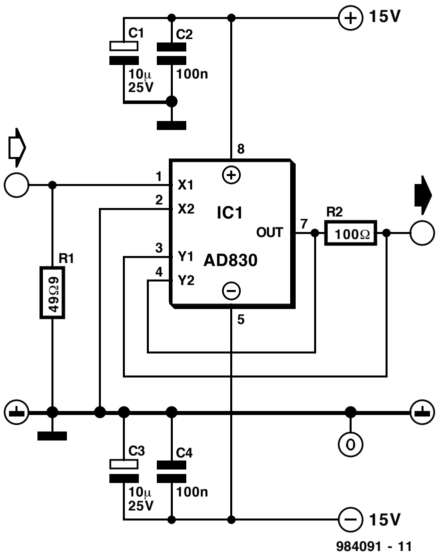

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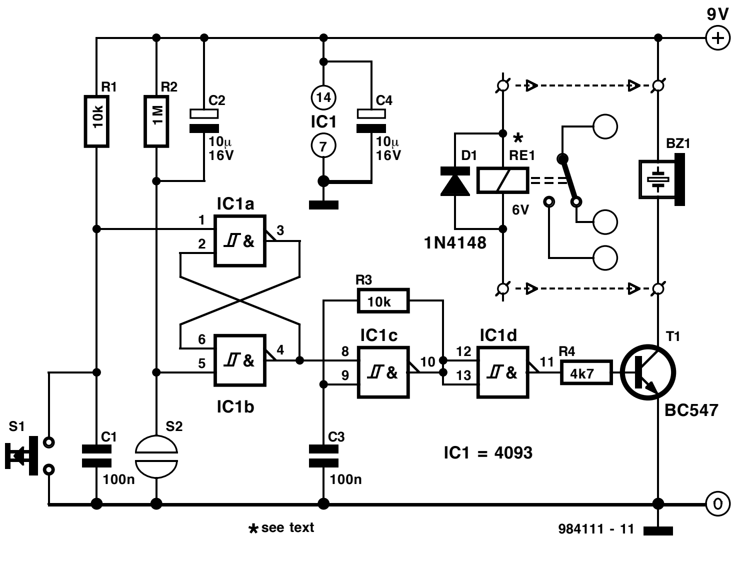


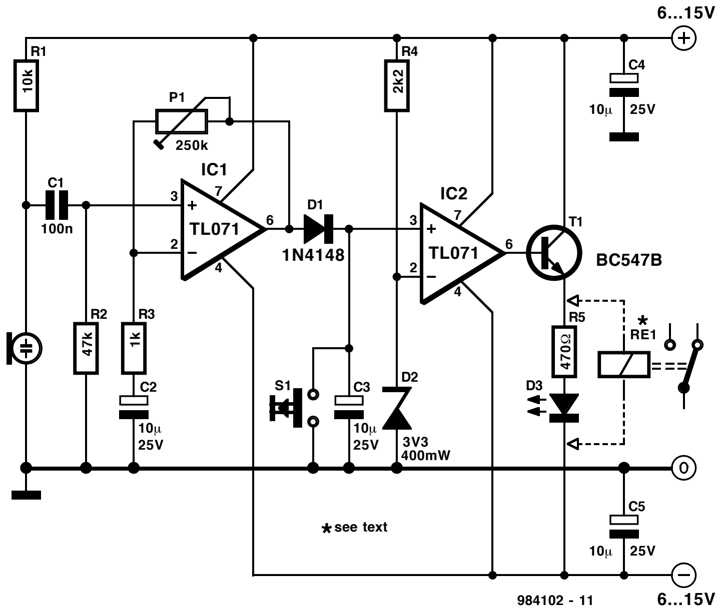
factor of the coil voltage is about 0.5, the relay should be a type with a coil voltage which is lower than the supply voltage. A 6-volt type is suggested if the circuit is powered from a 9-volt supply. The circuit has a modest stand-by current consumption of between 4 and 5 mA. This rises to about 40 mA when the relay is actuated. The supply voltage is uncritical and may be anything between 3 V and 15 V. Note, however, that it may not be possible to use a relay if a supply voltage lower than about 8 V is employed. If the circuit is found to be too sensitive, the value of resistor R2 may be decreased. (984111-1, W)

038 ambient-noise monitor

Design: T. Giesberts

Excessive noise is bad for your health and bad for your surroundings. It cannot be said too often: too many young people go prematurely deaf because of prolonged exposure to loud sounds. There cannot be any pleasure in excessively loud music: it hurts and, like skin cancer, the terrible effects do not immediately become noticeable.

The monitor in the diagram gives a visible warning when the ambient noise is at a dangerous level or it actuates a relay.



The noise sensor is a two-terminal electret microphone that is powered via R_1. The audio signal is applied to op amp IC_1, whose input resistance is fixed at 47 kΩ by R_2. The signal amplification can be set from unity to ×250 with P_1.

Operational amplifier IC_2 functions as a comparator which likens the amplified signal with a reference voltage of 3.3 V. If the signal at the non-inverting input of the op amp exceeds the reference voltage, the output of IC_2 changes state (goes high), whereupon T_1 is switched on. When this happens, the relay is energized or the LED lights. The relay contacts may be used to operate a warning light or buzzer, or to switch the noise source off. In the latter case, C_2 prevents the circuit returning to its original state (which would cause the noise source to come on again). The capacitor is charged to the peak value of the signal. Owing to the presence of D_1, it cannot be discharged via the output of IC_1, but only, and very slowly, via the high-resistance input of IC_2.

The monitor is reset with S_1.

[984102]

039 up/down drive for tone control

Design: T. Giesberts

The up/down drive is intended primarily for use with the tone controls described elsewhere in this issue.

The tone controls use electronic switches that are operated by a multi-position selector. The present circuit is intended as a replacement for this selector and has facilities for operating the tone controls via an up and a down key. A third key enables the user to switch over rapidly to a preprogrammed position of the relevant tone control.

The electronic switches are driven by a BCD-to-decimal decoder Type 4028 (IC_3), which in turn is controlled by a 4-bit preset up/down counter (IC_2). The counter uses the three lowest bits only. The MSB of decoder IC_3 is permanently low. Only the eight lowest outputs of the decoder are used and these are linked via K_1 to the control inputs of IC_1 and IC_2 in the tone controls.

The circuit is operated with S_1 and S_2. Switch S_3 is the earlier mentioned preset key. The data for the preset inputs are set with DIP switch S_4. Capacitor C_3 ensures that when the supply voltage is switched on, the preset data are automatically adopted by the counter.

Each of switches S_1 and S_2 drives an S/R bistable (US: flip-flop), which determines the level at the U/D input of counter IC_2.

Networks R_3-C_1 and R_4-C_2, in conjunction with Schmitt trigger IC_{1b} provide a thorough debouncing and at the same

factor of the coil voltage is about 0.5, the relay should be a type with a coil voltage which is lower than the supply voltage. A 6-volt type is suggested if the circuit is powered from a 9-volt supply. The circuit has a modest stand-by current consumption of between 4 and 5 mA. This rises to about 40 mA when the relay is actuated. The supply voltage is uncritical and may be anything between 3 V and 15 V. Note, however, that it may not be possible to use a relay if a supply voltage lower than about 8 V is employed. If the circuit is found to be too sensitive, the value of resistor R2 may be decreased. (984111-1, W)

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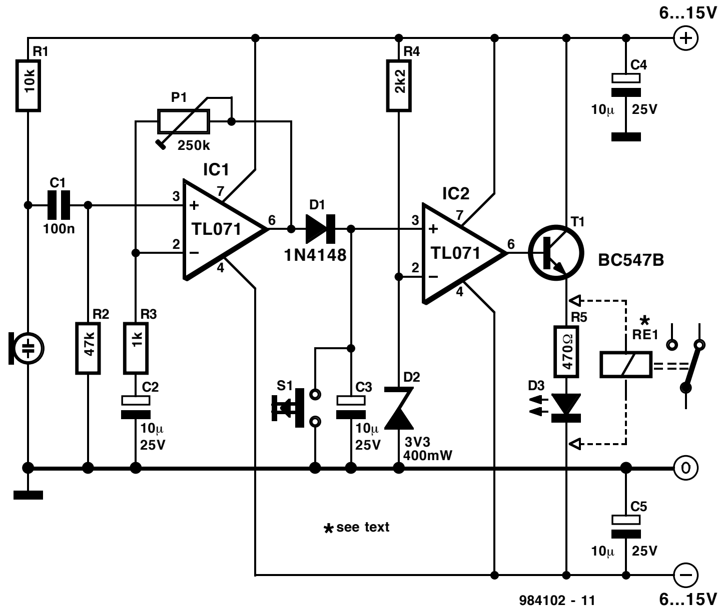
The noise sensor is a two-terminal electret microphone that is powered via R_1. The audio signal is applied to op amp IC_1, whose input resistance is fixed at 47 kΩ by R_2. The signal amplification can be set from unity to ×250 with P_1.

Operational amplifier IC_2 functions as a comparator which likens the amplified signal with a reference voltage of 3.3 V. If the signal at the non-inverting input of the op amp exceeds the reference voltage, the output of IC_2 changes state (goes high), whereupon T_1 is switched on. When this happens, the relay is energized or the LED lights. The relay contacts may be used to operate a warning light or buzzer, or to switch the noise source off. In the latter case, C_2 prevents the circuit returning to its original state (which would cause the noise source to come on again). The capacitor is charged to the peak value of the signal. Owing to the presence of D_1, it cannot be discharged via the output of IC_1, but only, and very slowly, via the high-resistance input of IC_2.

The monitor is reset with S_1.

[984102]



039 up/down drive for tone control

Design: T. Giesberts

The up/down drive is intended primarily for use with the tone controls described elsewhere in this issue.

The tone controls use electronic switches that are operated by a multi-position selector. The present circuit is intended as a replacement for this selector and has facilities for operating the tone controls via an up and a down key. A third key enables the user to switch over rapidly to a preprogrammed position of the relevant tone control.

The electronic switches are driven by a BCD-to-decimal decoder Type 4028 (IC_3), which in turn is controlled by a 4-bit preset up/down counter (IC_2). The counter uses the three lowest bits only. The MSB of decoder IC_3 is permanently low. Only the eight lowest outputs of the decoder are used and these are linked via K_1 to the control inputs of IC_1 and IC_2 in the tone controls.

The circuit is operated with S_1 and S_2. Switch S_3 is the earlier mentioned preset key. The data for the preset inputs are set with DIP switch S_4. Capacitor C_3 ensures that when the supply voltage is switched on, the preset data are automatically adopted by the counter.

Each of switches S_1 and S_2 drives an S/R bistable (US: flip-flop), which determines the level at the U/D input of counter IC_2.

Networks R_3-C_1 and R_4-C_2, in conjunction with Schmitt trigger IC_{1b} provide a thorough debouncing and at the same

factor of the coil voltage is about 0.5, the relay should be a type with a coil voltage which is lower than the supply voltage. A 6-volt type is suggested if the circuit is powered from a 9-volt supply. The circuit has a modest standby current consumption of between 4 and 5 mA. This rises to about 40 mA when the relay is actuated. The supply voltage is uncritical and may be anything between 3 V and 15 V. Note, however, that it may not be possible to use a relay if a supply voltage lower than about 8 V is employed. If the circuit is found to be too sensitive, the value of resistor R2 may be decreased. (984111-1, W)

038 ambient-noise monitor

Design: T. Giesberts

Excessive noise is bad for your health and bad for your surroundings. It cannot be said too often: too many young people go prematurely deaf because of prolonged exposure to loud sounds. There cannot be any pleasure in excessively loud music: it hurts and, like skin cancer, the terrible effects do not immediately become noticeable.

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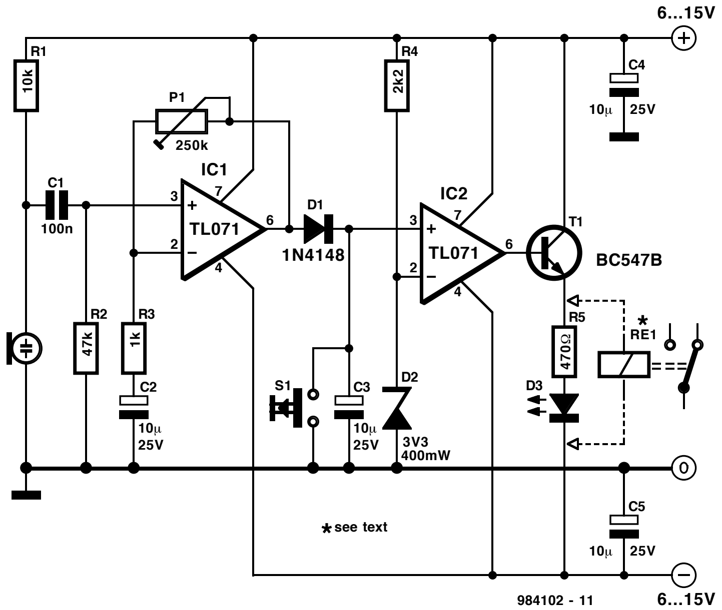

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Operational amplifier IC_2 functions as a comparator which likens the amplified signal with a reference voltage of 3.3 V. If the signal at the non-inverting input of the op amp exceeds the reference voltage, the output of IC_2 changes state (goes high), whereupon T_1 is switched on. When this happens, the relay is energized or the LED lights. The relay contacts may be used to operate a warning light or buzzer, or to switch the noise source off. In the latter case, C_2 prevents the circuit returning to its original state (which would cause the noise source to come on again). The capacitor is charged to the peak value of the signal. Owing to the presence of D_1, it cannot be discharged via the output of IC_1, but only, and very slowly, via the high-resistance input of IC_2.

The monitor is reset with S_1.

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Networks R_3-C_1 and R_4-C_2, in conjunction with Schmitt trigger IC_{1b} provide a thorough debouncing and at the same

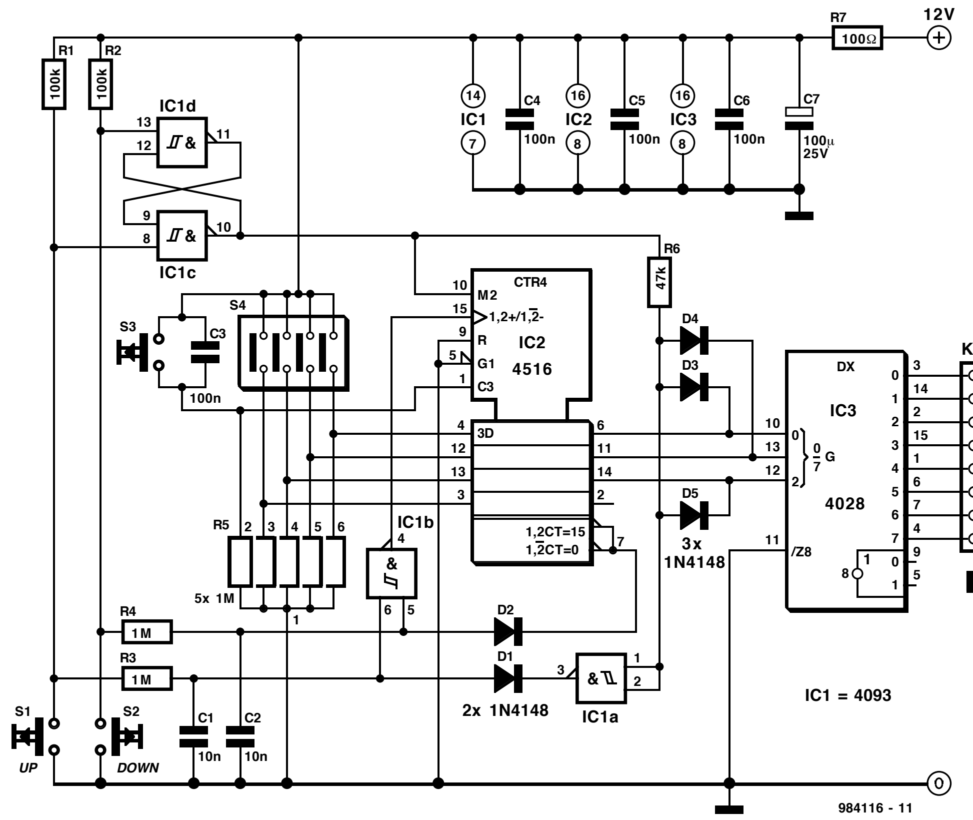
delay the clock pulse slightly. This delay guarantees that the clock pulse (output of IC_{1b}) arrives after the state of the counter has been defined.

To prevent the counter jumping from minimum to maximum or vice versa, the clock pulse is disabled in the outermost positions. In the minimum state, this is achieved simply by use of the carry-out terminal (pin 7) of the counter. In the maximum state, an auxiliary network, consisting of R_6, D_3, D_4, D_5/IC_{1a}, and D_1, was found necessary.

Diode D_2 ensures that pin 5 of IC_1 remains low when the minim state is reached until S_1 is pressed. The same is achieved by diode D_1 in regards of pin 6 of the IC when the maximum state is reached. Resistor R_6 serves to reset the clock disabling during down counting; when the down key is pressed, the output of IC_{1a} goes high again.

If an indication is desired of the actual state of the up/down drive, eight high-efficiency LEDs may be added at the output of IC_3 (anodes to the output, cathodes via a common 10 kΩ resistor to ground).

An indication whether



amplification or attenuation occurs may be given by an additional LED at the output of IC_{1c} or IC_{1d}.

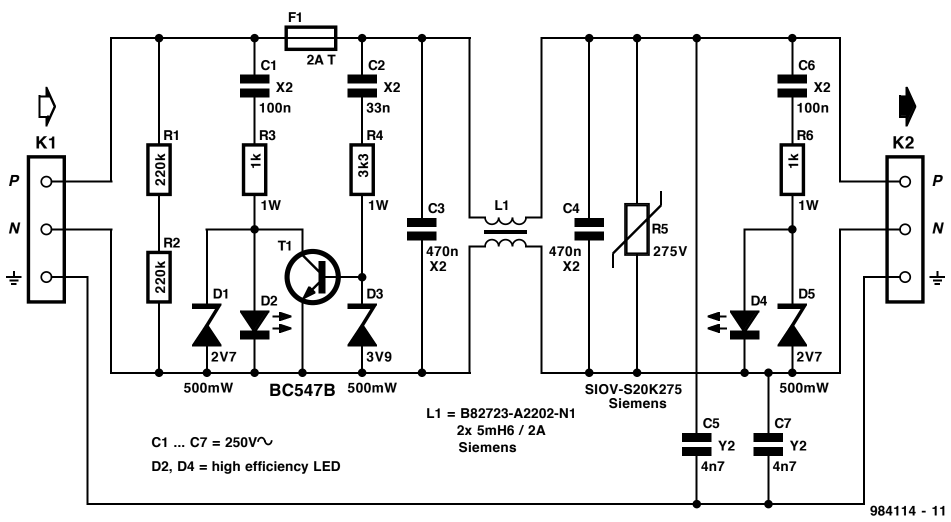
During quiescent operation, the circuit draws a current of 20 µA, which rises to about 140 µA when S_1 or S_2 is pressed. Network R_7-C_7 provides effective decoupling of the digital circuit from the analogue supply. [984116]

040 mains filter revisited

Design: I. Fietz

The mains filter described in 'PC Topics' (April 1998) could do with two useful additions: a fuse monitor and a voltage indicator. The filter proper consists of capacitors C_3–C_7 and inductor L_1. Its operation is described fully in the April 1998 article. The present article deals only with the additional features.

The fuse monitor indicates by the lighting of an LED that the fuse has blown. For this purpose, two capacitive potential dividers have been added: C_1-R_3-D_1 and C_2-R_4-D_3. When the fuse is intact, the potential at the base of T_1 is 3.9 V higher than that of the (N)eutral line. The transistor is on and short-circuits D_2. When the fuse blows, T_1 is off. Its collector potential, owing to D_1, is 2.7 V, so that D_2 lights. Note that this action is only possible when



the (L)ive line is positive w.r.t. the neutral line; this means that the LED flashes in rhythm with the mains frequency (since this is 50 Hz, it cannot be discerned by the human eye).

A second potential divider, C_6-R_4-D_4-D_5, at the output of the filter ensures that the LED (D_4) lights when the mains voltage is present at the output. [984114]

delay the clock pulse slightly. This delay guarantees that the clock pulse (output of IC_{1b}) arrives after the state of the counter has been defined.

To prevent the counter jumping from minimum to maximum or vice versa, the clock pulse is disabled in the outermost positions. In the minimum state, this is achieved simply by use of the carry-out terminal (pin 7) of the counter. In the maximum state, an auxiliary network, consisting of R_6, D_3, D_4, D_5/IC_{1a}, and D_1, was found necessary.

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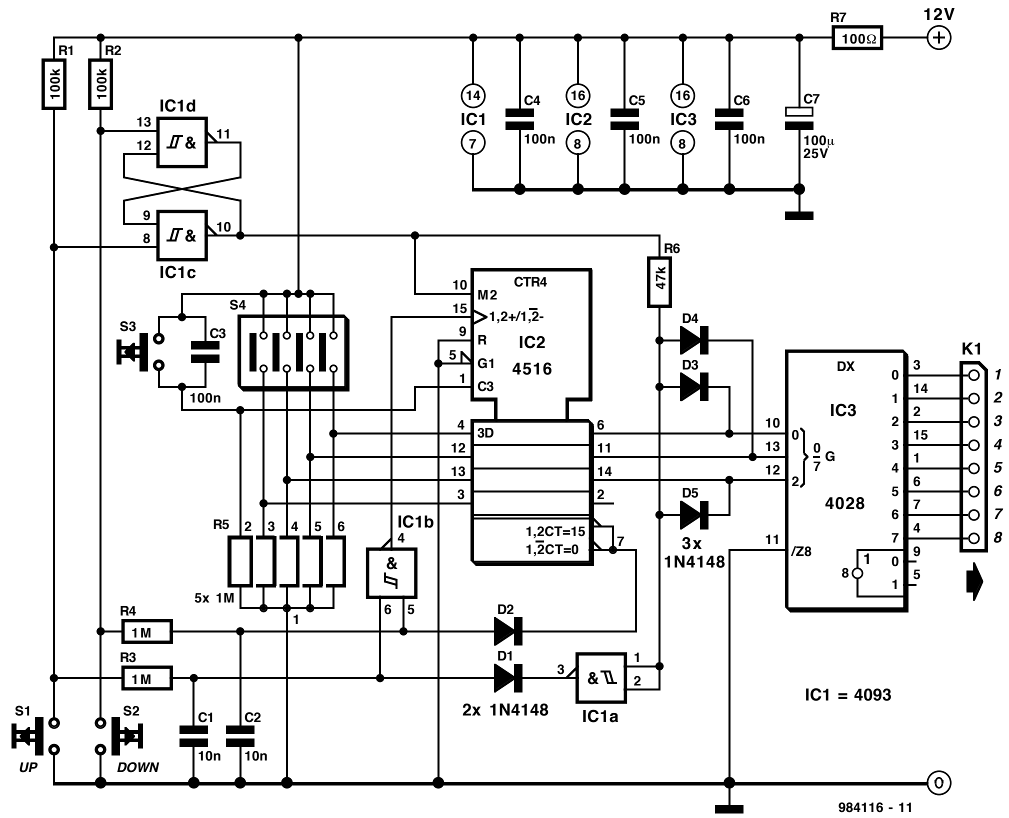

An indication whether amplification or attenuation occurs may be given by an additional LED at the output of IC_{1c} or IC_{1d}.

During quiescent operation, the circuit draws a current of 20 μA, which rises to about 140 μA when S_1 or S_2 is pressed. Network R_7-C_7 provides effective decoupling of the digital circuit from the analogue supply.

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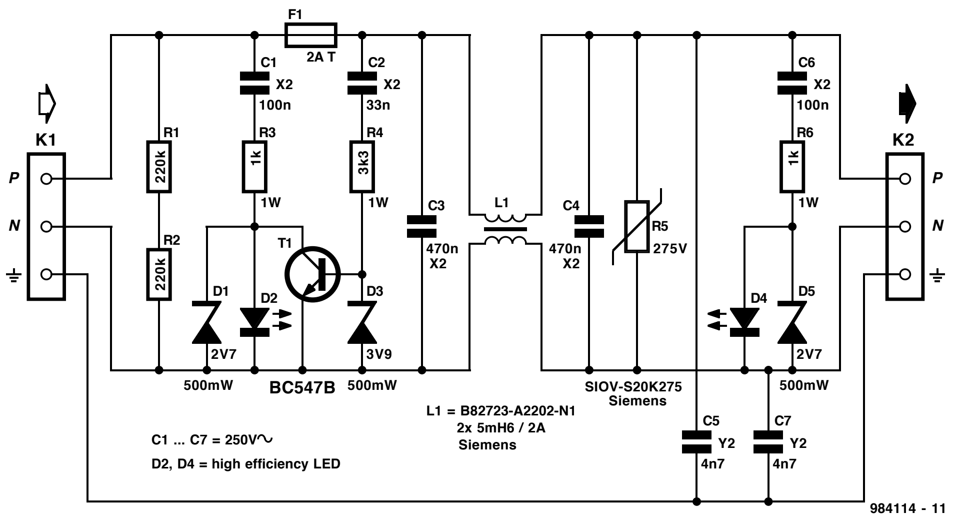


the (L)ive line is positive w.r.t. the neutral line; this means that the LED flashes in rhythm with the mains frequency (since this is 50 Hz, it cannot be discerned by the human eye).

A second potential divider, C_6-R_4-D_4-D_5, at the output of the filter ensures that the LED (D_4) lights when the mains voltage is present at the output.

[984114]

041 LED barometer

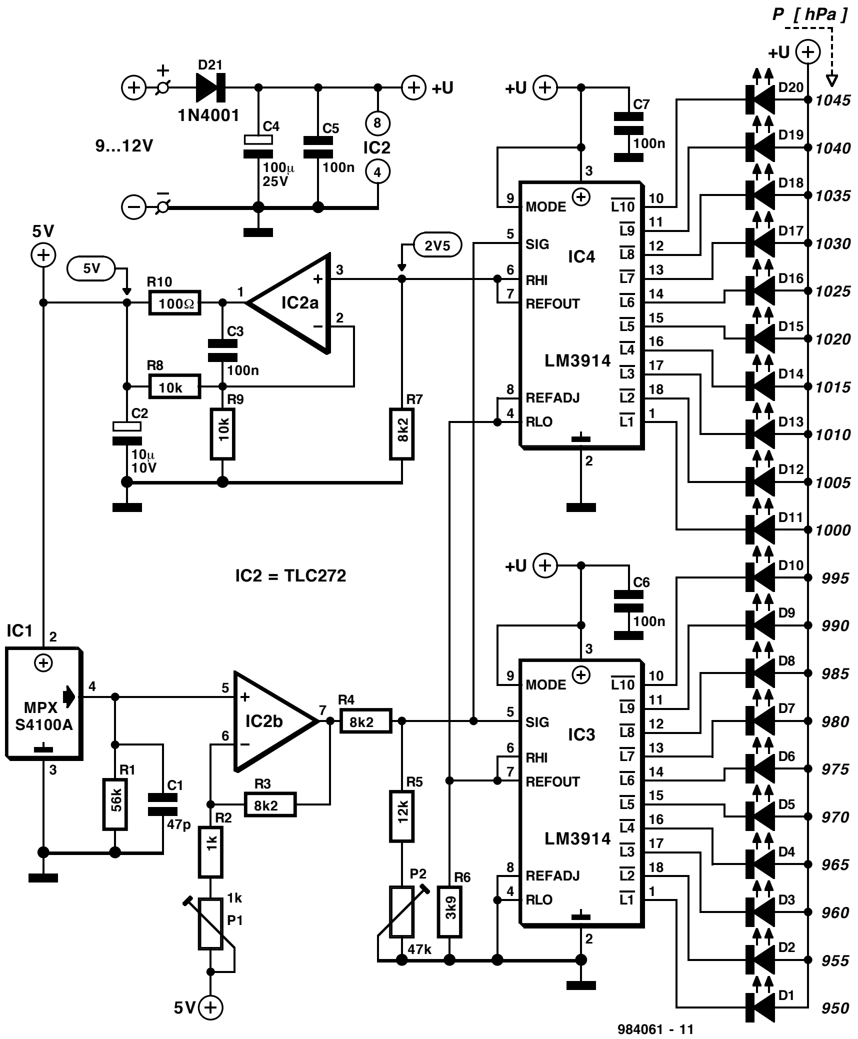
Being an all-solid-state design you don't have to tap on this barometer to get the latest air pressure reading!

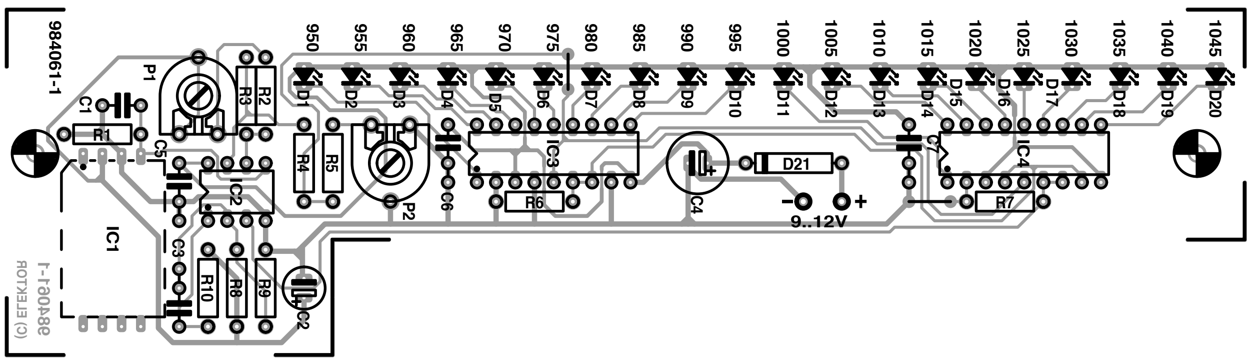
The main components in the circuit are air pressure transducer IC1, an MPXS4100A from Motorola, and two LM3914 LED bargraph drivers, IC3 and IC4. Both LED drivers generate a reference voltage of 1.25 V. The reference supply of IC3 is created with respect to ground. By connecting the RLO and REFADJ inputs of IC4 to the reference voltage created by IC3, the REFOUT pin of IC4 then supplies 2.5 V with respect to ground. In this way, the LED drivers are cascaded to give a scale of 20 LEDs each representing an air pressure increase of 5 hPa.

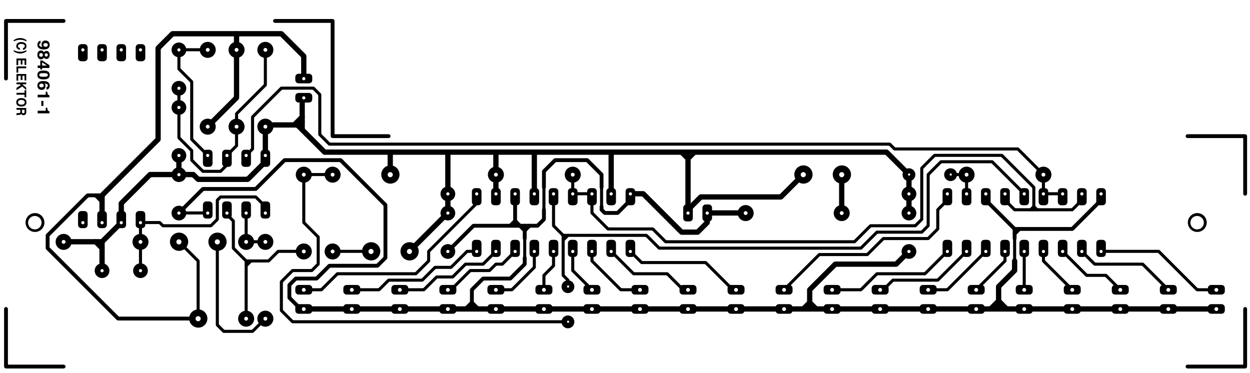
Because the output voltage of the pressure sensor follows any change in the supply voltage, a very stable 5-volt supply is required. This is provided by opamp IC2a which doubles the 2.5-V REFOUT potential from IC4. The sensor output voltage is expressed by the equation

$U_o = (0.001059*P - 0.1518)*5$ [V] (P [hPa])

Because we want an indication range of 945 hPa (all LEDs off) to 1045 hPa (all 20 LEDs on),







COMPONENTS LIST

Resistors:
R1 = 56kΩ
R2 = 1kΩ
R3,R4,R7 = 8kΩ2
R5 = 12kΩ
R6 = 3kΩ9
R8,R9 = 10kΩ
R10 = 100Ω
P1 = 1kΩ preset H
P2 = 47kΩ preset H

Capacitors:
C1 = 47pF ceramic
C2 = 10µF 10V radial
C3 = 100nF MKT (Siemens)
C5,C6,C7 = 100nF ceramic
C4 = 100µF 25V radial

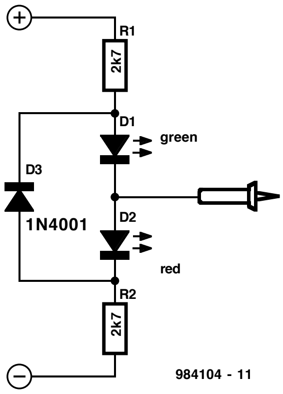
Semiconductors:
D1-D7 = LED, red, 3mm, high efficiency
D8-D13 = LED, yellow, 3mm, high efficiency
D14-D20 = LED, green, 3mm, high efficiency
D21 = 1N4001
IC1 = MPXS4100A (Motorola, Conrad)
IC2 = TLC272CP
IC3,IC4 = LM3914N

Miscellaneous:
PCB, order code 984061-1

the following minimum and maxim sensor output voltages can be calculated:
945 hPa ≅ 4.245 V = U_{low}
1045 hPa ≅ 4.774 V = U_{high}
The required gain, A, between the sensor output and the input of the readout is then
$A = U_{ref}/(U_{high} - U_{low}) = 2.5/(4.774 - 4.245) = 4.726$
In addition to this gain we also need a negative offset of 4.245 V, so that the output voltage is 0 V at an air pressure of 945 hPa. Components IC2b, P1, P2, R2, R3, R4 and R5 provide the gain and offset compensation. The 5-V reference voltage, IC2b, P1, R2 and R3 are the 'ingredients' to cancel the offset and at the same time provide a gain of 6.65 times. This gain may be reduced to the above mentioned value by adjusting P2.
The simplicity of the circuit is achieved at the cost of a fairly complex calibration procedure. Because preset P1 not only determines the offset but also the gain of IC2b, there is no way of avoiding multiple two-point calibration. In practice, you use an existing barometer or air-pressure information from your national or regional Met Office, and adjust the circuit several times at different air pressures. Alternatively, if you have access to a pressure vessel in which the pressure can be accurately controlled to 945 hPa, P2 is initially set to mid-travel, and P1 is adjusted until the output of IC2b supplies 0 V. Next, increase the pressure in the vessel to 1045 hPa, and adjust P2 until LED D20 just lights.
The printed circuit board of which the templates are shown here is available ready-made from the Publishers.

(984061-1, Bo)

042 12/24/48 V d.c. tester



Design: W. Mannertz
The present tester is intended primarily for testing the 24 V electrical circuits found on most pleasure craft. However, if the resistors are given different values, the circuit may, of course, be used for other voltage ranges. For 12 V, the value of the resistors should be 1.2 kΩ, and for 48 V, 4.7 kΩ.

The tester should be connected to the +ve and –ve voltage rails with test clips or crocodile clips, whereupon the test probe is placed on the point to be tested. When the potential at the point is positive, the red LED lights; if it is negative, the green one does.

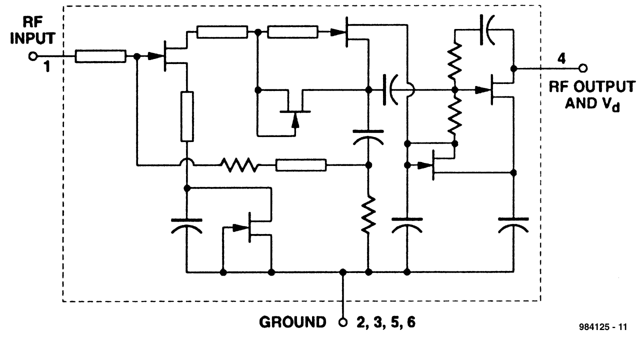
If the supply is not connected to earth, the tester may be used as ground-leak tester. In this situation, one of the LEDs lights when the test probe touches a point at earth potential and there is a leakage.

[984104]

043 0.5–6 GHz low-noise amplifier

By G. Kleine
The MGA-86563 from Hewlett-Packard is a three-stage, GaAs, MMIC (monolithic microwave integrated circuit) that offers low noise figure and excellent gain for applications from 0.5 GHz to 6 GHz. The device uses internal feedback to provide wideband gain and impedance matching. It is housed in an ultra-miniature SOT-363 package, which requires half the board space of the SOT-143.

The MGA-86563 may be used without impedance matching as a high performance 2 dB NF (noise figure) amplifier. Alternatively, with the addition of a simple shunt-series inductor at the input, the noise figure can be reduced to 1.6 dB at 2.4 GHz. For 1.5 GHz applications and above, the output is well matched to 50 Ω. Below 1.5 GHz, gain can be increased by using conjugate matching.



The input of the circuit in **Figure 1** is fixed tuned for a conjugate power match (maximum power transfer or minimum VSWR – voltage standing

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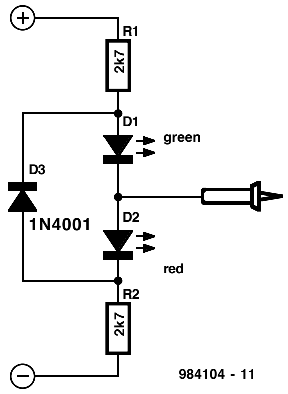
(984061-1, Bo)

042 12/24/48 V d.c. tester



Design: W. Mannertz

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The tester should be connected to the +ve and –ve voltage rails with test clips or crocodile clips, whereupon the test probe is placed on the point to be tested. When the potential at the point is positive, the red LED lights; if it is negative, the green one does.

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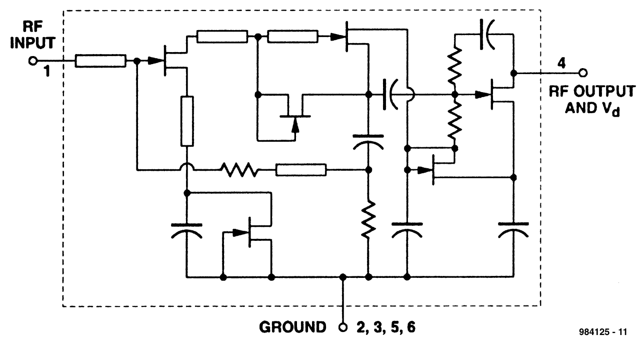

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(984061-1, Bo)

042

12/24/48 V d.c. tester

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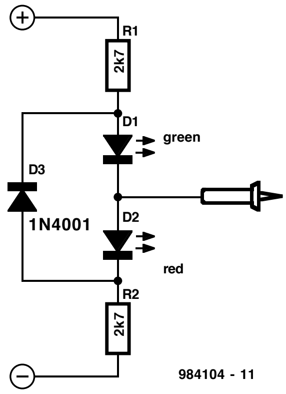

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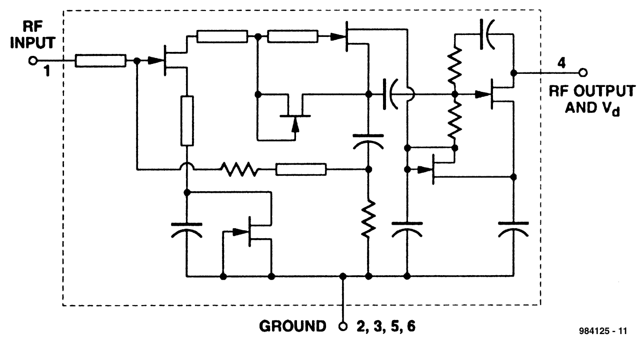
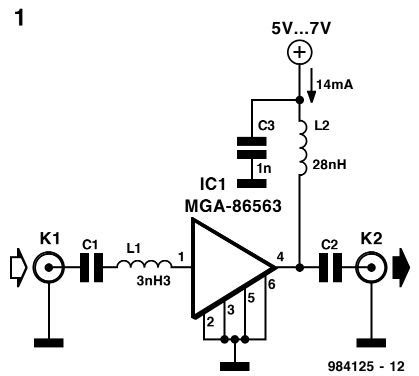
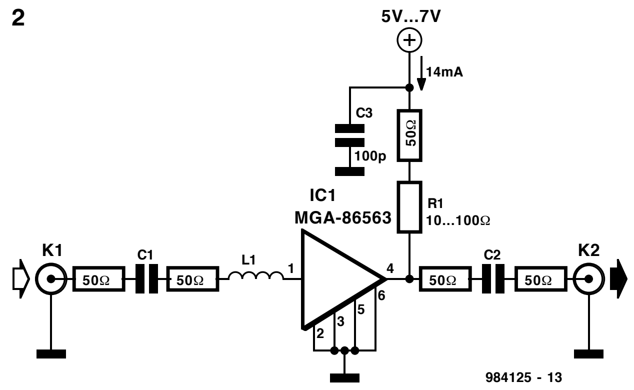
By G. Kleine
The MGA-86563 from Hewlett-Packard is a three-stage, GaAs, MMIC (monolithic microwave integrated circuit) that offers low noise figure and excellent gain for applications from 0.5 GHz to 6 GHz. The device uses internal feedback to provide wideband gain and impedance matching. It is housed in an ultra-miniature SOT-363 package, which requires half the board space of the SOT-143.

The MGA-86563 may be used without impedance matching as a high performance 2 dB NF (noise figure) amplifier. Alternatively, with the addition of a simple shunt-series inductor at the input, the noise figure can be reduced to 1.6 dB at 2.4 GHz. For 1.5 GHz applications and above, the output is well matched to 50 Ω. Below 1.5 GHz, gain can be increased by using conjugate matching.



The input of the circuit in **Figure 1** is fixed tuned for a conjugate power match (maximum power transfer or minimum VSWR – voltage standing





wave ratio) at 2 GHz.

The 3.3 nH inductor, L_1, in series with the input of the amplifier matches the input to 50 Ω at 2 GHz.

Inductor L_2 prevents any tendency to resonance over the operating range (2 GHz). When operation takes place at lower frequencies, its value may have to be increased accordingly.

A circuit for operation up to 6 GHz is shown in **Figure 2**. A 50 Ω microstripline with a series d.c. blocking capacitor, C_1, is used to feed r.f. to the MMIC. The input of the device is already partially matched for noise figure and gain to 50 Ω. The use of a simple input matching circuit, such as a series inductor, will minimize the amplifier noise figure.Since the impedance match for NF_0 (minimum noise figure) is very close to a conjugate power match, a low noise figure can be realized simultaneously with a low input VSWR.

DC power is applied to the MMIC through the same pin tat is shared with the r.f. output. A 50 Ω microstripline is used to connect the circuit to the following stage.

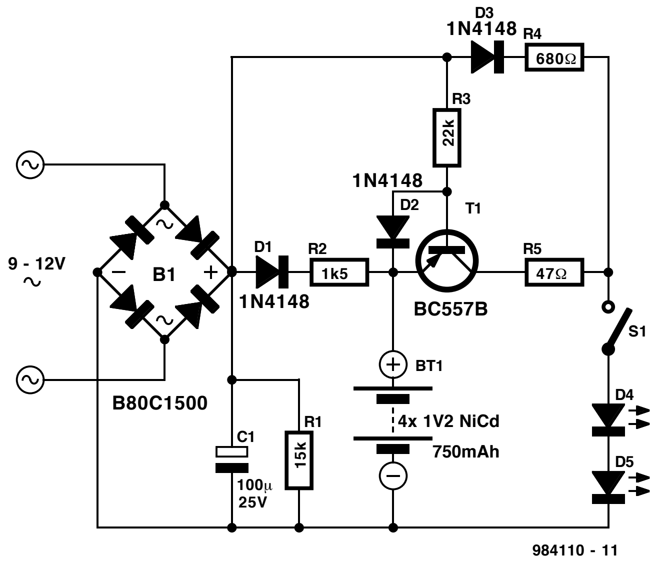
[984125]

044 LED lighting for consumer unit cupboard

Design H. Bonekamp

The consumer unit (or 'electricity meter') cupboard in some older houses is a badly lit place. If the bell transformer is also located in this cupboard, it may be used to provide emergency lighting by two high-current LEDs. These diodes are powered via a small circuit that switches over to four NiCd batteries when the mains fails.

The output voltage of the bell transformer is rectified by bridge B_1 and buffered by capacitor C_1. The batteries are charged continuously with a current of about 7.5 mA via diode D_1 and resistor R_2. The base of transistor T_1 is high via R_3, so that the transistor is cut off.



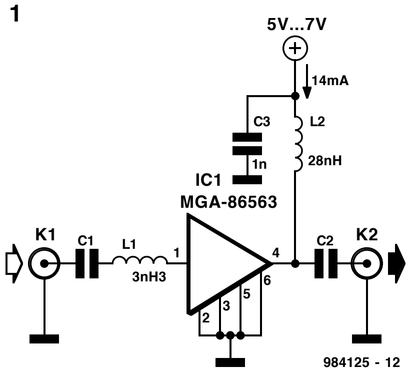
When the mains voltage fails, C_1 is discharged via R_1; when the potential across it has dropped to a given value, the battery voltage switches on T_1 via R_3 and R_1, provided switch S_1 is closed. When T_1 is on, a current of some 20 mA flows through diodes D_4 and D_5. The light from these LEDs is sufficient to enable the defect fuse or the tripped circuit breaker to be located.

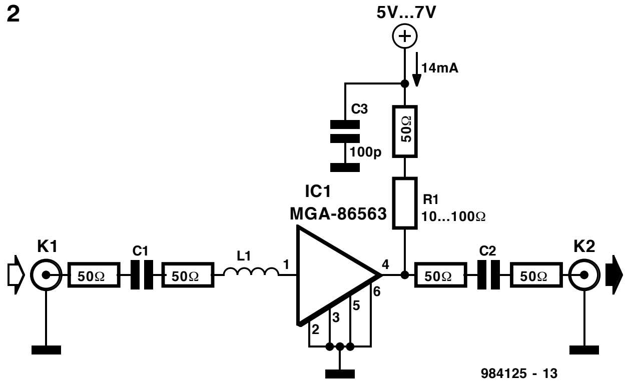
[984110]

045 ATU for 27-MHz CB radios

Design: G. Baars

This antenna tuning unit (ATU) enables half-wavelength (½λ) or longer wire antennas to be matched to the 50-Ω antenna input of 27-MHz Citizens' Band (CB) rigs. The ATU is useful in those cases where a wire antenna is less obtrusive than a roof-mounted 'vertical' or ground-plane. It is also great for 'improvised' antennas used by active CB users on camping sites and the like because it allows a length of wire to be used as a fairly effective antenna





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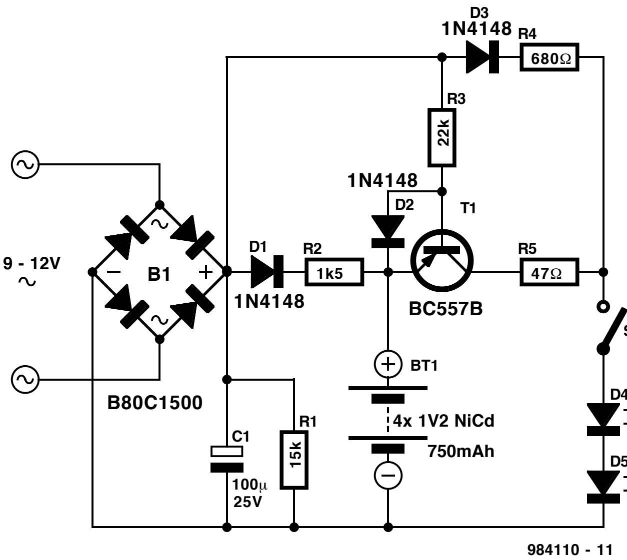

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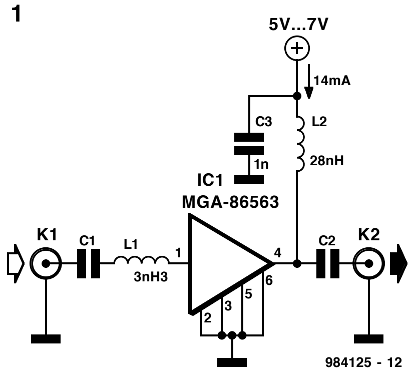

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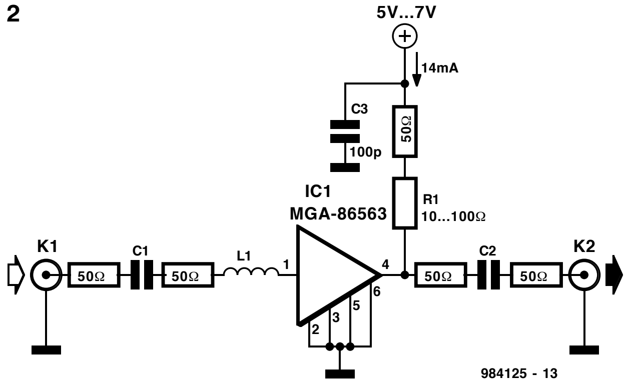

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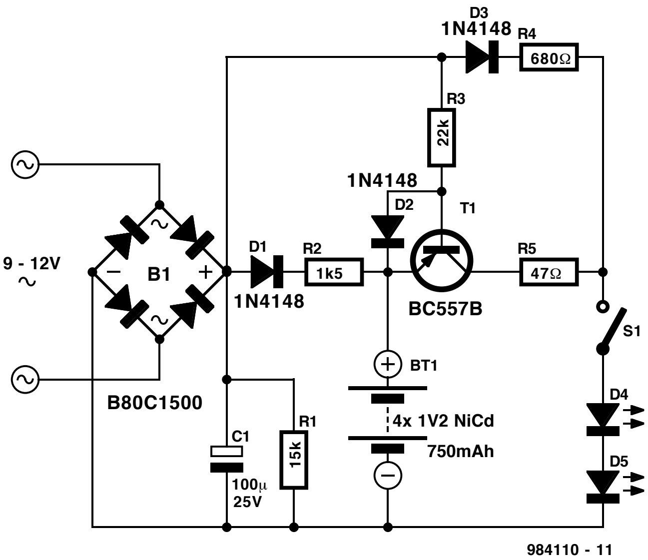

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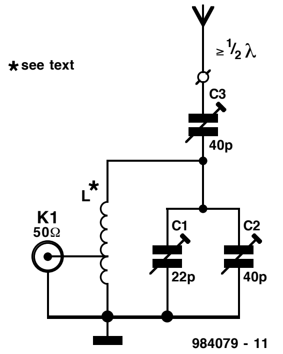
The coil in the circuit consists of 11 turns of silver-plated copper wire with a diameter of about 1 mm (SWG20). The internal diameter of the coil is 15 mm, and it is stretched to a length of about 4 cm. The tap for the antenna cable to the CB radio is made at about 2 turns from the cold (ground) side. Two trimmer capacitors are available for tuning the ATU. The smaller one, C1, for fine tuning, and the



larger one, C2, for coarse tuning.

The trimmers are adjusted with the aid of an in-line SWR (standing-wave ratio) meter which most CB enthusiasts will have, or should be able to obtain on loan. Select channel 20 on the CB rig and set C1 and C3 to mid-travel. Press the PTT button and adjust C2 for the best (that is, lowest) SWR reading. Next, alternately adjust C3 and C2 until you get as close as possible to a 1:1 SWR reading. C1 may then be tweaked for an even better value. No need to re-adjust the ATU until another antenna is used. In case the length of the wire antenna is exactly ½ λ (5.5 metres), then C3 is set to maximum capacitance.

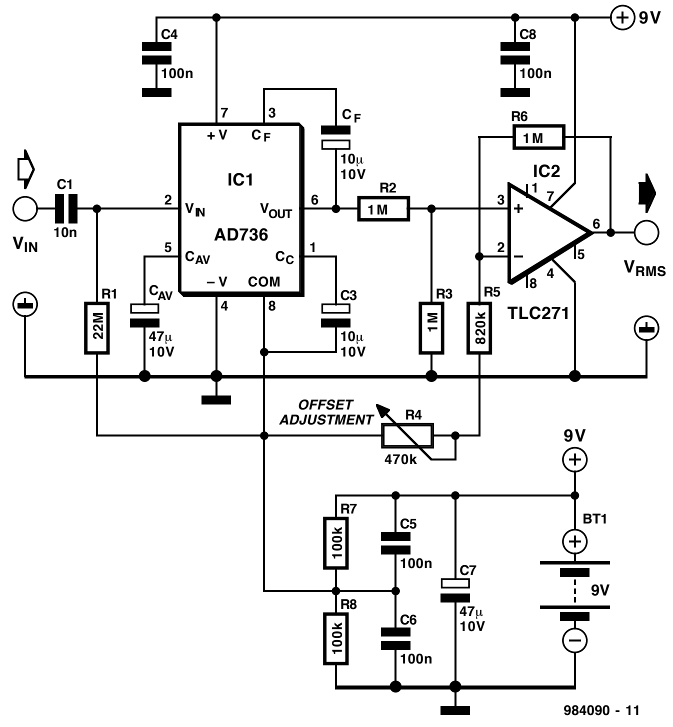
Although the ATU is designed for half-wavelength or longer antennas, it may also be used for physically shorter antennas. For example, if antenna has a physical length of only 3 metres, then the remaining 2.5 metres has to be wound on a length of PVC tubing. This creates a so-called BLC (base-loaded coil) electrically shortened antenna. In practice, the added coil can be made somewhat shorter than the theoretical value, so the actual length is best determined by trial and error. Finally, the ATU has to be built in an all-metal case to prevent unwanted radiation. The trimmers are than accessed through small holes. The connection to the CB radio is best made using an SO239 ('UHF') or BNC style socket on the ATU box and a short 50-Ω coax cable with matching plugs.

(984079-1; LL)

046 single-supply operation of the AD736

An Analog Devices Application

In dual-supply operation, the output (pin 6) of the AD736 is at 0 V, that is, halfway between the supply lines. But in single-supply operation, the output is at 1/2 V_{CC}. By adding a single-supply op amp as a differential amplifier, however, a true '0 V out for o V' single-supply circuit with a ground-referenced output as shown in the diagram. For this circuit, V_{RMS} = 0 V when V_{IN} = 0 V, and V_{RMS} = 200 mV d.c. when V_{IN} = 200 mV r.m.s.



In the circuit, a single 9 V positive supply powers the AD736. Resistors R_7=R_8= 100 kΩ form a potential divider across the 9 V battery that establishes a local 'ground' rail at 1/2 V_{CC}, or 4.5 V. The AD736's 'common' pin, its 22 MΩ input bias resistor, and the inverting input of U_2 (via R_4 and R_5) are all connected to this rail. The quiescent output voltage of the AD736, which is referenced to its 'common' pin, is 4.5 V.

A single-supply op amp, IC_2, is arranged as a unity-gain differential amplifier. Large value feedback resistors, R_2–R_5, are used to minimize loading of the 4.5 V rail. The op amp amplifies the difference between local ground at 4.5 V and the output of the AD736, which is also at 4.5 V for 0 V r.m.s. input. As the r.m.s. input to the AD736 increases from 0 mV to 200 mV, the AD736's output increases from 4.5 V to 4.7 V. The output of op amp IC_2 is the difference between the AD736's output and 4.5 V, or 0 mV to 200 mV d.c.

The remainder of the circuit works as follows. The AD736's output is a.c. coupled; R_1 provides a path for the BiFET op amp's input bias current (typically 1 pA) to flow. The offset voltage caused by the bias current flowing through R_1 is negligible.

Capacitor C_3 between pins 1 and 8 of IC_1 provides a low frequency cutoff of 2 Hz. Other cutoff frequencies, f, can be calculated from

$$f=1/2\pi RC$$

where f is in Hz, R is in ohms, and C is in farads.

Optional capacitor C_F, in parallel with an 8 kΩ feedback resistor, fixed internally by IC_1, forms a single-pole low-pass filter with a 2 Hz cutoff frequency. The value of C_F in farads is given by

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[984090]

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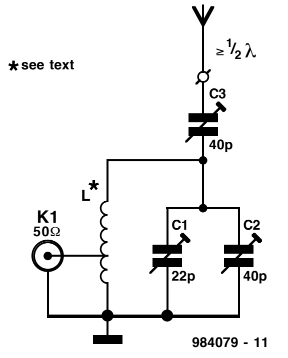

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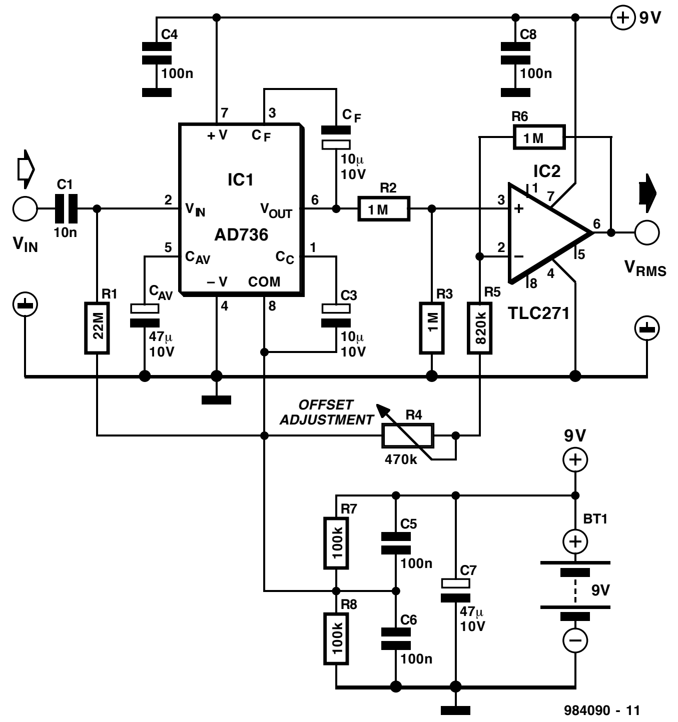

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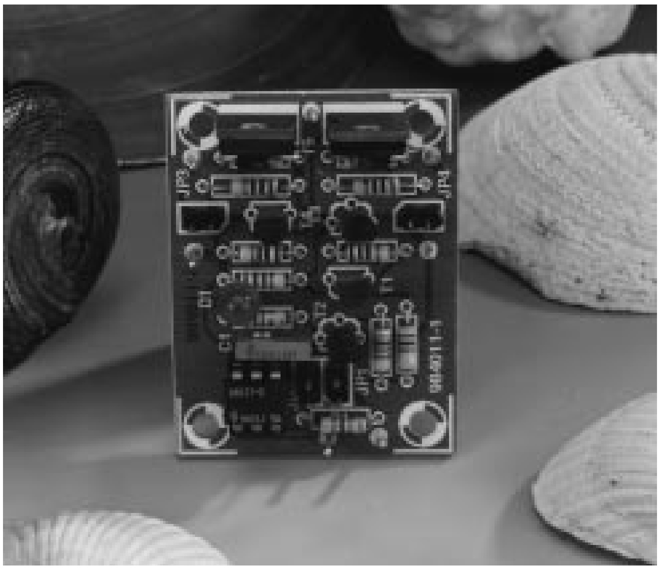
[984090]

047 digital output with sink/source driver



Design: R. Veltkamp

When it comes to using a PC or a microprocessor system to control 'real-world' loads like lamps, relays, and motors, there are basically two camps: programmers and hardware specialists. The combined species seems to be rare! Anyway, this article is aimed at the latter group. The circuit diagram shows a one-channel power driver with an (optional) electrically isolated input and a power output capable of sinking as well as sourcing current.

If galvanic isolation is not required at the input, **omit the optocoupler and fit the two jumpers.** In that case, the circuit is driven by a TTL-compatible logic signal. In case the optocoupler is used, the driver responds to a current-loop signal with strength of between 10 mA and 20 mA.

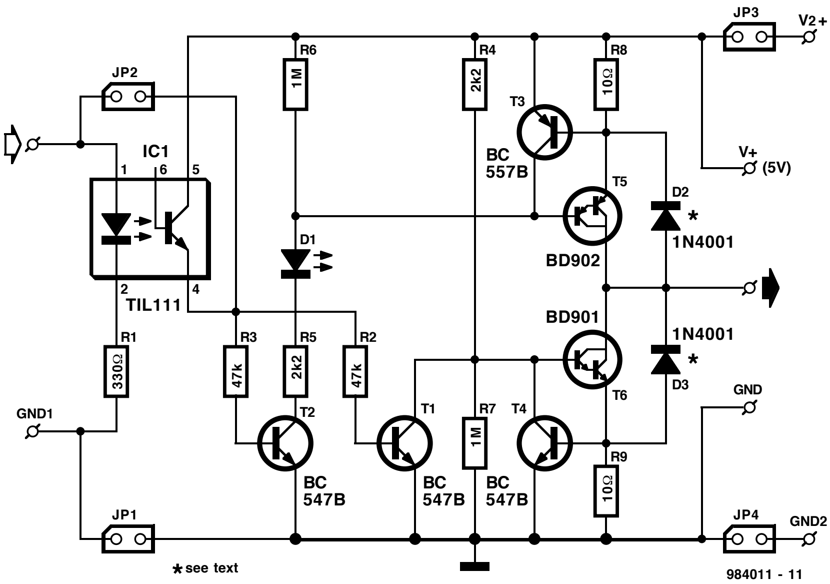
An LED, D1, is inserted in the collector line of amplifier stage T2 to provide a visible 'channel active' indication. The symmetrical sink/source power driver consists of a pair of complementary BD901/902 Darlington transistors with associated current limiting resistor, R8-R9. Resistor R8 determines the maximum source current, and R9, the maximum sink current. Both currents are calculated from $I = 0.65\ \text{V}/R$. The driver board itself, by the way, has a current consumption of just a few milli-amps.

Diodes D2 and D3 are only required if inductive loads like relay coils are controlled, and different Darlington pairs like the BD911/912 are employed. As opposed to the BD901/902, the BD911/912 complementary pair does not have internal anti-surge diodes across the collector-emitter path. As a matter of course, the Darlington transistors have to be cooled depending on the currents they sink or source.

Jumpers JP3 and JP4 have to be fitted if the controlled load(s) are not already connected to a supply line. If the jumpers are fitted, then the load current will be drawn from the driver board.

This driver is pretty fast: it can handle switching frequencies up to about 3 kHz without problems if the TIL111 optocoupler is used (as suggested in the circuit diagram). Higher frequencies may undoubtedly be achieved if a faster optocoupler is employed. If you need to control more channels than just one (say, four), you just build as many driver boards as you need. Unfortunately, the printed circuit board whose artwork is shown here is not available ready-made from the Publishers.

(984011-1)



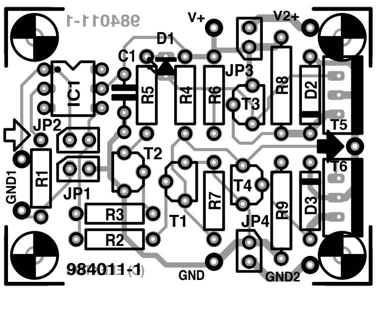
COMPONENTS LIST

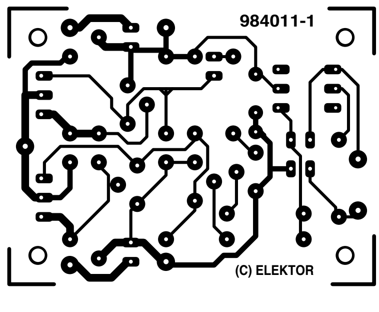
Resistors:
R1 = 330Ω
R2,R3 = 47kΩ
R4,R5 = 2kΩ2
R6,R7 = 1MΩ

Capacitor:
C1 = 100nF

Semiconductors:
D1 = LED
D2,D3 = 1N4001 (optional, see text)
T1,T2,T4 = BC547B
T3 = BC557B
T5 = BD902 or BD912 (see text)
T6 = BD901 or BD911 (see text)
IC1 = TIL111 or 4N35 or CNY17-2

Miscellaneous:
JP1-JP4 = 2-pin SIL header with jumper
Heatsinks for T5/T6, as required





Converting parallel information into serial format is a function often performed by complex logic or other dedicated hardware. The approach shown in this article proves that there is a good alternative based on a handful of discrete parts. The reverse process, serial to parallel conversion, is of similar simplicity.

Design by G. Visschers

parallel-to-serial converter

and the other way around, with just four ICs

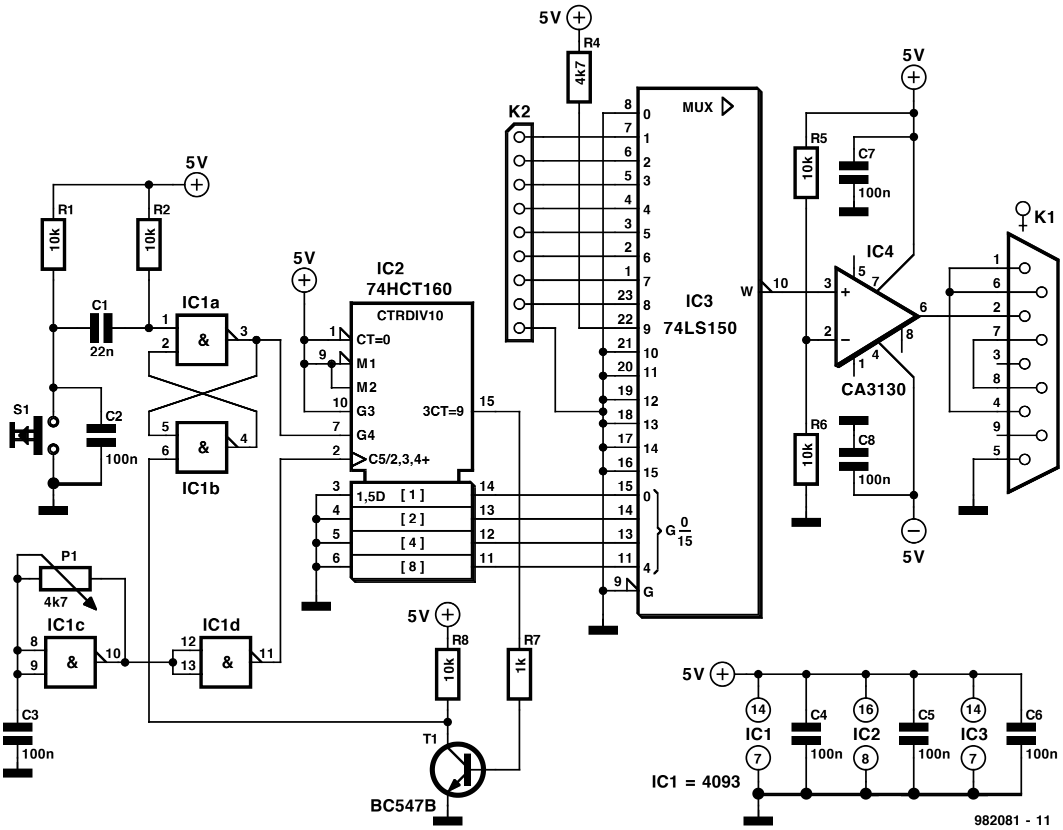


Figure 1. Circuit diagram of the parallel-to-serial converter. Using just four simple TTL ICs, parallel data is converted into serial format at a bit rate of 9600 per second.

Let's start by mentioning that the present circuit is *not* a state-of-the-art superfast parallel-to-serial converter. However, it may be just the right circuit or sub-circuit if you were looking for a simple and rather clever solution.

This article describes how a couple of common-or-garden TTL ICs are employed to convert parallel data into serial format, using a hardware-defined baudrate (speed) of 9600 bits per second. The transmission format is very common: 8 databits, 1 stop bit and no parity bit — in practice this setting will be fine for all but the most exotic cases.

The circuit diagram of the bidirectional converter may be found in **Figure 1**. The heart of the circuit is IC3, a type 74LS150. This IC is responsible for the actual parallel-to-serial conversion. Eight of the 16 inputs of this multiplexer are connected to K2, the parallel input of the converter. Input E0 of the IC represents the start bit, and E1 through E8, the databits. Input E9, finally, is used to generate the stop bit.

The inputs of the 74LS150 are scanned by a 74LS160 BCD counter. Every time S1 is pressed, the 74LS160

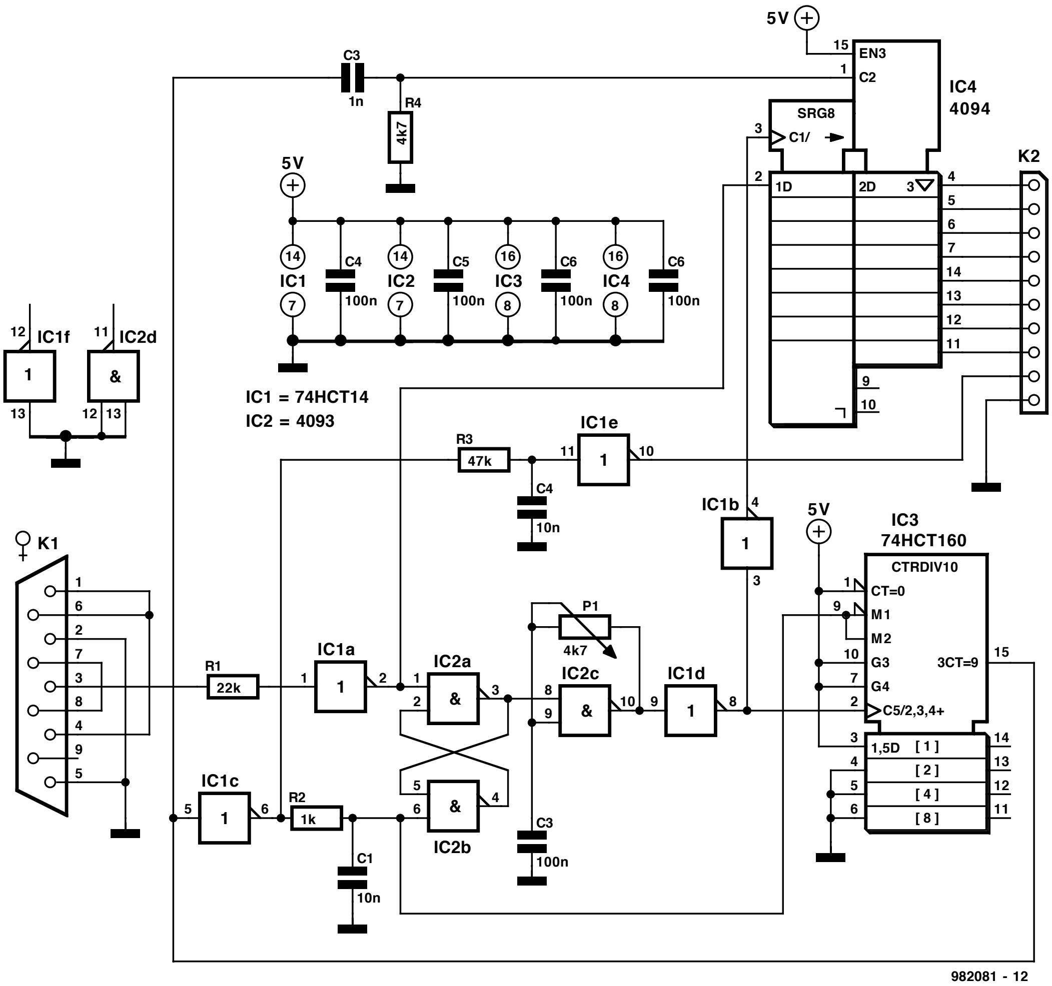


Figure 2. The other way around is no problem either. This simple circuit converts serial information into parallel.

counts up from 0 to 9 and so applies the associated BCD code to the A-D inputs of IC3. Because of the action of capacitor C2, this also happens when the supply voltage is first switched on. One byte is then converted and transmitted.
If the circuit is used as a sub-assembly in a larger unit, components R1, R2, C1 and S1 may be omitted. The input of IC1a is then connected to the driving circuit.

The operation of the rest of the circuit is should be easy to understand because a really simple counter circuit is used. The flip-flop built around IC1a and IC1b may be set with S1, and reset by the BCD counter at the end of the serial code transmission. Once the flip-flop is set, the BCD counter is enabled, and each clock pulse then causes a new bit to be placed on the serial output line. A simple RC clock generator is built around buffers IC1c and IC1d. It is dimensioned such that a bit rate of 9600 per second is achieved. The exact bit rate is set with the aid of preset P1. For lower bit rates, capacitor C3 has to be increased accordingly. For a bit rate of 2400, for example, a 1-μF capacitor is a good choice. In this way, the circuit can be 'tweaked' for nearly every bit rate you may want to use — all you have to do is modify the oscillator as required.

RS232, step by step

The only missing element is the line interface. For this purpose we call in the help of a symmetrically powered CA3130 opamp. This opamp, configured as a comparator, converts the TTL signal received from the multiplexer into a serial signal which toggles between +5 V and –5 V. In this way, we strive to meet the electrical requirements defined for the RS232 interface. Only one line of the serial interface is actually used: TxD (transmitted data). On the connector, the handshaking lines RTS (request to send) and CTS (clear to send) are interconnected, as well as the triplet DSR (data set ready), DCD (data carrier detect) and DTR (data terminal ready). In this way, the RS232 port is 'enabled', and capable of serial communication.
A simple power supply with +5 V and –5 V outputs is sufficient for this project.

The other way around

So far we've only discussed the conversion from parallel to serial format. The reverse process, serial to parallel, is also implemented in a very simple manner.
The relevant circuit is shown in **Figure 2**. Connector K1 is connected to the serial port on the PC. The connector has a number of links to make sure hardware handshaking is disabled. By way of inverter IC1a, the serial signal (TxD) arrives at the D (data) input of IC4, a binary counter.
IC2a and IC2b together form a SR (set-reset) flip-flop. In conjunction with an oscillator built around IC1d and IC2c, and a counter type 74HCT160, they act as the heart of the circuit, that is, as far as timing is concerned.
When data is received at the serial input, it is converted to TTL level (R1, IC1a), and then applied to the input of the SR bistable. This bistable starts the oscillator whereupon oscillator (clock)

Now hear this

The RS232 port in an MS-DOS computer has to be set up (or 'configured') to make sure it is in the right mode (not mood) for reception of data. The command to use is

```
mode com2:9600,n,8,1
```

Next, a simple program may be employed to read data. The following QBASIC program shows the way

```
start:
IF INP(&H2FD)>96 THEN PRINT INP(&H2F8)
GOTO start:
```

In this example, 2F8H is the address of the COM port, and 2FDH that of the status register which may be read to see if there is new data. If another COM port is used, then these addresses have to be changed accordingly.

The serial to parallel converter is also simple to test. The program printed below continually sends the number sequence 00 through 25r to the parallel port:

```
FOR X = 0 TO 255
OUT &H2F8,X
FOR Y = 1 TO 1000: NEXT Y
NEXT X
```

The line

```
FOR Y = 1 TO 1000: NEXT Y
```

has been added to slow down the resulting datastream if a fast computer is being used. Thanks to this setting, you can even see that a new value is transmitted on every loop iteration.
Both programs make use of communications port COM2. If you want to use another COM port, then the port address has to be modified accordingly.

pulses are transmitted to the clock input of the counter (IC3) and the shift register (IC4). Eventually, the shift register shifts out the bits one by one.
One the oscillator has produced nine clock pulses, the SR flip-flop is reset again via the signal at the RCO output of IC3 (which is inverted by IC1c). The RC network consisting of R2 and C1 lengthens the last pulse. If that was not done, there would be a fair chance of the shift register missing the last pulse, mainly because IC4 (a CMOS IC) is considerably slower than IC3 (a HCT IC). The RC network consisting of C3 and R4 supplies the strobe pulse which enables data to be read into the output register of IC4. This signal is supplied by the RC0 output on IC3. Data will remain present on the output until the next strobe pulse appears. A peripheral device connected to the parallel port is furnished with a strobe pulse via R3 and C4.
When properly dimensioned the circuit is suitable for serial signals travelling at a rate of 9600 bits per second. By increasing C2 to 470 nF, the bit rate may be dropped to 2400. Preset P1 allows the bit rate to be accurately adjusted. Unfortunately, adjusting the clock oscillator is not as easy as we would like it to be. The problem is that the oscillator is only active when serial data is being received. For the purpose of aligning the circuit, this 'problem' may be solved by temporarily connecting pin 8 of IC2c to the +5 V line (i.e., temporarily break the link between pin 3 of IC2a and pin 8 of IC2c). Next, the clock frequency may be measured at pin 8 of IC1d (2400 Hz for 2400 bits/s, or 9600 Hz for 9600 bits/s).

(982081-1)

Most soundcards have a 'line' input and one for an electret (condenser) microphone. To be able to connect an inductive tape-recorder head or a dynamic microphone, an add-on preamplifier is needed.

Design by M. Wenzel

preamplifier for soundcard

for inductive pick-up elements and dynamic microphones

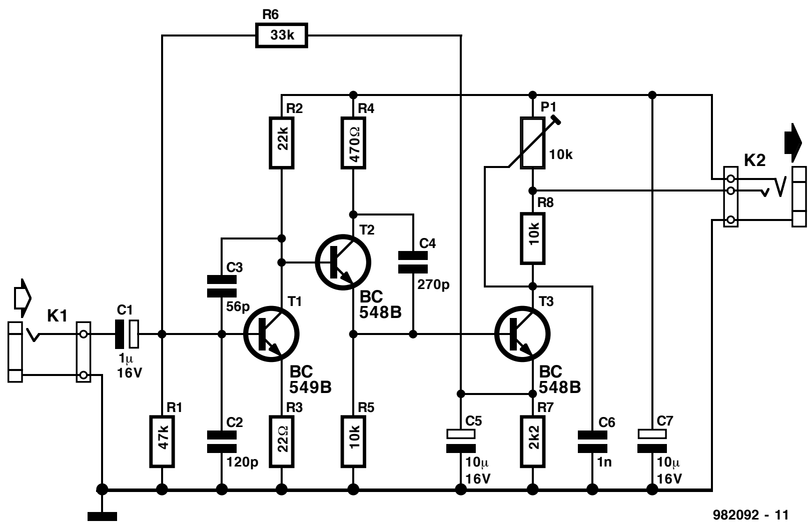


Figure 1. Circuit diagram of the preamplifier for PC soundcards.

Even in this day and age of integrated microelectronics, a transistorised circuit built from discrete part has a right of existence. The preamplifier described in this short article goes to show that it will be some time before discrete transistors are part of the silicon heritage. The preamplifier is suitable for use with a soundcard or the microphone input of a modem.

As you will probably know, most soundcards have input sockets for signals at line level (stereo), as well as one for a (mono) electret microphone. For the applications we have in mind, connecting-up an inductive pick-up element or a dynamic microphone, both inputs are in principle suitable, provided the source signal is amplified as required.

The author eventually chose the microphone input on the soundcard. Firstly, because the line inputs are usually occupied, and secondly, because the bias voltage supplied by the microphone input eliminates a separate power supply for the preamplifier.

The microphone input of a soundcard will typically consist of a 3.5-mm jack socket in stereo version, although only one channel is available. The free contact is used by the soundcard to supply a bias voltage to the mono electret microphone. This voltage is accepted with thanks by the present preamplifier, and conveniently obviates an external (mains adaptor) power supply.

A classic design

In true transistor-design fashion, the preamplifier consists of three stages. Capacitor C1 decouples the signal received from the microphone or pick-up element, and feeds it to the input of the first stage, a transistor in emitter configuration, biased to provide a current amplification of about 300 times. Together with the source impedance of the microphone or pick-up element, capacitors C2 and C3 form a low-pass filter which lightly reduces the bandwidth. In addition, the output low-pass, R2-C3, reduces the dynamic collector resistance at higher frequencies. In this way, the filter reduces the gain in the higher part of the frequency spectrum and so helps to eliminate any oscillation tendencies.

The first, high-gain, stage is terminated by T2. Unlike T1, this transistor does not add to the overall gain, because the output signal is taken from the emitter (common-collector circuit). T2 thus acts as an impedance converter, with C4 reducing any tendency to oscillation.

The output stage around T3 is a common-emitter circuit again. In it, preset P1 determines the voltage amplification. T3 is biased by means of a direct-current feedback circuit based on components R7 and C5. To this is added an 'overruling' dc feedback path back to the input transistor, via R6. This measure guarantees good dc stability in the preamplifier.

The circuit is small enough to be built on a piece of veroboard or stripboard, and yet remain reasonably compact. To prevent interference from external sources, the completed board should be mounted in a properly screened (metal) enclosure, with the connections to the input source and the sound card made in screened cable.

The preamplifier provides a frequency-linear response. In case the source signal is marked by frequency correction (e.g., RIAA), then a matching linearization circuit should be used if the relevant signals are used by the computer.

(982092-1)

Today's top-line PCs are fully equipped and ready to function as true multimedia machines. Many PC users extend their machines with a TV card to be able to view TV pictures on the PC monitor, and digitize pictures for further processing. Such a combination of a PC and a TV card offers even more possibilities, however, such as reception and storage of Teletext pages. Recent developments in this field also allow Internet pages to be received from cable TV networks. This is a free service, as no telephone company or Internet Service Provider is involved. In Europe, the German TV station 'ZDF' leads the way with Intercast.

By our editorial staff

Intercast

free Internet pages via TV signals

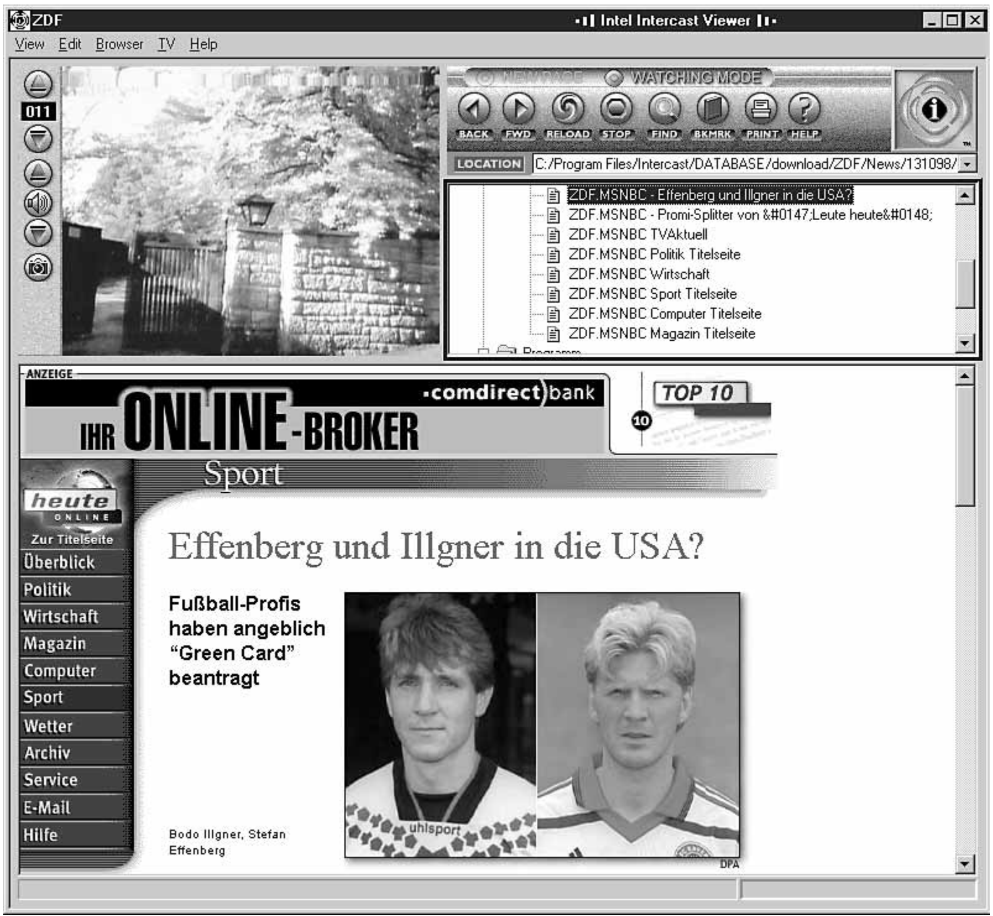


Figure 1. In the top left-hand corner, the Intercast Viewer shows the TV picture. This is flanked by an overview of received Intercast pages. The lower part of the Window shows the selected page.

Functionally, the TV set and the PC seem to be in a merging process. Using special Internet set-top boxes it is possible to surf the Net using a TV set, while the computer is being transformed into a TV set by the addition of a TV tuner insertion card.

Traditionally, information suppliers ('broadcasters') have been busy developing several methods to enable extra information to be conveyed via existing communication channels. As far as our regular TV channels are concerned, most of us are used to having Teletext, a facility that allows broadcasters to transmit lots of (text) information that reaches us by way of a couple of (normally invisible) TV lines in the vertical blanking interval.

Recently, a number software and hardware manufacturers came up with a method of information supply which is more up to date than the Teletext system. Using a number of TV lines in the vertical blanking interval, data is transmitted that allows HTML pages to be built in a suitable receiver. In this way, an Internet-like display is created. The advantage of these systems is that anyone with a TV antenna or a cable-TV connection is in a position to receive these signals, and that the service is normally free of charge. Note, however, that this is basically 'one-way' traffic. For a 'real' Internet link, you will need two-way communication in the form of a telephone connection or a two-way cable-TV link, and that is not the case with the recent developments. The inherent disadvantage is, therefore, that the transmitted information is always the same, i.e., it can not be adapted to reflect the requirements of individual users.

In spite of this handicap, there seems to be a lot of interest in broadcasting Internet-like pages by way of regular TV channels. After all, you only have to

switch on your PC, start the right program, and the information is delivered to your hard disk, automatically and free of charge. After some time, you can browse around for information you find useful. Meanwhile, two different systems have emerged that seem to have a fair chance of survival: Intercast (developed by Intel) and WaveTop (from US based WavePhore). The latter system seems to be used in the USA only, and will not be discussed in detail here.
Intercast technology goes back to 1995, and NBC, CNN and MTV have been transmitting Intercast pages for quite some time. In Europe, the German broadcaster ZDF started to put Intercast transmissions on the air in 1997. This initiative was followed by a another German TV station, DSF, in August 1998. If and when other broadcasters jump the bandwagon remains to be seen.

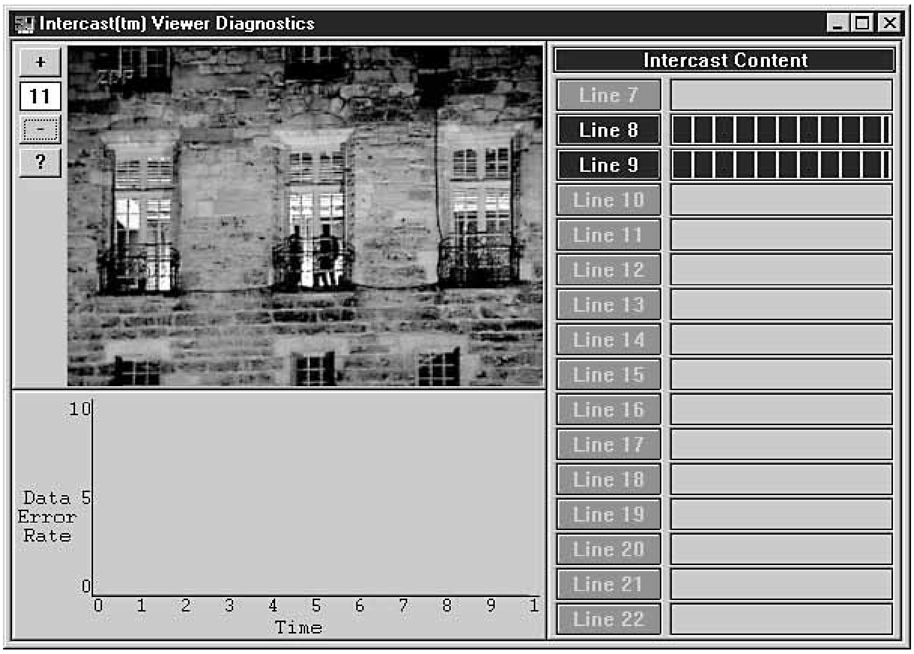


Figure 2. This utility program that comes with the Viewer may be used to gauge the reception quality of the Intercast service.

How does Intercast work?

The name ‘Intercast’ is a contraction of the words ‘Internet’ and ‘broadcast’, and so indicates the ‘merging’ of TV pictures and Internet pages. At the transmitter side, the HTML pages to be broadcast are first built and then added to the TV signal. The HTML information is conveyed in a number of free lines in the vertical blanking, which in the PAL TV system lasts 22.5 lines. Although up to 10 of these TV lines may be reserved for Intercast, the actual number will depend on any services which already be available in different countries (Teletext, VIT lines, etc.) Each line in the vertical-blanking period allows up to 10 kbits of data to be transmitted. A (maximum) capacity of 10 lines therefore allows a data rate of up to 100 kbits/s to be achieved.
At the receiver side, we require a PC and TV tuner card which is compatible with the Intel Intercast program, for example, a card from Hauppage or Miro (Pinnacle). After starting the program you are presented with a three-part window (**Figure 1**). In the top left-hand corner you see the TV picture, flanked to the right by an overview of received pages. Below these two sub-windows is an area showing a page selected from the overview, or the currently received page.
The user creates a cache of, say, 25 Mbytes on the hard disk to enable all received data to be stored. As long as the computer is switched on with the Intercast program running, a continuous flow of data will be received and stored. Important pages, for example, news headlines, are updated at regular intervals, so that it is not necessary to leave the PC switched on all day to ‘catch’ a certain page. Special overviews are transmitted showing transmission schedules for specific subjects.
The received HTML pages allow hyperlinks to be entered that point the way to Internet sires. If the user clicks on one of these hyperlinks, the internet browser is automatically launched, and a connection to the Internet is established. Currently, this works with Microsoft Internet Explorer only.
Windows 98 comes with a module called WebTV, which has to be installed before you can receive Intercast. But that’s not all, because you also have to install (manually!) the program iit22020.exe which may be found in the folder drivers/webtv/intercst on the Windows 98 CD-ROM. It should be noted, however, that this version is not suitable for PAL TV signals. When we write this, version 2.0 of Intercast has just been released by Intel, and it is available for downloading from www.intercast.de. This version is suitable for Windows 95 as well as Windows 98 (older versions are only compatible with Windows 95).

WaveTop

In the US of A, WaveTop pages are being broadcast by a large number of TV stations. WaveTop resembles Intercast in that vertical blanking lines are employed to convey data. The essential difference between the two systems is that Intercast supplies information to go with the programs of a specific transmitter. Consequently, the broadcaster determines which subjects appear on the Intercast pages.

Intercast on the Internet

If you are interested in Intercast and would like to know more about this interesting technology, there a number of excellent sources of information available on the Internet.
Although there is an official organization involved in Intercast, co-ordinating the activities of all companies with an interest in the technology, the relevant web site, *www.intercast.org*, has little of interest to us.
An extensive story on Intercast may be found at *www.fhr.ch/~rvogt/intercst*. Well worth having a look at!
If you are after information about compatible TV cards, we suggest stopping by at Hauppage (*www.hauppage.com*, *www.hauppage.co.uk* and *www.hauppage.de*) and Miro (*www.pinnaclesys.com* and *www.pinnaclesys.de*).
For further information on Intercast broadcasts in Europe, consult the German ZDF web site at *www.zdf.de/programm/intercast/index.html*.

By contrast, WaveTop supplies a specific programme (divided in topics such as news and sports) which is set up centrally. In this system, every TV station supplies the same pages.

Windows 98 comes with a WaveTop viewer which may be installed as a component of WebTV. In spite of this, we doubt if this system will ever make it to general acceptance in Europe.

Other systems

Traditionally, Germany has been the leader in the development of other technologies that enable data to be transmitted along with TV signals. Way back in 1986, the WDR Computer Club came up with their 'Videodat' decoder, and the system has been in use ever since for the free software-over-air service linked with the relevant TV programme.

Deutsche Telekom have also teamed up with Dresden's Technical University for the development of Broadcast Online TV. As opposed to WebTop and Intercast, this system employs the horizontal blanking interval (sync pulse, front and rear porch) to convey data. A major disadvantage, however, is that a special receiver card is needed.

Figure 3. MTV also broadcasts Intercast data in the US; alas, not yet in Europe.

For the time being, we will have to make do with what's available on a larger scale, and in Europe that means Intercast. If you can pick up an Intercast-savvy TV station, and you have a TV card in your PC, do give it a try.

(982091-1)

A PC, most of us would say, consists of a complete computer case with a keyboard and a monitor, all wired up and installed in the workplace. And yet, the functionality of a PC is increasingly employed for so-called embedded control systems. Compactness is the buzzword in this field. US-based ZF Microsystems recently introduced a chip that reduces all major functions of a PC to a single component with the size of a credit card.

PC-on-a-chip

ZF Microsystems present an integrated solution

Anyone who has ever looked inside an PC must have come to the conclusion that the machine consists of a number of integrated circuits (or 'chips'), insertion cards, a power supply and cable bundles. The inherent disadvantage of using so many separate components is that they have to be linked by connectors!

In the common-or-garden variety PC installed in our homes and offices, the screw links used to secure individual components can be relied upon to provide the necessary mechanical stability for quite some time. This is in stark contrast with a PC used at the heart of an embedded control system, mainly because of the much stricter requirements in respect of mechanical loading. If control systems are fitted inside machines, mechanical reliability becomes a major issue, requiring a totally different structure of the PC and the interfaces connected to it.

The OEMmodule486 (OEM = original equipment manufacturer) from the US company ZF Microsystems has been specially developed for rugged embedded-control applications. The heart of a PC has been integrated into a single functional module (SCC, Single Component Computer) suitable for surface-mount assembly. To be able to connect the SCC to standard PC extension cards, the system provides support for the ISA bus, thus securing a direct link to the reliable and widely used PC/104 interface. This interface combines the versatility of the ISA bus with a compact and reliable connection system which is also suitable for stacking interfaces.

A closer look at the SCC

The single-chip computer developed by ZF Microsystems is designed around an 80486SX CPU clocked at 100 MHz to which is added the full complement of I/O functions PC users would expect to get.

Because the entire system is compatible with the PC/AT Industry Standard Architecture (ISA), it integrates the following functions: a DRAM controller, an ISA bus interface, a keyboard interface, two serial ports and one parallel port, a connection for a floppy disk drive and an EIDE interface for two devices, for example, a hard disk and a CD-ROM drive. Apart from these functions, the CSS also contains an AT compatible BIOS ROM, and a ROM in which Caldera's embedded DR-DOS is stored. The standard DRAM memory has a size of 2 Mbytes. The system may be extended with external RAM up to a size of 64 Mbytes using standard 3.3-V EDO RAM (70 ns).

With the chip powered up and a display interface connected, your PC monitor will show the DOS prompt. In addition to the standard system software, the flash memory has sufficient room for the storage of system-specific routines. Standard DOS software may be used without problems in this environment.

Interestingly, purchasing an OEMModule486 means that you automatically get a free licence for the use of the DOS and the BIOS. Traditionally, these two components represent a major investment when it comes to developing an embedded control system.

The circuit is suitable for use with a single 5-V supply, and consumes about 2.5 watts of power at a clock frequency of 100 MHz. The complete circuit measures approximately 56 x 76 x 12 mm, and has 240 connections arranged at a pitch of 0.04 inch.

(982094-1)

For further information, contact:
ZF Microsystems, 1052 Elwell Court,
Palo Alto, CA 94303 USA:
toll-free: 800-683-5943;
tel.: 650-965-3800; fax: 650-965-4050; e-mail: info@zfmicro.com;
Internet: www.zfmicro.com.

With the introduction and general proliferation of the Soundblaster card and its derivatives, almost any PC can be beefed up with a number of new multimedia functions. While some functions like the wavetable synthesizer and the sound-sampler are only found internally in the PC, others are available that are typically used in combination with peripheral equipment. The interface described in this article enables signals on the multifunctional joystick connection to be converted into a number of standard connections and signal levels. Any joysticks and MIDI equipment you may have available may then be hooked up in a simple manner.

Design by L. Lemmens

joystick and MIDI interface for Soundblaster cards

expand those soundcard connections!

A few years ago, the PC achieved a decent sound option thanks to the Soundblaster expansion card. In addition to the requisite analogue inputs and outputs, this card also came with connections for a joystick and MIDI equipment. Once other manufacturers of soundcards started to use this configuration as an example, the Soundblaster soon became the *de facto* standard.

Once a soundcard has been fitted into a PC, a synthesizer function and a sound-sampler become available. Furthermore, the card contains all logic to connect two joysticks, while a complete MIDI interface is also supplied as a standard.

Users of a rather simple soundcard will usually not be satisfied with the capacity of the built-in FM synthesizer. A real wavetable synthesizer as available on up-market soundcards, produces much more natural sounds. An external MIDI expander or MIDI keyboard may fill this obvious blank, provided the computer has a standardised MIDI interface with the full complement of connections. As far as hardware is concerned, most soundcards meet this requirement — only the interface is not equipped with the standardised MIDI connection and associated signal levels. That's why it is not possible to directly connect a keyboard or other MIDI hardware. *Ergo*, a special converter cable has to be made, complete with signal level adaptors. This solution is not only awkward, it also creates new problems because the MIDI signals are found on the joystick connection, and the plug you want to use will obviously preclude the connection of the joysticks! Fortunately, this little problem is easy to eliminate using some creativity and a small circuit.

Transistors and gates

The circuit diagram of the joystick/MIDI interface is shown in **Figure 1**. As already mentioned, the main function of the interface is to convert already *available* connections and signal levels into *standard* connections and matching levels.
This level conversion is relevant to the MIDI signal. With a standard MIDI interface, use is made of current loops. Consequently, the transmitter side has a current source, and the receiver side, an optocoupler. On the soundcard, however, the MIDI signals are only available at TLL level.
Looking at the connectors typically found on MIDI equipment, a MIDI input is usually complemented by two MIDI outputs and a MIDI 'thru' connection. All four connections make use of 5-pin DIN plugs. The two MIDI outputs supply identical signals, while the MIDI-thru connections supply a copy of the signal applied to the MIDI-in socket.
The soundcard only supplies electrical signals for *one* MIDI output and *one* MIDI input. The standard circuit described here ensures that all four previously mentioned connections become available, based on these two signals only. The circuit diagram clearly shows that this is by no means a formidable task. The entire circuit is built around inverters from the 74HCT14 integrated circuit. These inverters have a TTL compatible input and a TTL output which is turned into a current source with the aid of a couple of resistors.
Using gate IC1d the MIDI output from the computer is converted and used, among others, to make LED D3 light. When this indicator lights, MIDI signals are being transmitted.
Although buffers IC1b and IC1c are connected in parallel, each of them drives its own output. The two 220-Ω resistors are found in any MIDI interface, and so turn the TTL output into a current source.
The MIDI input is a traditional configuration based on an optocoupler. Resistor R10 and diode D1 protect the optocoupler against reverse and/or excessive input voltages. The resistor limits the current through the optocoupler. Received MIDI signals cause LED D2 to light.

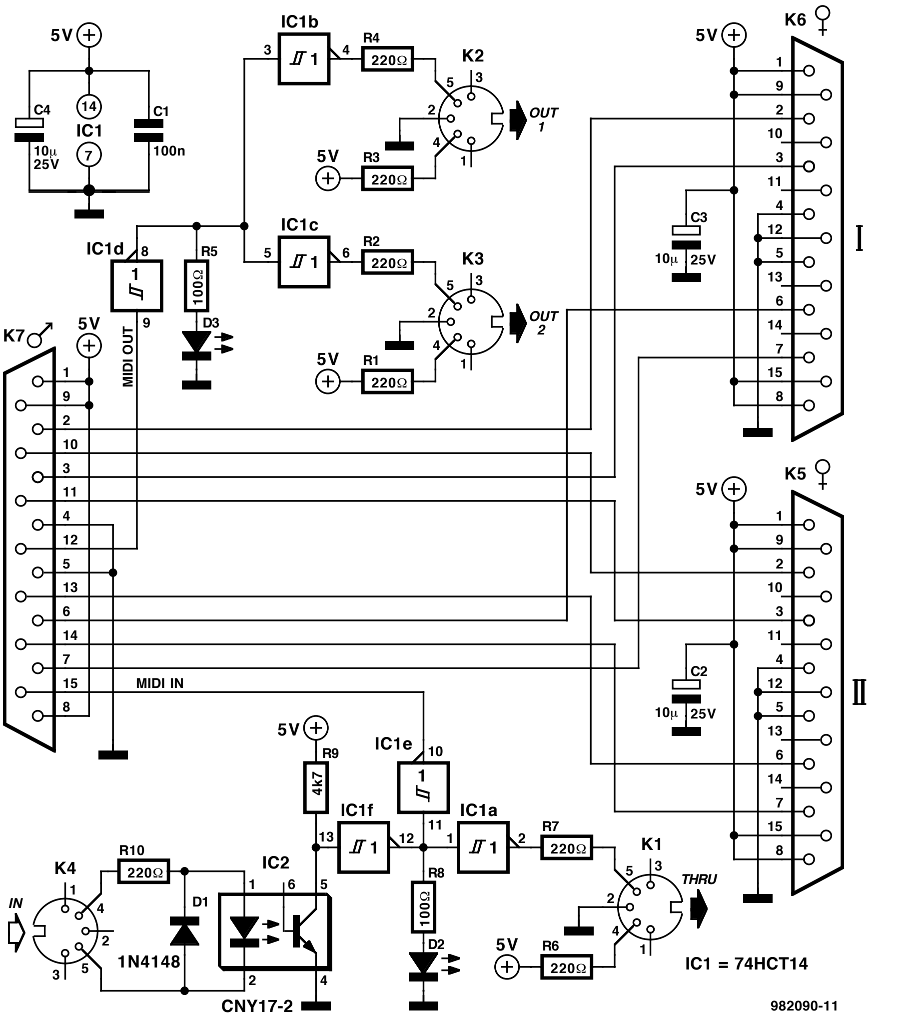
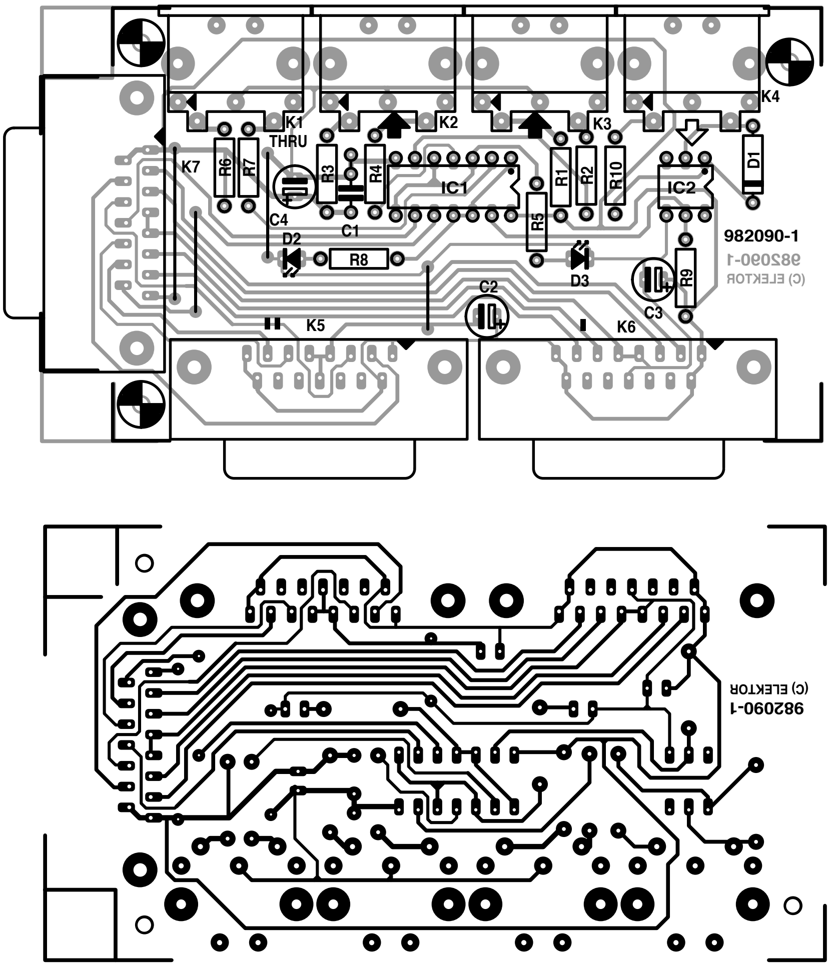


Figure 1. Circuit diagram of the joystick & MIDI interface. The circuit converts signals on the joystick connector of a soundcard into standardised connections and signals.



COMPONENTS LIST

Resistors:
R1-R4,R6,R7,R10 = 220Ω
R5,R8 = 2kΩ7
R9 = 4kΩ7

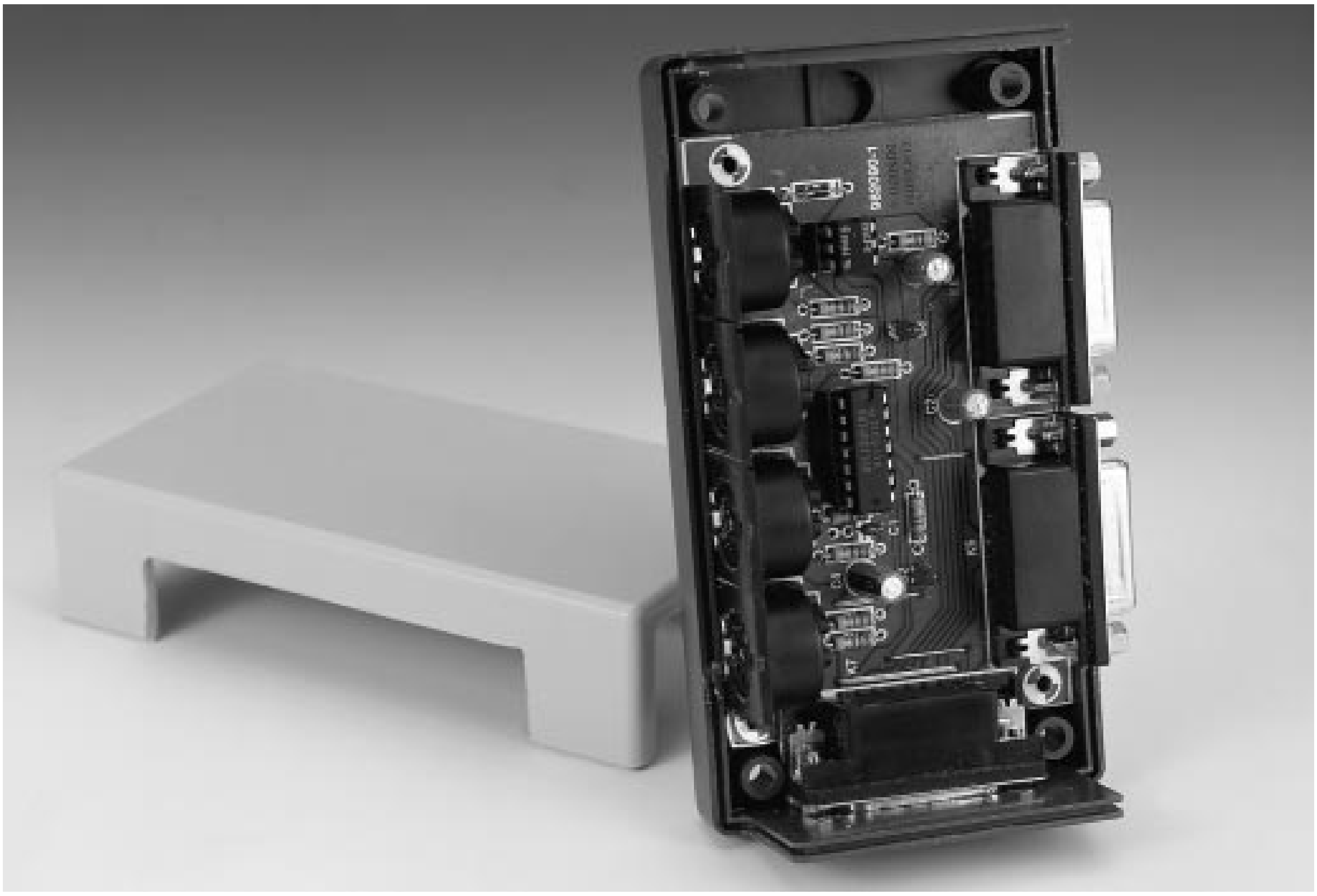
Capacitors:
C1 = 100nF
C2,C3,C4 = 10μF 25V

Semiconductors:
D1 = 1N4148
D2,D3 = LED
IC1 = 74HCT14
IC2 = CNY17-2

Miscellaneous:
K1-K4 = 5-way DIN-socket, PCB-mount
K5,K6 = 15-way sub-D socket (female), angled, PCB-mount
K7 = 15-way sub-D plug (male), PCB-mount
Case: 120x64x40mm, e.g., Bopla type E430.
PCB, order code 982090-1, see Readers Services page.

Figure 2. Copper track layout and component mounting plan of the PCB designed for the extension circuit (board available ready-made).

Figure 3. One of our built-up prototypes. The board fits exactly in a Bopla plastic case Type E430.



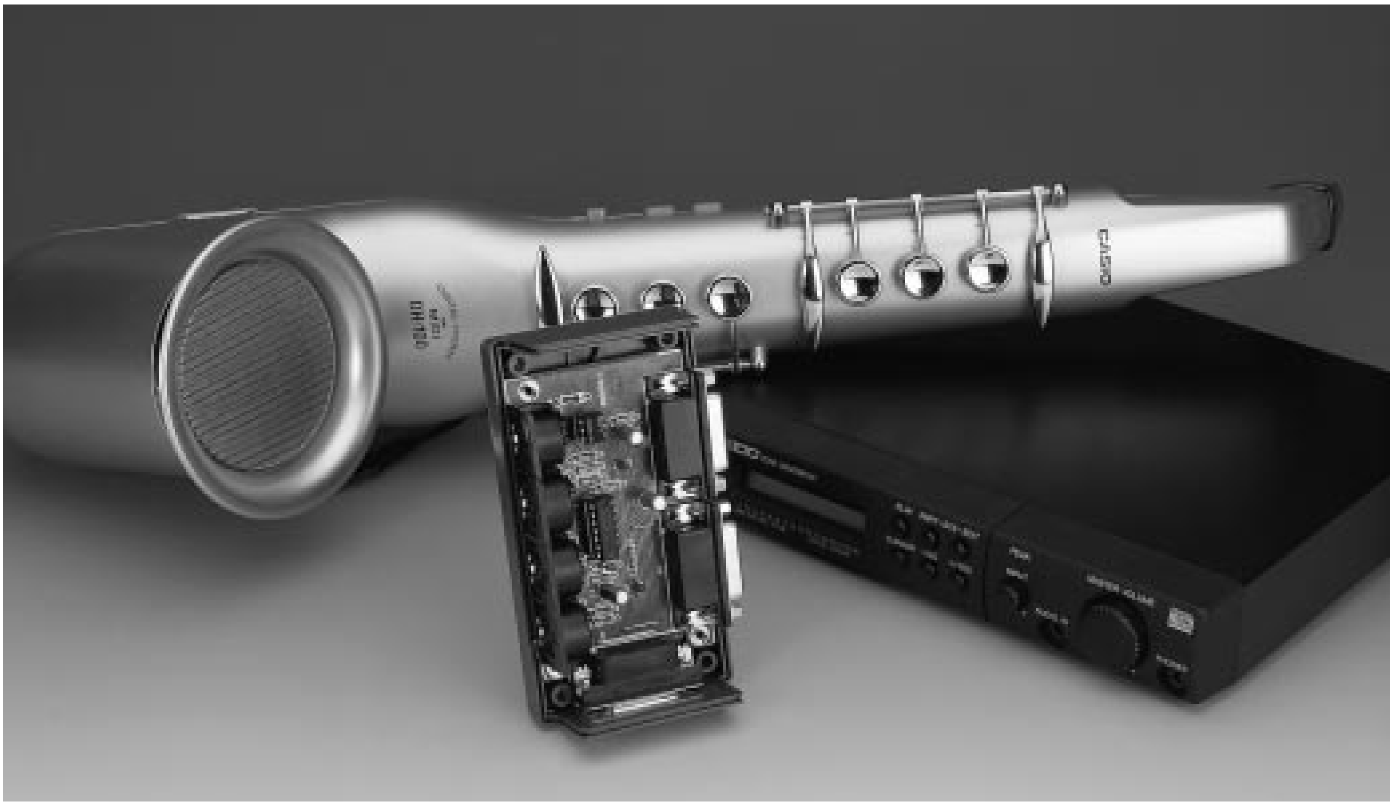


Figure 4. The MIDI connections allow commonly used MIDI instruments to be hooked up, like this expander from Yamaha, or Casio's electronic saxophone.

Because the supply voltage is taken from the joystick connector, the circuit does not require a separate power supply. So, it's all a matter of connecting it all up and start using it! In Microsoft lingo: *plug and play*.

Two joysticks

Although the joystick interfaces of most soundcards are complete, it should be noted that all signals are combined on a single connection, where they sit together with the MIDI signals! Our aim is therefore to 'untangle' the signals for the two joystick ports, and direct each set to its own connector.
Now, on the connectors we find two pins for the Fire buttons (a and b) as well as the connections for the potentiometers that control the movements in the horizontal and vertical directions (x and y respectively). To these should be added the positive power supply voltage and, of course, ground.
The design of the printed circuit board developed for this project may be found in **Figure 2**. The PCB has a clear layout with four sockets for the MIDI interface arranged at one side, and the two joystick connectors, at the other. Finally, there's a 15-way connector for the cable link with the PC.
The single-sided PCB contains a couple of wire links which have to be installed first so that they are not overlooked later. Next, you fit the connectors and the remaining components.
Build the circuit as neatly as you can, and connect it to the joystick port on the PC via a 15-way flatcable. That's it — you have an interface which allows two joysticks and a number of MIDI devices to be connected up to your PC, which is, dare we say it, *functionally extended*. (982090-1)

MIDI hardware — a closer look

MIDI is an acronym for Musical Instrument Digital Interface. Essentially, it is an interface that allows electronic musical instruments to communicate with each other. The standard was defined in the early 80's. Over the past decade or so, the computer industry has gradually adopted the relevant interface, which is currently available on almost any PC equipped with a soundcard. Thanks to the MIDI interface, software may be employed to control electronic musical instruments. The reverse is also possible: using a keyboard and a sequencer, a piece of music may be stored in a computer.
The MIDI interface hardware typically found on PC soundcards is derived from the RS232 interface. It comprises a kind of current loop which is used by the transmitter to send information to the receiver. At the receiver side, an optocoupler is used with a switching level of 5 mA. The switching times should be shorter than 2 μs. The bit rate is defined as 31.25 kbits/s ($\pm$1%). The asynchronous communication is based on one start bit, eight databits and one stop bit. The physical connection is made using 5-way 180-degree style DIN plugs and sockets, of which only pins 4 and 5 are used. Pin 4 is connected to +5 V via a 220-Ω resistor, and pin 5 to the output of the driver, also via a 220-Ω resistor. Pin 2 may be used as a ground terminal for the MIDI cable screening. Pins 1 and 3 are not used. MIDI cables are screened and have twisted wires. Their length should not exceed 15 m.

Together with a small program running under DOS, the circuit discussed in this article acts as an ADC-based voltmeter that allows you to calculate the capacity of a rechargeable battery. The PC also lends a helping hand by displaying graphs showing the instantaneous battery capacity and terminal voltage.

Design by F. Simonnot

battery capacity measurement by PC

run true-capacity tests on rechargeable batteries

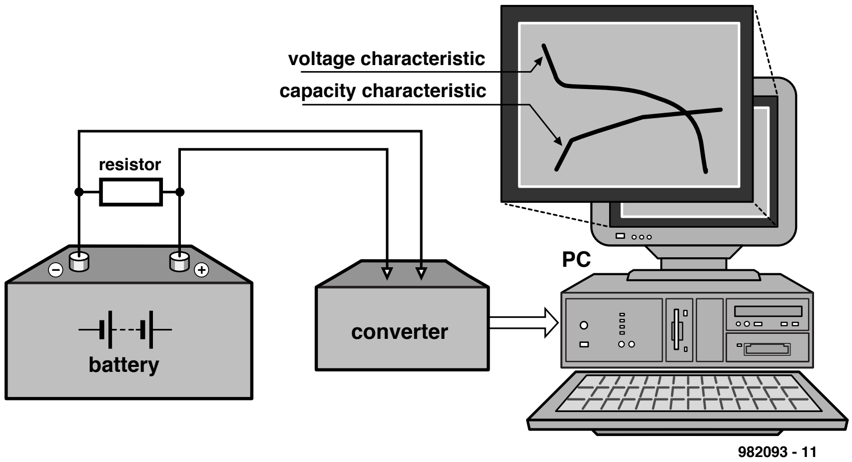


Figure 1. This block diagram of the system shows the basic elements: battery, converter and PC.

If you have ever considered giving an old PC any sort of useful function and a second lease of life, this project is a great opportunity because of its modest hardware requirements: a 386 PC with a VGA display and a classic printer port (EPP not required) is basically all you need as far as the PC is concerned. However, if your PC does not have a maths co-processor, the control program we're about to describe may have to be recompiled.

Principle of operation

As shown in the block diagram in **Figure 1**, we're talking about a measurement circuit based on an inexpensive and widely available converter. The converter is connected to the battery under examination by means of a resistor. Using a certain measurement interval, the PC, by way of its printer port, requests measurement data from the converter.

By computing the V/R ratio, the control program determines the current in the battery-resistor circuit, as well as the battery charge, and multiplies the latter quantity with a factor representing the discharge time. The result of this operation is added to the mAh (milli-ampère hour) counter. In this way, the proposed system acts like an integrator.

Hardware

The circuit diagram of the converter is shown in **Figure 2**. At the input of the circuit we find a voltage divider consisting of resistors R1 through R6, and an associated 6-way rotary switch. The step size of the input voltage is 5 V, the range is 5-30 V, and the lower resistance in the ladder is connected to the input of an analogue-to-digital converter (ADC) type ADC0804. This IC should not be a problem to obtain locally as it is produced by several semiconductor giants like Harris, National Semiconductor and Philips Semiconductors.

The ADC0804 is flanked by a type 74HC257 quadruple 2-to-1 line multiplexer with 3-state outputs.

The ADC0804 is a CMOS 8-bit A/D converter based on the 'successive approximation' principle. Its differential analogue input is marked by excellent common-mode rejection characteristics, and allows the 'zero' level of the analogue input to be given an offset. If necessary the reference voltage may be made adjustable to suit any voltage range smaller than the one normally allowed by the available 8-bit resolution.

The 6-way rotary switch at the input of the ADC acts as a voltage divider and allows one of the input voltage ranges between 5 and 30 V to be selected, the step size being 5 V. The 5.1-V zener diode in this part of the circuit has a protective function.

The multiplexer at the output of the ADC0804 serves to split the 8-bit digital data at chip outputs DB0 through DB7 into two 4-bit words which are sent to the PC printer port by way of dataline D0 (pin 2) and control lines Error (pin 15), Select (pin 13), Paper End (pin 12) and Acknowledge (pin 10).

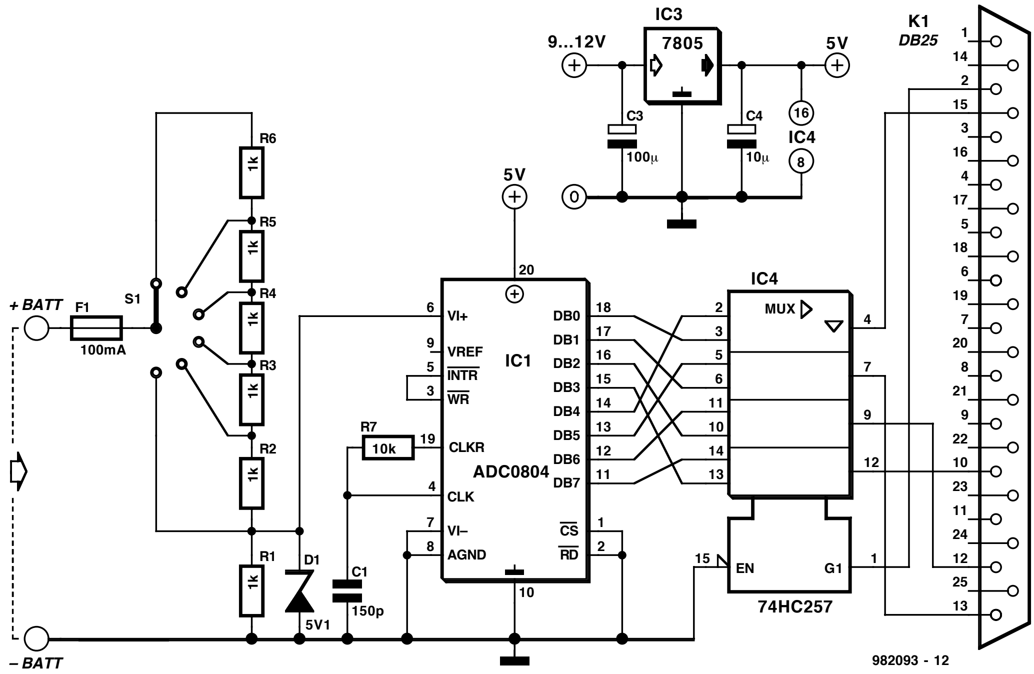


Figure 2. Simple hardware for the PC-aided battery capacity meter.

The clock frequency in the circuit is determined by RC network C1-R7, according to the following equation:

$F_{CLK} = 1 / 1.1RC$

which gives approximately 606 kHz using the respective 150-pF and 10-kΩ components.

The conversion time of the ADC is smaller than 100 μs.

The input of the converter is protected by a 100-mA fuse.

The capacity measurement is performed by connecting a resistor across the + and – terminals of the battery. The value of this measurement resistor will depend on the battery type, and the program can help you when it comes to determining the battery characteristics.

Software

You have to inform the program about the value of the battery load resistor, and the voltage range set on the analogue-to-digital converter. Based on the measurement values obtained from the ADC, the program first calculates the battery tension. Next, the load current is computed using the resistor value and the battery voltage. The result, multiplied by the duration of the measurement, yields the instantaneous battery capacity.

Now, let's have a look at how this works in practice. The control program is called Accbench.exe, and may be found on the project diskette with order number **986034-1**, along with the source code file written in Turbo

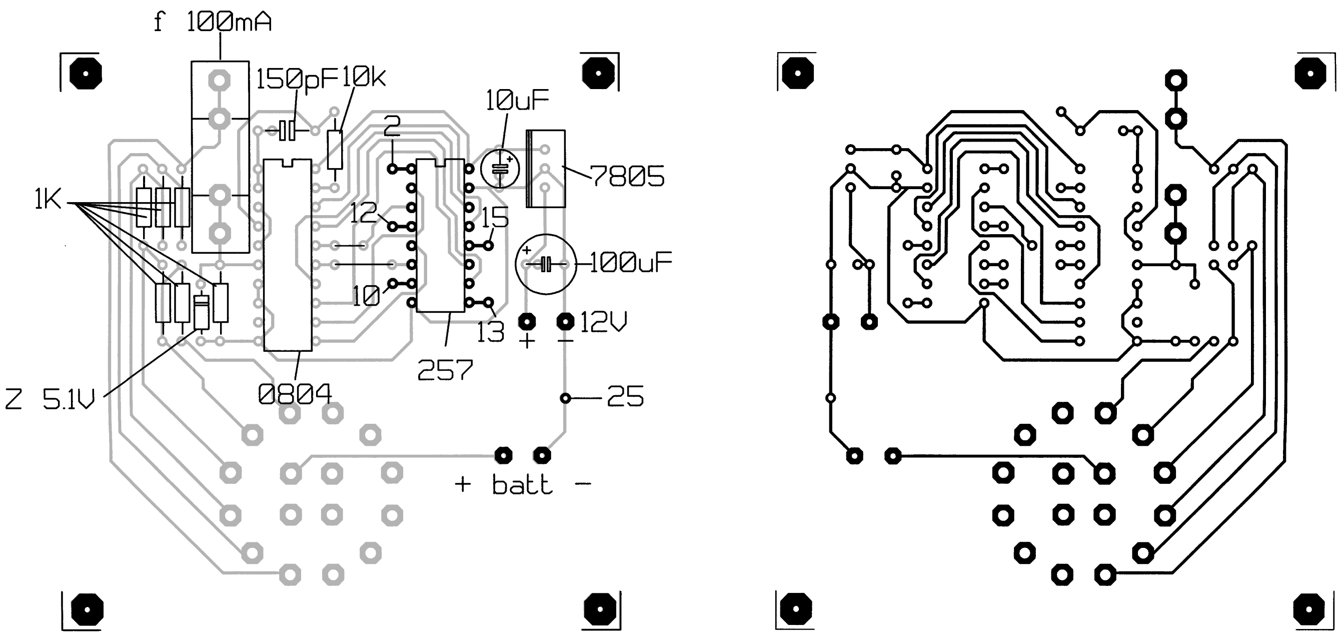


Figure 3. PCB artwork designed by the author — copper track layout and component mounting plan.

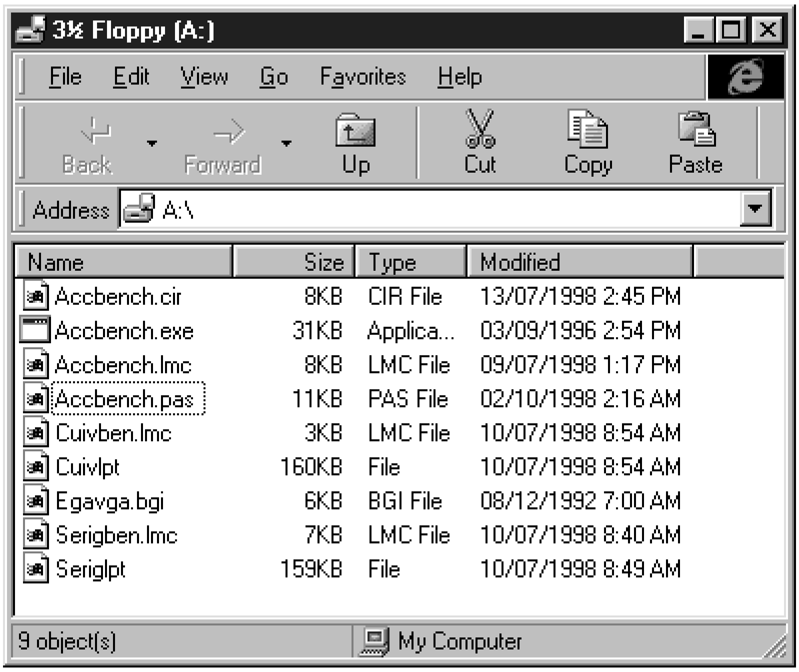


Figure 4. Screendump showing the files found on the project disk.

Pascal 7.0. Although parts of the source code are written in French, the relevant file will still be of use, we reckon, to those of you wishing to make modifications to the program.
Once launched, the Accbench program asks you if you want to do the calculations to determine the various parameters. At this point you may enter the nominal battery voltage, nominal battery capacity and the value of the load resistor. Using these input parameters the program calculates the discharging current, the amount of time it takes to fully discharge the battery, the power dissipated by the load resistor, and the estimated duration of the discharging cycle, by means of sampling.
It should be noted that the program uses a number of maximum values as far as the battery voltage and the battery capacity are concerned — the relevant values are 30 V and 12,000 mAh. If you enter higher values, the program will notify you by producing an error message.
The more audacious among you may want to have a go at modifying the 'converter control' section of the control program, starting with the procedure called 'convert', in order to make the program work with MAX187 used in the CPU Thermometer published in the October 1997 issue of *Elektor Electronics*.

Construction

Figures 3a and 3b show the copper track layout and component-mounting plan of the printed circuit board designed for this project by the author. Note that this PCB is not an *Elektor Electronics* design, and that it is not available ready-made through our Readers Services.
Fitting the components on the board should not present problems. Although it is necessary to refer to the circuit diagram to be able to locate the components on the PCB, that should not cause difficulties because there are only a couple of parts to mount, and their values are printed on the board.
If you want to avoid any risk of damaging the ADC0804 and 74HC257 integrated circuits, we suggest using good quality IC sockets.
The completed circuit board may be fitted in a small plastic case. The case is drilled to accept a 25-pin sub-D socket, like the one normally used for the PC printer port. This connector is linked to the PC's printer port via a standard cable. Two other wires leave the case: these are for the connection to the battery under test. The free ends of these flexible wires are fitted with 'crocodile (croc)' clips for easy connection to the battery terminals. The relevant connections on the converter board are marked '+ batt -'.
Pins 2, 10, 12, 13, 15 and 25 of the sub-D socket are connected to the '257 outputs, G1 and ground, as indicated in the circuit diagram.
The regulated supply voltage of between 9 and 12 V is connected to the solder pads labelled +12V and –12V; these are found near the 7805 voltage regulator.

In conclusion

It goes without saying that the present circuit may be used to perform pretty exhaustive capacity tests on all sorts of rechargeable batteries.
The approach chosen, discharging by a fixed resistor (as opposed to constant-current discharging which is also frequently employed), allows a measurement process to be used that closely resembles 'real life' conditions. For instance, it allows you to get a fairly good idea of the life expectancy of a rechargeable battery used in a torchlight, provided you know the characteristics of the lamp used.
As shown by the screendump **Figure 4**, the disk supplied for this project contains the following files

accbench.cir
Simulation file for Microcap V.

accbench.exe
Compiled, executable file.

accbench.lmc
Circuit diagram, Layo1 format.

accbench.pas
Source code file in Turbo Pascal 7.0.

cuivben.lmc
Copper track layout of PCB, Layo1 format.

cuivlpt
Print file for the PCB copper track layout, for LaserJet or DeskJet printers (300 dpi).

egavga.bgi
a DOS driver that allows any IBM-compatible PC to display the battery graphs.

seriben.lmc
Print file for the component overlay, Layo1 format.

Serglpt
Print file for the PCB component overlay, for LaserJet or DeskJet printers (300 dpi).

The disk also contains a number of authentication files.

(982093-1)

Eye pattern meter
PC Topics Supplement, March 1999, p. 13. (992002)
The moving coil meter shown in Figure 3 should have a sensitivity of 100-200 µA f.s.d.

Electronics Freeware
May 1999, PC Topics Supplement, p. 4 (990011-1)
The correct url for Digital Works is

http://www-scm.tees.ac.uk/users/d.j.barker/digital/digital.htm

Battery capacity measurement by PC
PC Topics Supplement, December 1998, p. 14-16. (982093)
With reference to the circuit diagram, a number of logic gates in IC4 have been transposed to improve the PCB layout. Functionally, this is of no consequence.
However, one track on the board is missing: that between pin 2 and pin 8 of IC1 (ADC0804). If this link is added, C1 is effectively connected and the circuit will work as described.

Sealed lead-acid battery charger
May 1999, p. 26-31. (990037-1)
In Table 2 (Component Values), the two formulas for R6 should read

0.45 / I [ohms].

D9 is missing from the parts list. As indicated in the circuit diagram, this diode is a type 1N5401. If the charger always supplies currents smaller than about 1 A, diode D9 may also be an 1N4001 or similar.

General Coverage Receiver
January & February 1999 (980084).
In the preselector section, the upper varicap diode, D14, has no dc path. A suggested method of improving the behaviour of the varicap (without modifying the PCB) is to replace capacitor C83 (220pF) with a wire link.

Flash Designs – address information
April 1999, New Products, p. 73.
In the New Products section, the address and telephone number of Flash Designs should be changed to read

Flash Designs, Ltd.,
North Parade House,
North Parade,
Bath BA2 4AL.
Tel. (01225) 448630.

We extend our apologies to Flash Designs and our readers for any inconvenience caused by the incorrect address information.

Electronic Spirit-Level
July/August 1998, p. 36 (984038).
In the circuit diagram, all LEDs (D2 through D10) should be reversed. The PCB layout is all right.