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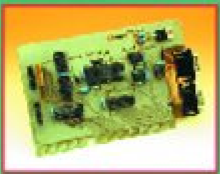

CAN Interface
for the ISA bus

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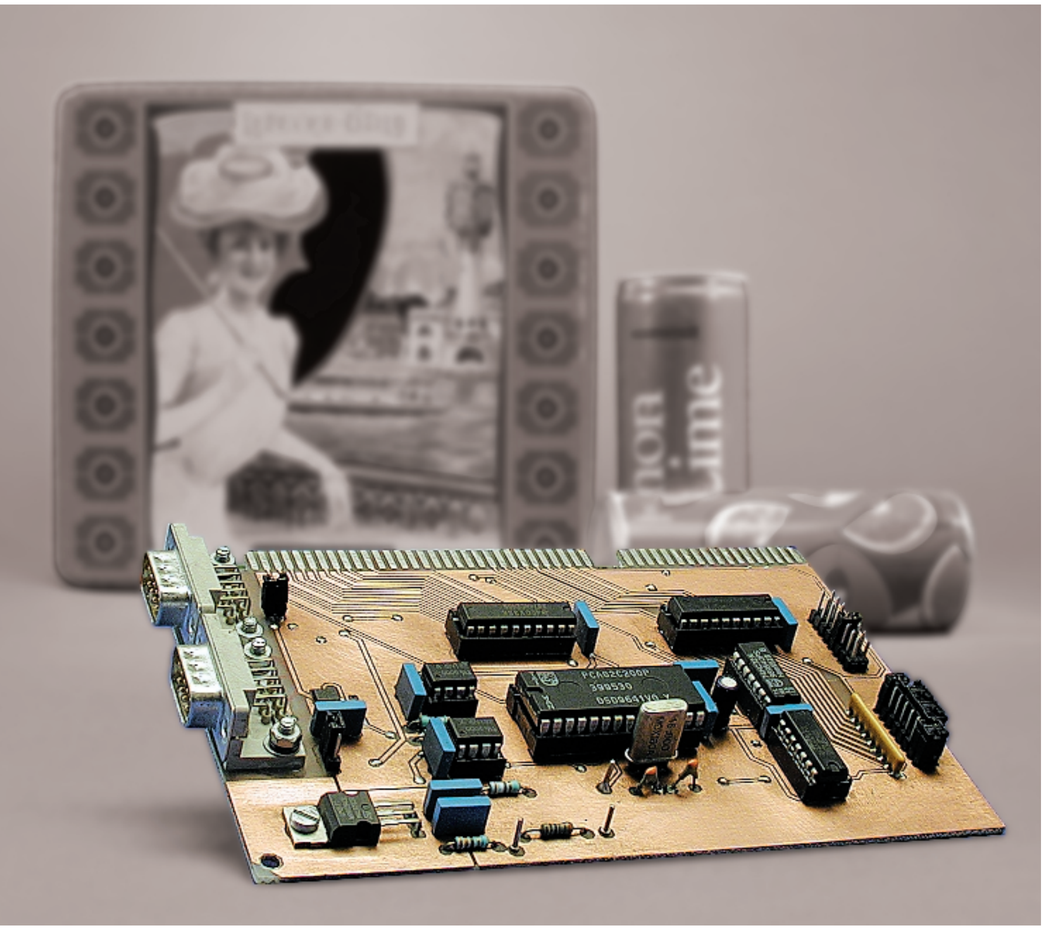
CAN Adapter for ISA Bus

A plug-in card for PCs

Design by B. Bouchez

CAN is currently the de facto standard for bus systems in the automobile industry. The fact that sales of CAN-related components have surpassed the most optimistic sales estimates is a clear indication of the relative success of the CAN bus. The CAN interface card described here is intended for installing into a free ISA expansion slot in your PC. Great for experiments in getting acquainted with CAN.



When it comes to developing an application employing the CAN bus in one way or another, a PC is an indispensable 'tool' for debugging and testing. Assuming you want to develop 100% bug-free application (and who doesn't?) the PC and a suitable interface card will allow you to visualise and record CAN frames travelling up and down the bus, as well as supply test frames. To show you the basics of how it's done, we've developed a small test application running under Windows. This little program will not fail to demonstrate the main features of the card.

At the same time, the interface described here may be the heart of a complex automation system because it is compatible with current industry standards like CANOpen and DeviceNet. We have to emphasise, however, that the present adapter card does not have any on-board intelligence, so that all processing of whatever protocol has to be handled by the PC. In certain cases, the resultant software overhead may be significant.

Windows being the current stan-

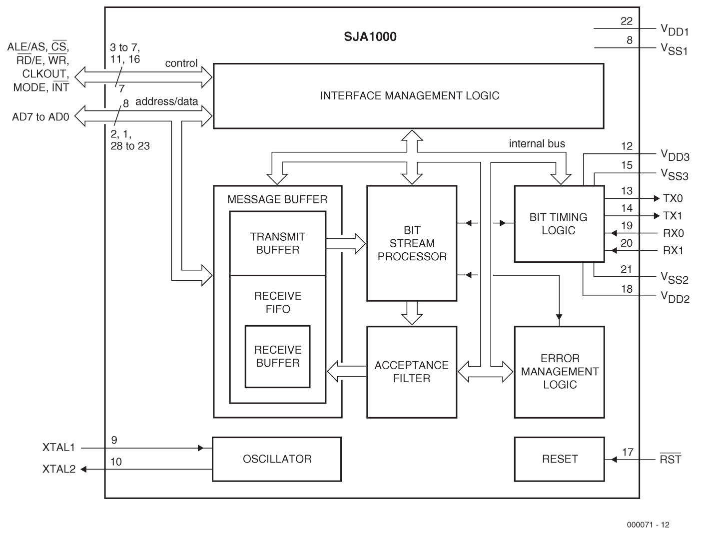


Figure 1. SJA1000 CAN Controller architecture.

dard in terms of operating systems for PCs, we have developed a driver which handles the most 'delicate' part of the adapter support. This driver is easily integrated into the majority of today's compilers, including Delphi, C++ Builder, Visual BASIC and Visual C++. The driver and the associated source code files are available on a diskette with order code **000071-11** which may be obtained through our Readers Services. The project software is also available from the Free Downloads section of our website.

At the CAN side

The present CAN adapter is based on one of the most powerful CAN controller chips available today, the SJA1000 from Philips Semiconductors. This chip, whose internal architecture is shown in **Figure 1**, comes in a 28-pin case. The SJA1000 is the successor of the 82C00 and pin compatible with it. However, the SJA1000 is CAN2.0 compliant (frames with a 29-bit identifier), it supports higher quartz crystal frequencies and, most important of all, it integrates a FIFO (First in, First Out) register in the CAN receiver section. As we will see a bit further on, the FIFO is an extremely welcome addition when it comes to interfacing the SJA1000 to a PC running Windows.

The SJA1000 consists of three functionally independent blocks having their own power supplies. That explains the presence of no fewer than six power supply pins on the chip: VDD1, VDD2, VDD3, VSS1, VSS2 and VSS3. Because we do not make use of certain functions available for direct connection of the chip (i.e., the output buffer and the input comparators), the three power supply voltages are distributed without separate processing.

The electronics

To get a better picture of how it all works, let's have a look at the circuit diagram in **Figure 2**.

At the heart of the circuit we find the SJA1000 CAN controller.

The function of the PCA82C250, IC2 in the circuit, is to convert the TTL signals processed by the optocouplers into ISO11989 compliant levels for the CAN bus. The 82C250 contains a power driver which is resistant to different types of short-circuit and overload conditions which may occur on the CAN. The RS pin of IC2 allows the slew rate of the output amplifier to be controlled and, consequently, the amount of electromagnetic components radiated by the CAN wires. In our case, this pin is tied to ground — this selects the highest possible slew rate, assuring reliable communication at data rates up to 1 Mbit/s.

In order to preclude any earth loops between the PC and equipments hooked up to the CAN, the bus is electrically isolated from the PC.

The circuit to achieve electrical isolation is classic and based on fast optocouplers (IC3 and IC4). Don't forget that the CAN bus supports a maximum data rate of 1 Mbit/s!

The link to the 'physical' CAN bus is by way of connectors K1 and K2, whose pin assignments follow the 'CAN in Automation' recommendations. The two connectors are wired in parallel to enable the PC to be actually 'inserted' into the CAN system. Because of the high throughput required at times by the CAN, an inserted configuration is to be preferred over a kind of 'shunt' position on the network.

If you decide to connect your card to a CAN bus carrying its own supply voltage (which is the most common configuration in industrial automation systems) you can reap the benefits of the internal electrical isolation afforded by the card. In that case, all you have to do is remove jumper JP1 and install JP2 in the 'high' position indicated in the circuit diagram. Voltage regulator IC1 is then powered by the CAN bus, while the 'bus' section of the card is effectively isolated from the 'control' section. The card will work from any CAN supply voltage between 9 V and 35 V (again, in accordance with the 'CAN in Automation' recommendations).

If you believe you can do without electrical isolation, or if your CAN bus does not carry its own supply voltage (as in most vehicle-mounted CANs), you simply install JP1 and move JP2 to the 'low' position indicated in the circuit diagram. In that configuration, the 'bus' section of the circuit will be supplied from the PC's +12 V rail. When dealing with extended buses (i.e., large distances between CAN devices), you have to be cautious about earth loops between the PC and other equipment.

At the PC side...

The SJA1000 has been designed to interface in the simplest possible way with the majority of currently available microcontrollers, including Intel 805x and Motorola 68xx devices. The selection between the two 'industry-standard' families is made by the logic level applied to the MODE pin of IC8 (+5 V = Intel, 0 V = Motorola). Both microcontroller families employ a multiplexed address/data bus, which is not the case on the ISA bus in the PC. Fortunately, the SJA1000 has an internal demultiplexer to separate the address and data components in accordance with the bus requirements.

Because the ISA bus is not multiplexed, it

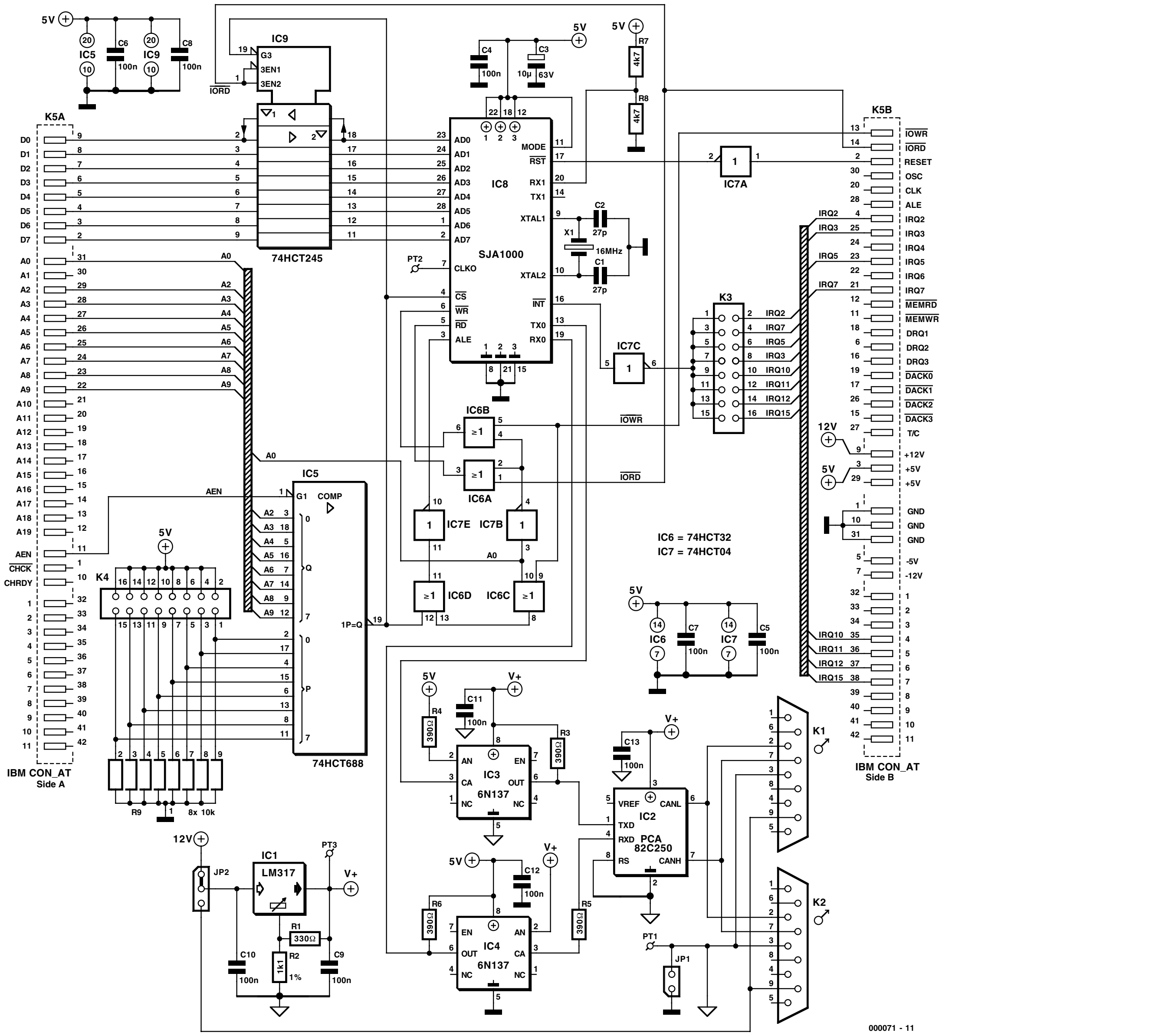


Figure 2. Circuit diagram of the CAN adapter for the PC ISA bus.

is necessary to restore the correct signal timing before pins AD0-AD7 of the SJA1000 can be driven in the proper fashion. In order to catch each component (data or address), the card occupies two contiguous addresses in the PC's I/O area. In fact, to prevent cluttered hardware, the card occupies four addresses, of which only the first two are used.

Circuit IC5, a 74HCT688, determines the I/O range taken by the CAN card. When the eight most-significant (MS) I/O address bits match the setting on K4, the card is enabled, whereupon bus buffer IC9 is actuated, as well as the logic network built from the gates in IC6 and IC7. To select an I/O range starting at, for example, 330h, you install a jumper over pin pairs 15-16, 13-14, 7-8 and 5-6 of K4. This sets up the code 11001100xx at the input of IC5.

Once the PC starts to write to the even-numbered address in the user-defined I/O area (bit A0 at 0), gates IC6C, IC6D and IC7E allow the write pulse to arrive at the ALE pin which controls the demultiplexer inside the SJA1000.

Let's look at a real-life example. If the ISA insertion card is to reside at address 330h and you wish to access the clock register of the SJA1000 at 1Fh, all you have to do is have the following instructions executed by the PC (a bit further on we'll get to the particular meaning of these instructions under Windows):

```
MOV DX, #330h
MOV AL, #1Fh
OUT DX, AL
```

Once a register has been selected in the SJA1000, we can read or write its contents by employing the function of the $\overline{WR}$ and $\overline{RD}$ pins in the classic way. These pins are enabled by IC6b and IC6A respectively, which can not be activated unless address line A0 is at logic High (even-numbered address). To read the contents of the selected register by means of the routine shown above, you program

```
MOV DX, #331h
IN AL, DX
```

Similarly, to write, for example, the value 5Ah to the selected register, you run

```
MOV AL, #5Ah
MOV DX, #331h
OUT DX, AL
```

Gate IC7A serves to invert the Reset signal for the PC to ensure it is at the proper level for the SJA1000.

Similarly, gate IC7C inverts the polarity of the SJA1000 interrupts for the sake of the PC. The interrupt channel employed by the card is simply selected by means of a jumper placed on K3. If interrupts are not required (as in the cases of our driver for Windows), it is

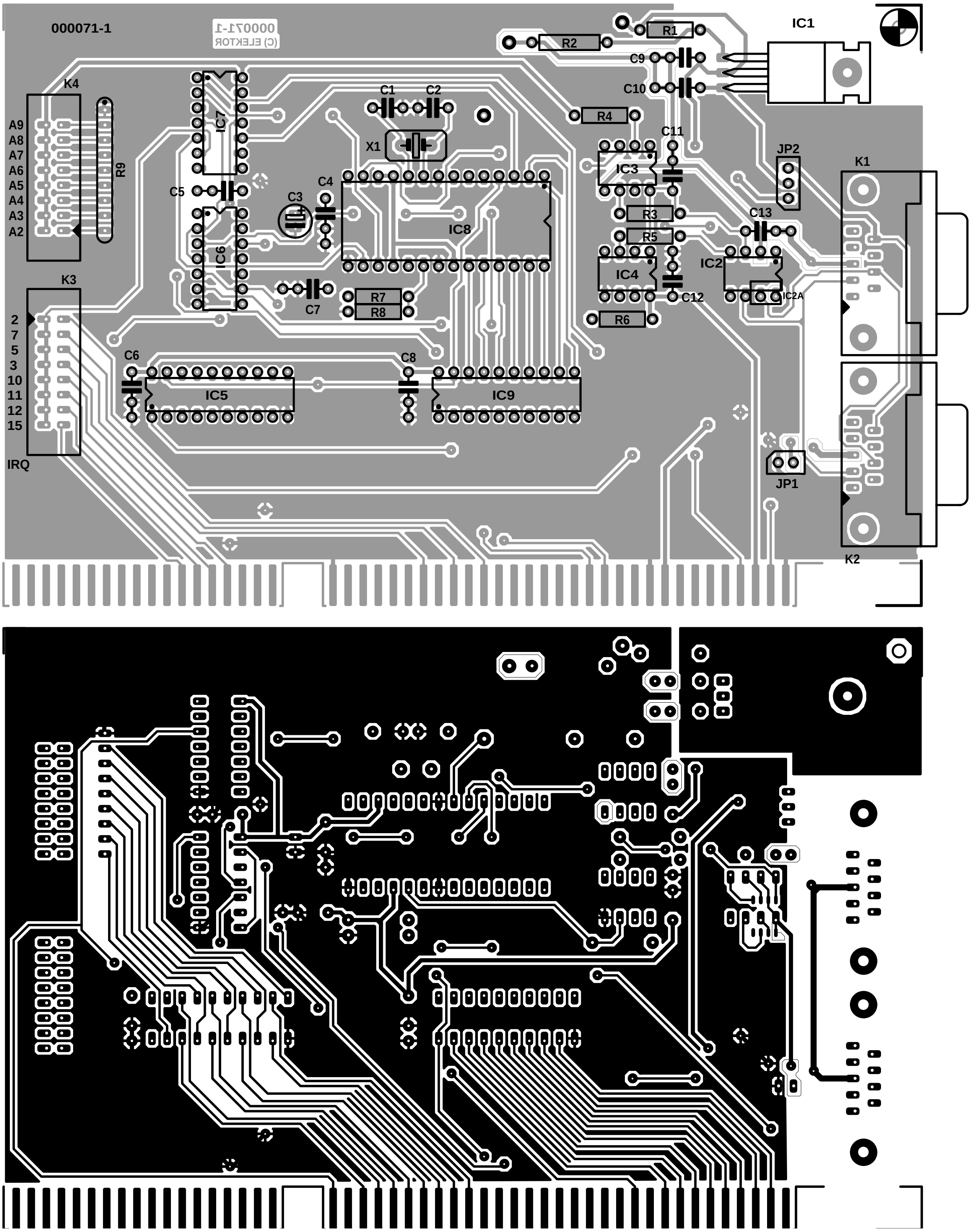
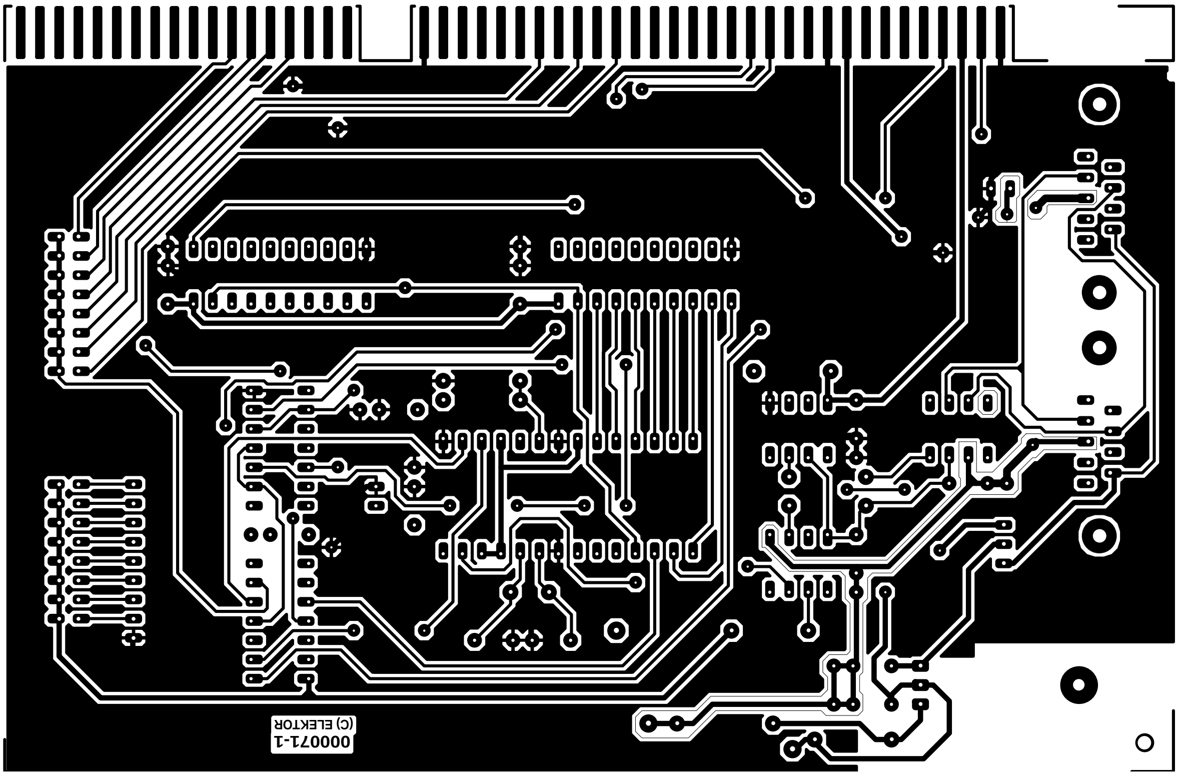


Figure 3. Copper track layout and component mounting plan of the PCB designed for the CAN adapter card (board available ready-made).



sufficient to remove the jumper from K3.

Finally, the SJA1000 employs its own clock source built around X1. Changing the crystal frequency is possible, but you will need to recalculate the division factors stored in the internal registers, using the information supplied by Philips. In all fairness, this would seem to be a job for true CAN specialists only! If you want to use the driver written by the designer of this circuit, stick to a quartz crystal frequency of 16 MHz.

Construction

From a constructional point of view, the present card is simple to build (by almost any standard) and contains just a small number of components. The copper track layout and component mounting plan are given in **Figure 3**.

You will notice that circuit board has two footprints for IC2, one a DIL8 shape, the other, a SO8 (SMA style). Unfortunately, these days the PCA82C250 seems to be available in the SO8 case only, which makes soldering by hand a tad difficult but not impossible. Some of you may be lucky enough to get their hands on the leaded version in the DIL8 case.

The order in which the components are soldered onto the board is traditional: resistors and small capacitors, sockets, connectors and only then the taller and/or larger parts. Please pay attention to the orientation of SIL array R9. Voltage regulator IC1 has to be mounted with its pins bent at right angles (90 degrees) to allow its tab to be secured to the PCB, which will also act as a minimal yet adequate heatsink.

The software

The relative simplicity of the hardware (including the construction phase as described above) is in stark contrast with the complexity of programming the SJA1000. The bare essentials are described in the SJA1000 datasheets, which we'd say are a must for every serious user of the present card. The datasheets may be downloaded in the form of a pdf document (235 kB) from
http://www.semiconductors.com/pip/SJA1000#datasheet
Mind you, just having this document safely stored on your PC is no guarantee for the eventual production of bug-free drivers!

Because the program listings explaining the operation of the SJA1000 in polled interrupt mode are simply too large to be reproduced in this article, we have posted them in the Free Downloads section (items December 2000) of our website at
http://www.elektor-electronics.co.uk
Of course, these listings have to be adapted to the PC and operating system in question. Note that the listings apply to Basic CAN mode only (11-bit identifier). Fortunately, they should be easily adapted to implement Extended CAN (29-bit identifier).

DOS, Windows and the rest

If you decide to use the ISA CAN adapter card under DOS (yes it is still around and in fact experiencing a come-back in industrial applications, despite Windows CE), writing a driver remains relatively simple thanks to the source files on the project diskette and the SJA1000 flowcharts.

Under DOS, you get virtually direct access to all resources offered by the card. In fact you can exploit it to the full because you are not 'hindered' by the operating system.

Using the adapter card under Windows 3.x (a.k.a. 16-bit Windows) requires a DLL to be written which is similar to the one described on the project disk. The help file supplies the necessary details on modifications to carry out on the 32-bit source code to slim it down to 16-bit use. Win 3.x, by the way, tolerates DLLs accessing the interrupt household, a fortunate fact which should enable you to achieve performance and throughput not much lower than under DOS.

COMPONENTS LIST

Resistors:
R1 = 330Ω
R2 = 1kΩ1
R3 à R6 = 390Ω
R7,R8 = 4kΩ7
R9 = 8-way SIL array, 10 kΩ

Capacitors:
C1,C2 = 27pF
C3 = 10µF 63V radial
C4-C13 = 100nF

Semiconductors:
IC1 = LMT317T (TO-220 case)
IC2 = PCA82C250 (DIL8 or SMA) (Philips)
IC3,IC4 = 6N137
IC5 = 74HCT688
IC6 = 74HCT32
IC7 = 74HCT04
IC8 = SJA1000, PCA82C200P (DIL28)
IC9 = 74HCT245

Miscellaneous:
JP1,JP2 = 2-way pinheader with jumper
K1,K2 = 9-way Sub-D plug (male), PCB mount
K3,K4 = 16-way boxheader or pinheader
X1 = 16MHz quartz crystal
PC1-PC3 = solder pin
PCB, order code **000071-1** (see Readers Services page)
Project disk, order code **000071-11** (see Readers Services page)

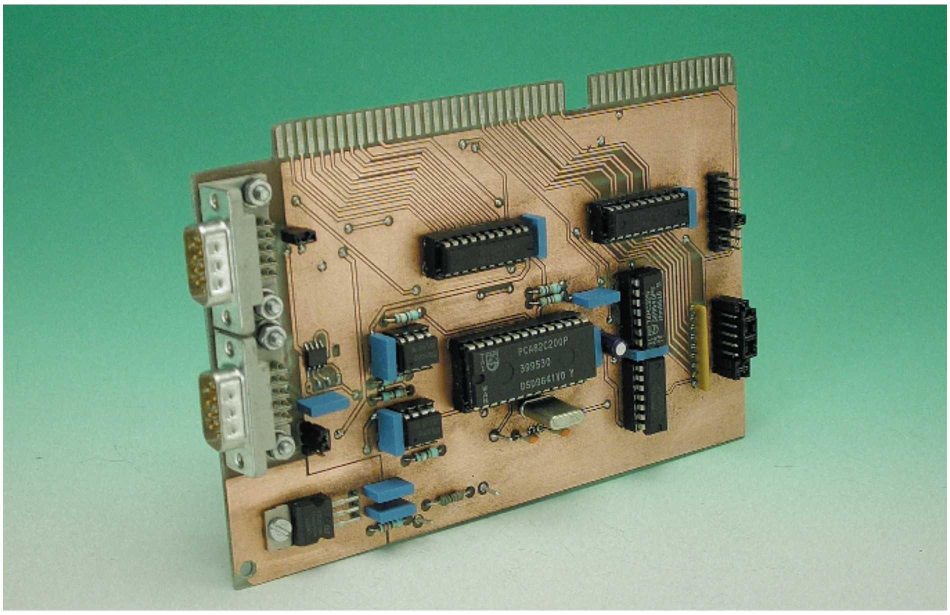


Figure 4. Fully functional prototype of the CAN adapter.

However, the real excitement begins with the 32-bit cores of Windows 95, 98 and NT, in which the processor runs in protected mode. Two problems should be addressed: access to the I/O space and support for interrupts.

It should be reiterated that none of the 32-bit Windows versions allow an .EXE program to modify the interrupt resources. Without going into too much detail, the applications run at protection Level 3, while it is necessary to be at Level 0 to be able to reprogram the interrupt controller. The only way of getting there is to write a system driver. A difficult task, because the relevant utilities supplied by Microsoft are by no means models of simplicity. What's more, Windows 95 requires you to write VXD virtual drivers, Windows 98, WDM's, and Windows NT, drivers of the SYS type. So it's goodbye dear interrupts…

In practice, there is a glimpse of hope. After all, the SJA1000 contains a FIFO RAM that allows the processor load to be lightened by stocking data received while in read mode. By reading the SJA1000 receiver registers at regular intervals, we can make do without interrupts. This method has been adopted and implemented in the driver developed for the project — received data is read at regular (millisecond) intervals.

The FIFO having a length of 64 bytes, the driver will be able to continuously process 64 × 100 bytes per second, corresponding to a bus data rate of 500 kbits/s — reasonable we'd say for the application at hand. Given the fact that the bus will rarely be required to operate at full speed (load) the driver will be capable of supporting 1 Mbit/s traffic because the pause intervals will be sufficient. Advanced users should also note that suitable control of the data filters in the SJA1000 will allow the processor load to be reduced considerably.

The real problem under Win-32 is the theoretical impossibility of getting direct access to I/O or memory. In principle, if you attempt to execute an IN or OUT instruction at protection Level 3, you will unwittingly trigger a 13h interrupt which flags 'protection violation'.

Fortunately, the protection is not actually 'armed', while Microsoft have recognized the problem. In fact the latest versions of Windows 98 do support I/O access. What's more, the I/O area reserved for 'experimental' I/O (330h to 33Fh) is no longer ferociously guarded by Dear Bill *c.s.*

Our driver has a modular construction. By default, it employs a simple but effective layer, **DIRECTIO.DLL** to access hardware directly without a driver, at Level 0. If you insist on using a virtual driver, we have included **HWPORT95.DLL** which allows freeware program HWPORT95 to be used, which, in turn, is based on a VXD. This freeware program is also found on the project diskette. To use it, simply modify the **CanTest.INI** file by replacing parameter 'directt95.dll' by **HWPORT95.DLL**.

Under Windows NT, things are becoming more complex because the protection at Level 3 is active. Using the files on the project disk, the driver is not able to make the card operate under Windows NT. The solution is to write your own DLL based on the model DIRECTIO.DLL, and including a call to a driver SYS for NT.

The CANTEST program

The program **CANTEST** (of which the source code and executable files are found on the project disk) is a 'CAN Monitor'. It allows the data frames travelling up and down the CAN bus and complying with hew criteria set up by the 'Acceptance Code' and 'Acceptance Mask' registers of the SJA 1000 to be visualised. In addition to the Monitor function, CANTEST also allows user-defined CAN data frames to be transmitted.

The CAN timing registers may be freely modified to evaluate the performance of a CAN bus at data rates between 10 kbits/s right up to 1 Mbits/s. It is even possible to test at 1.6 Mbits/s, although that data speed is well outside the CAN specs.

Testing the CAN adapter

The simplest way to test the CAN adapter card is to run the CANTEST program which contains auto-detection and test algorithms. If after launching the program the window appears correctly, you may safely assume that the card has been properly detected at the address indicated in the CANTEST.INI file (don't forget to edit it to your requirements before running the test program), and that the card responds to control commands.

If CANTEST pops uo a window reading '–4 : CAN Adaptor not found', you should look at these possible causes:

1. The I/O address occupied by the card does not match the relevant allocation in CANTEST.INI. Modify the .INI file or reconfigure the jumpers on the card to ensure equal hardware and software settings.

2. The I/O address is already in use by another peripheral device. The preferred I/O range for the CAN card is 330h - 33Fh. To minimize the risk of conflicts, this is the area reserved for 'experimental applications'.

3. The card contains a hardware fault which you obviously have to trace and eliminate. Another cause of problems could be the ISA bus settings in the BIOS, so check these as well. A too high ISA bus speed may prevent the SJA1000 from responding correctly (rare though because the SJA1000 is a pretty fast device).

When forced to probe deeper into problems with the CAN card, simply resort to the use of good old 'DEBUG' found in your DOS folder. Launch it and use the instruction O to write different values to the clock divider registers in the SJA1000 (register 31, or 1F in hex), referring to the datasheets for valid numbers.

Assuming that the card sits at I/O address 330h in the PC you'd type

```
O 330 1F    <return>
O 331 07    <return>
```

By reading back the same register you should get back what you wrote in it:

```
I 331       <enter>
>07
```

By 'scoping the test point connected to CLKOUT, you should observe a change in frequency of the rectangular signal as a function of values written to the register.

If this test is okay, you may be assured of the correct functioning of the SJA1000 and its PC interface.

Testing the opto-isolated CAN interface is more difficult. In principle, the only useful test consists of linking the card to a bus which is known to function at it should, and then run CANTEST to verify that data are being received. If problems persist, start by checking if the 82C250 and the optocouplers are properly supplied at +5 V. If not, check the configuration of the jumpers that define the supply voltage of the LM317 voltage regulator. If you have selected the bus supply option, inspect the cabling between the sub-D connectors and a dc power supply set to a voltage between 12 V and 24 V.

If you are in doubt about the correct functioning of this section of the circuit, you may remove the SJA1000 from its socket and connect a wire between the socket pin for TX (linked to the transmit optocoupler). Alternately connect the other end of the wire to 0 V and then +5 V. You should observe a change in the differential voltage between CAN_H and CAN_L on the output connectors. The input channel may be tested in the same way by connecting up a regulated power supply set to about 2.5 V. By alternating the polarity of this voltage source you should be able to produce level changes on the RX pin of the SJA1000, this is connected to the output of the receive optocoupler.

(000071-1)

For further reading:

Controller Area Network (CAN), *Elektor Electronics* September 1999 through January 2000.

On project disk # 000071-II

Copyright.txt : author's copyright notice
contents.txt : disk contents
readme.txt : late breaking information
cantest.dof : CANTEST compilation options for Delphi
cantest.dpr : CANTEST Delphi source code
cantest.dsk : DELPHI environment for CANTEST (optional)
cantest.exe : ready-compiled executable
cantest.ini : CANTEST parameters. Edit to select CAN adapter I/O base address and driver type
cantest.res : CANTEST binary resources (Delphi) (optional, file contents depends on Delphi version)
fichetest.pas : CANTEST main window source code
fichetest.dcu : CANTEST main window, compiled module
fichetest.dfm : source code of Windows objects in main window
direct95.dll : link library for direct access to PC I/O resources under Windows95/98
direct95.dof : Delphi compilation options for DIRECT95 library
direct95.dpr : Delphi source code of DIRECT95 library
direct95.dsk : Delphi environment for DIRECT95 library (optional)
direct95.res : DIRECT95 library binary resources (Delphi) (optional, file contents depends on Delphi version)
err_code_iolib.pas : Include file to distribute identical error codes among CAN applications
hwport95.dll : compiled library for freeware HWPORT95 instead of DIRECT95 library
impcanlib.pas : PCANLIB driver Include file for Delphi applications
pcanlib.dof : Delphi compilation options for PCANLIB driver
pcanlib.dpr : Delphi source code of PCANLIB driver for use with CAN adapter
pcanlib.dsk : Delphi environment for PCANLIB driver
pcanlib.res : Delphi binary resources (optional, file contents depends on Delphi version)
pcanlib.dll : link library for CAN adapter

Universal Mobile Telephone Service (1)

Part 1: From System 1 to UMTS

By G. Kleine

The frequency allocation sell off for the next generation of mobile phones caused a stir this April when the U.K. government scooped £22.48 billion. More recently the German government netted a staggering £31.8 billion in its licence sell off. What new features can we expect from this next generation of mobile phone?

The ETSI or European Telecommunications Standards Institute has defined a technology standard for the third generation of mobile phones called IMT-2000 or *International Mobile Telecommunications 2000* These proposals have been approved by the ITU and also go by the name UMTS or Universal Mobile Telecommunications System. The main aim of this standard is to ensure global harmonisation of broadband mobile communications for 3G or Third Generation phones including frequency allocation and data transfer protocols. This will ensure that

UMTS handsets will be compatible in whatever country adopts the network. One of the main goals of the standard is this so called global roaming. You should be able to make calls using a UMTS/IMT2000 mobile phone whether you're in Bratislava or Bridlington. This means that when you sign up for a UMTS phone and are assigned a phone number, it will be a global number. No need to prefix with country codes — you can always be reached on this number.

The theoretical maximum data rate of UMTS is 2 Mbit/s but this can only be achieved under optimum propagation conditions for example if the mobile is operating in a city and is not moving (immobile?). If the mobile is moving relatively slowly (i.e. walking pace) then transmission rates of up to 384 kbit/s are possible. In any case the minimum guaranteed data rate is 128 kbit/s even with the mobile travelling at vehicular speeds.

Abbreviations

2G	Second Generation (Mobile System)
3G	Third Generation (Mobile System)
AMPS	Advanced Mobile Phone System (US analogue cell network)
ANSI	American National Standards Institute (USA)
ARIB/TTC	Association of Radio Industry and Business (Japan)
BW	Bandwidth
CDMA	Code Division Multiple Access
DCS	Digital Cellular System (DCS1800)
DECT	Digital European Cordless Telephone
EDGE	Enhanced Data Rate for GSM Evolution
ETSI	European Telecommunications Standardisation Institute
FDD	Frequency Division Duplexing
FDMA	Frequency Division Multiple Access
FPLMTS	Future Public Land Mobile Telephone System
GMSK	Gaussian Minimum Shift Keying
GSM	Global System for Mobile Communications (originally called: Groupe Speciale Mobile)
GPO	General Post Office (State run predecessor to BT)
GPRS	General Packet Radio Service
HSCSD	High Speed Circuit Switched Data
IMT	International Mobile Telecommunications
ITU	International Telecommunications Union
LNA	Low Noise Amplifier
LO	Local Oscillator
MSS	Mobile Satellite Service
NMT	Nordic Mobile Telephone (Scandinavian analogue cellular net)
NTT	Nippon Telecom & Telegraph Corp. (Japanese analogue cellular net)
PA	Power Amplifier
PCS	Personal Communications System
QPSK	Quaternary Phase Shift Keying
8PSK	8 Phase Shift Keying
SIM	Subscriber Identity Module
SMS	Short Message Service
TACS	Total Access Communication System (U.K.)
TCP/IP	Transport Control Protocol / Internet Protocol
TD/CDMA	Time Division CDMA
TDD	Time Division Duplexing
TDMA	Time Division Multiple Access
TTA	Telecommunications Technologies Association (South Korea)
UMTS	Universal Mobile Telephone Service
UTRA	UMTS Terrestrial Radio Access
WAP	Wireless Access Protocol
W-CDMA	Wideband Code Division Multiple Access, Wideband-CDMA
WRC	World Radiocommunication Conference

New UMTS features

Tomorrow's globally roaming UMTS mobile phone will offer many more features than we see on current handsets. Along with SMS and WAP there will be a multitude of multimedia facilities available. At last you should see the introduction of true video phones and video conferencing over a UMTS radio link. High quality music tracks will be downloaded in an instant and short video postcards may be sent to friends. Interactive games with other phone users will be introduced and shopping methods will also be transformed, your UMTS handset will deal with authorisation and payment of purchases. Company reps and service personnel will find that while on the move they have a direct link to the company's intranet. As a UMTS handset moves, its approximate location will also be logged so that when you request information on local restaurants for example or ask for a street map if you are lost you will receive information relevant to your current location. Anyone dealing with Fleet management or deployment of emergency services and those who need to know the whereabouts of the mobile will find this feature useful.

Predicting how a UMTS handset will look is difficult but anyone who has sent an email using a WAP phone will realise that it will need to be equipped with a more convenient method of data entry and will probably evolve along the lines of a portable PC or palmtop.

A glimpse back into the past

Commercial mobile communications systems in the U.K. really began with *System 1* operated by the G.P.O. (see **table 1**). The system had very poor coverage and was not cellular. Each radiophone requires one frequency for sending and one for receiving i.e. four channels in total for a call between two radiophones. With a voice channel bandwidth of 100kHz each call needed 400kHz of the entire network bandwidth. *System 2* was developed but never implemented. *System 3* was later introduced with a reduced voice bandwidth of 25 kHz. Despite the increased capacity it could never keep up with the demand for this system.

The big breakthrough came with the introduction of the first cellular networks. Although the fundamentals of the cellular network had been theorised, it was in 1947 that D.H. Ring of Bell Labs in the USA formalised the cellular concept in a paper. The basic idea is that instead of using powerful transmitters to cover a large area, the area is divided up rather like the hexagonal panels of a patchwork quilt. Low power base stations are placed at the centre of each panel or 'cell'. Base stations in adjacent cells are not allowed to use the same frequencies, but these frequencies can be re-used in other non-adjacent cells. The beauty of the system is that if the number of subscribers in a particular area is large, for example in central London, the number of cells can be increased by reducing the cell size and using more, lower powered base stations, thereby increasing the network capacity.

The U.S. company Bell made trials of the first cellular system in 1978 in Chicago but did not start deployment of their *AMPS* (Advanced Mobile Phone System) until about 1983. Meanwhile the Scandinavian countries had installed the first commercial cellular network in 1981. Working at 450 MHz, this first generation analogue mobile phone system was known as *NMT* (Nordic Mobile Telephone) and offered roaming. It was later superseded by a 900 MHz version. In the early 80's the U.K. introduced *TACS* (Total Access Communication System). This is based on the American *AMPS* but is, needless to say, not compatible with it.

The UK government took an unusual step by deciding that the network licences should be awarded to commercial companies rather than a state owned monopoly like the G.P.O. Cellnet and Racal Vodafone were the successful bidders.

TACS operates in the 900 MHz band and uses analogue FM. The frequency bands are divided up using FDMA upper and lower bands at 935-960 MHz and 890-915 MHz allow 1000 channels per band with a nominal channel spacing of 25 kHz. The channels are employed as duplex pairs so that if a mobile is receiving a signal in channel 3 of the upper band, it will be transmitting on channel 3 of the lower band. The send and receive signals will therefore always be separated by 45 MHz. Dedicated control channel pairs are included to enable the network to handle mobile registration, 'handover' as the mobile passes from one cell boundary to another, output power adjustment, call tariff logging and more.

A typical cell size has a 1 km radius in urban and 15 km radius in rural environments. Maximum base station power is 100 W and mobiles are restricted to 10 W output power. To handle the ever increasing subscriber base, extra frequencies were later introduced to boost the network capacity. This system became known as *ETACS*. At the time of writing, Cellnet have marked the discontinuation of its analogue network and Vodafone plans to follow suit in the summer of 2001.

WWW Information sources

Organisations:

UMTS-Forum	http://www.umts-forum.org
UMTS-Licence information	http://www.umts-forum.org/licensing.htm
International Telecommunication Union	http://www.itu.int/imt
European Telecommunications Standards Institute	http://www.etsi.org/umts
UK Regulatory authority	http://www.spectrumauctions.gov.uk
Universal Wireless Communications Consortium	http://www.uwcc.com
Third Generation Partnership Project	http://www.3gpp.org

Manufacturers and network operators:

BTCellnet	http://www.btcellnet.net
Ericsson	http://www.ericsson.co.uk
Nokia	http://www.nokia.com
Alcatel	http://www.alcatel.co.uk
Motorola	http://www.motorola.co.uk
Nortel	http://www.nortelnetworks.com
Nippon Telecom & Telegraph Corporation	http://www.nttdocomo.com
Orange	http://www.orange.co.uk
Philips	http://www.pcc.philips.com
Sagem	http://www.Sagem.com
Vodafone	http://www.vodafone-retail.co.uk

During 1990 the GSM Phase 1 specification appeared after its 11 year incubation period with the ETSI. This was the beginning of the 2G or second generation mobile phones and uses digitised speech. One year later Vodafone launched the first U.K. GSM network, and by 1992 coverage was beginning to extend from the large cities and airports into rural areas. The frequency bands used are identical to the TACS system (400 TACS channels were deliberately reserved for the introduction of GSM). In 1993 and 1994 One2one and Orange respec-

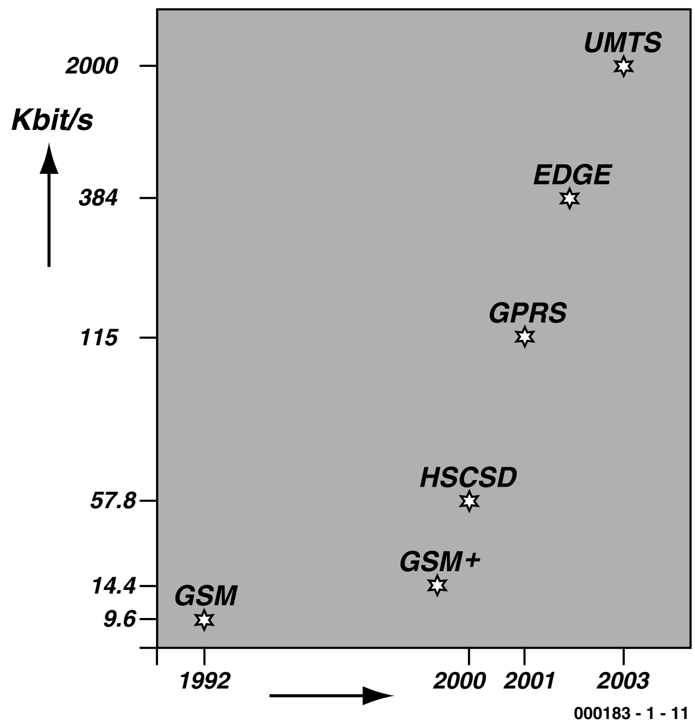


Figure 1. Evolution of GSM to UMTS.

tively launched their GSM system, this time using GSM1800 (otherwise known as PCN or DCS1800) operating in the 1.8 GHz band. 1994 also saw the introduction of the BT Cellnet GSM network in the 900 MHz band. Most current mobiles offer dual-band operation and are therefore compatible with both. This dual-band capability is useful if you intend using the mobile overseas in one of the 110 countries that have adopted GSM, or if your network provider decides in the future to use both bands to solve capacity problems.

In North and South America and in Canada things are a little different. Here GSM1900 operates at 1.9 GHz (also known as PCS1900). Some dual-band mobiles work at GSM900 and GSM1900. Alternatively a tri-band mobile will cover every eventuality. Like TACS, GSM 900 uses FDMA to divide the frequency bands. This time however each of the two bands is divided into 125 channels with a bandwidth of 200 kHz per channel. The 45 MHz duplex spacing is maintained. GSM1800 has 374 channels with a duplex spacing of 95 MHz. In addition to using FDMA to divide up the frequency band into channels, TDMA is used to chop up each channel into contiguous time frames of 4.615 ms. Each frame is further divided into 8 time slots of 0.577 ms. When a mobile is making a call it will be allocated one slot in each frame for sending its digitised voice. Likewise one slot in the receive band is claimed for receiving digitised voice. Mobiles in stand-by mode will only need to listen to certain frames in a paging channel.

Modern GSM mobiles comprise of two to four highly integrated chipsets. A fully integrated base band chip is under development and coupling this with an HF stage and a few components will allow handsets to shrink in size even further. A relatively new concept is the Dual-mode mobile. This is a GSM mobile phone that can automatically switch into a DECT cordless phone when you are in range of a home base unit (or in the office through the company PABX), thereby making cheaper calls through your cordless phone base unit over a land line. Phones offering this feature have been introduced by Sagem and BT Cellnet.

The success of GSM has been much greater than first anticipated, current estimates indicate approximately 400 million GSM subscribers worldwide. Forecasts for the year 2010 indicate 1.8 billion subscribers, the greatest growth expected in Asia. In Europe in the coming years the increasing use of multimedia features means that the relatively low data rate of GSM will not be sufficient. We shall look at some of the interim solutions that have been developed to increase the GSM data rate. Its eventual replacement by UMTS will of course allow a maximum data rate of 2 Mbit/s.

One characteristic of multimedia applications is that the volume of data flowing to and from a subscriber is not symmetrical. Unlike speech communication, browsing the Internet will produce far more information passing from the base station to the subscriber (downlink direction) than in the uplink direction. Also unlike speech, the transfer of the data is not real-time critical, it can be sent in

Table 1. Mobile radio networks in the U.K.

Generation	Network	Operating	Frequency range	Channel pairs	Channel spacing	Duplex spacing	Modulation
-	System 1	1965	163 MHz	-	100kHz	-	FM, analogue
-	System 3	c.1970	163 MHz	-	25 kHz	-	FM, analogue
1 G	TACS	1985-2000	890 - 960 MHz	1000	25 kHz	45 MHz	FM, analogue
2 G	GSM	1991	880 - 960 MHz	173		45 MHz	GMSK, digital
	PCN(GSM1800)	1993	1710 - 1880 MHz	374	200 kHz	95 MHz	GMSK, digital
3 G	UMTS	2002 ?	1885 - 2025 MHz, 2110 - 2200 MHz	12	5 MHz	190 MHz	WCDMA, digital

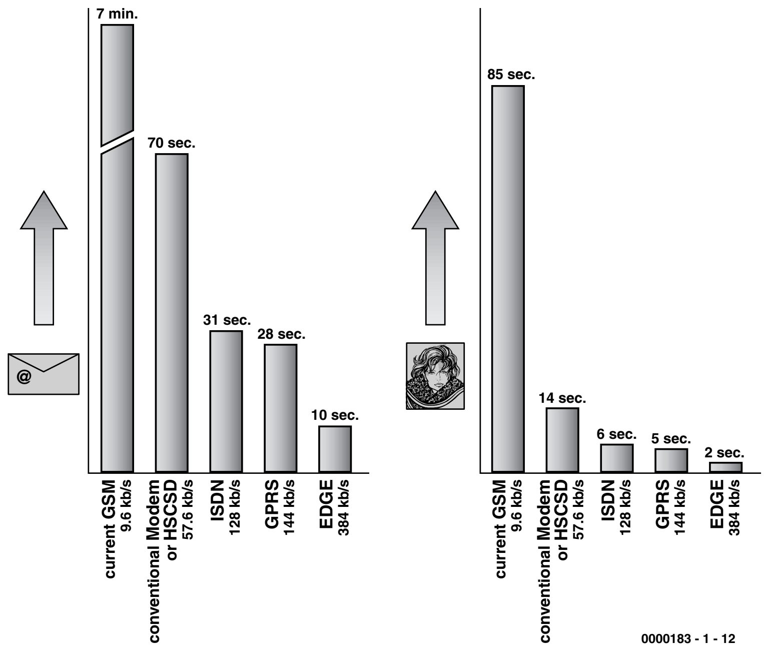
packets which do not necessarily need to be transferred in the correct order. The capacity of the uplink and downlink can be adjusted according to needs.

From GSM to UMTS

The thirst for increased data capacity becomes more apparent year by year. WAP phones are the current fashion but they do not offer any increase in data capacity over the standard 9.6 kb/s. Network providers are implementing various techniques to increase data capacity and bridge the gap between the 9.6 kb/s offered by the 2G GSM phones and the 2Mb/s offered by future 3G phones (see **Figure 1**). Handsets that can use these techniques are known as 2.5G phones. Each of these methods is compatible with the GSM network and does not cause problems with existing handsets.

A standard GSM channel achieves a data rate of 9.6 kb/s with just one time slot in the frequency channel containing 8 timeslots. Altering timeslot allocation is just one method that can be used to increase data capacity. Using a technique called *HSCSD* or *High Speed Circuit Switched Data*, is the simplest and cheapest method for the network operator to increase capacity. It requires only a software update at the base stations. Firstly a more efficient data coding method is used to increase the data rate of each GSM channel from 9.6 kb/s to 14.4 kb/s (so called *GSM+*). Next, four of the eight TDMA timeslots are allocated to one subscriber. This gives a data rate of 57.6 kbit/s. For the user with an HSCSD compliant mobile however this technique has a few disadvantages. Firstly, it takes about one minute to set up the HSCSD call. Secondly, once the connection is established, it acts as a quasi fixed-line in that the four channels are allocated to the subscriber for the duration of the call. The call tariff takes this into account and this method turns out to be expensive for the subscriber. This technique has not been widely deployed in the U.K. but Orange are currently using this method on some of its services.

Another 2.5G system is the *GPRS* or *General Packet Radio Service*. This is a specification produced by the ETSI to provide packet data transfer in a GSM network. This technique can provide a maximum data rate of 171.2 kb/s i.e., three times faster than current modems using a land line! The increased speed is achieved by utilising all 8 slots in a GSM TDMA frame. These would normally be assigned to eight different subscribers. GPRS is actually a universal packet switched data service which overlays the GSM system. It is responsible for packetising data, spreading it



out over free time slots and reassembling it all again at the other end i.e., in the mobile.

A novel feature of this system is its totally different approach to phone use and call billing. The phone is typically switched on in the morning to log on to the network and remains connected for the whole day. Call charges are based on the volume of data sent rather than the connect time. This 'always-on' feature allows almost instantaneous email and web browsing.

BT Cellnet launched their GPRS network in June 2000. Offering a data rate of 28 kb/s, this should be up to 96 kb/s by the new year. Nation-wide coverage was anticipated by September 2000. Orange and Vodafone plan to begin their service before the year end and many European countries have announced similar timescales so GPRS roaming in Europe will be possible. Detractors of the system have pointed out that the claimed data rate of this method is overly optimistic. What happens to all those other 7 subscribers whose timeslots you stole? Well, it seems as though multimedia data flow is typically not continuous but 'bursty'. A screenful of data or a file for example will be downloaded in a burst of activity and then … silence as we scratch our head and decide which option to select. The silences are used to service other subscribers.

Further enhancement to GSM can also be achieved by using a more efficient method of signal modulation. GSM uses GMSK or Gaussian Minimum Shift Keying, changing this to 8-PSK or eight times Phase Shift Keying increases the capacity of each channel from 9.6 kb/s to 48 kb/s. This technique is used for *EDGE*, or *Enhanced Data rates over GSM Evolution*. Peak data rates of 384 kb/s can be achieved with this system because like GPRS it will use all 8 time slots to send data. GSM/EDGE transceivers will be able to switch between the two modulation modes dynamically to ensure backward compatibility for existing handsets. EDGE provides an evolutionary path from GPRS to UMTS by combining the packet data transfer of GPRS with the modulation method that will be used in the future with UMTS. Integration of this system will be less costly than building the completely new network necessary for UMTS. This system will be offering high speed data links for GSM users before and after the introduction of UMTS and we can be sure to see dual band/dual mode handsets operating between GSM and UMTS in the near future.

(000183-1)

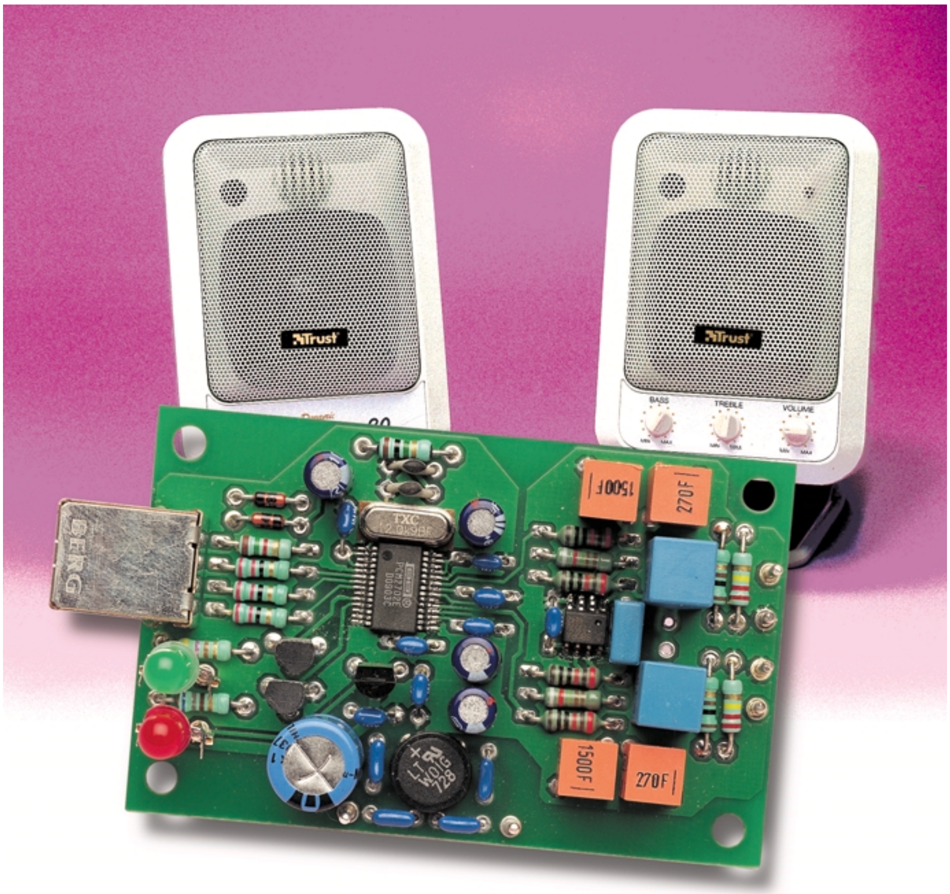
In the second and final part of this article we look at the frequency allocation, licensing and technical background to UMTS.

USB Audio-DAC

A mini PC Sound Card

Design by T. Giesberts

This circuit is pre-eminently suitable for PCs and laptops that are not equipped with a built-in sound card but do have a USB interface. However, the 'mini PC sound card' can also be useful for other applications and various experiments.



The introduction above provides a nearly complete character outline for the circuit under consideration. In brief, it is a 16-bit stereo version of a digital to analogue converter with USB interface. The converter may be used for a range of audio applications. But the first thing that comes to mind is an external sound card for PCs and laptops without such a card, or as an expansion card in the event a second one is desired.

The circuit is conspicuously compact and simple by design, because it consists mainly of an integrated D/A-converter supplemented by two opamps. This simplicity however, does not imply that the quality of this 'USB audio-DAC' has been compromised. On the contrary, we would like to say. A short profile:

The DAC possesses an integrated USB interface that complies with version 1.0 of the standard. It accepts 16-bit stereo and monaural USB data streams and is equipped with an 8× oversampling digital filter. The circuit contains an 'Enhanced multi-level delta-sigma modulator' and is compatible with sampling rates of 32, 44.1 and 48 kHz. In addition, it has a digital attenuator, a soft-mute function as well as optical Suspend- and Playback-indicators. And finally, there is no need for additional drivers when used with Windows 98 (and later).

The description above sounds all very nice, but it is the end result that this all leads to, that is more important. Well then, in a separate table, all the measuring results are summarised. Examining the results with more than a cursory glance, you will have to reach the conclusion that the performance is very reasonable.

PCM2702

The heart of the circuit consists of a PCM2702, a two-channel single-chip D/A-converter with integrated USB interface controller. **Figure 1** shows the block diagram detailing the internals of the IC. For more information you may refer to the (condensed) datasheet printed else-

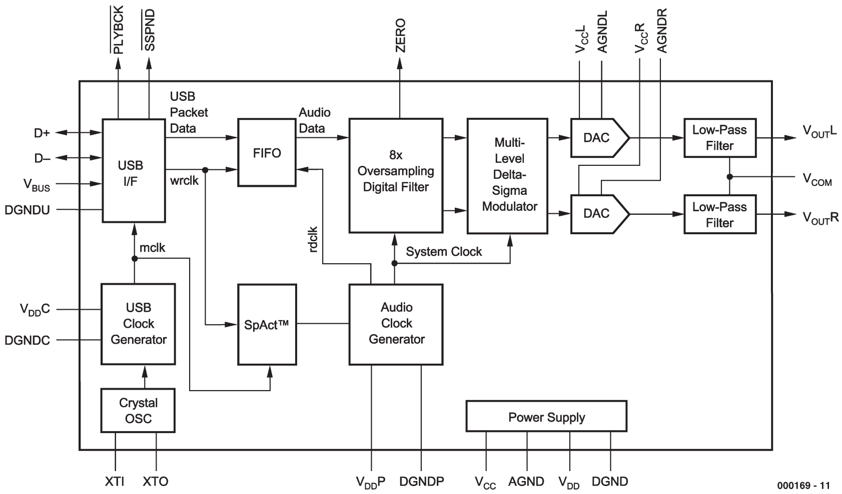


Figure 1. Block diagram of the single-chip D/A-converter PCM2702.

where in this issue.

Control and audio data are both transferred to the PCM2702 via pins D+ and D–. All data to and from the IC are transferred at full speed. V_{BUS} (pin 8) and DGNDU (pin 9) are also connected to the USB bus. V_{BUS} does not consume any power and is used only to detect the connection to the USB bus.

The PCM2702 has two interfaces. Each interface is constructed by some specific setting. Interface #0 has only one setting. This setting describes the standard audio interface

The most important three connections are the input and output terminals, and the, so-called Feature Unit.

The input terminal is defined as 'USB stream'. This input accepts two channel audio data streams. The output terminal is defined as a 'speaker'. The Feature Unit supports volume and mute control.

The built-in digital volume control can be adjusted from 0.0 dB to –64.0 dB in steps of 1 dB. Each channel can be set independently, but master control of both channels simultaneously is also possible. Mute is only available in master control.

Interface #1 has three alternative settings. Setting #0 is the 'Zero Bandwidth' setting (a kind of mute function). Setting #1 selects 16-bit stereo and #2 is the 16-bit monaural setting.

The PCM2702 requires a 12 MHz clock for USB and audio functions.
The on-chip oscillator with an external crystal (which is how it is used here) may generate this clock. Or, if desired, an external clock signal may be applied to pin XTI instead.

The IC includes an internal power-on reset circuit, which automatically initialises the digital logic when the power supply is switched on. The PCM2702 is ready to receive audio data after completion of the reset sequence and a connection to the USB bus.

The Circuit

The circuit (**Figure 2**) has largely been copied from the evaluation board DEM-PCM2702 from Burr-Brown. As can be observed, besides the PCM2702, the active components are limited to a dual opamp, a couple of general-purpose transistors and a voltage regulator. A handful of standard parts completes the circuit.

We refrained from fitting a power supply connector and mini-jack on the PCB. As a consequence, either a mains power adapter or a small PCB transformer (voltage >8 V, power >0.5 VA) may be used. A mains power adapter with regulated output is preferred, because the ripple suppression ratio of the 5 V regulator 78L05 (IC3) on the PCB is only 50 dB.

To connect active PC speakers, for example, a mini-jack or cinch connectors may be used. Cinch connectors permit the connection of a normal power amplifier without the need for a special adapter cable.

Because of the relatively large supply current (approximately 60 mA) it was not considered appropriate to power the circuit from the USB bus (since the maximum current permitted is only 100 mA, a second device would increase the current consumption to unacceptable levels). The PCM2702 has good power supply rejection which makes addi-

Measurement Results

(measured at 0 dB, 44.1 kHz unless otherwise indicated)

input signal		USB audio data
nominal output voltage		62 % V_{cc} (1.096 V_{eff} bij 5.000 V)
bandwidth (10 kΩ load)		5 Hz to $f_s/2$ (f_s = 32/44.1/48 kHz)
amplitude at 20 kHz		–0.25 dB
analogue filter bandwidth		30 kHz (2nd order 0.25dB-Chebyshev)
output impedance		100 Ω
signal to noise ratio		> 101 dBA
THD+N	(1 kHz, B = 80 kHz)	< 0.0035 %
	(20 kHz, B = 80 kHz)	< 0.025 %
IMD	(60 Hz/7 kHz = 4:1)	< 0.006 %
channel separation		> 116 dB
stopband attenuation digital filter		> 82 dB
current consumption		< 60 mA

The distortion figures are, without exception, very good. The numbers for signal to noise and channel separation are no worse. The frequency response (measured using a test CD) of the sound card is remarkably flat. For the benefit of the purists, this graph has been reproduced here. The only irregularity consists of a small amplitude increase that is caused by the Chebyshev output filter. It appears like a large 'bump', but when closely examining the scale of the graph it is realised that the ripple is less than 0.25 dB!

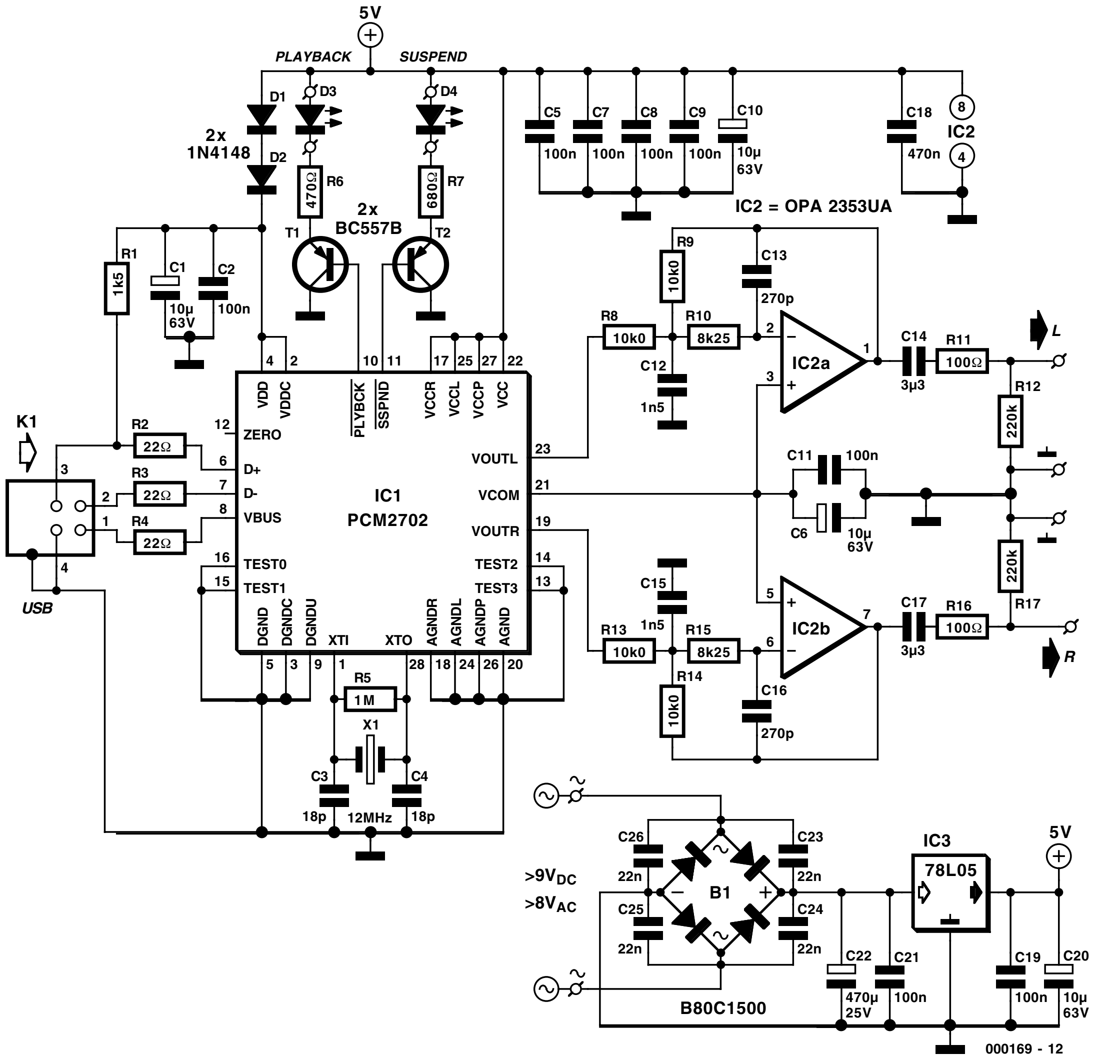


Figure 2. The complete schematic of the USB audio-DAC is remarkably simple.

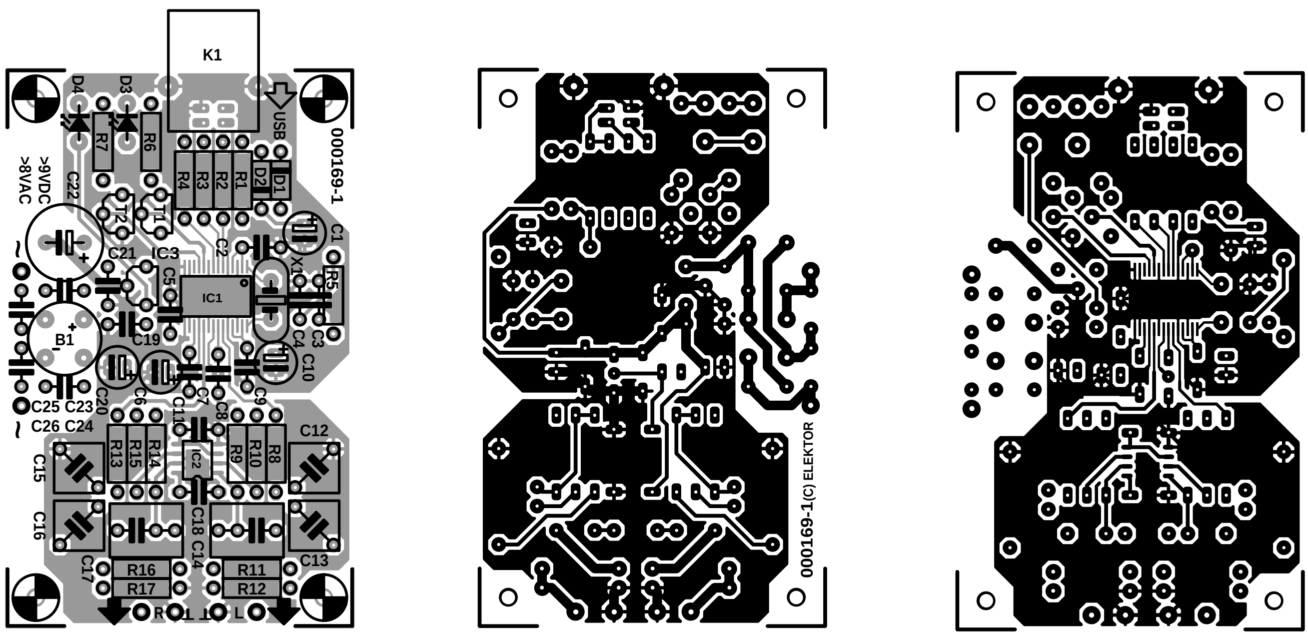


Figure 3. The PCB is very compact, in part because IC1 and IC2 are SMD devices.

tional SMD capacitors unnecessary. RF decoupling is provided by standard 100 nF ceramic capacitors (with 5 mm lead pitch).

With the crystal oscillator, keep in mind that C3 and C4 act as C_{load} for the attached crystal (here C_{load} = C3/2 + $C_{parasitic}$ (when C3=C4); if necessary, the frequency may be measured at the XTO pin and is allowed to deviate by at most 500 ppm).

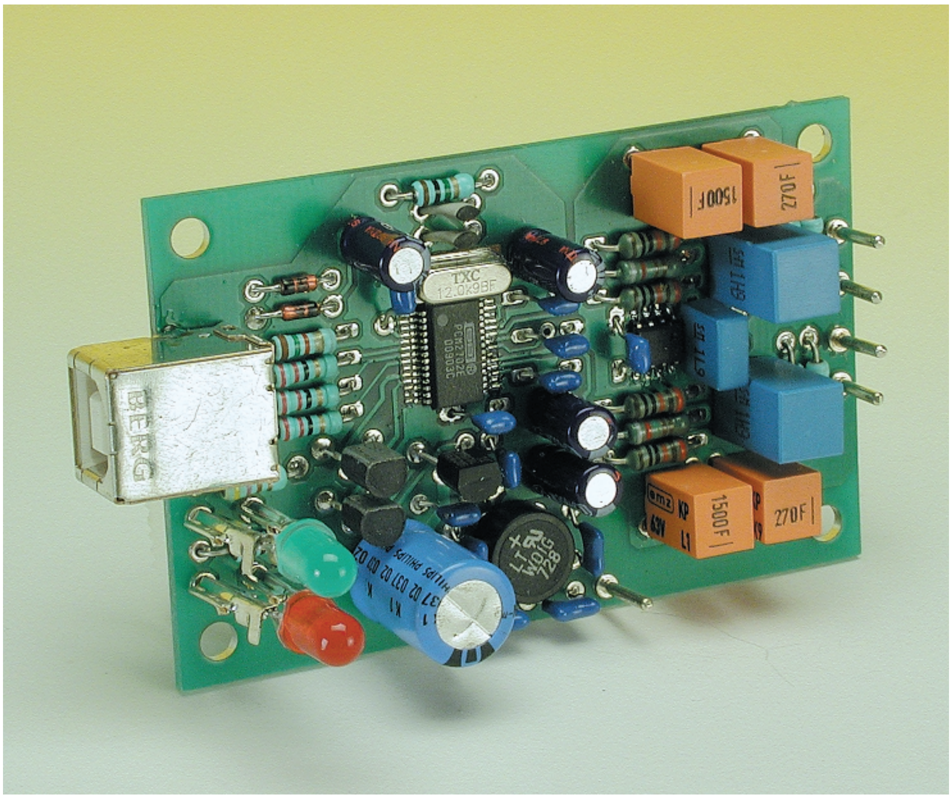
This audio DAC, for a change, does not use Sallen-Key filters, but uses Multiple Feedback filters (MFB) instead. The output filters are DC coupled. This is possible because the Vcom output biases the opamps to exactly half of the power supply rail. An additional advantage of the MFB filters is the relative insensitivity to component tolerances. C11 provides RF decoupling of the bias voltage. Because these are only 2nd order filters, the Chebyshev type with 0.25 dB ripple was selected. This is a compromise between the smaller bandwidth of the analogue output filter (better suppression of the mixing products in the audio band) and quality.

The amplifier selected for the output filter is an OPA2353UA. This is a high speed, single supply, rail to rail, low noise CMOS opamp with a maximum output current of 40 mA. As a result of using this amplifier, an additional power supply is avoided and the output filter can simply be powered from the same 5 V that is already present.

The outputs have to be DC decoupled, of course. This is achieved with C14/C17 (3.3µF MKT; check that the pitch is either 5 or 7.5 mm!), R11/R16 ensure that capacitive loads will never be able to cause trouble.

R12/R17 make sure that C14/C17 are always charged. This prevents a loud 'pop' if the loudspeakers are connected to the circuit after this has already been powered up.

Two LEDs indicate the status of the circuit.



COMPONENTS LIST

Resistors:
R1 = 1kΩ5
R2,R3,R4 = 22Ω
R5 = 1MΩ
R6 = 470Ω
R7 = 680Ω
R8,R9,R13,R14 = 10kΩ0 1%
R10,R15 = 8kΩ25 1%
R11,R16 = 100Ω
R12,R17 = 220kΩ

Capacitors:
C1,C6,C10,C20 = 10µF 63V radial
C2,C5,C7,C8,C9,C11,C19,C21 = 100nF ceramic, lead pitch 5mm
C3,C4 = 18pF
C12,C15 = 1nF5 1% polystyrene/polypropylene
C13,C16 = 270pF 1% polystyrene/polypropylene
C14,C17 = 3µF3 MKT (Siemens), lead pitch 5 or 7.5mm
C18 = 470nF, lead pitch 5mm
C22 = 470µF 25V radial
C23-C26 = 22nF ceramic, lead pitch 5mm

Semiconductors:
D1,D2 = 1N4148
D3 = green high-efficiency LED
D4 = red high-efficiency LED
B1 = B80C1500 (80V piv, 1.5A peak), round case
T1,T2 = BC557B
IC1 = PCM2702E (SSOP-28 case) (Burr-Brown)
IC2 = OPA2353UA (SO-8 case) (Burr-Brown)
IC3 = 78L05

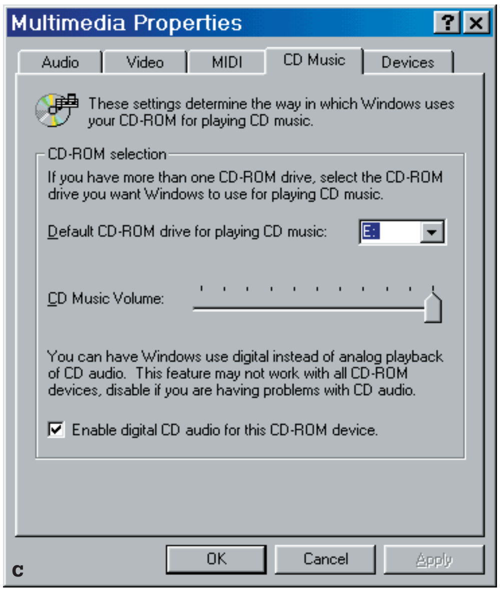
Miscellaneous:
K1 = USB connector, receptacle style B (PCB mount)
X1 = 12MHz quartz crystal
PCB, order code **000169-1** (see Readers Services)

Filter software

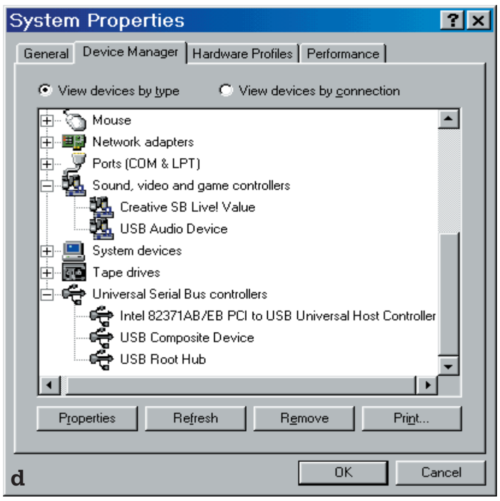
On the Internet, at the Burr-Brown website (http://www.burrbrown.com/applications) the program FilterPro may be found. The program that may be downloaded is called filter.zip, which contains, among other things, filter2.exe. This is a DOS program that enables the calculation of the well-known Sallen-Key filters as well as the MFB filters that are used here. Several parameters may be adjusted and the exact values of the E-96 series resistors can be calculated. The capacitors are automatically adjusted for the selected scaling resistors (E-12 values), but it is also possible to choose your own values, this is particularly useful when using previously measured capacitors. Those who disagree with the choice of the Chebyshev filter for this circuit, may modify the filter themselves to, for example, a Butterworth or a Bessel type.

a

b



c



d

Figure 4. The installation on the PC is not particularly difficult. The text provides a number of useful hints.

D3 is lit up when the PCM2702 is playing back audio data (Playback). D4 is lit up when the USB interrupts the audio data stream to the PCM2702 (Suspend).

Construction

The PCB layout for the USB audio DAC is shown in **Figure 3**. It is remarkably compact, having dimensions of only 46 × 74 mm.

Both the PCM2702 and the dual opamp OPA2353VA are surface mount devices. A considerable amount of soldering skill is required. The problems are only minor when soldering IC2, but for IC1 a soldering iron with very small tip is absolutely necessary (the pins have only a 0.65 mm pitch!). Use desoldering wick to remove any excess solder. To prevent overheating of the IC, allow it to cool down from time to time. This also applies to the soldering operation.

Once both SMD ICs have been carefully soldered to the PCB, the remainder of the parts may be fitted. These are, without exception, standard components of which there are only a very small number, so no problems are anticipated here.

Installation

We installed the circuit on a PC with Windows98SE. Windows98 (and later) contains drivers for USB audio playback as standard.

For testing purposes it is possible to power the circuit using a 9 V battery. The red LED (D4) should light. Once the connection to the USB cable is made (cable type A to B), the red LED should extinguish immediately and a window should pop up with the message 'Burr-Brown Japan PCM2702'. This is followed by the window 'Add New Hardware Wizard' which will want to install the drivers for a 'USB Composite Device' (refer to **Figure 4a**). Once the drivers are installed, the Wizard window will reappear which now will want to install a 'USB audio Device' (**Figure 4b**). Having completed all of this, there should be, in Control Panel, under System Properties in the Device Manager under 'Sound, video and game controllers' a 'USB audio device' and under 'Universal Serial Bus controllers' a 'USB composite device' (**Figure 4c**). If another sound card is already installed, then using Control Panel under Multimedia Properties in the Audio tab it is now also possible to select 'USB Audio Device' as the Preferred device for Playback.

We assume that every PC these days is equipped with a CD-ROM drive. With a regular sound card, the analogue output from the CD player is normally directly connected to this card. This permits the listening to, or the processing of, audio-CDs using the PC. If you would like to listen to audio CDs using the USB audio DAC, then, in Multimedia Properties in the 'CD Music' tab tick 'Enable digital CD audio for this CD ROM device' (refer **Figure 4d**). There are now four controls in the 'Volume Controls' window: one for 'speaker' (master), one for 'CD Player', one for 'Wave' and one for 'SW Synth' (for MIDI, a Software Wavetable Synthesiser).

Practical Hints

There are older generations of PC motherboard (not ATX) that do possess a USB controller but do not have the appropriate connectors fitted. Separate USB-dual-brackets are available (standard with two USB connectors type A) with an 8 or 10 way socket that may replace an unused expansion slot cover. If no expansion slots are available, then it may be possible to remove the bracket with the 25 way printer port and 9 way serial port and move these to the break-out openings in the computer case instead. The USB bracket may then be fitted in the newly created space.

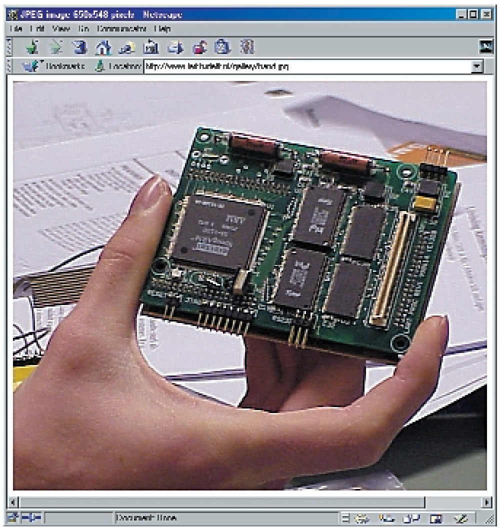
There are three variations (there could be more) of the socket for this bracket, so pay close attention to the header pinout. The connections in the socket are usually easily swapped around. By carefully lifting up the plastic locking tabs and simultaneously pulling on the wires, the individual connectors can be removed from the socket.

(000169-1)

Linux Advanced Radio Terminal (LART)

Build your own mini computer

In 1998, as part of the MMC project, researchers at the Technical University of Delft, the Netherlands, started to design a small but powerful computer system for wireless multimedia experiments. The requirements: low power consumption, inexpensive and with plenty of computing power.



The acronym MMC means Mobile Multimedia Communication and embraces a multidiscipline project at the Technical University of Delft in the Netherlands. One of the results of the research programme was the LART computer which was designed simply because no ready to go computer system could be found that could meet the requirements posed by the MMC project. Since then, LART has grown into a full-blown mini computer with its own Internet website, extension boards and software.

LART stands for Linux Advanced Radio Terminal. The LART mainboard has a size of just 10 × 7 cm and contains a Digital SA-1100 StrongArm CPU ticking at 220 MHz, 32 Mbytes of EDO RAM and 4 Mbytes of Flash memory. Power consumption is less than one watt while the computing power is said to exceed 200 Mips. The LART board can run in stand-alone mode, starting the operating system from Flash memory. The 4-Mbyte Flash memory is large enough to hold a bootstrap loader, a compressed Linux kernel and a compressed RAM disk. Thanks to high-efficiency on-board voltage converters, the board can work off any supply voltage between 3.5 V and 16 V. All essential connections to the board are available on two connectors. To complete the design, there is a serial connector that can be configured as an RS232 port.

All LART hardware and software may be found on the dedicated LART website. Here you find circuit diagrams, PCB layouts, software and extension modules.

Besides the main board, there are two important extensions: a kitchen Sink Board (KSB) and an Ethernet Board. The latter is a 10-Base T adapter for the LART board, designed around the CS8900A Ethernet chip from Crystal. The KSB contains all hardware extensions to turn the LART into a full-blown multimedia computer. On the KSB you'll find a stereo 16-bit audio output (44.1 kHz), an IDE/ATA interface, two PS/2 connectors for mouse and keyboard, and connectors IRDa, USB, video, etc.

The software consists mainly of Linux sources and binaries. Knowledge of Linux is a prerequisite. Besides all hardware and software, the LART site also contains an FAQ list and an overview of projects in which a LART has been applied.

Those of you who are now eager to start building a LART straight away should be warned that the most hardware is built from SMA parts which may not be easy to source locally. What's more, the board consists of five layers which make it almost impossible to produce as a one-off. Fortunately, email addresses of people willing to assist you may be found in the LART FAQ.

(005154-1)

LART pages:

http://www.lart.tudelft.nl

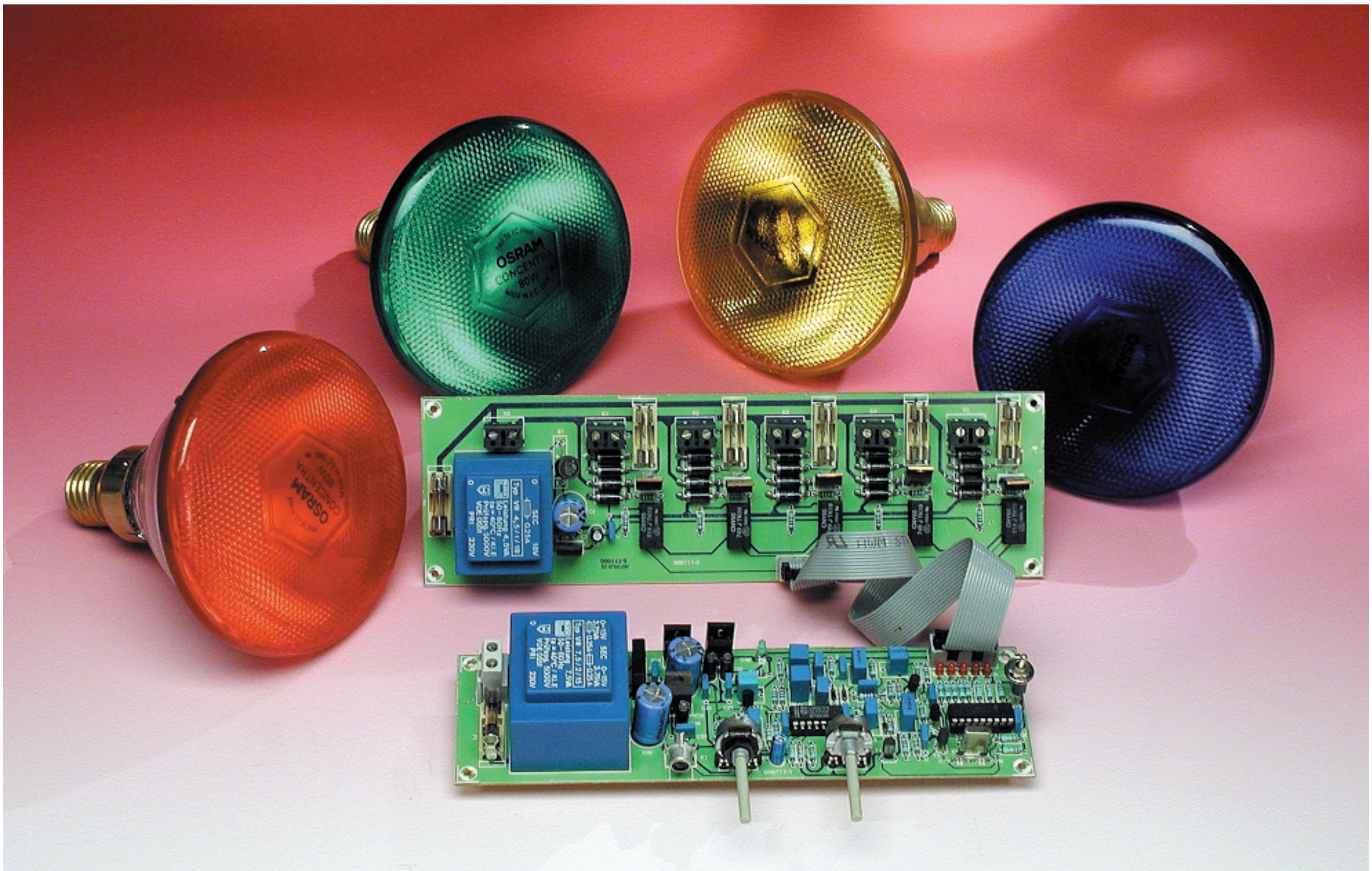
Sound-to-Light PLUS

Microprocessor-driven lights effects

Design by Ron Wouters

Text by David Daamen

The use of lights effects units in conjunction with music is presently very popular. Of course we don't want to stay behind at Elektor. This design uses a PIC16F84 and has sound-to-light, running lights modes and 'beat detection'.



Please note: prototype shown here differs slightly from the PCB layouts shown in Figures 3 and 4.

Most people will remember the running light units which typically used a 4017, a digital decade counter whose outputs go high in turn when a clock signal is applied. And what about the classic sound-to-light unit: three lights, each flashing according to the sounds in different bands of the audio spectrum. Those were sound-to-light units in their most elementary form. But in the age of the microprocessor it all has to be faster, better and offer more functionality. And that is exactly what this circuit does.

This circuit certainly offers a lot of functionality, since it combines the running lights and sound-to-light modes. The running lights do more than just go forwards and backwards, there is also a choice of several pre-programmed patterns. The unit also has a beat-detection circuit.

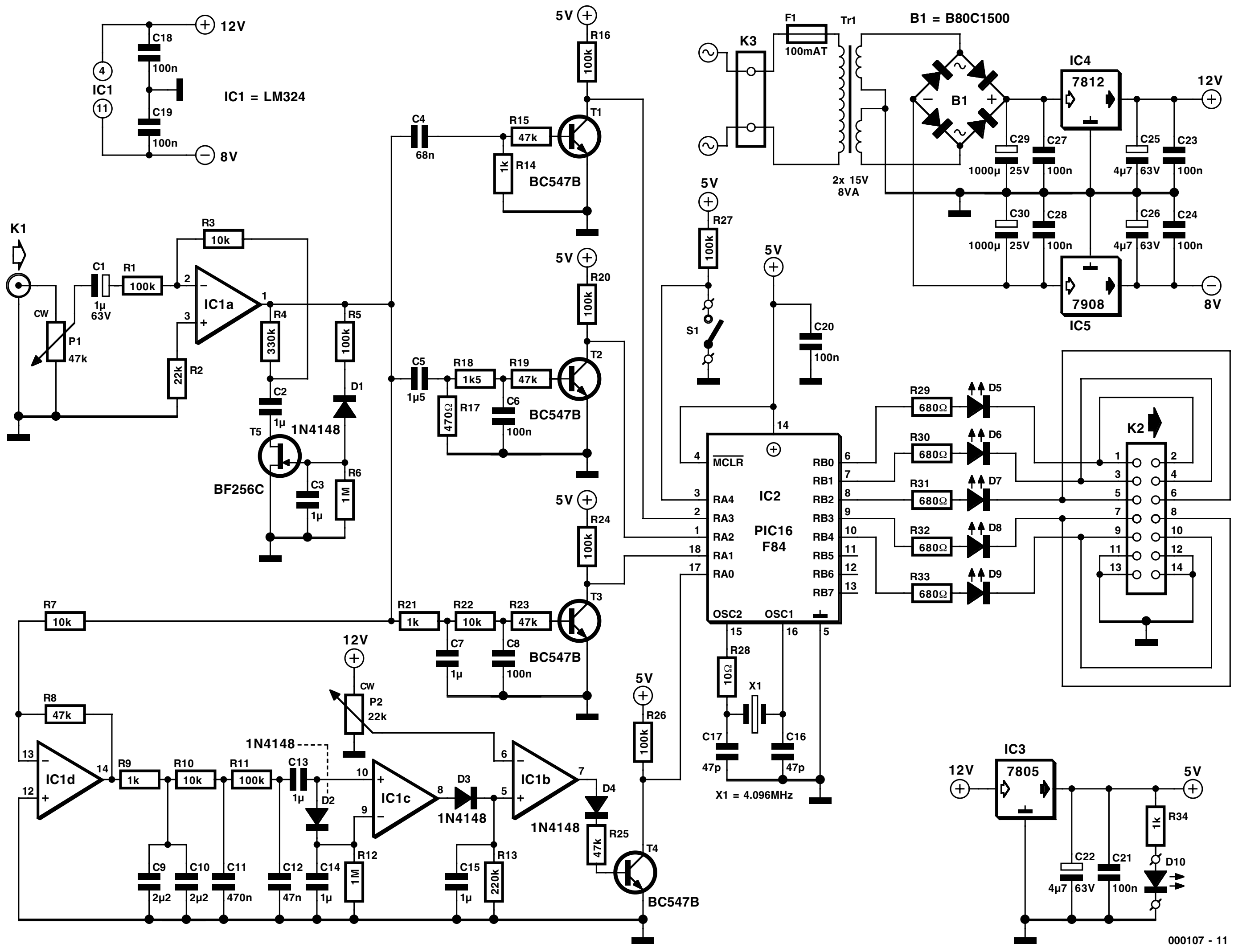


Figure 1. The sound processing and control section.

The change of pattern of the running lights is determined by the beat of the music. All this can be can be realised in a simple manner by using a PIC16F84 to control the lights. This microcontroller is inexpensive, fast, has a large number of I/O pins and can be programmed easily.

Operation

As we mentioned earlier, the circuit consists of a running-lights as well as a sound-to-light unit. The program has been implemented such that the PIC switches between these two modes every four seconds. In the sound-to-light mode, the lights flash to the beat in three bands in the spectrum of the connected audio signal. The same audio signal is also used in the second mode of operation. Here the beat of the music causes the lights to step through five pre-programmed patterns.

Sometimes the music can have little or no beat to it. When the detection-algorithm cannot find the beat during six seconds, it causes the unit to switch automatically to the sound-to-light mode. In this case, the unit is also prevented from switching back to the running lights mode. The running lights mode will only be enabled again when the beat returns.

It's also possible that no audio signal is present at all. When this is the case there is obviously no point in having the sound-to-light mode on. When no music is detected for a period of 30 seconds the unit starts flashing the lights in a random pattern. Of course the controller will revert to its normal mode of operation as soon as the music signal returns. This way we can guarantee that some lights effects will always be produced.

And finally, we can choose to have the unit work only in the sound-to-light mode.

Circuit

The circuit of the sound-to-light unit consists of two sections. **Figure 1** is the section which deals with the signal processing and which provides the signals for the lamp-drivers. The sound source (for example a CD player or the REC-out connection of an amplifier) is connected to the circuit via K1. P1 is used to set the sensitivity of the input. The amplitude of the signal should be between 0.7 and 1.5 V_{pp}. After P1 the signal is amplified by IC1a and T5. This combination of opamp and FET forms an AGC (automatic gain control) unit that keeps the signal at a more or less constant level. The values of R5 and C3, which determine the behaviour of the AGC, have been chosen such that the beat of the audio signal is kept intact.

The resulting signal is then fed to three separate filters: a low pass filter at 160 Hz (R21, C7, R22 and C8), a band pass filter set between 225 and 1060 Hz for the mid frequencies (C5, R17, R18 and C6) and a high pass filter at 2340 Hz (C4 and R14). Following each filter is an NPN transistor with base and collector resistors, which converts the signal to TTL levels. These signals can then be fed to the PIC for processing.

The signal from the AGC doesn't only go to the three filters for the sound-to-light effect, but also to a beat-detection circuit. IC1d amplifies the signal by a factor of 4.7 and is followed by a low pass filter with a roll-off of 18dB. This filter is designed with a cut-off frequency of 34 Hz, using R9, C9, C10, R10, C11, R11 and C12. The output of the filter goes to IC1c, which is the actual beat-detector. The inverting input of the opamp is driven by a slightly delayed signal via C14 and R12, which causes a series of pulses to appear at its output, with a frequency depending on the audio signal. If the right type of signal is present, the output will follow the beat of the music.

These pulses are integrated by D3 and C15 and then go to IC1b. This opamp is used as a comparator, where P2 sets the sensitivity of the beat-detector. The pulses from the comparator drive T4 via D4 and R25. This transistor also converts the signal to TTL levels.

So now there are four signals available, of which the first three are used in the sound-to-light mode. The fourth signal determines when the lights change pattern in the running lights mode.

The program in IC2 selects the active mode. A ready programmed PIC can be ordered from *Elektor Electronics* Readers Services (order code **000107-41**). For those of you who wish to program the PIC themselves, a diskette is available containing the program (order code **000107-11**). Please note that the project software is not available as a Free Download from our website because the author receives royalty payments on every copy sold of the disk and programmed PIC.

The processor section is very simple. The only external components used by the PIC are R28, X1, C16 and C17, which are used to generate its clock signal. S1 is used to switch between the fixed sound-to-light mode or alternating between the two modes. R27 is used to keep the signal at the input

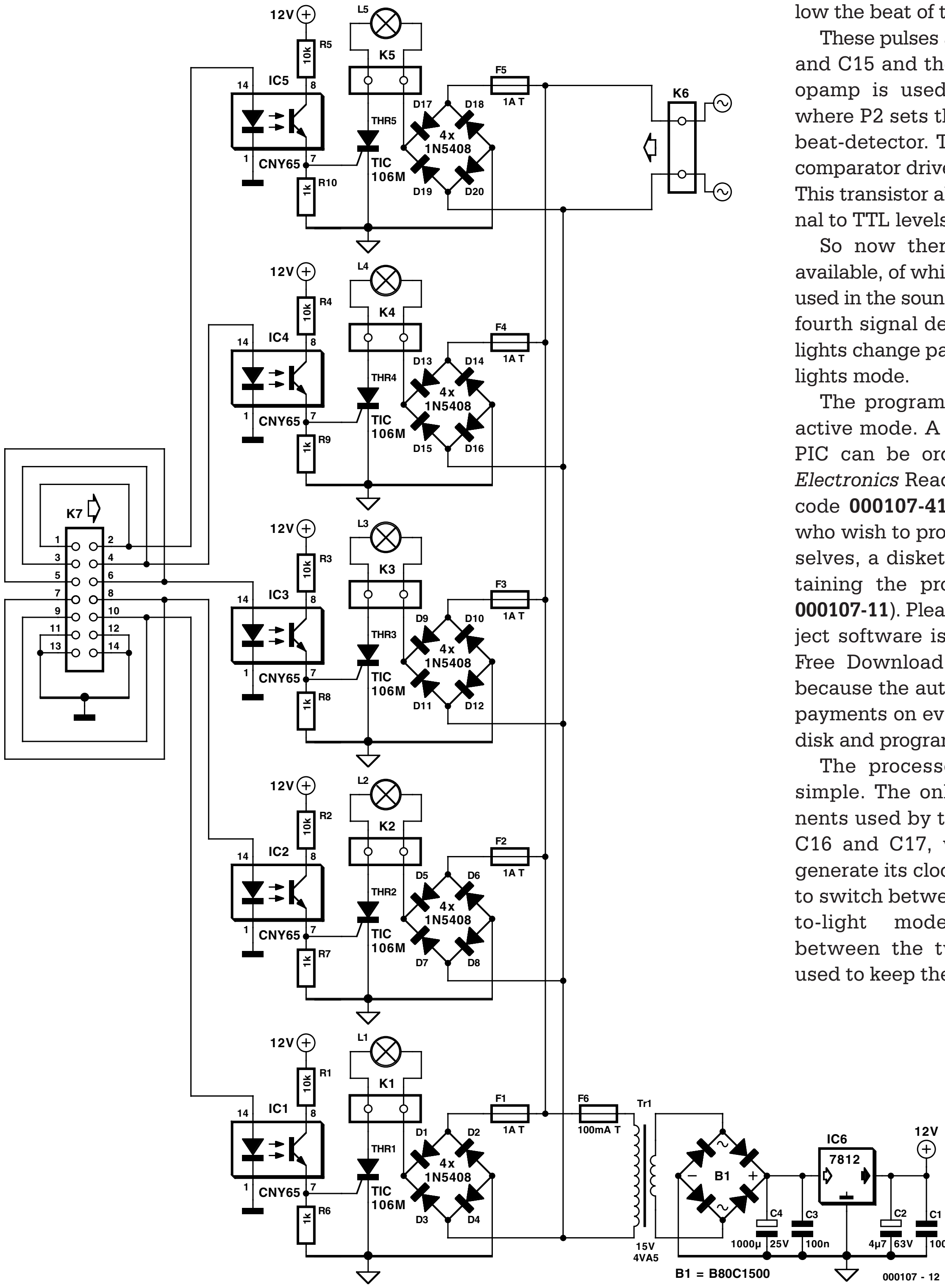


Figure 2. The lamp-driver section.

of RA4 (pin 3) at a defined level when the switch is open.

Five outputs of the PIC (RB0-RB4) are used to drive the lamps. Each output is connected to K2 via a resistor and LED. These signals are connected to a separate board that finally drives the lamps. We've designed a separate board since it is more convenient to place it near the lamps, or even in the same enclosure as the lamps. An ordinary low-voltage cable (with a minimum of 6 cores) can be used to make the connection.

The lamp-driver board (**Figure 2**) has an optocoupler (IC1-IC5) for each channel. In the event of a fault the control board and any connected audio equipment will remain isolated from the deadly mains potential.

When the control board is connected via K7, the LED in each optocoupler will be connected in series with the LED on the control board. LEDs D5-D9 will therefore not only show which lamp should be on, but also that

PIC16F84

Yet again, a microcontroller of the PIC family from Microchip is used in this circuit. It's not surprising that it is a 16F84. This part is ideal for design and development work, because it can be (re)programmed very easily. There is no need to erase the device under UV light. The flash memory is simply overwritten when the device is in its program mode. Once the design has been completed, a different processor can be used for mass production (for example a cheaper version without the flash memory). This is something that we purposely didn't do. It is always interesting to experiment a little and this circuit is perfect for that. There are enough spare inputs left to use. And you're not restricted to switching lamps; many other mains-powered appliances can be connected. You could connect a microphone to the input and use the high pass filter and some modified code to make a whistle detector to switch appliances on and off. And there are probably many other (and more original) applications. For those who want to find out more about the 16F84 or other Microchip processors, their website is at www.microchip.com. On www.thepicarchive.cwc.net you'll find a wide range of PIC-programmers, programming software, program examples and projects. There are also descriptions of 'in-circuit' programmers, so you no longer have to remove the PIC from the circuit in order to program it.

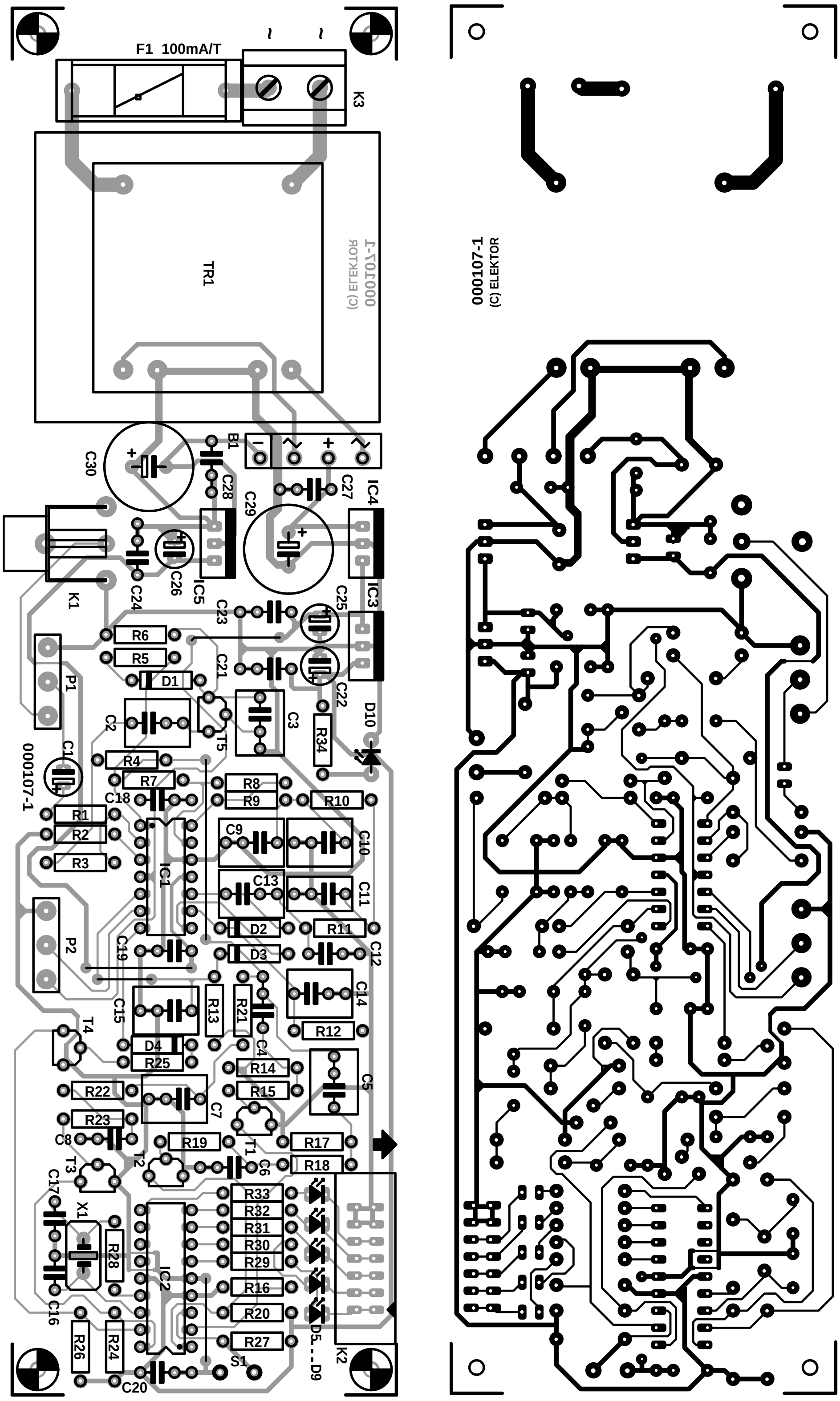
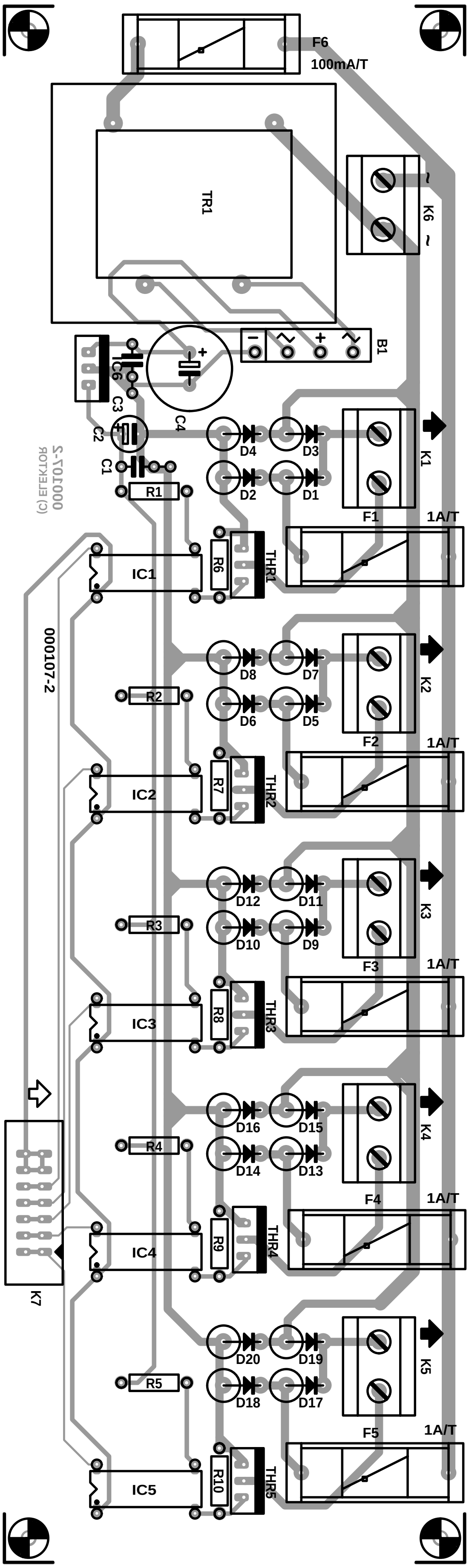


Figure 3. Copper track and component layout of the control PCB.


F6
100mA/T
TR1
K6
B1
C6
C3
C4
C2
C1
R1
D4
D3
D2
D1
K1
F1
1A/T
R6
THR1
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000107-2
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D8
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D6
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K2
R2
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R7
THR2
IC2
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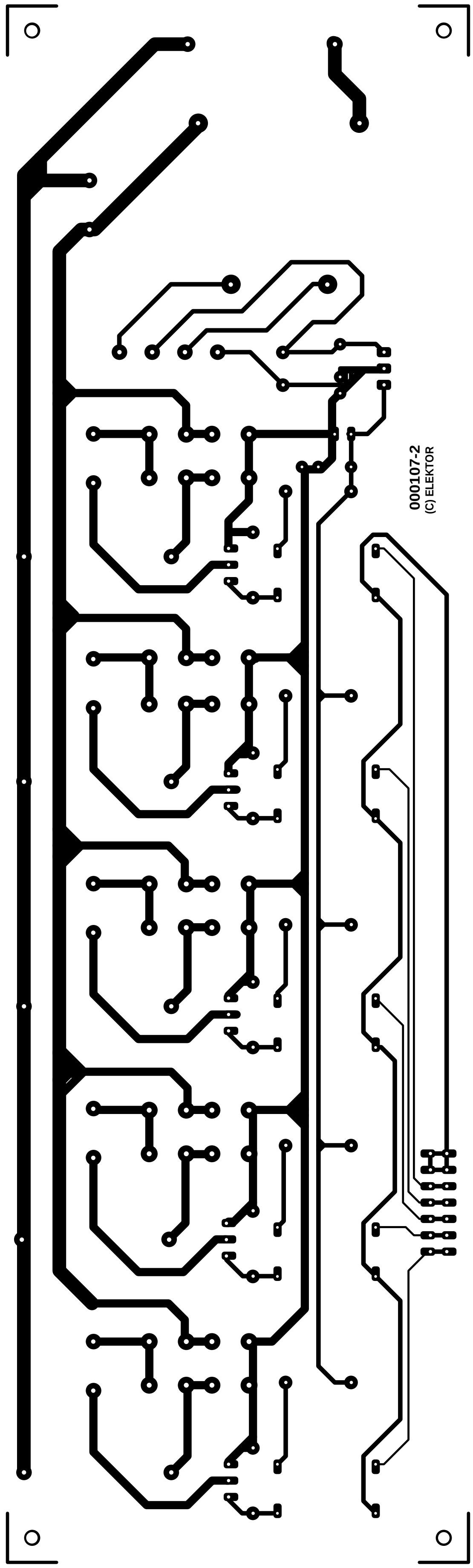
IC5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the connection between the boards is intact.

The optocoupler directly drives the thyristor (Thr1-5) that switches the mains voltage. Since a thyristor can only conduct in one direction, the mains voltage is rectified (D1-D4 and D17-D20). Without these, the lamps would only work at half their power. A fuse finally protects each channel.

Each PCB has its own voltage supply, using standard 78- and 79-series voltage regulators. The negative voltage produced by IC5 is only used by the opamps in IC1.

Figure 4. Copper track and component layout of the lamp-driver PCB.

COMPONENTS LIST

Control Board

Resistors:
R1,R5,R11,R16,R20,R24,R26,R27 = 100kΩ
R2 = 22kΩ
R3,R7,R10,R22 = 10kΩ
R4 = 330kΩ
R6,R12 = 1MΩ
R8,R15,R19,R23,R25 = 47kΩ
R9,R14,R21,R34 = 1kΩ
R13 = 220kΩ
R17 = 470Ω
R18 = 1kΩ5
R28 = 10Ω
R29-R33 = 680Ω
P1 = 47kΩ linear potentiometer, mono
P2 = 22kΩ linear potentiometer, mono

Capacitors:
C1 = 1µF 63V radial
C2,C3,C7,C13,C14,C15 = 1µF MKT, lead pitch 5 or 7.5 mm
C4 = 68nF
C5 = 1µF5 MKT, lead pitch 5 or 7.5 mm
C6,C8,C18-C21,C23,C24,C27,C28 = 100nF
C9,C10 = 2µF2 MKT, lead pitch 5 or 7.5 mm
C11 = 470nF
C12 = 47nF
C16,C17 = 47pF
C22,C25,C26 = 4µF7 63V radial
C29,C30 = 1000µF 25V radial

Semiconductors:
D1-D4 = 1N4148
D5-D9 = red LED
D10 = high-efficiency LED, green
B1 = B80C1500 in rectangular case (80V piv, 1.5A peak)
T1-T4 = BC547B
T5 = BF256C
IC1 = LM324
IC2 = programmed PIC16F84-10/P (order code **000107-41**)
IC3 = 7805
IC4 = 7812
IC5 = 7908

Miscellaneous:
K1 = PCB mount cinch socket (e.g., Monacor/Monarch type T-709G)
K2 = 14-way boxheader
K3 = 2-way PCB terminal block, lead pitch 7.5 mm
S1 = on/off switch, 1 contact
X1 = 4.096MHz quartz crystal
Tr1 = mains transformer 2x15V/8VA (e.g., Monacor/Monarch type VTR8215)
F1 = fuse, 100mA, time lag (T) with PCB mount holder

Lamp Driver Board

Resistors:
R1-R5 = 10kΩ
R6-R10 = 1kΩ

Capacitors:
C1,C3 = 100nF
C2 = 4µF7 63V radial
C4 = 1000µF 25V radial

Semiconductors:
D1-D20 = 1N5408
B1 = B80C1500 in rectangular case (80V piv, 1.5A peak)
IC1-IC5 = CNY65
IC6 = 7812

Miscellaneous:
K1-K6 = 2-way PCB terminal block, lead pitch 7.5 mm
K7 = 14-way boxheader
La1-La5 = 240-V light bulb with socket
Thr1-Thr5 = TIC106M
Tr1 = mains transformer 15V/4VA (e.g., Monacor/Monarch VTR4115)
F1-F5 = fuse 1A, time lag (T) with PCB mount holder
F6 = fuse 100mA, time lag (T) with PCB mount holder
PCB, order code **000117-1** (see Readers Services page)
Disk, project software, order code **000107-11** (see Readers Services page)

Construction

Figures 3 and 4 show the artwork for both PCBs. *Elektor Electronics* supplies them in one piece, with a perforated join. The 14-way connectors are directly opposite each other, making it easy to make the connections if you want to leave the boards together.

Mounting the components shouldn't present any difficulties. The ICs on the control PCB can be mounted in sockets. The optocouplers on the lamp-driver PCB are best mounted directly onto the board. Furthermore, the rectifier diodes on the lamp-driver PCB have to be mounted vertically. The connection between the two boards is made with a 14-way cable with IDC headers at each end. This connection may be up to several metres long.

The two potentiometers and switch S1 are mounted on the enclosure of the control circuit, making them easily accessible.

Take great care that you choose a suitable enclosure for the lamp-driver PCB. It would help if you first read the Electrical Safety Page that is published regularly in Elektor. For the lamp outputs you can use Euro style mains outlet chassis sockets. If you're using earthed sockets, it becomes advisable to build the enclosure to the Class I standard, where the case is connected to the supply protective earth. The mains inlet can be a Euro style chassis plug.

You should stick a copy of the equipment label on to the enclosure, so that it becomes clear which type of fuses should be used (the label shown here applies to an enclosure that houses both boards).

Keep in mind when using the Sound-to-Light Unit that each light output can drive a maximum load of 200 W.

(000107)

On project disk # 000107-II

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Valved RIAA Preamplifier

using an ECL86

Design by G. Haas

In spite of CDs and digital audio signal processing, there are still people who are devoted to the vinyl phonograph record. And they enjoy their analogue treasures the most in the company of contemporary valve technology.

For nearly a century, vinyl phonograph records played an important role as a storage medium for voice and music. Although nowadays the CD has taken over the leading role, analogue records are still widely distributed. Hundreds of millions of records are present in archives and private record cabinets. These include irreplaceable documentary recordings, collectors' items and many commonplace recordings that reflect the spirit of a particular era and the technology available at that time, as well as recordings that are technically and artistically outstanding. For friends of analogue recordings, there is still a market for phonograph records, in which re-pressings or new pressings can be obtained. Due to the small quantities produced, as well as the high quality, the prices are fairly high.

In a direct comparison on a good stereo installation, a high-quality phonograph record can easily match or surpass the quality of a CD. The only disadvantages of records are the mechanical noise from the pick-up needle and the crackling caused by dust and electrostatic discharges. Consequently,

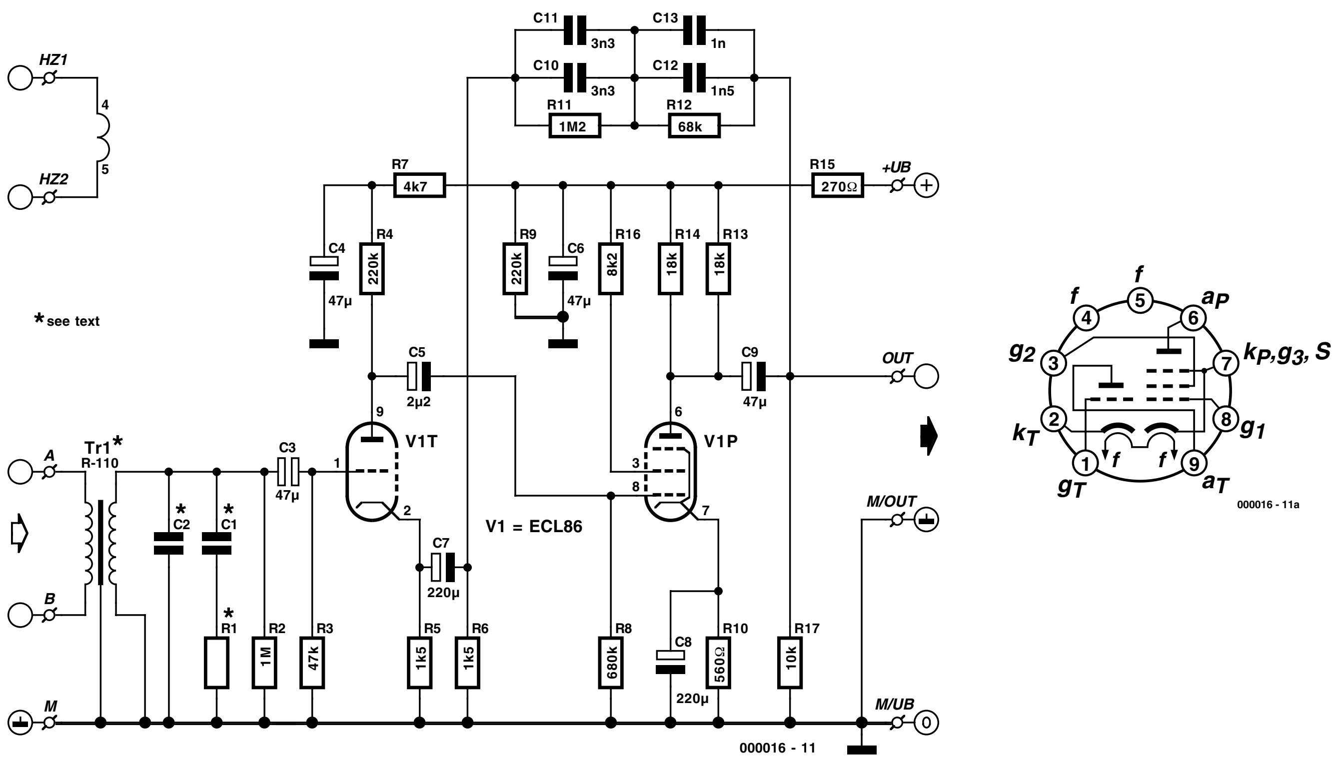


Figure 1. Schematic diagram of the RIAA preamplifier with the optional input transformer.

there is a steady demand for good equalising preamplifiers, including those based on valve technology. This is where the circle closes. Before the coming of semiconductor technology, valve technology was used for everything. This put its stamp on the recording technology and the 'sound'. If we revert to using valves to amplify the phonograph signal, we come closer to the sound experience of earlier times. That is the *raison d'être* for this construction project.

A valved opamp

An equalising preamplifier using valves cannot be compared to one built with semiconductors. Modern opamps have very high open-loop gains, so there is lots of reserve gain available for negative feedback. It is very costly and difficult to achieve equally high levels of open-loop gain with a valve amplifier.

Dual triodes or audio-frequency input-stage pentodes are usually used in these designs. However, here again we chose to take a somewhat unconventional approach and use an ECL86, as can be seen from the schematic diagram of the preamplifier in **Figure 1**. A high-gain audio triode combined with a high-current audio output pentode can be regarded as a valve opamp (high gain, high input impedance and low output impedance). In order to better understand the operation of this circuit, let's first look at its functions in detail.

If the negative feedback network formed by R11, R12 and C10–C13 is disconnected, the open-loop gains of the valves are determined by their associated circuitry. The cathode resistor of the pentode is bridged by a large electrolytic capacitor. This means that the gain of the pentode is equal to $A_v = R_a \times G_m$ (see the box 'Gain Calculations'). With the triode, the DC operating point is determined by R5. Capacitor C7 represents a short circuit for AC voltages, so resistors R5 and R6 are effectively connected in parallel. Due to negative feedback applied via the cathode resistor, the triode will not achieve its full theoretical gain level. Without this negative feedback, the triode would have a gain of 77 in this configuration, while the pentode would have a gain of 90. The combined gain would be 6930, which is around 76.5 dB. The actual gain of the triode is significantly less than 77, due to the negative feedback. In addition, we have to take into account the range of variation in the characteristics of individual valves. Consequently, the basic gain is set to around 35 dB at 1 kHz. This is not breathtakingly high, but it has the advantage that you do not have to use a selected valve, but instead can use any one you wish.

It is not necessary to describe the schematic in more detail, since the functions of the individual components are evident. The operating principle of the ECL86 in this configuration has been thoroughly described in the Valve Preamplifier article. The quality of the components must match what is stated in the components list. If you adhere to this rule, you will have no trouble achieving performance figures equal to those given in the technical specifications.

With this circuit, the old debate about electrolytic capacitors in the signal path will flare up once again. The main advantage of electrolytic capacitors is that they provide very high capacitance in a very small space. This makes it possible to have low-impedance signal coupling, even at very low frequencies. In addition, a lot of technical progress has been made in the construction of electrolytic capacitors in the last few years — more than

Technical Specifications

Stabilised operating voltage		330 V
Nominal output voltage	into 100 kΩ	200 mV
Output impedance	at 20 Hz	2 kΩ
	at 1 kHz	150 Ω
	at 20 kHz	25 Ω
Current consumption	approximately	20 mA

MM (source impedance 750 Ω)

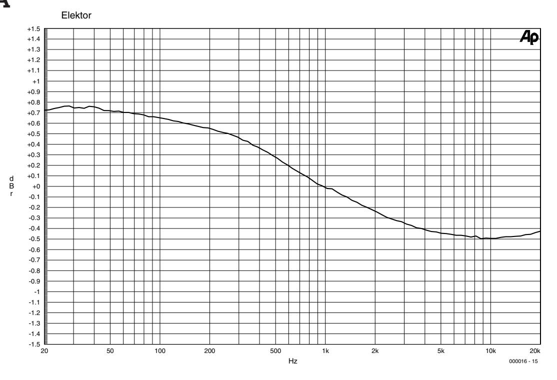
THD+N	BW= 80 kHz, 1 kHz	< 0.06 %
THD+N	A-weighted	< 0.014 %
S/N	22 Hz – 22 kHz	> 65 dB
S/N	A-weighted	> 76 dB
Gain	1 kHz, U_N = 3.5 mV	35 dB

MC (source impedance 25 Ω, using an R-110 input transformer)

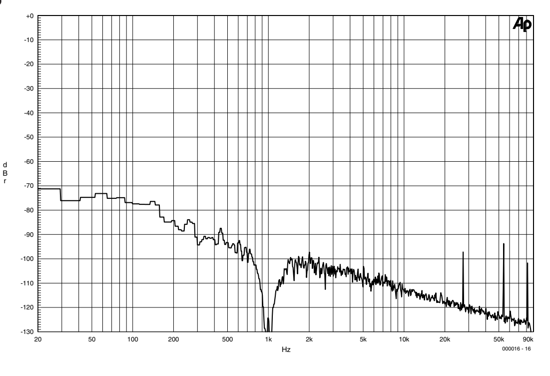
THD+N	BW = 80 kHz, 1 kHz	< 0.07 %
THD+N	A-weighted	< 0.018 %
S/N	22 Hz – 22 kHz	> 63 dB
S/N	A-weighted	> 74 dB
Gain	1 kHz, U_{IN} = 0.37 mV	55 dB

Figure A shows the deviation from the RIAA curve. The relative deviation remains within +0.8 dB / –0.5 dB with a load of 47 kΩ. Due to the high output impedance, departures from these results may occur with other loads, particularly at low frequencies.
Figure B shows the frequency spectrum for a 200-mV, 1-kHz signal with a 100-kΩ load. It can be seen that the THD+N value consists almost exclusively of the noise component. The peaks at 25 kHz and above are generated by the switching power supply (soon to appear in Elektor Electronics!). Lying at –90 dB, they are not only so low as to have no effect on the measurement, they are also far outside the range of human hearing.

A



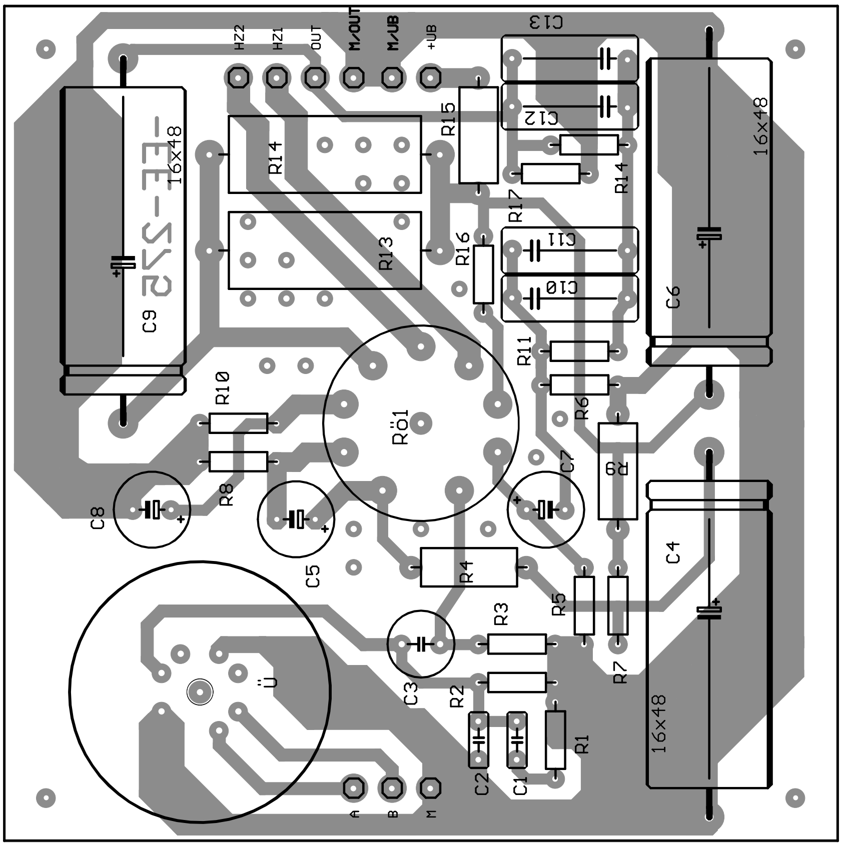
B



with other types of passive components. We suggest that you first build the circuit and listen to the results. After this, you are free to experiment with other types of capacitors.

MM or MC?

Now we come to an interesting circuit detail at the input. If a conventional MM (moving-magnet) cartridge is used, it is connected directly to C3. Resistor R3 then provides the standard termination impedance of 47 kΩ. Resistor R2 functions only as a bleeder resistor for DC voltages, and C2 can be used as needed. A particular capacitive load is prescribed for each type of MM cartridge, in order to obtain a linear frequency response. Usually, the connecting cable of the phonograph is dimensioned so that it provides the correct load in combination with the stray circuit capacitance. If this is not sufficient, C2 must be used to provide compensation. A ceramic capacitor is usually used here, with a value ranging between 10 pF and several hundred picofarads.
If an MC (moving-coil) cartridge is to be used, it is both helpful and worthwhile to employ an audio input transformer. Moving-coil cartridges have significantly better reproduction characteristics than moving-magnet cartridges, due to their operating principle. Their disadvantage is that the output voltage is approximately a factor of ten lower. This can be compensated by using an R-110 input transformer, which is a toroidal-core transformer enclosed in Mu-metal. It boosts the low signal voltage by a factor of 10 (20 dB), with practically no noise and a very small distortion component. This type of transformer can only produce odd harmonics. The distortion factor depends on the primary voltage level and the frequency. The lower the frequency and the higher the voltage, the higher the distortion factor. The characteristics of the R-110 transformer are measured at 1 mV, which is a good standard value for MC cartridges.
The Mu-metal enclosure for the transformer is relatively expensive, but it is absolutely necessary. Without it, all induced magnetic disturbances would be correspondingly amplified along with the signal and would degrade the signal-to-noise ratio. Using the transformer allows us to avoid trying to get the utmost out of amplifier technology, with all the associated disadvantages. In order for the input transformer to work perfectly, certain basic considerations must be taken into account. The transformer used here changes the voltage by a factor t = 10. The voltage is transformed upwards by a factor of 10, but the impedance is transformed by a factor equal to t^2. If the transformer is terminated with 47 kΩ, for example, the termination impedance R seen by the moving-coil cartridge is (R3 / t^2) = (47 kΩ / 10^2) = 470 Ω.
In order for this low impedance to have an effect at the amplifier input, C3 must have a high capacitance. The 3-dB corner frequency lies at around 0.07 Hz in this case. In addition, the amplifier input sees the low source impedance, which has a beneficial effect on the noise behaviour. The transformer



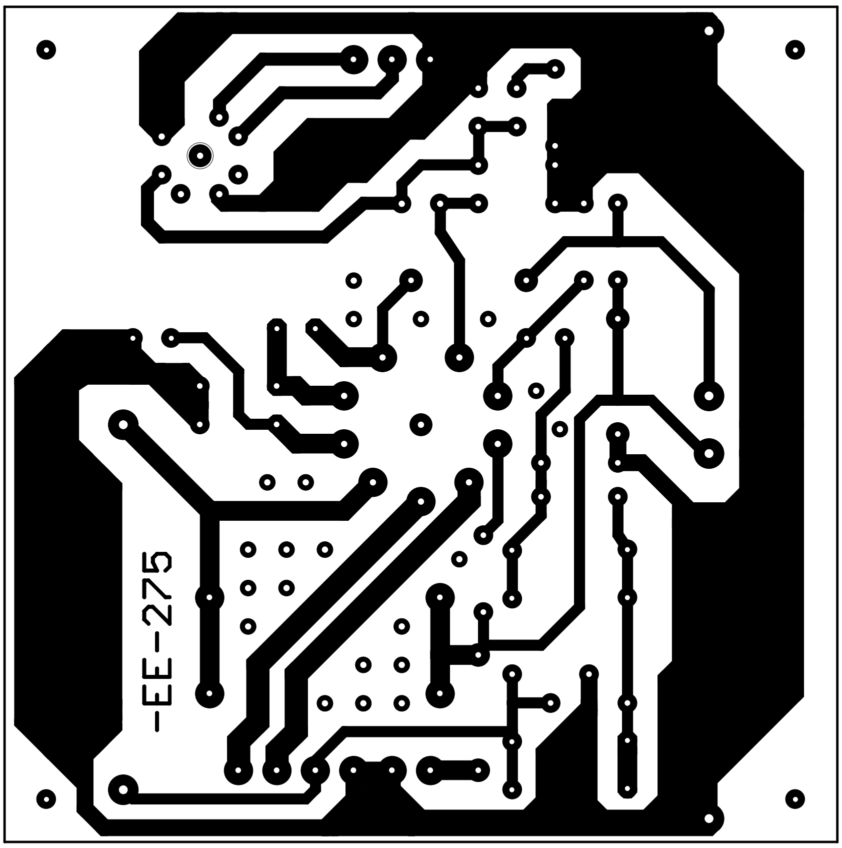


Figure 2. The printed circuit board, which supports one (monophonic) channel, is available from the author.

has a screen between the primary and secondary windings, which diverts interference to ground.

In order to use a moving-coil cartridge correctly, you must observe the specifications given on its data sheet. For example, if a termination resistance of 1 kΩ is specified, R3 must be increased to 100 kΩ. In this way, the cartridge sees the required termination resistance, while the source resistance seen by the preamplifier is the internal resistance of the moving-coil cartridge. R2 can be omitted if an input transformer is used, since the transformer winding provides the DC path. The frequency response can be corrected using C2, C1 and R1. C2 represents the nec-

COMPONENTS LIST

Resistors:

(metal film, 0.7 W, 1% tolerance,
MO = 5%, tolerance)
R1 = see text
R2 = 1MΩ (see text)
R3 = 47kΩ
R4 = 220kΩ, MO, 2W
R5,R6 = 1kΩ5
R7 = 4kΩ7
R8 = 680kΩ
R9 = 220kΩ, MO, 2 W
R10 = 560Ω
R11 = 1MΩ2
R12 = 68kΩ
R13,R14 = 18kΩ, MO, 4.5 W
R15 = 270Ω, MO, 2W
R16 = 8kΩ2
R17 = 10kΩ

Capacitors:

C1,C2 = see text
C3 = 47µF 35V bipolar
C4 = 47µF 450 V axial
C5 = 2µF2 400V, lead pitch 5mm
C6 = 47µF 450V axial
C7,C8 = 220µF 25V, lead pitch 5mm
C9 = 47µF 450V axial
C10,C11 = 3nF3, 2.5 % polypropylene, min. 100V
C12 = 1nF5, 2.5 %, polypropylene, min. 100V
C13 = 1nF, 2.5 %, polypropylene, min. 100V

Miscellaneous:

V1 (Rö1) = ECL86
1 noval (9-pin) socket, ceramic, for board mounting
1 PCB, epoxy resin strengthened, 70µm copper layer
1 Moving-Coil transformer, type R-110, see text
Solder pins

essary capacitive load, which depends on the cartridge employed and the stray circuit capacitance. Depending on the configuration, there may be a resonance peak in the audible range. This can be suppressed using R1 and C1, in order to make the frequency response linear. If C1 and R1 are needed, their values will lie in the range of 22 pF to 1 nF and 5 kΩ to 20 kΩ.

Construction tips

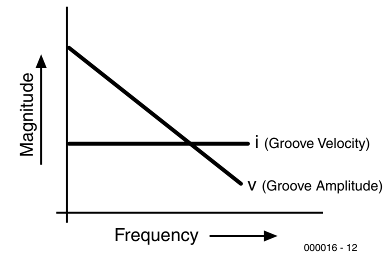
The MC cartridge should be connected as symmetrically as possible using XLR connectors. This is the only way to conduct the very small pickup signal to the amplifier input without introducing interference. Connect the screen braid to pin 1 in the XLR connector. The following small table shows how everything is arranged:

RIAA Cutting

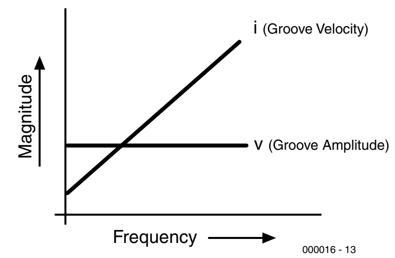
RIAA is an acronym that stands for ‘Record Industry Association of America’. This organisation has specified how phonograph records are to be cut (mastered) and reproduced, so that every record can be played on every phonograph anywhere in the world. What is the purpose of the well-known RIAA curve?

When a record is recorded, the first thing that has to be done is to decide on the groove spacing. The narrower the spacing, the more program material that can be placed on each side. The price of narrow spacing is reduced lateral excursion, which means reduced dynamic range. With a given groove width, the maximum excursion must be limited, since otherwise groove overcutting will occur. This is especially true in the low frequency range (below around 500 Hz), where the largest amplitudes appear.

A



B



A phonograph record is cut using an electrodynamic process. The signal currents for the individual channels are passed through separate coils that drive the cutting stylus, similar to how a dynamic loudspeaker is driven. A spring/mass system is driven by a coil/magnet system. The driven cutting stylus is free to move along only two axes. The impedance of the coils is frequency-dependent. If constant-amplitude drive is used, the cutting velocity rises with increasing frequency (see **Figure A**). With constant-cutting-velocity drive, the amplitude drops with increasing frequency (see **Figure B**.) Since the frequency range that must be handled runs from 20 Hz to 20 kHz, the amplitude ratio with constant cutting velocity is 1:1000 (60 dB).

Such an extremely large dynamic ratio would mean that no usable signal-to-noise ratio would be left at high frequencies. For the optimal utilisation of the surface area of the record, constant-amplitude recording is ideal, but this means that the cutting velocity rises with increasing frequency. When the record is played back using a coil/magnet system, the law of induction thus says that the output voltage will rise with increasing frequency.

C

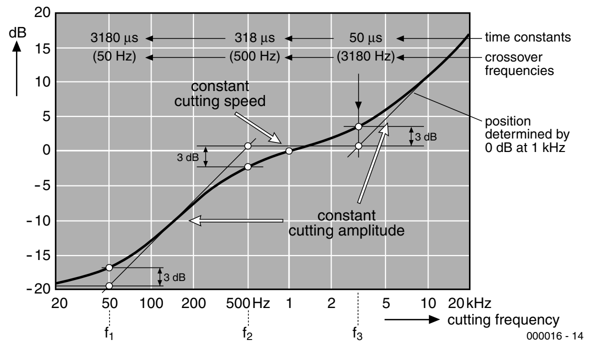
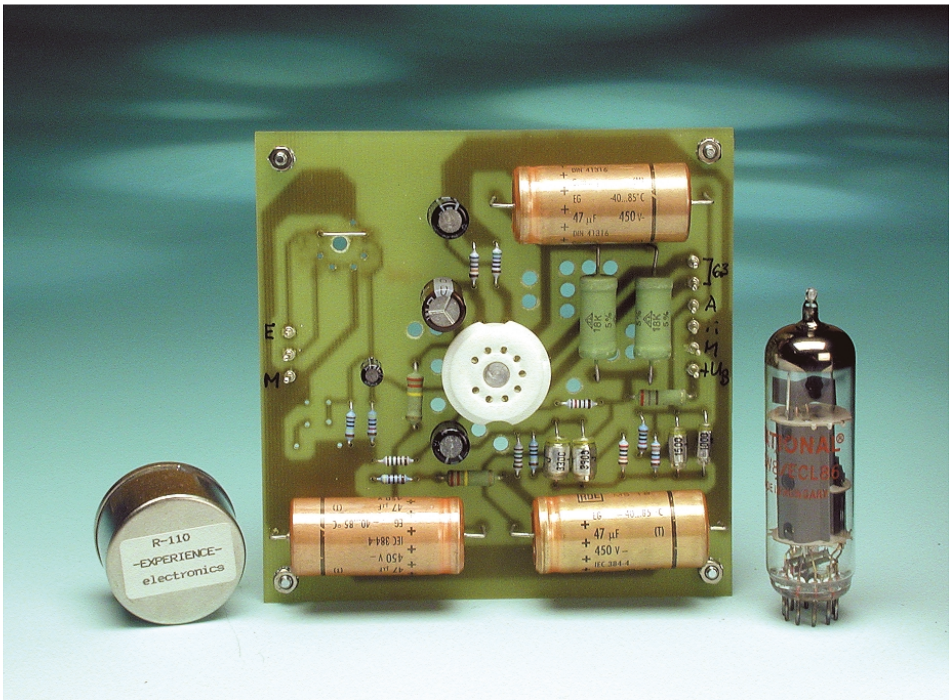


Figure C shows the RIAA recording curve with its corner frequencies and associated time constants. Constant-amplitude recording is used in the frequency range between f1 and f2, in order to limit the maximum excursion and thus avoid groove overcutting. This can be considered to be the same as bass attenuation. In the region between f2 and f3, the behaviour of the coil/magnet system is taken into account and constant-velocity recording is used. Above f3, up to the end of the range, constant-amplitude recording is again used. The net result is that the bass frequencies are attenuated enough to avoid groove overcutting, while the surface area of the record is efficiently exploited. The mid-range frequencies are handled neutrally in comparison to the low frequencies, while the high frequencies are emphasised, which leads to a significant improvement in the signal-to-noise ratio. In the frequency range from approximately f3 upwards, the human ear has increased sensitivity to noise components. During playback, the high frequencies are reproduced too loud, so they must be attenuated. This is precisely the effect that improves the signal-to-noise ratio. The bass frequencies, by contrast, must be emphasised, which leads to increased sensitivity to induced mains-frequency signals and their harmonics. If an equalising preamplifier can maintain the complementary reproduction curve within less than ±1 dB, it is considered to be high-end equipment. The numerical values in the table, which represent the RIAA curve with reference to 1 kHz = 0 dB, are a useful aid for making measurements.

Hz	dB	Hz	dB
20	+19.3	800	+0.7
30	+18.6	1k	0.0
40	+17.8	1.5k	–1.4
50	+17.0	2k	–2.6
60	+16.1	3k	–4.8
80	+14.5	4k	–6.6
100	+13.1	5k	–8.2
150	+10.3	6k	–9.6
200	+8.2	8k	–11.9
300	+5.5	10k	–13.7
400	+3.8	15k	–17.2
500	+2.6	20k	–19.6

Standard lead colour	XLR pin	RIAA-Preamp
white	2	a left
blue	3	b left
screen	1	m left
red	2	a right
green	3	b right
screen	1	m right



If you do not use a moving-coil input transformer, you should use a quasi-symmetric lead arrangement with Cinch connectors, instead of the usual asymmetric arrangement. Connect the screen and the blue or green lead together inside each Cinch connector. This keeps the screens free of signal voltages in this case as well.

The amplifier circuit is relatively simple, but it has very good basic characteristics. Very high-quality construction is absolutely necessary if these are to be fully realised. The amplifier circuit is designed to be monaural, so separate printed circuit boards (as shown in **Figure 2**) are needed for the two channels. The circuit board is not available from Readers Services, but it can be obtained from the author. Mount the two circuit boards in a generously sized, screened enclosure, well separated from interference sources. A high degree of channel separation can be maintained by spacing the two boards widely apart, or by placing a sheet-metal screen between the two boards if space is tight. This makes a lot of difference to the sound. The standard requires >26 dB at 1 kHz, which lies within the realm of what is possible.

The power supply does not belong in the amplifier enclosure. Only the amplifier boards are mounted in the non-magnetic metallic enclosure, while the power supply is housed externally and placed at a sufficient remove from the equalising amplifier. The well-filtered, floating supply voltages (filament and high voltage) are fed to the amplifier enclosure via separate cables. This avoids the superimposition of the filament and high-voltage currents, which could increase the base noise level. The input and output sockets of the amplifier are mounted isolated from ground. The negative poles of the high voltage and filament supplies are connected together, along with the enclosure and signal ground, at a single point. This results in a non-grounded power supply and avoids hum loops. If the signal leads in the phonograph are isolated from the chassis, the phonograph chassis must be connected to the amplifier enclosure via the lead provided for this purpose (black).

The ECL86 has a filament voltage of 6.3 V at 0.66 A. The filaments must be connected in series to match the DC filament voltage of 12.6 V.

Once you have carefully built this preamplifier, you are ready to enjoy unimpeded listening pleasure. Even old monophonic records will sound distinctly better, since a stereo cartridge has significantly better tracking characteristics than a mono cartridge. The value of your record collection will be distinctly enhanced by this low-noise, low-distortion amplifier using primarily K2 components.

(000016-1)

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Gain Calculations

The gain of a triode is determined by its transconductance, its internal resistance and the value of the anode resistor that is used. The gain formula is $A_v = R_a \times g_m$, where g_m is the dynamic transconductance. The values of the static transconductance and internal resistance can be obtained from the valve data book. The value of the dynamic transconductance, which ultimately determines the gain of the triode circuit, must be calculated using the formula

$$g_m = G_m \cdot \frac{R_i}{R_i + R_a}$$

where

g_m	=	dynamic transconductance
G_m	=	static transconductance
R_i	=	internal resistance
R_a	=	anode resistance

For pentodes, the formula is $A_v = R_a \times G_m$. A tolerance of 5% for the anode resistor is considered to be very precise, since many parameters of active components often shows a wide range of variation. For example, if the data book gives a value of 1.6 mA/V for the transconductance of an ECC83, a variation of ±30% is easily possible. The internal resistance amounts to 62.5 kΩ. You can easily calculate how much the actual gain can vary for a given value of the anode resistor. The ECL86 consists of half of an ECC83 together with an audio output pentode, which has a transconductance of 10 mA/V.

1.5 V AF Amplifier

An optimised two-stage design

By Stephan Weber Stephan.weber@connect.com

The brief: design a tiny AF amplifier powered by a single 1.5 V cell, yet capable of driving an 8 Ω loudspeaker. How can these requirements best be met?

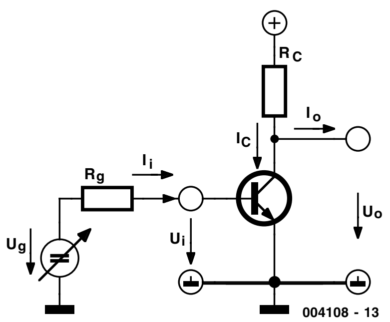


Figure 1. In the common-emitter configuration the emitters of the transistors are connected to a fixed potential.

A high power output cannot be expected of an amplifier running on a 1.5 V power supply. For example, the quiescent output voltage of the amplifier is ?·V_{CC} = 0.75 V, and so, unless an output bridge is used, the maximum possible output swing U_s is only ±0.75 V. This is reduced to ±0.55 V by the saturation voltage of the output transistors. If the battery is no longer fresh, or is under load, this might be further reduced to say U_s = ±0.4 or 0.5 V. The RMS output power into an 8 Ω load is then

$$P = U_s^2 / 2{\cdot}R \approx 15 \text{ mW}.$$

This is not especially loud, but, as long as a high-efficiency loudspeaker is used, entirely adequate for normal indoor listening.

Two Amplifier Stages

Since both the output voltage and the control voltages of the discrete transistors are limited by the low supply voltage, the output transistors must be connected in **common-emitter configuration (Figure 1)**. In many AF power amplifiers with a higher supply voltage, the well-known push-pull emitter-follower circuit is preferred.

In this design, the maximum output current is approximately 60 mA. If the transistor has a current gain of 150, its maximum base current will

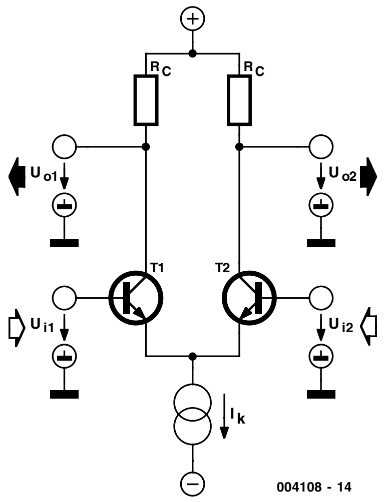


Figure 2. Schematic of the differential amplifier.

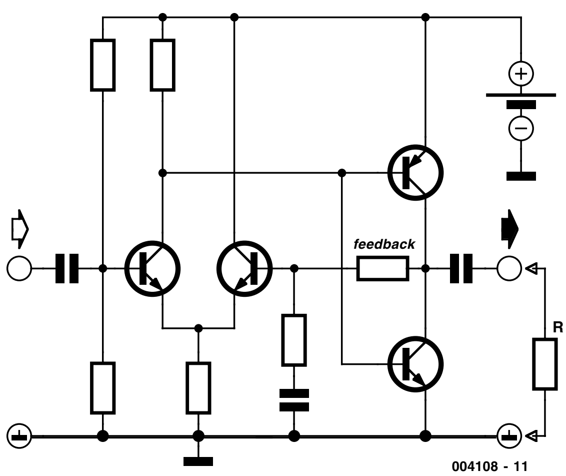


Figure 3. Theoretical circuit of the 2-stage amplifier.

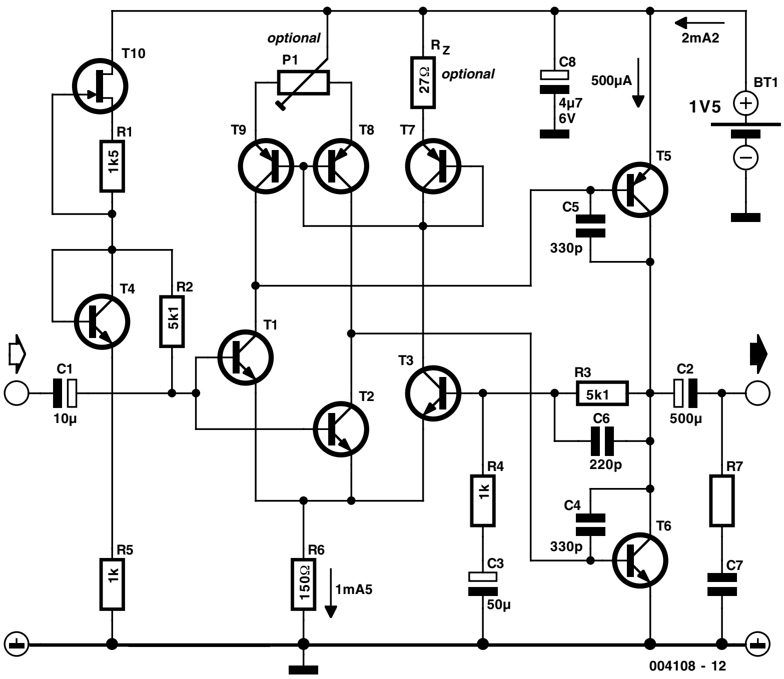


Figure 4. Practical amplifier circuit.

then be around 400 µA, too high for the input to an amplifier. A preamplifier stage is therefore required. With the use of negative feedback, the noise performance of the amplifier can be improved in comparison to a one-stage design. An IC designer might employ three or four stages, but our discrete circuit needs to be as compact as possible.

If we were to connect our common-emitter output stage directly to an emitter-follower stage, the base voltage of the emitter-follower would be $2 \cdot U_{BE} \approx 1.4$ V, hardly practical with a one-cell power supply. A better solution is to use a common-emitter configuration in both the preamplifier and output stages.

Negative Feedback

The use of two common-emitter stages in a row makes the provision of negative feedback more difficult: each stage inverts the signal, and so the overall effect is non-inverting. However, for negative feedback we require an inverted signal. We have two alternatives: either provide negative feedback on each stage separately, or use a **differential amplifier** (**Figure 2**). A differential amplifier has similar properties to the common-emitter circuit, but also provides a non-inverting input. The bases of the transistors in the differential amplifier form its inputs, the collectors its outputs.

The constant current I_k (the sum of the two emitter currents) determines the quiescent input current, while the internal resistance of the current source sets the common mode rejection ratio. Setting the operating point is made rather easier by the use of a differential amplifier, since the base-emitter voltages of the two transistors are compensated for.

Theory and Practice

The theoretical circuit in **Figure 3** shows the two-stage configuration with differential amplifier, common-emitter output stage, and the provision of negative feedback using a voltage divider at the inverting input to the differential amplifier.

A number of details of the theoretical circuit must be worked out in order to arrive at the practical circuit of **Figure 4**. Particularly critical is the way the bases of the output transistors are connected. The base voltage of the upper transistor must be V_{CC}–U_{BE}, that of the lower U_{BE}. These conditions can only be simultaneously satisfied for a narrow range of supply voltages, and small variati-

Construction tips:

Choice of transistors

We have used types BC549C (NPN), BC559C (PNP) and BF245A (FET). The most important characteristics are a high current gain and a low saturation voltage. All the transistors used in the preamplifier stage should be from the same series. The JFET should have a threshold voltage of less than 1.5 V: hence our use of the A-type. Other suitable devices include the BF244A and BF256A.

Can I use a 4 Ω loudspeaker?

Yes. The output transistors will need to be replaced by higher-power devices, such as the BC327-40 and BC337-40.

How can I improve the sound quality?

In general, choosing a higher-quality loudspeaker and improving its mounting will make more difference to the sound quality than modifications to the circuit. However, the bass cut-off frequency can be adjusted by changing the coupling capacitors. The quiescent current of the amplifier stages can also be increased. This increases the loop gain and therefore reduces the total harmonic distortion (THD). The same effect can be obtained by replacing R6 by a transistor current source.

What can I use it for?

The amplifier would be a good partner for the Midget MW Radio described in the March 2000 issue of Elektor Electronics. Alternatively, it could be used as an output amplifier for a crystal set. Another possibility is an audio test amplifier: a test probe can be connected to the input of the amplifier and used, for example, to check signal strength and frequency or look for distortion in other audio circuits. A loudspeaker or headphones will of course have to be connected to the amplifier's output. Remove the low-impedance load from the output: then the open-loop gain will increase and the distortion will fall. Then reduce C4 and C5 appropriately, and the amplifier will operate as a preamplifier for signals up to about 1 MHz. If RF transistors are fitted, a broadband HF amplifier operating up to 100 MHz can be constructed. And finally, as long as only a modest volume is required, the circuit could be used as a power amplifier for a siren.

ons will have a large impact on the quiescent current consumption. The collector resistors in the differential amplifier also have a great effect on the quiescent current of the output transistors.

In practice also, the operating point of the differential amplifier must be set to minimise the effect of variations in temperature and supply voltage on the circuit. The maximum base current of the output transistors has already been calculated as 400 μA, and so the quiescent current in the differential amplifier will need to be around 1 mA.

One side of the differential amplifier circuit is duplicated in order to provide better coupling between the two stages. Its outputs (i.e., the collectors of T1 and T2) drive the base connections of the two output transistors.

The collector resistance is set using a current mirror made from PNP transistors T7, T8 and T9. Transistor T4 in conjunction with FET T10 sets the quiescent current of the differential amplifier essentially independently of other effects.

No special precautions are required to set the quiescent current of the output stage. A value of 1 to 2 mA is adequate, and a reasonable range is acceptable. An adjustment can be provided by fitting a potentiometer between the emitters of the current mirror output transistors and V_{CC}. Negative feedback resistors R3 and R4 set the gain at approximately 5. Thanks to negative feedback, the input impedance is approximately 50 kΩ, and so in practice the total input impedance is determined by R2 (plus the input impedances of T2 and T10).

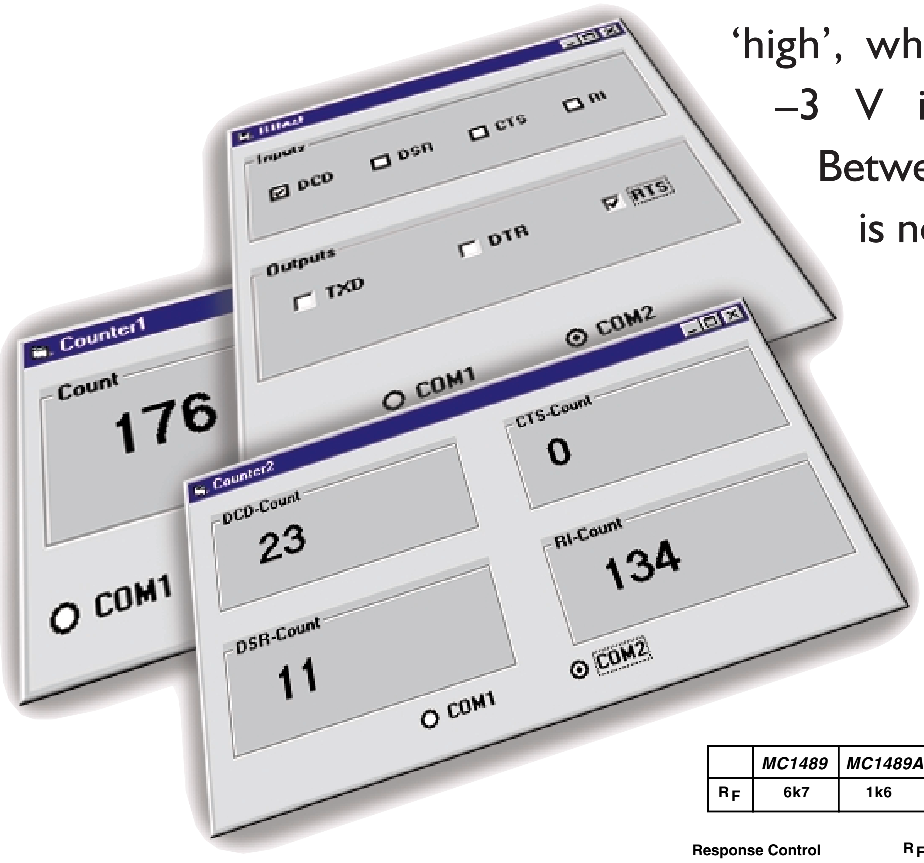
The remaining components stabilise the operation of the circuit. Amplifiers employing negative feedback often have a tendency to oscillate; here, with only two stages, it is less likely. The Miller capacitors C4 and C5 provide additional safety; their values are not critical. Capacitors C1 and C2 isolate the amplifier from DC offsets at the input and output. A reservoir capacitor (say 4.7 μF) across the power supply is essential. The series RC combination (22 Ω + 10 nF) at the output ensures a minimum load even at high frequencies, and thereby ensures the stability of the amplifier.

(004108-1)

PC Serial Peripheral Design (4)

By B. Kainka

So far we have studied the characteristics of the RS-232 outputs in detail. Now it is the turn of the inputs. The standard demands that an input voltage of more than +3 V is considered as 'high', while a voltage lower than –3 V is considered as 'low'. Between these voltages the level is not defined.



Typical RS-232 interface cards use a type 1489 receiver (**Figure 1**). This IC requires a single 5 V power supply. The schematic shows a simple switching stage consisting of three transistors. As can be seen, the

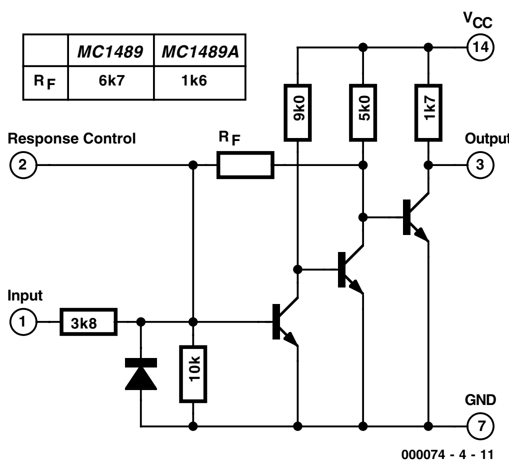


Figure 1. Schematic of 1489 (source: Motorola).

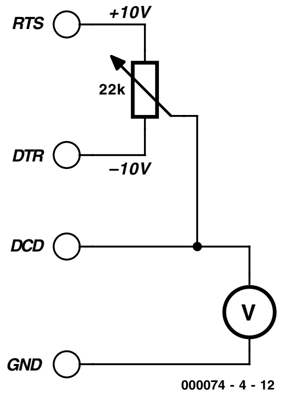


Figure 2. Measuring the input thresholds.

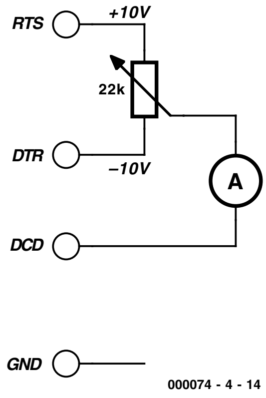


Figure 4. Measuring the input current.

Listing 1

The Counter1 program

```
Dim DSRold, Counter1

Private Sub Form_Load()
 i = OPENCOM("COM2,1200,N,8,1")
 If i = 0 Then
   i = OPENCOM("COM1,1200,N,8,1")
   Option1.Value = True
 End If
 If i = 0 Then MsgBox ("COM
Interface Error")
 TXD 1
 RTS 1
 DTR 1
 Counter1 = 0
 DSRold = DSR()
 Timer1.Interval = 20
End Sub

Private Sub Timer1_Timer()
DSRNew = DSR()
If DSRNew > DSRold Then
   Counter1 = Counter1 + 1
   Label1.Caption =
Str$(Counter1)
End If
DSRold = DSRNew
End Sub
```

threshold voltage will not be very different from the base-emitter threshold of about 0.6 V. Taking into account the voltage divider at the input (consisting of a 3.8 kΩ and a 10 kΩ resistor) we arrive at a figure of about 0.8 V. A further resistor R_F links the output of the second transistor stage back to the base of the first, providing feedback. The circuit therefore behaves as a Schmitt trigger. There are therefore two switching thresholds, one when the input voltage is rising and one when it is falling. With the input in between these two levels the output will remain in its previous state. According to the datasheet the lower threshold for the 1489 is 1 V, while the higher is 1.25 V, a difference of 0.25 V. The 1489A, in contrast, has a lower resistor R_F, giving thresholds of 1 V and 1.95 V.

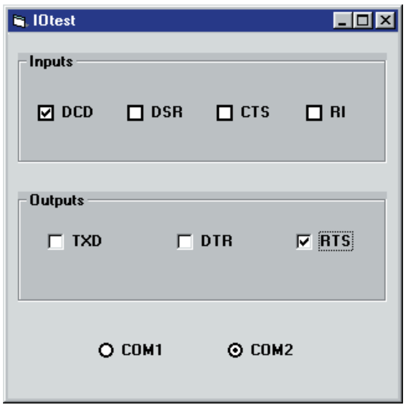


Figure 3. Observing the state of the inputs.

Measurements

The 1489 sets the *de facto* standard for RS-232 inputs. In modern PCs the receiver is normally integrated into a more complex IC, giving the manufacturer the choice of adhering to the quasi-standard, or to interpret the RS-232 standard differently. It is therefore interesting to determine the exact behaviour of the inputs on a particular PC. For this, an adjustable voltage source is required: this does not mean that we will need a piece of laboratory equipment, since the interface itself provides the voltages we shall need. A simple potentiometer suffices to help make our measurements (**Figure 2**).

The IOTEST program from the first instalment of this series is used to make the measurements. The RTS signal is turned on, and the DCD signal is monitored (**Figure 3**). Readings are taken from the voltmeter as the potentiometer is adjusted. The author's PC gave the following results:

Lower threshold: 1.0 V
Upper threshold: 2.0 V

The input circuit in the PC is therefore a 1489A rather than a 1489.

These measurements can be repeated for all four inputs on the serial interface, and similar readings will be obtained. Knowing the upper threshold is useful for some possible applications. For example, it is not possible to detect a 1.5 V battery connected to an input, whereas a direct connection between an RS-232 output and an RS-232 input will always be detected correctly. However, if an LED is connected, the results may be different: it can happen that the LED will glow, but the input will still read as zero.

For some experiments it is important to know the input current rather than the threshold voltage. Here again we can make some simple measurements (**Figure 4**), and, again, we observe a hysteresis. We measured:

Upper threshold: 0.18 mA
Lower threshold: 0.36 mA

From these results we can deduce that an RS-232 output can easily drive, in parallel, more inputs than the four available. From the voltages and currents measured we can calculate an input resistance of 5.6 kΩ. This result seems plausible, looking at the schematic of the 1489.

An important result from these measurements is that a negative voltage is not required for an input to read as zero, even though the RS-232 standard prescribes a voltage below –3 V. Many experiments can therefore be carried out with only a single power supply. There may, of course, be the odd PC which behaves differently from our example. In particular, some laptops require negative input voltages. This must be taken into account in some of our experiments.

Listing 2

Pulse counter with four inputs

```
Private Sub Timer1_Timer()
 DCDnew = DCD()
 DSRnew = DSR()
 CTSnew = CTS()
 RInew = RI()
 If DCDnew > DCDold Then
   Counter1 = Counter1 + 1
   Label1.Caption =
Str$(Counter1)
 End If
 If DSRnew > DSRold Then
   Counter2 = Counter2 + 1
   Label2.Caption =
Str$(Counter2)
 End If
 If CTSnew > CTSold Then
   Counter3 = Counter3 + 1
   Label3.Caption =
Str$(Counter3)
 End If
 If RInew > RIRold Then
   Counter4 = Counter4 + 1
   Label4.Caption =
Str$(Counter4)
 End If
 DCDold = DCDnew
 DSRold = DSRnew
 CTSold = CTSnew
 RIold = RInew
End Sub
```

Reading the state of a switch

It is easy to use the serial interface to read the state of up to four switches (**Figure 5**). One output, for example DTR, is required: this is set high, in order to generate the required voltage. The switches can be connected via a cable of practically any length. Up to three typical 'reset-switches' can be fitted to the circuit board (**Figure 6**), in which case three input signals are used.

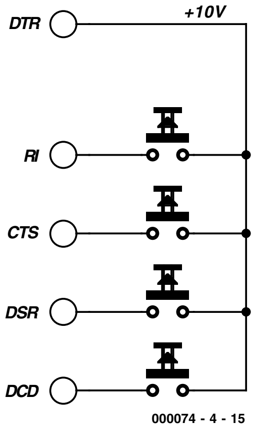


Figure 5. Up to four switches can be connected.

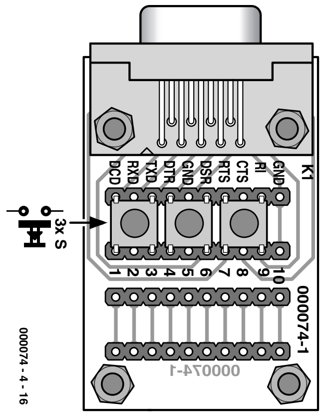


Figure 6. Three buttons can be fitted directly to the circuit board.

Pulse counter

Building a counter in digital hardware is a relatively complicated job. When a PC is available, however, it is easy. Here a pulse counter is constructed in Visual Basic. The DSR signal serves as an input: a simple push-button can be used to provide pulses. Any other sensor could be used, as long as it can provide the correct voltage.

Listing 1 shows the program Counter1; the results on the screen are shown in **Figure 7**. Two global variables are used. DSRold stores the previous state of the DSR signal and Counter1 stores the count. The two variables are initialised in the first routine. The counter is set to zero, and DSRold is loaded with the state of the DSR signal. The outputs are also all turned on. This allows any of the outputs to be connected to the DSR input to provide a count signal.

The actual counting is done in the timer routine. Windows calls this routine roughly once every 20 ms. Each time the state of the DSR input is compared with the state on the previous call. If the new state is greater than the old state, a change from 0 to 1 must have happened: in other words, a rising edge. These edges are counted by incrementing the variable Counter1 by one each time. The value on the screen is only refreshed when the variable changes.

The pulses are counted reliably whether the button is pressed rapidly or slowly, up to about 5 pulses per second. When the pulse rate is faster than this, however, it will be found that some pulses may be missed. The exact limit depends on the PC. With a timer interval of 20 ms a signal with low and high periods of 20 ms should theoretically be read correctly. A total period of 40 ms corresponds to a frequency of 25 Hz (25 pulses per second). However, problems are observed at lower frequencies than this. From this we deduce that Windows cannot reliably maintain a timer interval of 20 ms. A similar observation was made when we discussed the LED flasher program. Even with a timer interval of 50 ms definite irregularity can be seen.

This counter program is certainly not the fastest possible. Furthermore, we can easily construct a four-way counter, as shown in **Figure 8**: we simply need to write out the code four times. Each time a different input is read and processed. **Listing 2** shows the modified timer routine.

(000074-4)

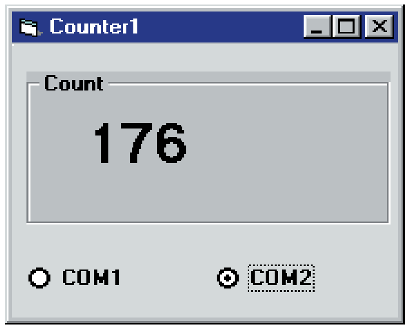


Figure 7. The Counter1 program.

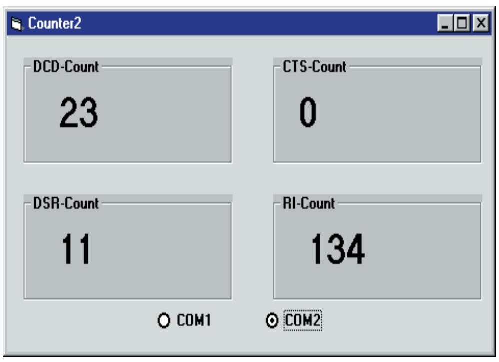


Figure 8. The four-way counter Counter2.

GBDT
GameBoy Development Tools

By Luc Lemmens

The articles on the GameBoy Digital sampling oscilloscope (GBDSO) did not fail to convince many of you that the Nintendo game console is a veritable gateway to more serious applications. This month we'll look at the tricks of the trade, or what goes into creating your GB application.

Since its introduction on the toy market, the immensely successful Gameboy computer game console has been used applications 'totally different' from than those envisaged by its manufacturer, Nintendo. The original game cartridges containing no more than a ROM (sometimes accompanied by a RAM), illegal copies soon appeared on the black market. The next step was to develop Flash-RAM cartridges which could be loaded over and over again (from a PC) with different games. The Internet abounds with websites where this (illegal) activity is practised.
On a more serious (and certainly less illegal) front, people are creating their own games for the GameBoy. This deserves more respect than the copying activity because one needs to know the internal workings of the Gameboy, and be able to program in the processor language. The GBDSO takes this one step further by adding not only software but also new hardware to the GameBoy, enabling the 'toy' to perform a task which is completely different from what it was originally designed for! Not surprising, really, because the GameBoy is a compact system with its own control software, a graphics display, processor, pushbuttons and an attractively styled plastic case, all a price which defeats home construction.

Like us, you'll soon get lost if you try to follow all GameBoy links, threads and other references on the Internet. In this article we will endeavour to inform you about 'places to visit' if you are interested in development tools for the GB. Alas, in view if the huge volume of GameBoy-related information on the web, it is impossible to cover it all.

Inside the GameBoy

The heart of the GameBoy is a Z80-ish processor. The Z80 has been featured in many *Elektor Electronics* projects, and is an offshoot of the good old Intel 8080. The instruction set, however, misses out on some instructions, while others have been added. Furthermore, the flag register looks slightly different because the sign and overflow bits have been omitted.

Developing your own software requires a system memory map like the one shown in **Figure 1**. The address space of the Gameboy comprises several RAM banks (regular internal memory, video memory, sprite memory), ROM banks, memory-mapped I/O and three registers that allow other memory banks to be selected. In this way, the physical memory can be much larger than 64 kBytes as originally specified for the Z80 processor.

At the end of this article you'll find a reference to the document called 'GMB-SPEC.SC' in which all relevant hardware aspects are discussed at length. A must for anyone

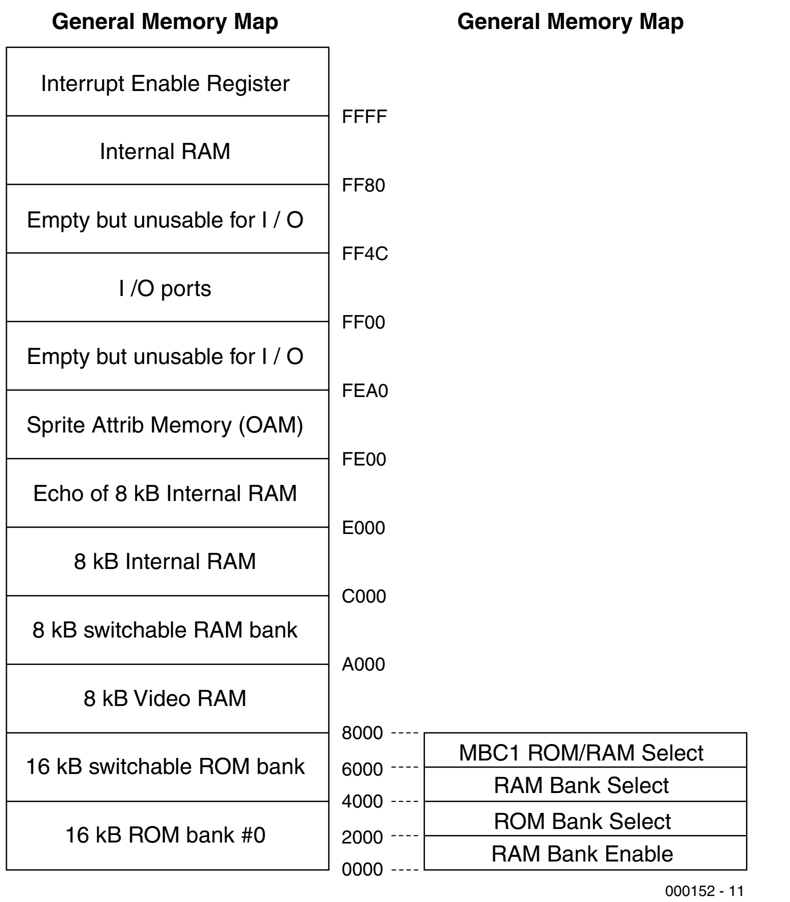


Figure 1. GameBoy memory map.

serious about developing GameBoy cartridge hardware.

Development Software

While there are various superb assemblers for the Z80, you'll like the news that there also is a real C Compiler for the GameBoy: the GameBoy Development Kit (GBDK). None the less, many developers will stick to their assemblers, mainly because the C Compiler fails to generate code which is optimised for compactness. A number of assemblers float around on the Internet, including RGBDS, TASM (Table Assembler — **not** Turbo ASsembler), ADVancedGBIDE and the assembler included with the GBDK. Professionals typically use ISAS, an assembler recommended by Nintendo. Unfortunately, it is only made available to licensed designers.

Each assembler has its own pros and cons, and your choice will ultimately be governed by personal taste and requirements.

The graphics

The Gameboy works with tiles, which in graphics terms translates into rectangles of 8?6 pixels. The next large unit, a 'map' consists of 32?32 tiles. Graphics screens may be designed on the PC using tile editors or map editors like Tile Buddy, GB Tile Designer (GBTD) and GB Map Builder (GBMB), all suitable for running under Windows 95. DOS users may employ TILE256. Besides these editors, there are also programs that enable graphic files to the BMP, PCX, GIF or TIF format to be converted to GameBoy pictures.

Hardware or software?

Several options are available when it comes to testing applications. Of course, you could use a cartridge containing an EPROM or a Flash memory, but for your first (faltering) attempts a GameBoy emulator is highly recommended. Alternatively you could resort to an EPROM emulator, although that requires an adapter board to be made to enable it to be connected. Whatever method you use, the final test will have to be on real hardware, and that is especially true for project designed on emulators, as well as for designs involving new hardware added to a cartridge.

In most cases, you'll also need a helper program to add the special Nintendo header to the ROM cartridge and in addition computes the right checksum. The job is best handled by a program like RGBFIX. At power-on, the GameBoy first runs a couple of routines from internal ROM to check if the cartridge contains the correct checksum, and the Nintendo copyright message is stored in the right location in the cartridge memory. If these two conditions are not satisfied the GameBoy will not get past its own start screen. If everything is okay, it is up to the ROM in the cartridge to make the little computer do the necessary.

Shopping list

First of all you need a GameBoy, at least that's what you would think. True and not true. On the Internet you may find several emulators that allow GameBoy software to be tested on many different computer types, before you even get your hands on the real thing. However, especially when designing new hardware, you really have to get your hands on a GameBoy to make sure your application actually works.

Besides the hardware development tools, you'll also find an integrated development system useful, like 'GameBoy Development Studio'. This wonderful product bundles all components of the development software in one graphics-based package. Finally, the GameBoy Development Kit (GBDK) is highly recommended, if only for the extensive collection of examples included.

(000152-1)

Internet addresses:

Jeff Frohwein's Technical Page:
http://www.devrs.com/gb/

Pascal Felber and Michael Hope's GBDK:
http://gbdk.sourceforge.net/

Ian James GBDS:
http://www.geocities.com/Eureka/9827/

Dr Pan's Gameboy technical documentation:
http://www.gbdev.org/news/dl.html

Paul Robson's GB97 emulator:
http://users.aol.com/autismuk/gameboy.htm

Multi-Purpose IR Receiver

For 'Running Text Display'

Design by F. Wohlrabe

This universal IR receiver works with the IR keyboard transmitter for the Running Text Display described in the February 2000 edition of Elektor Electronics. Its versatility allows it to perform simple switching tasks or complex control of test equipment remotely from a standard PC keyboard.

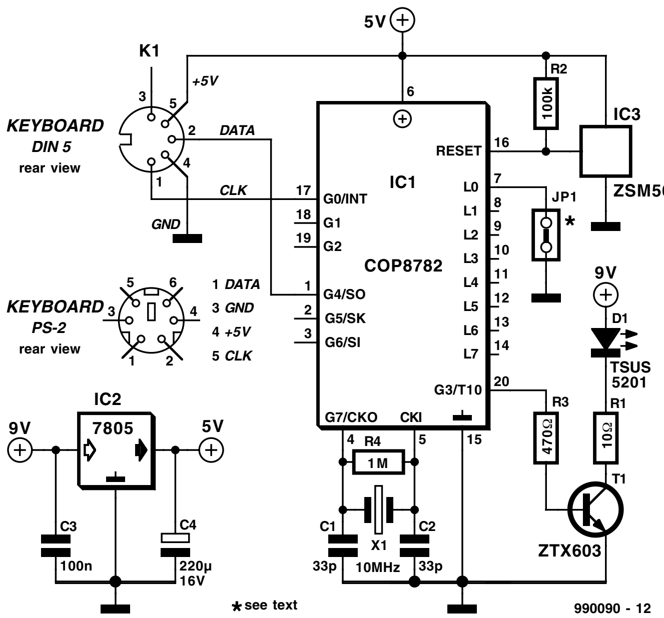


Figure 1. The IR keyboard transmitter from the February edition.

There are two basic methods used for sending data over an infrared link. The simplest is to switch the transmitting diode on and off to represent 1's and 0's in the data stream. To save energy, these bits are made as short as possible. This however makes it more difficult for the receiver to detect the signal, especially if there are high levels of ambient light. The accepted method of overcoming this problem is to switch the transmitter diode with a carrier frequency that is modulated by the data. The receiver electronics are relatively simple and can be easily integrated along with the IR receiving diode to produce a compact device with good sensitivity and immunity to ambient light levels.

There are many integrated receivers available and they generally operate with a carrier frequency in the range of 30 kHz to 40 kHz. Much remotely controlled household equipment employs the RC5 encoding standard, this has a carrier of 36 kHz. We will also use this frequency in our design.

The circuit diagram for the IR keyboard transmitter is reproduced here as **Figure 1**. It was fully described in the February 2000 edition of *Elektor Electronics*, but just to re-cap; a 36 kHz carrier frequency is generated by the microcontroller, this is modulated by the code of the pressed key. This signal then drives the transmitting diode D1. Using a standard receiver, it is possible to achieve a line of sight range of about 10 m. Most walls and ceilings offer good reflectivity to Infrared so a line of sight link is not strictly necessary.

The communications protocol used for this project does not comply with any IR communication standard. It was chosen as a compromise between low current consumption, good data security and simplicity of decoder design. Each keypress generates a message comprising 1 start bit, 8 data bits, 1 parity bit and 1 stop bit. The parity bit ensures that the number of 1's in the data (including the parity bit) is always an even count. A typical message generated by a keypress is shown in **Figure 2.**

The infrared transmitting diode is

driven by Darlington transistor T1. This transistor produces pulses of current to drive the transmitting diode from the control signal. To increase the range of the unit, it is possible to reduce the value of the current limiting resistor R1 and replace D1 with a high intensity IR LED. In principle, any IR LED can be used for this transmitter. The short pulses of current are not sufficient to generate excessive power dissipation in the transistor.

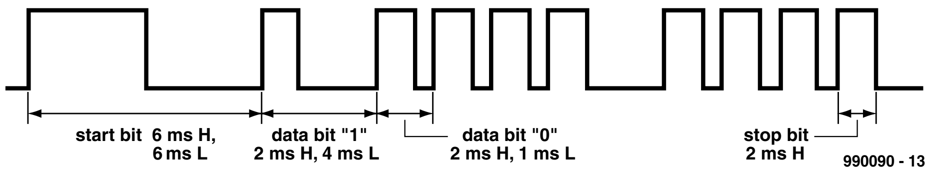


Figure 2. Modulation signal format.

Receiver

The receiver is shown in **Figure 3**. It decodes the IR signal and outputs the corresponding data to its eight output lines. These lines can be used to control or switch external equipment. Alternatively the receiver can be configured to output the hex code for each key press of the keyboard.

Applications for this design range from simple control of remote lamps or devices to a remote front panel for some test equipment. Building keypads into test equipment can often make the equipment design too complicated. This remote keyboard could also help to simplify testing procedures.

The IR signal is received by a sensitive IR detector (IC4). This device includes an IR photodiode, filter, amplifier and demodulator. The output signal is fed directly to an input of the microcontroller (IC2).

A low-pass filter formed by R1 and C1 removes noise and ripple from the supply voltage.

In order to make the receiver as versatile as possible it can be configured to output the key presses in two ways:

Mode 1

This configuration is used for direct control. The eight output lines L0 to L7 are switched when one of the 1 to 8 keys is pressed on the keyboard. For example, one press of key 1 will set output L0 high. Pressing key 1 again will cause L0 to be reset low. LED's D1 to D8 will switch on when the corresponding output is

Microcontroller

The OTP-Microcontroller from National Semiconductor is particularly suited for this application and is used in the transmitter as well as in the receiver:

- 4096 x 8 OTP EPROM
- 128 Bytes RAM
- 1 μs cycle at 10 MHz
- 16 Bit Timer with the operational modes: Timer with Auto-Reload, External Event Counter and with Capture Function
- 16 I/O, of these 14 can be individually programmed as input or output
- Selectable Pin configuration: TriState, push-pull or pull-up
- Microwire interface
- Interrupt source: External Interrupt with selectable edge, Timer Interrupt, Software Interrupt

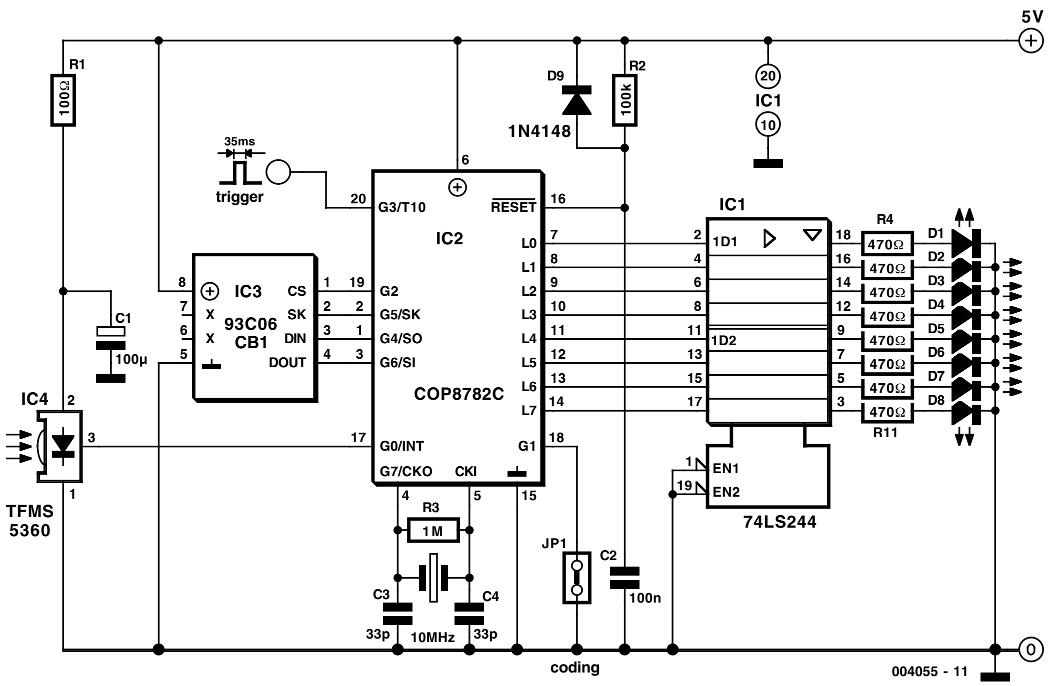


Figure 3. The IR receiver also makes use of the COP8-Microcontroller from National Semiconductor.

high. When the Enter key is pressed all eight outputs will be reset low.

Mode 2

This configuration will cause the controller to output the hexadecimal code for the pressed key. The output code corresponds to the scan code of the pressed key as explained in the February 2000 edition of *Elektor Electronics.*

For example:

	Key	Code
	Esc	08H
	Return	5AH
	Space	29H
	A	1CH
	S	1BH

This hexadecimal code will appear on the output pins L0 to L7 and can be used for your chosen application. Each time a new key is pressed, it will simply overwrite the last code at the output.

The article 'PC Keyboard Encoding' in the February 2000 edition of *Elector Electronics* gives an insight into how versatile a PC keyboard can be and how it can be reconfigured for different uses. In our application, the controller in the IR transmitter configures the keyboard to use Scan Code Set 3. The keyboard will therefore output only one keycode when a key is pressed, for example, pressing key 1 will output the value 16hex. This simplifies decoding in the receiver.

Pin G1 of the microcontroller is used to switch between the two operating modes of the receiver. If no jumper is fitted the pin will be pulled high by an internal pull-up resistor, and the receiver will operate in mode 1. A jumper fitted to G1 will pull this pin low and this will signal the receiver controller to operate in mode 2.

Output G3 provides a trigger signal whenever a key is pressed on the transmitter keypad. This can be used by external equipment to capture and latch the new key value. It has a pulse width of approximately 35 ms and is output in both modes of operation.

R2, C2 and D9 generate a power-up reset for the microcontroller.

Bus driver chip IC1 buffers the output from the controller and can supply a useful 15 mA from each output pin. The controller itself can only supply 3 mA per pin, but in applications where this is sufficient, IC1 can be omitted. In this case the output LEDs will not be fitted either.

The receiver also includes a small, economical serial EEPROM that can store up to 32 bytes of information. Only one byte of this memory is used. When the receiver is switched off, it will store the state of its output pins L0 to L7 in this EEPROM and read them out at power-up to ensure that the output pins remember their last state. This device can also be omitted if this feature is not important in you application

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Project Software

The software (source code) for both COP8 controllers may be found on a diskette with order code **004055-11.**

Ready-programmed controllers are also available from Elektor Electronics:

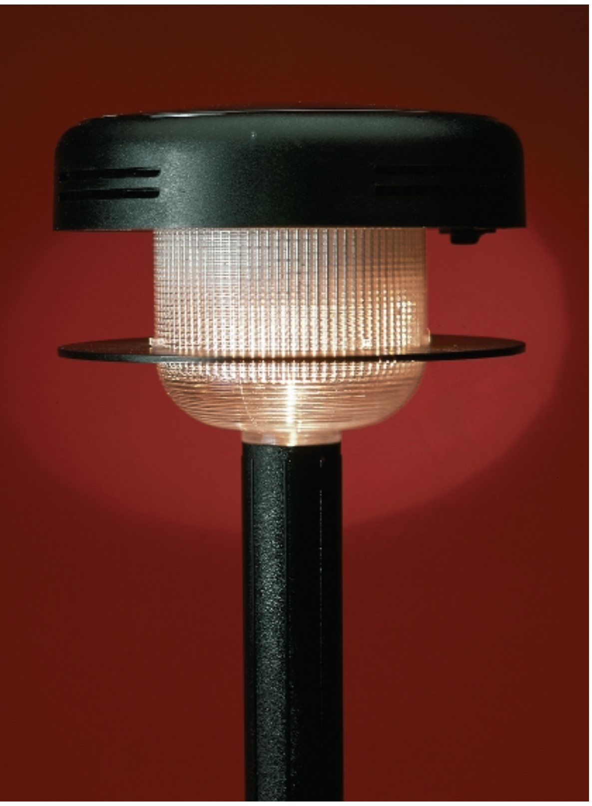
996527-1 (transmitter)
004055-41 (receiver)

Garden Lighting

using solar cells

by K. Walraven and S. van Rooij

Completely 'self-supporting' garden lamps using solar cells as their energy source are gradually becoming more and more common. How do they actually work? We took one apart to find out.



From environmental and technical considerations, buying such a solar-cell garden lamp has a lot to recommend it. It's a great thing that the energy necessary for the lamp to burn in the evening can be drawn from the sunlight that is available for free during the day. In addition, such a lamp is enormously practical, since you can place it in any desired location in the garden without having to dig a trench through the lawn or flowerbeds. You are also free to change your mind about the best location for the lamp — something that would have unpleasant consequences with ordinary garden lamps.

What makes up a typical solar-cell garden lamp? A certain number of elements are in any case necessary for it to function.

It's clear that there must be a *light bulb* and some *solar cells*. However, the bulb is naturally not powered directly from the solar cells, so there must be a *storage battery* and a suitable *charging circuit* to allow the battery to be charged by the solar cells. In addition, the idea is that the lamp should only burn during the evening and the night, and that needs a *twilight switch* with a light-sensitive cell. It's not necessary to do anything to switch off the lamp, since that happens automatically as soon as the battery is fully discharged.

Some of the more luxurious models have a small fluorescent tube in place of a normal light bulb, and in this case a small converter is also necessary. However, the model that we examined contained a small 2.5-V/75-mA halogen bulb, and thus did not need a converter. As far as the electronics are concerned, the whole thing can thus remain very simple.

Simplicity wins out

Our garden lamp consists of a simple plastic structure. Eight solar cells are mounted at the top, and inside there are a small halogen bulb, two penlight NiCd cells and a small printed circuit board for the electronics. As can be seen from **Figure 1**, there isn't all that much inside. This lamp costs around 15 pounds, and it can be found in several different shops.

The electronics also turn out to be extremely simple. **Figure 2** shows the complete schematic of the internal circuitry. The twilight switch is on the left, and its output controls the lamp via transistor T4. To the right are the on/off switch, a diode and the eight solar cells.

Charging

During the day, as long as there is sufficient light, the voltage generated by the solar cells is 8 × 0.45 V under ideal conditions, with a current that depends on the size of the cells — in this case, approximately 140 mA. With less light, less current is supplied. The charging circuit consists simply of a single Schottky diode (D1). The current generated by the solar cells passes through this diode, with its typical low voltage drop of 0.3 to 0.4 V, and charges the NiCd cells.

There is no overcharge protection. It is not actually necessary, since all NiCd cells can handle a continuous charging current equal to 1/10 of their capacity (60 mA in this case), while modern cells are so robust that

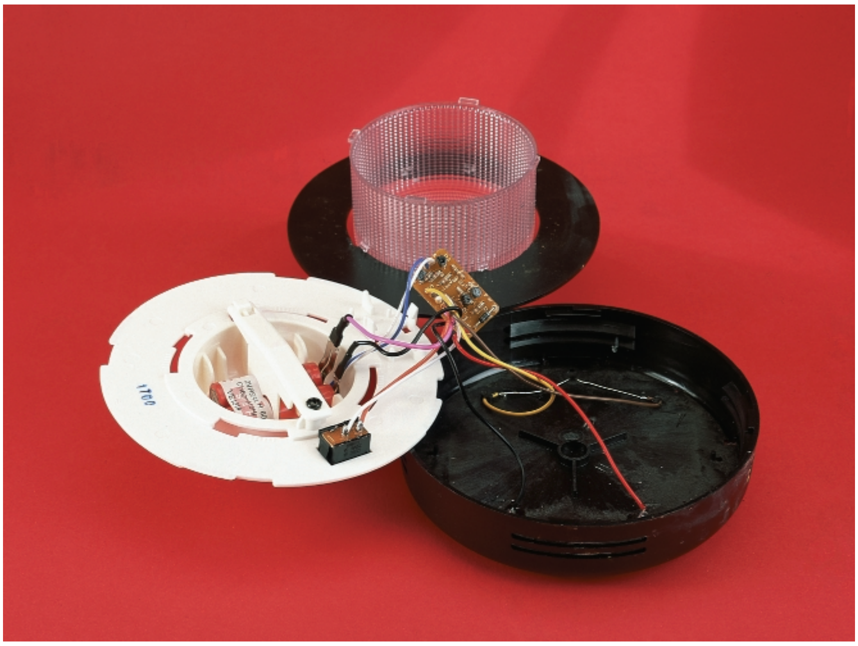


Figure 1. Once the lamp has been opened up, you can see that there's not very much inside.

twice this amount of current does not cause any problems. The advantages of using a somewhat higher charging current are naturally that the battery is already fully charged after several hours of sunlight, and that a certain amount of charging takes place even on rainy days and during the winter. Solar cells act as light-dependent current sources, so the more light there is, the more current they produce. The voltage is determined by the load, but it can never be higher than the previously mentioned 0.45 V per cell.

Approximately 2.8 V is necessary to charge two NiCd cells. If we add the voltage drop across D1, we arrive at a required voltage of 3.2 V. This is 0.4 V per solar cell.

Charging takes place continuously, even when switch S1 is off. It is important to make sure that both NiCd cells are fully charged the first time. Otherwise, one cell may become fully discharged before the other one when they are discharged. As a result, this cell may have a reverse-polarity voltage applied to it, which will shorten its useful life. Therefore, when first putting the lamp into service, you should place it outside with S1 switched off for at least one day in full sunlight, or two days if the weather is cloudy.

Burning

When S1 is closed, voltage is applied to the part of the circuit containing the light bulb.

An LDR is used to determine whether it is light or dark outside. During the day, the resistance of the LDR is low, and the voltage on the base of T1 is also low, so that it is cut off. T2, T3 and T4 are then also cut off, so that the bulb is not illuminated. As soon as it becomes dark, the resistance of the LDR increases, and the voltage on the base of T1 rises. T1 starts to conduct when the voltage is around 0.65 V. This causes T2, T3 and T4 to conduct as well, and the lamp starts to burn. T1 then receives a bit of extra current via R4, so that positive switching takes place when the circuit is sitting 'on the edge'. This is called hysteresis. It means that a threshold is set such that the light level has to drop a bit more before the lamp will switch on again once it is off, and vice versa. This means that the circuit does not react to every passing cloud or insect that is flying around.

As long as it remains dark, the lamp continues to burn until the battery is fully discharged. A fully charged battery has a capacity of 600 mAh, which is enough to supply the 75-mA bulb for approximately eight hours. This is sufficient for the evening and a large part of the night. In the winter, this is not possible, since the battery will probably not be fully charged due to a lack of sunlight.

When the battery becomes fully discharged, its voltage drops. If the voltage drops below 1.25 V, T2 and T3 are cut off, since their base-emitter junctions are in series and thus need at least this amount of voltage. The lamp is then switched off, and the battery is not further discharged.

In the long term

NiCd batteries usually have a lifetime of around 500 to 1000 charge/discharge cycles. After two to three years of continuous use, therefore, the two penlight cells of the garden lamp will probably be ready for replacement. However, these cells are presently so inexpensive that this is not a serious disadvantage. Naturally, there is also a limit on the lifetime of the light bulb, but here again, making a replacement is quick and inexpensive.

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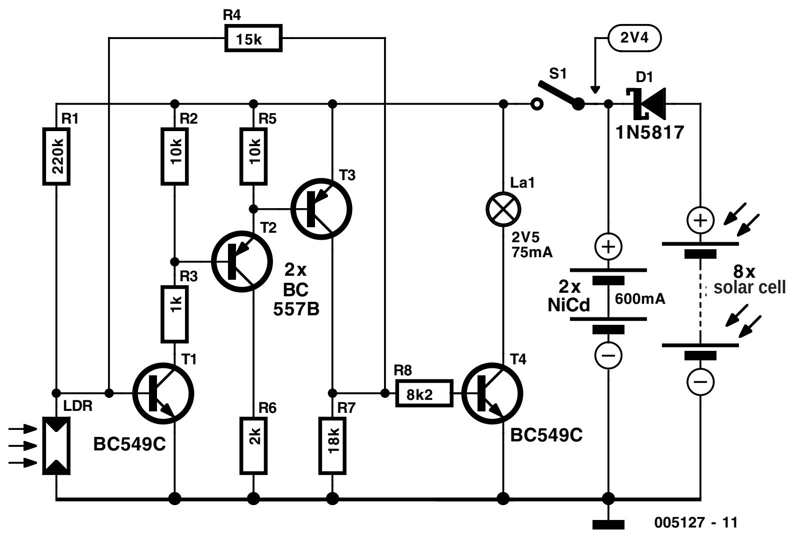


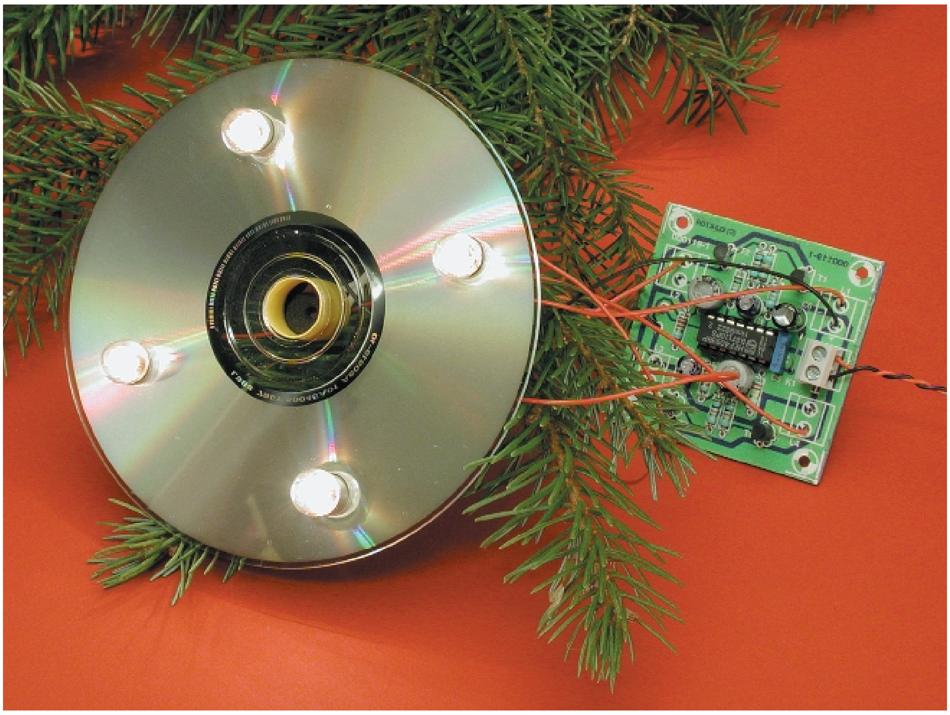
Figure 2. The necessary functions are provided by some very simple electronics.

A CD for Christmas

tasteful lighting effects

By S. van Rooij (text) and P. Lay (design)

By now, loyal readers probably expect that every year around the holiday season, *Elektor Electronics* will come up with a playful circuit that provides an original touch in or around the Christmas tree. This year, our starting point is a compact disc.



Admittedly, this is naturally nothing more than 'fooling around'. Still, we know that most electronics enthusiasts enjoy doing something unusual for Christmas decorations. Some of the standard lighting ornaments that you can find in the department store bins are attractive enough, but everyone already has them. If you are an electronics hobbyist and you want to make a good impression, you will have to think of something somewhat original. In order to stimulate this a little bit, *Elektor Electronics* regularly provides some suggestions.

The example described here is a very simple design. There is no talk of processors or other difficult-to-use components. Although we make use of a modern medium in the form of a CD(ROM), if you read on you will quickly see that we use the CD in a rather improper manner.

Reflector

Practically every hobbyist should have a few surplus CDs lying about, and the basic idea is to use one of them as a sort of reflective screen. If a lamp is held in front of a CD, colour fringes appear in the reflection, and this is naturally an effect that we can happily put to good use in a Christmas decoration.

So what's it all about? You can get an impression from the photo accompanying this article. Four holes are drilled in the CD, and four small incandescent lamps are mounted in these holes. We have intentionally used lamps here, since they emit a broader spectrum of light than LEDs and thus provide more colourful reflections from the silver disk. The lamps are switched on and off rhythmically, each at a different rate. This creates an especially unusual effect. An LDR is placed in the middle opening of the CD. This forms part of a twilight switch, which ensures that the ornament lights up only when the room is dark.

Schematic

A glance at **Figure 1** clearly shows that the design of the electronics that must provide the lighting effects is very simple. Four oscillators, four transistors and four lamps — that's about all there is to it.

The twilight switch consists of nothing more than a voltage divider (R7/R8). When it becomes dark, the resistance of LDR R8 increases, and

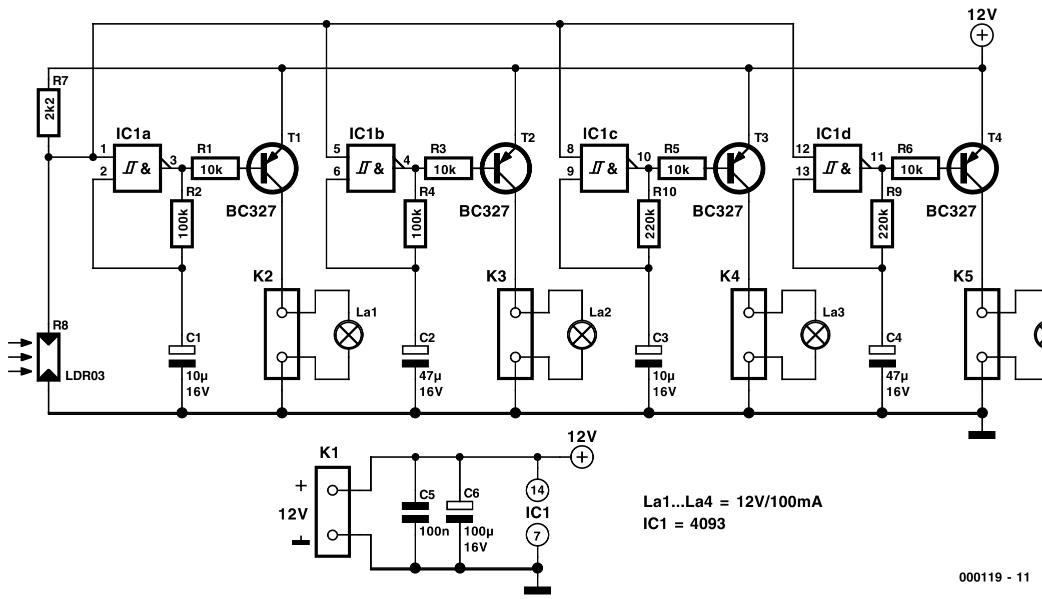


Figure 1. Four oscillators and four small incandescent lamps are all you need for festive lighting effects.

the four 4093 oscillators are enabled. The four lamps are then driven in flashing rhythms that are determined by the RC networks R2/C1, R4/C2, R10/C3 and R9/C4. Since all four networks have different component values, the lamps will flash completely asynchronously. In practice, this proved to give the nicest effect. Naturally, you can adjust the pattern to suit your own taste by modifying the values of the frequency-determining components.

Construction

The printed circuit board shown in **Figure 2** has been developed for this circuit. In light of the simplicity of the electronics, the construction of the printed circuit board hardly needs any further explanation. The simplest way to mount the circuit board is to use four standoffs to fasten it directly to the back of the CD. The lamps can be connected using short lengths of wire. There is room on the circuit board for four terminal blocks, but the lamp wires can of course be soldered directly to the appropriate locations on the board instead.

Two connector pins are provided on the board next to R7 for connecting the LDR. As already mentioned, LDR R8 is fixed in the centre hole of the CD. A small cardboard tube can be used to screen the LDR against being affected by undesired light.

A standard 12-V mains adapter (500 mA) can be used as a power source. If you make things robust enough, you may even be able to hang the combined CD and circuit board from its power-supply wires as a sort of oversized Christmas ball. Merry Christmas!

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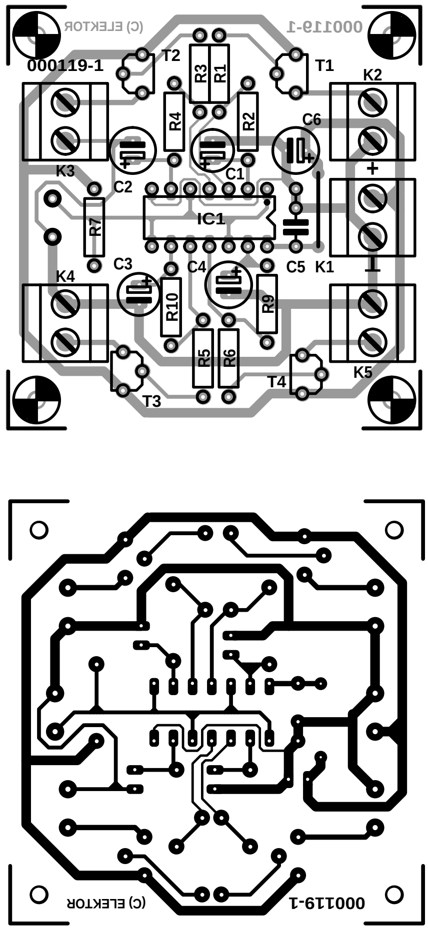

Figure 2. Printed circuit board layout for the 'Christmas CD' (not available ready-made).

COMPONENTS LIST

Resistors:
R1,R3,R5,R6 = 10kΩ
R2,R4 = 100kΩ
R7 = 2kΩ2
R8 = LDR03 (Eurodis, Bolton)
R9,R10 = 220kΩ

Capacitors:
C1,C3 = 10µF 16V radial
C2,C4 = 47µF 16V radial
C5 = 100nF
C6 = 100µF 16V radial

Semiconductors:
T1-T4 = BC327
IC1 = 4093

Miscellaneous:
K1-K5 = 2-way PCB terminal block, lead pitch 5mm
La1-La4 = miniature bulb, 12V/100mA
Surplus CD

Notation:

Article title . . . .[month — page number]
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Power Supplies & Battery Chargers

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[continued overleaf]

Radio, TV & Communications

Test & Measurement

Datasheets

Corrections & Updates, Letters (P.O. Box 190)

Specifications			At T_A = +25°C, V_{CC} = $V_{CC}L$ = $V_{CC}R$ = $V_{CC}P$ = 5.0V, V_{DD} = $V_{DD}C$ = 3.3V, f S = 44.1MHz, signal frequency = 1kHz and 16-bit data, unless otherwise specified.			
PARAMETER		**CONDITIONS**	**PCM2702E**			**UNITS**
			MIN	**TYP**	**MAX**	
RESOLUTION				16		Bits
HOST INTERFACE			Supports USB Revision 1.0, Full Speed			
DIGITAL AUDIO FORMAT						
Audio data Format			USB ISOCHRONOUS OUT			
Audio Data Bit Length				16		
Audio Data Channel				1, 2		
Sampling Frequency (fS)				32, 44.1, 48		
DIGITAL INPUT/OUTPUT						
Input Logic Level	V_{IH} (1)		2.0			VDC
	V_{IL} (1)				0.8	VDC
	V_{IH} (2)		0.7 V_{DD}			VDC
	V_{IL} (2)				0.7 V_{DD}	VDC
Input Logic Current	I_{IH} (1)	$V_{IN} = V_{DD}$		+65	+100	µA
	I_{IL} (1)	V_{IN} = 0V			±10	µA
	I_{IH} (2)	$V_{IN} = V_{DD}$			±10	µA
	I_{IL} (2)	V_{IN} = 0V			±10	µA
Output Logic Level	V_{OH} (3)	I_{OH} = –1mA	2.8			VDC
	V_{OL} (3)	I_{OL} = +1mA			0.5	VDC
DYNAMIC PERFORMANCE (4)						
THD+N at V_{OUT} = 0dB				0.002	0.005	%
THD+N at V_{OUT} = –60dB				1.2		%
Dynamic Range		EIJA, A-Weighted	96	100		dB
Signal-to-Noise Ratio		EIJA, A-Weighted	100	105		dB
Channel Separation			98	103		dB
DC Accuracy						
Gain Error				±1.0	±3.0	% of FSR
Gain Mismatch, Channel-to-Channel				±1.0	±3.0	% of FSR
Bipolar Zero Error		V_{OUT} = 0.5 V_{CC} at BPZ		±30	±60	mV

PCM2702

Integrated Circuits
Special Applications

ELEKTOR ELECTRONICS

DATASHEET 12/2000

PCM2702

16-Bit Stereo Digital-to-Analog Converter with USB Interface

Manufacturer

Burr-Brown, P.O. Box 11400, Tucson, AZ 85734, USA. Internet: http://www.burr-brown.com

Features

Integrated USB interface:
- Full-Speed Transceiver Supports 12Mbps Data Transfer.
- Fully Compliant with the USB 1.0 Specification.
- Adaptive Mode for Isochronous Transfer.
- Self-Powered Device.

Accepts 16-bit stereo and mono USB audio data streams.

Analog performance (V_{CC} = 5V):
- Dynamic Range: 100dB (typ. at 16-bit)
- SNR: 105dB (typ.)
- THD+N: 0.002% (typ. at 16-bit)
- Full-Scale Output: 3.1V_{pp}

8x oversampling digital filter:
- Passband: 0.454f_S
- Stopband: 0.546f_S
- Passband Ripple: ±0.002dB
- Stopband Attenuation: –82dB

Sampling rate (f_s): 32kHz, 44.1kHz, 48kHz

On-chip clock generator with single 12MHz clock source

Multi-Functions:
- Digital Attenuator: 0dB to –64dB, 1dB/step
- Soft Mute
- Zero Flag
- Suspend Flag
- Playback Flag

Dual Power Supplies:
- +5V for Analog portion
- +3.3V for Digital portion

Package: SSOP-28

Application Example

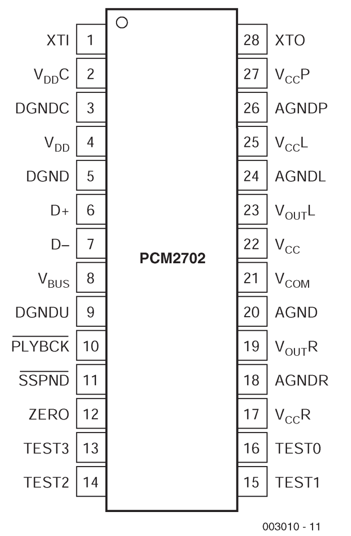
USB Audio DAC, *Elektor Electronics* December 2000.

Applications

Stand-alone USB audio speakers
CRT/LCD integrated USB audio speakers
USB audio amplifiers
Other USB audio applications

Description

The PCM2702 is a single chip digital-to-analog converter offering two D/A output channels and an integrated USB 1.0 compliant interface controller. The newly developed SpAct™ (Sampling Period Adaptive Controlled Tracking) system recovers a stable, low-jitter clock for internal PLL and DAC operation from the USB interface audio data.
The PCM2702 is based upon Burr-Brown's Enhanced Multi-level Delta-Sigma Modulator, an 8x oversampling digital interpolation filter, and an analog output low-pass filter.
The PCM2702 can accept a 48kHz, 44.1kHz and 32kHz sampling rates, using either 16-bit stereo or monaural audio data. Digital attenuation and soft-mute features are included, and are controlled via USB audio class request.



PCM2702

Integrated Circuits
Special Applications

ELEKTOR ELECTRONICS

DATASHEET 12/2000

PIN	NAME	TYPE	DESCRIPTIONS
1	XTI	IN	Crystal Oscillator Input. (1)
2	$V_{DD}C$	—	Digital Power Supply for Clock Generator, +3.3V.
3	$D_{GND}C$	—	Digital Ground for Clock Generator.
4	V_{DD}	—	Digital Power Supply, +3.3V.
5	D_{GND}	—	Digital Ground.
6	D+	IN/OUT	USB Differential Input/Output Plus.
7	D–	IN/OUT	USB Differential Input/Output Minus.
8	V_{BUS}	IN	USB Bus Power (this pin NEVER consumes USB bus power). (2)
9	$D_{GND}U$	—	Digital Ground for USB Transceiver.
10	$\overline{PLYBCK}$	OUT	Playback flag, active LOW (LOW: playback, HIGH: idle).
11	$\overline{SSPND}$	OUT	Suspend flag, active LOW (LOW: suspend, HIGH: operational).
12	ZERO	OUT	Zero flag (LOW: Normal, HIGH: ZERO.)
13	TEST3	IN	Test pin 3. Connect to digital ground. (2)
14	TEST2	IN	Test pin 2. Connect to digital ground. (2)
15	TEST1	IN	Test pin 1. Connect to digital ground. (2)
16	TEST0	IN	Test pin 0. Connect to digital ground. (2)
17	$V_{CC}R$	—	Analog Supply for R-channel, +5V.
18	$A_{GND}R$	—	Analog Ground for R-channel.
19	$V_{OUT}R$	OUT	Analog Output for R-channel.
20	A_{GND}	—	Analog Ground.
21	V_{COM}	—	DC Common-Mode Voltage for DAC.
22	V_{CC}	—	Analog Supply, +5V.
23	$V_{OUT}L$	OUT	Analog Output for L-channel.
24	$A_{GND}L$	—	Analog Ground for L-channel.
25	$V_{CC}L$	—	Analog Supply for L-channel, +5V.
26	$A_{GND}P$	—	Analog Ground for PLL.
27	$V_{CC}P$	—	Analog Supply for PLL, +5V.
28	XTO	OUT	Crystal Oscillator Output.

Notes: (1) 3.3 V tolerant. (2) Schmitt trigger input with internal pull-down, 5V tolerant.

PCM2702

Integrated Circuits
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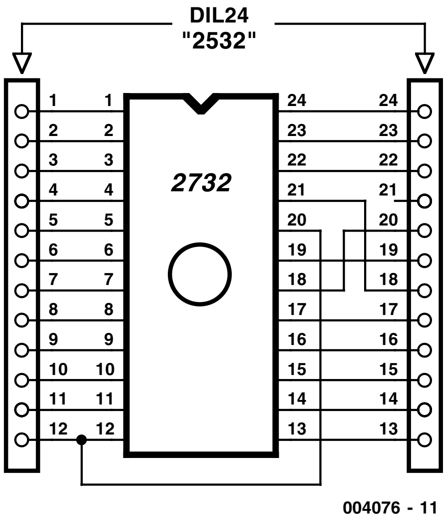
ELEKTOR ELECTRONICS

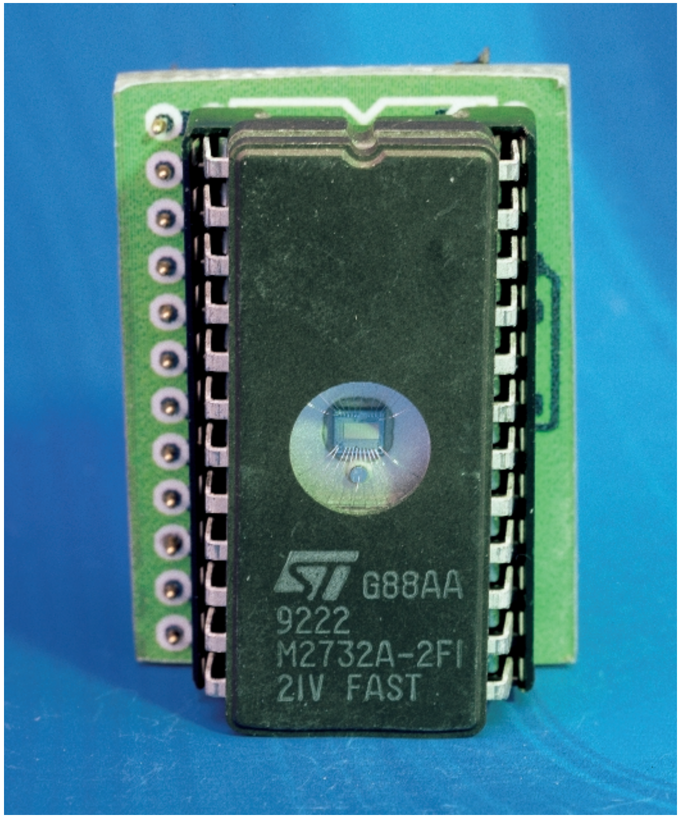
DATASHEET 12/2000

PARAMETER		CONDITIONS	PCM2702E			UNITS
			MIN	TYP	MAX	
ANALOG OUTPUT						
Output Voltage		Full-Scale (–0dB)		62% of V_{CC}		V_{p-p}
Center Voltage				50% of V_{CC}		V_{DC}
Load Impedance		AC-Load	5			kΩ
DIGITAL FILTER IMPEDANCE						
Passband		±0.002dB			$0.454f_S$	
Passband		–3dB			$0.490f_S$	
Stopband			$0.546f_S$			
Passband Ripple					±0.002	dB
Stopband Attenuation		Stopband = $0.546f_S$	–75			dB
Stopband Attenuation		Stopband = $0.567f_S$	–82			dB
Delay Time				$34/f_S$	11	s
ANALOG FILTER PERFORMANCE						
Frequency Response		At 20kHz		≠0.02		dB
POWER SUPPLY REQUIREMENTS						
Voltage Range	V_{DD}, $V_{DD}C$		±3.0	±3.3	±3.6	VDC
	V_{CC}, $V_{CC}L$, $V_{CC}R$, $V_{CC}P$		±4.5	±5.0	±5.5	VDC
Supply Current	I_{DD}	$V_{DD} = V_{DD}C = 3.3V$		22	30	mA
	I_{CC}	$V_{CC} = V_{CC}L = V_{CC}R = V_{CC}P = 5.0V$		18	25	mA
Power Dissipation	P_D	$V_{DD} = V_{DD}C = 3.3V$ and $V_{CC} = V_{CC}L = V_{CC}R = V_{CC}P = 5.0V$		165	225	mW
TEMPERATURE RANGE						
Operation Temperature			0		70	°C
Storage Temperature			–55		+125	°C
Thermal Resistance, Θ_{JA}		SSOP-28		100		°C/W

NOTES: (1) Pins 8, 13, 14, 15, 16: VBUS , TEST3 TEST2 TEST1, TEST0. (2) Pin1: XTI. (3) Pins 10, 11, 12, 28: PLYBCK, SSPND, ZERO, XTO. (4) The dynamic performance is based upon ideal host signal quality, and may vary according to the system. Dynamic performance specifications are tested using a Shibasoku #725 THD Meter with 400Hz HPF, 30kHz LPF, Average Mode, and 20kHz Bandwidth limiting. The load connected to the analog output is 5kΩ, or larger, via AC coupling.

001 2532/2732 EPROM Adapter





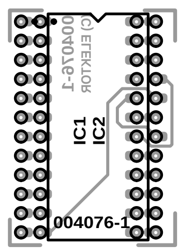
L. Lemmens

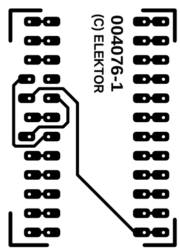
Many old pieces of equipment contain type 2532 EPROMs, which are currently difficult or impossible to obtain. If the software has to be modified or the EPROM fails, it is in principle possible to replace the EPROM with a modern 2732, but this has a different pinout. A handy hobbyist can of course make the necessary modifications to the circuit board, but it is more convenient to use an adapter board that can be fitted in place of the 2532. The main circuit board can then remain in its original state.

The printed circuit board shown here simply swaps the connections to a few pins. It connects pin 18 of the 2732 to pin 20 of the 2532, pin 20 to pin 12, and pin 21 to pin 18. The remaining pins are connected 1:1 to each other.

This board is not available from Readers Services, so you will have to etch it yourself. Mount two 14-pin strip connectors on the underside of the board, so that the adapter circuit can be plugged into the original 2532 socket. On the top side of the adapter, fit a normal 24-pin IC socket for the 'new' 2732.

(004076-1)





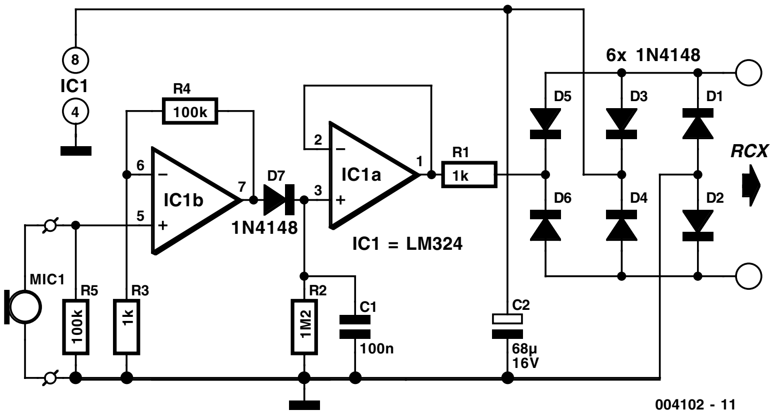
002 Electronic Ear for Lego RCX Module

H. Steeman

One interface that is missing in the Lego MindStorms system is an electronic ear. We don't mean that the RCX should respond to spoken commands (which requires a large amount of electronics and software), but it would be handy if it could respond to basic sounds (or sound levels). The circuit presented here allows the RCX to sense different sound levels. The sound is picked up by a crystal microphone, which is an inexpensive component that is available in every electronics shop. The signal from the microphone is converted into a variable quasi- resistance value. The RCX, in turn, can use this value to determine if a particular sound level has been exceeded. If the trigger threshold is set to the right level, the RCX will then react to a previously set sound level. The RCX input must be configured as a light sensor input for this function.

The operation of the circuit is simple. IC1, which is wired as a non-inverting amplifier, amplifies the microphone signal by a factor of 100. The output signal from the opamp is rectified by D1 and smoothed by C1. Resistor R2 allows the capacitor to discharge. The resulting DC voltage drives IC2, which acts as a buffer. The output of this opamp is connected to the sensor input of the RCX via a 1-kΩ resistor (R1).

Just as with the analogue input adapter described elsewhere



in this issue, the RCX sees a variable resistance value at the sensor input, and it converts this into a measurement value between 0 and 100. In the idle state, when no sound is sensed, the measurement value lies between 90 and 100. The louder the sensed sound, the lower the measurement value. You can use the light-sensor routine of the Lego software to set the responses to various sound levels. If you use a threshold value of around 85, then a level under 85 will be sensed as a sound signal, while a level above 85 will be sensed as silence. If you clap your hands near the sensor, the circuit will detect this. If you use these 'observations' to increment a counter, it is even possible to measure the number of sound pulses within a defined interval, and then to carry out some action based on the result.

(004102-1)

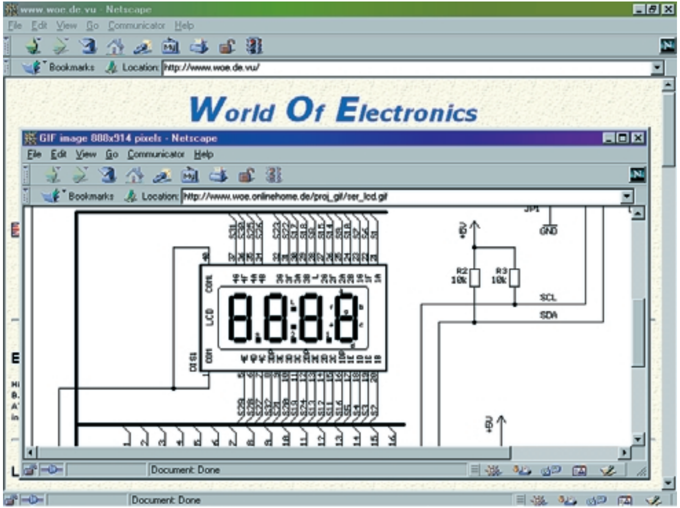
003 Display Info

H. Baggen

The website 'World of electronics' (www.woe.de.vu), from the German Michael Gaus, makes a tidy and well-organised impression. It offers the electronics engineer or hobbyist a lot of information, and it is certainly worth taking the trouble to surf to this site. The main component of this site is information about various types of displays. There are separate sections for LED displays, LCD modules and vacuum fluorescent displays. In addition, there is a separate page with a large number of links for searching for data sheets on the Internet. This is especially handy if you want to find information about a component that is not so well known.

There is a page with electronic projects, which at the time of writing contains a controller for a Nokia TV tuner, a LED running light display, a LED clock with a unique form, an RS232 to I^{2}C converter and an AT89C2051 board. There is also a list of component suppliers, but this is presently limited to German sources. You can view this site in German or in English.

(004110-1)



004

ECG Amplifier

H. Bonekamp

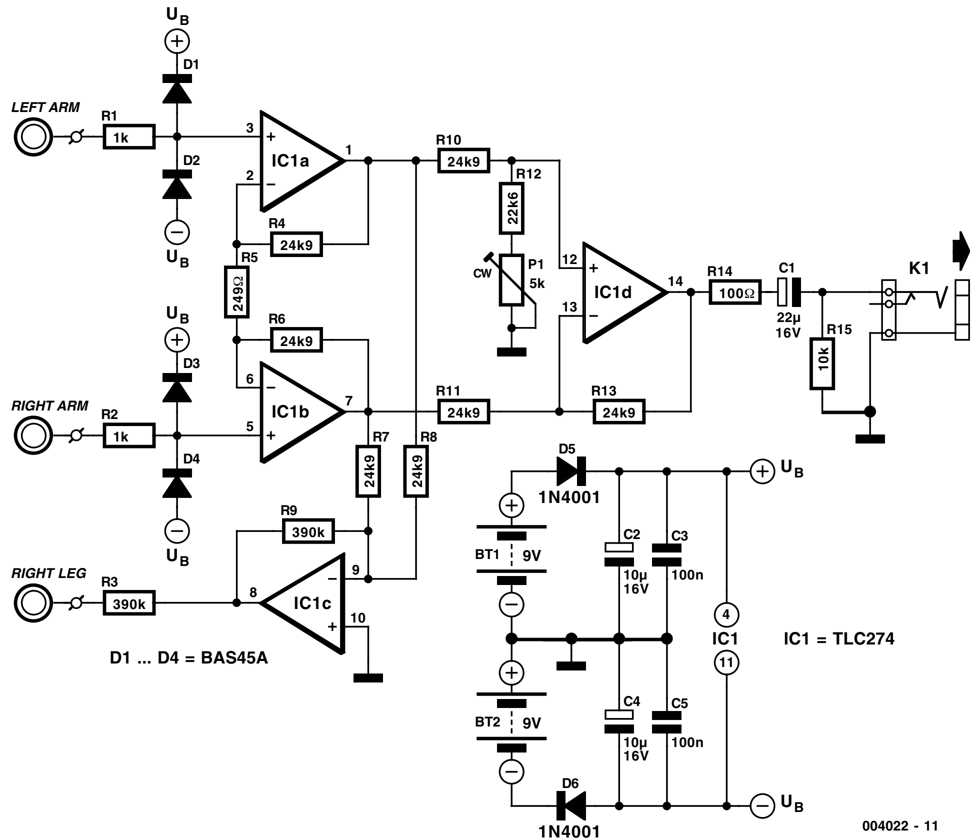
This circuit allows an ECG signal to be displayed on an oscilloscope. Opamps IC1a, b and d form an instrumentation amplifier with a gain of 201. IC1c amplifies the common-mode signal by a factor of 31, and supplies this signal to the right leg. The first consequence of this is that the body is brought to a defined common-mode level, so that the signal will not lie outside the range of the instrumentation amplifier. The second consequence is that negative feedback is applied to the common-mode signal, so that the amplitude of this (undesired) signal is reduced even further. Diodes D1 through D4, along with resistors R1 and R5, are added to the circuit to protect the inputs against damage from excessive electrostatic charges.
The CMRR (common-mode rejection ratio) of the instrumentation amplifier can be set using P1. To make this adjustment, connect both inputs of the instrumentation amplifier together,

and then connect a 100-mV, 50-Hz AC signal between the connected inputs and earth. Measure the output signal using an oscilloscope, and adjust P1 to minimise the level of the output signal.

It is important that the electrodes make very good contact with the skin. In our test measurements, winding three uninsulated copper wires several times around the index fingers (and the right leg) proved to be sufficient to provide a good signal. The amplitude of the ECG signal measured with this arrangement was 200 mV.

The current consumption of this circuit is only 2 mA, so the batteries should last a long time. This circuit must never be connected to a mains-operated power supply, in consideration of safety precautions that are necessary when making this sort of measurement on the human body.

(004022-1)



005 High-Voltage Difference Amplifier

A Burr-Brown Application

The INA146 is an integrated precision difference amplifier in an SO-8 style package. It can operate with a common mode input voltage in excess of the supply voltage to the IC. With a single sided +5 V supply this IC can handle common mode input voltages up to +40 V or ±100 V with a symmetrical ±15 V supply. The IC is ideally suited for use in applications such as high precision current shunt measuring in high voltage test equipment, as a sensor amplifier, as a monitoring device for powered vehicles or as a voltage controlled current source.

The internal circuit for this device makes use of on-chip precision resistors, laser trimmed to ensure accurate gain (achieving a gain error of only 0.025 %) and high (80 dB) common mode rejection. Excellent temperature coefficient of these on chip elements also ensures that the device maintains accuracy over a temperature range of –40-+85 °C.

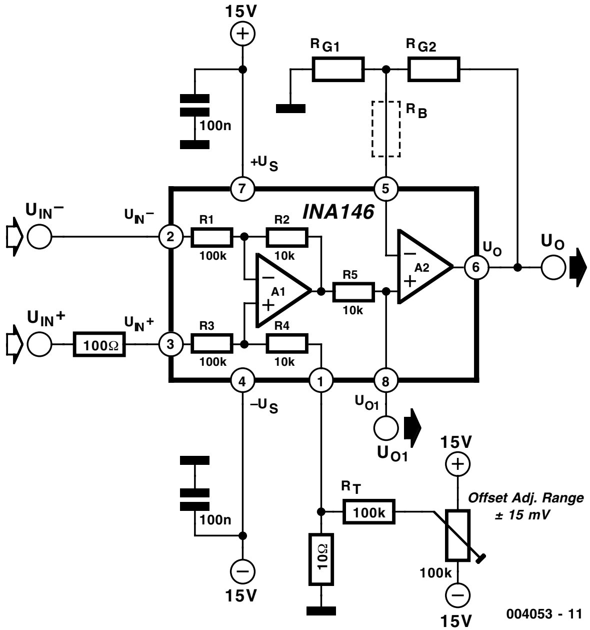
The INA146 has two internal amplifiers, the first difference amplifier has a gain of 0.1 V/V and the second buffer amplifier allows the gain of the device to be selected by fitting two external resistors R_{G1} and R_{G2}. Setting this gain to 1 will allow voltages up to 100 V to be measured. The output voltage of the device is given by:

$$V_O = (V_{IN}^+ - V_{IN}^-) \cdot 0.1 \cdot (1 + R_{G2} / R_{G1})$$

Commonly used values of gain are shown in the table.

To use the INA146 as a voltage controlled current source connect pin 1 to the output (pin 6) and connect a 10 kΩ resistor between pin 6 and pin 5. R_{G1} and R_{G2} are not fitted in this instance. The output current from pin 8 is given by:

$$I_{OUT} = (V_{IN}^+ - V_{IN}^-) / 10\ k\Omega$$



The 100 Ω resistor shown at the input (pin 3) and the network connected to pin 1 is optional and only required if you need to make a correction of the offset voltage. Using the values shown, the offset adjustment range is ±15 mV at the output. The offset correction network should only be used when it is necessary. To maintain good common-mode rejection, the source impedance connected to pin 1 should be less than 10 Ω and the resistor added to the positive input should be 10 times this value or 100 Ω.

The input impedance of the INA146 is determined by the input resistor network and is approximately 100 kΩ. The source impedance at the two input terminals must be nearly equal to maintain good common-mode rejection. A 12-Ω mismatch between these two impedances will result in a 8 dB degradation in the common-mode rejection ratio. Source impedances greater than 800 Ω are not recommended even if they are perfectly matched.
More information on the INA146 may be found on

Gain total (V/V)	Gain de A2 (V/V)	R_{G1} (Ω)	R_{G2} (Ω)	R_B (Ω)
0.1	1	-	10 k	-
0.2	2	20 k	20 k	-
0.5	5	12.4 k	49.9 k	-
1 10	11.0 k	100 k	-	
2 20	10.5 k	200 k	-	
5 50	10.2 k	499 k	-	
10	100	10.2 k	1 M	-
20	200	499	100 k	9.53 k
50	500	100	49.9 k	10 k
100	1000	100	100 k	10 k

http://www.burr-brown.com

(004053-1)

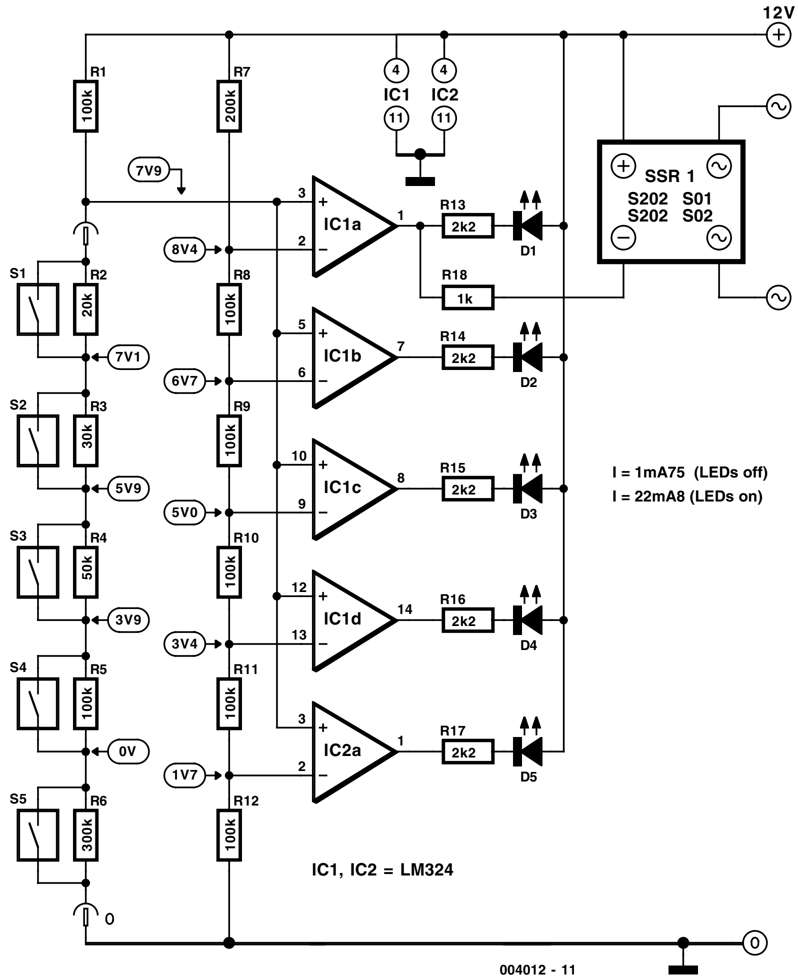
006 Rainwater Storage Gauge

Dr U. Pilz

Not only on ecologically grounds but also economically it makes sense to collect rainwater for use in the garden and increasingly for `grey water` domestic use. People who take rainwater collection seriously use large underground tanks for storage. The problem now arises, how can the water level be determined without lifting the tank hatch and peering in? One solution is to use float switches mounted at different heights in the water tank, and to use a row of LEDs mounted remotely to show the water level in the tank. Transferring this information to a remote display will involve long cable runs so it is of interest to reduce the number of cables to a minimum.

The circuit here shows how information about the water level in a tank can be sent over two wires to a remote LED display. R1 together with the resistor chain made up of R2-R6 form a voltage divider, float switches are wired across the resistors R2-R6 in one arm of the voltage divider. As water flows into the tank and the level rises, switch S5 closes followed by S4, etc. Each time a switch closes, it will short out its parallel resistor in the chain thereby changing the output voltage of the divider. When the tank is full, all five switches will be closed and all the LEDs will be on.

The voltage output from this divider chain is applied to the inputs of five op amps that are configured as comparators. A voltage chain comprising R7-R12 supplies the reference voltage for each of the five comparators. Both divider chains use the same supply so they will be insensitive to supply fluctuations. The maximum supply current for the circuit is less than 25 mA.



Some of the resistors chosen to make up the voltage dividers are not standard values but can be easily made up from combinations of 10 kΩ and 100 kΩ resistors.

If you need to expand this five level display to give a better resolution of the tank contents, it is a simple job to add more float switches and to expand the voltage divider chain. IC2 also has three spare op amps; these can be pressed into service as further comparators.

Underground tanks inevitably require a pump to move the water to where it will be used. An optional feature of this design is the pump protection circuit. When LED D1 goes off indicating that the tank is almost empty, the solid state relay SSR1 can be used to switch off the mains power to the pump. This will prevent damage to the pump when the tank runs dry. The S202 SE1 solid-state relay (SSR) from Sharp has an isolation voltage between its input and output of 3000 V (Class 1). It is important to note here that any mains equipment near the water tank installation must be supplied from an RCD safety socket for the sake of your own health!

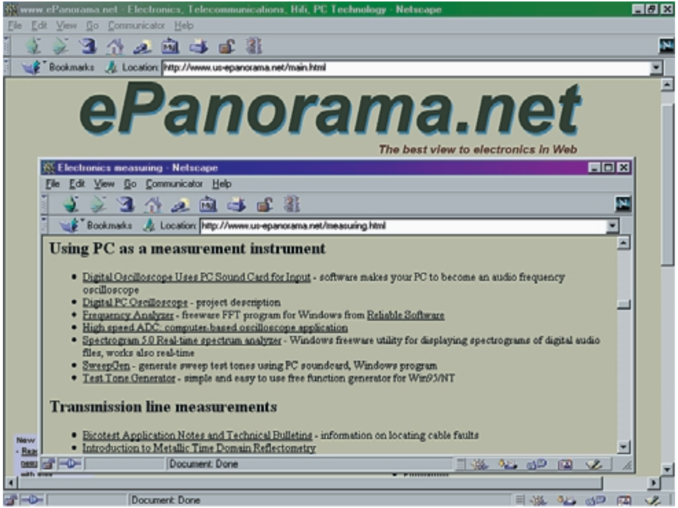
(004012-1)

007 ePanorama.net

H. Baggen

Many people visit the Internet electronics pages of Tomi Engdahl (*www.hut.fi/Misc/Electronics*). The master site is located on a fast server belonging to the Technical University of Helsinki. The pages with thousands of electronics-related links are the best known. A new site has now been opened for this link summary, with the name 'ePanorama'. You can find this at *www.us-epanorama.net/index.html*. The site has a very functional layout. On the opening page, you right away find an extensive list of electronics subjects, such as audio, books, cables, components, computers, digital signal processing, infrared, magazines, measuring, MIDI, oscillators, radio, soldering, standards, telecommunications and video. In total, there are presently around 60 headings. Each heading is broken down into subheadings. Under 'Measuring', for example, we found the subheadings 'Measuring and testing', 'General information', 'Oscilloscopes', 'PC as a measurement instrument', 'Transmission line measurements', 'Cable wiring testers', 'High voltage measurements', 'Capacitance measurements' and so on.

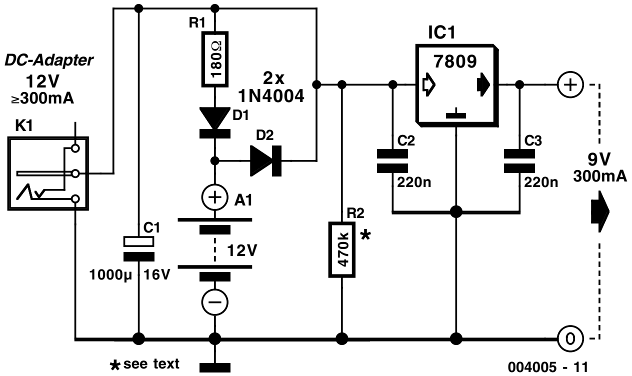
The number of available links on this site is truly enormous. If you are looking for something on a particular electronics subject, start here! There is a very good chance that you will quickly find what you need.



(004111-1)

008 DC Adapter with Battery Backup

Based on an idea from P. Lay



With just a low cost DC adapter and the circuit described here it is possible to build a low cost stabilised, uninterruptible 9 V supply. On the grounds of safety and economy, a simple unstabilised 12 V D.C. adapter is used as the power source, a universal adapter with its output set to 12 V will do equally well. The output voltage of an adapter under low load conditions (up to approximately 1/3 of the rated output current) is over 15 V, even at the rated output current, there will be sufficient voltage to supply a 9 V voltage regulator. The rating of the DC adapter should be chosen according to the output current required at 9 V. Common values are 300 mA, 500 mA and 1 A.

The 9 V voltage regulator used in this circuit has a built in thermal shutdown mechanism so that if too much current is drawn from the device, it simply turns off as it overheats and will not supply any current until the case temperature returns to normal. If the unit is intended to supply more than say 150-200 mA then to prevent thermal shutdown it will be necessary to fit a heatsink to the voltage regulator. The rule of thumb used to calculate the size of heatsink is that you should be able to touch it during operation at maximum load, without burning you finger.

When choosing the DC adapter, it is always better to select one with a higher current rating than is needed this will ensure that its output voltage is high enough to be able to also charge the 12 V cells. As long as mains voltage is on the DC adapter, the voltage across C1 will be higher than the voltage of the cells. Charging current will flow through R1 and D1 to the cells. Current also flows to the voltage regulator and out to the load connected at the output. Diode D2 in this situation will not conduct because the voltage at its cathode is greater than that at its anode

When the mains voltage fails or is turned off, diode D2 conducts and current will now flow from the Nickel Cadmium cells to the voltage regulator, thereby automatically keeping the output voltage at 9 V.

The value of resistor R1 is chosen so that a charging current to the cells is not greater than 1/10th of the cells capacity (if the cells are rated at 1100 mAh, the charging current must not exceed 110 mA). From the point of view of cell longevity it is better to reduce this charging current even further (1/20 or 1/50 C). When calculating this resistor, the value of the no-load voltage should be used. This will give the highest charging current. To calculate the charging current using R1 with a value of 180 Ω. The cells measure 13.8 V when fully charged and the no-load output voltage of the DC adapter is 17 V. Charging current is given by the formula:

(17 V – 13.8 V – 0.7 V) / 180 = 13.9 mA.

Substituting the actual measured values in this formula will enable you to calculate the value of R1 to give the correct charging current for the cells.

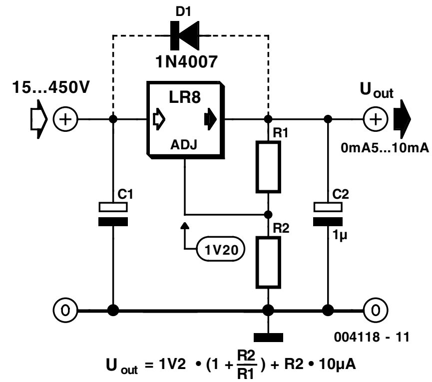
004005

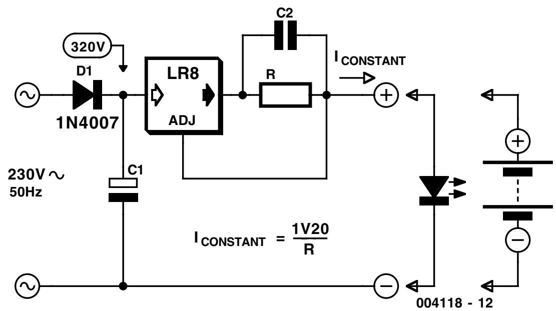
009 High Input Voltage Linear Regulator

G. Kleine

Commonly used 3-pin linear voltage regulators, for example the LM317, cannot handle input voltages in excess of about 30 V. The LR8A from Supertex Inc is a new, adjustable three pin regulator that can accept input voltages up to 450 V and can supply an output current of 0.5 mA to 10 mA. Using this device it is possible to work with rectified 230 VAC.

The LR8 has a wide input voltage range of +12 V to +450 V. Two external resistors (R1 and R2) allow the output voltage to be adjusted from 1.20 V to 440 V provided that the input voltage is at least 10 V greater than the output voltage. The LR8 adjusts the voltage difference between the V_{out} and ADJ pins to a nominal value of 1.20 V. This 1.20 V is amplified by the external resistor ratio of R1 and R2. An internal constant bias current of 10 µA is connected to the ADJ pin so that V_{out} is increased by a constant voltage of 10 µA times R2. The formula for calculating the output voltage is given next to the circuit diagram. To ensure stable operation of the regulator a minimum output current of 500 µA is necessary and a bypass capacitor of minimum 1.0 µF should be used. Protection circuits in the LR8 limit the output current to 15 mA typically and temperature protection ensures that the device temperature will not exceed 125ºC. When the device reaches its temperature limit, the output voltage/current will decrease to keep the junction temperature within limits.



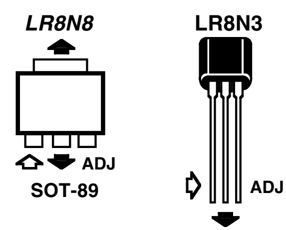


The two circuit diagrams show the LR8 used as a voltage regulator and as a constant current source. The current source can be used to a drive an LED. This configuration would give an LED with super-wide input voltage range, i.e., from +12 V to +450 V. The LR8 was originally designed to be used for switch mode supply start-up applications so it incorporates a feature which shuts down the LR8 when the output voltage exceeds the input voltage. Diode D1 is therefore necessary in the voltage regulator circuit diagram to prevent the output voltage exceeding the input voltage at any time.

The minimum value of the input capacitor C1 can be calculated from the following formula:

$$C1_{(min)} = (I_L\ t) / (V_{pk} - V_{out} - 10\ V)$$

Where I_L is the load current, and t the period between two voltage peaks. At 50 Hz, using one rectifying diode this will give a value t = 20 ms. V_{pk} is the peak input voltage, while V_{out} is the selected output voltage.

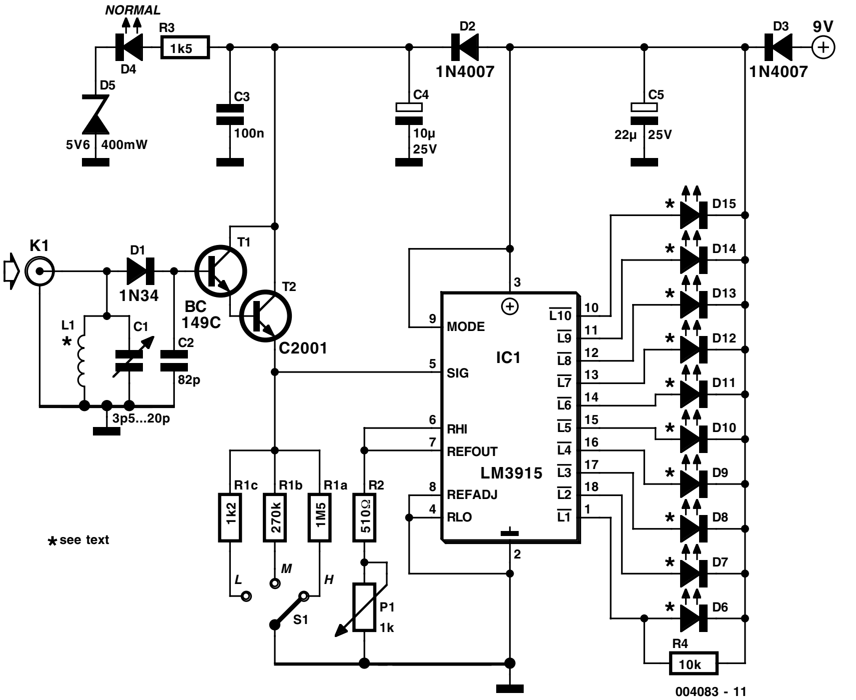


The LR8 is available in two package outlines. The LR8N8 is a SOT-89-SMD-package while the LR8N3 is the familiar TO-92-Transistor outline (e.g. BC 238). The TO-92 package can dissipate a maximum of 0.74 W while with suitable heatsinking, the SMD package can dissipate 1.6 W.

(004118)

Internet Address: www.supertex.com

010 Relative Fieldstrength Meter for 2-m Band



N.S. Harisankar, VU3NSH

A fieldstrength meter (FSM) is extremely useful for radiation gain/loss and radiation pattern measurements, and, in general, to discover why some antennas perform better than others.

This FSM is designed for work on 2-metre band (144-146 MHz) antennas, in particular, Yagi types which require quite a bit of tweaking for optimum performance. A tuned RF filter, L1-C1, is used at the input to make the FSM selective for the desired band. The heart of the circuit is the popular LM3915 3 dB/step analogue display driver from National Semiconductor. A continuous-scale readout is available in the form of a LED bar. Dot mode may also be selected by disconnecting IC1 pin 9 from the positive supply line. In 'dot' mode, current consumption will be much lower than in 'bar' mode when up to 100 mA may be drawn from the battery when all LEDs (bar segments) are on. The FSM is powered by a 9V (6F22) battery.

The inductor in the filter is home made, it consists of 2.5 turns of 22 SWG copper enamelled wire. No core is used, and the internal diameter is 7 mm. The two-transistor combination behind the filter forms a high-gain amplifier whose output voltage is developed across one of three resistors connected to ground by the L-M-H sensitivity selector, S1.

The circuit should be installed in a metal box. The three LEDs for the lowest levels are green. Then come three yellow LEDs, two orange LEDs and finally two red LEDs for the highest fieldstrength levels. The highest level LED is wired to IC1 pin 10, the lowest level LED, to pin 1.

The FSM is a breeze to adjust. Connect a rubber helical antenna to the FSM input and key your transmitter on 145 MHz, running no more than 500 mW output. Set up the FSM at a reasonable distance from the transmitting antenna. Increase or decrease the distance to the antenna until two or three segments light, then adjust C1 for the highest number of segments to light. If necessary, adjust the brightness control P1 to suit ambient light conditions.

(004083-1)

011 Scope Triggering Aid for Video Signals

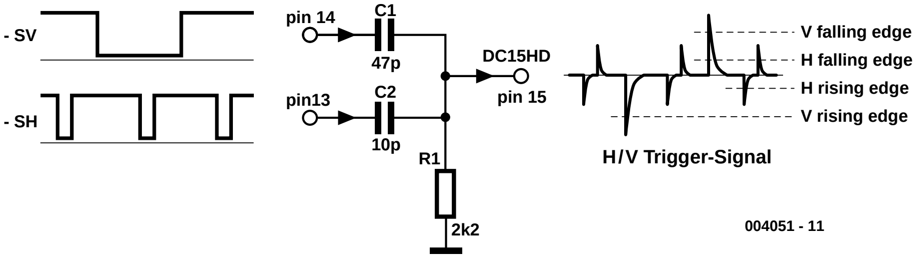
W. Foede

This small circuit is useful for investigation and testing video signals. It allows you to select different triggering points on a VGA synchronisation signal by simply adjusting the trigger level of the oscilloscope. The trigger points are as follows (from the highest to the lowest trigger level):

- Triggering on the back edge of the vertical sync signal.
- Triggering on the back edge of the horizontal sync signal.
- Triggering on the front edge of the horizontal sync signal.
- Triggering on the front edge of the vertical signal.

The circuit could hardly be simpler; it consists of two capacitors and one resistor. C1 and R1 form a differentiating net-

work with a time constant conforming to the vertical sync signal; likewise, C2 and R1 perform the same function on the horizontal sync signal. In the accompanying diagram, it can be seen how the combined trigger pulse output is derived from the edges of the two sync signals.
The pin numbers given in the diagram refer to a 15-pin PC video connector (DC15HD). It is possible to mount the three components of this circuit directly onto the PCB side of the graphic card video connector, in which case you can use pin 15 (normally unused) as the trigger signal output. (004051-1)



012

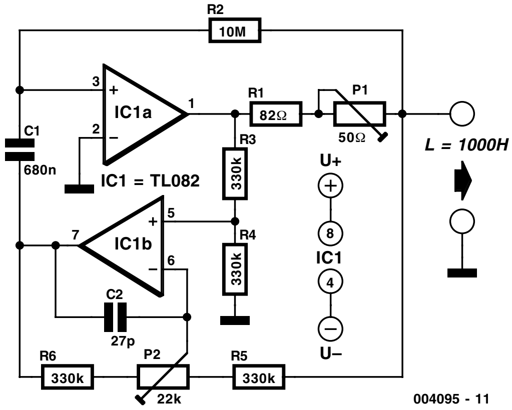
1 kH Synthetic Inductor

B. Kainka

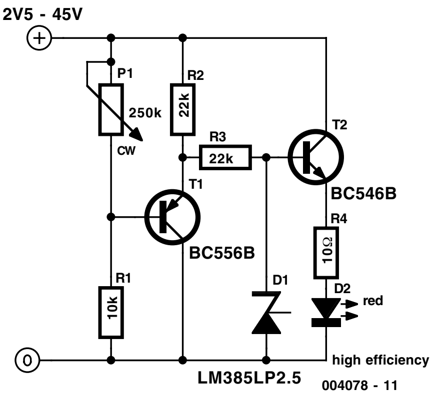
Inductors can be mimicked quite easily using operational amplifiers. The circuit shown here was developed to have an inductance of 1000 H (say, one thousand henry) with good damping. Using this design you can build a resonant circuit with a centre frequency of less than 1 Hz. The slow behaviour allows you to use conventional measuring instruments to investigate the circuit in real time. The circuit can also be used as part of a filter design.

Opamp 1 operates as an Integrator, Opamp 2 as a difference amplifier. The output voltage of Opamp 2 is equal to the voltage drop across R1 and P1, which is proportional to the output current. This voltage is differentiated by Opamp 1, C1 and R2. The net effect is that the circuit behaves as an inductor. P1 allows adjustment of the inductance value. P2 allows adjustment of the Q factor of the coil by altering the symmetry of the difference amplifier and with it the stability of the circuit.

(004095)



013 Simple Voltmeter



H. Bonekamp

This circuit provides a simple means to determine the voltage of a low-impedance voltage source. It works as follows. P1, which is a 1-W potentiometer, forms a voltage divider in combination with R1. The voltage at their junction is buffered by T1, and then passed to reference diode D1 via R3. D1 limits the voltage following the resistor to 2.5 V. An indicator stage consisting of T2, R4 and LED D2 is connected in parallel with D1. As long as the voltage is not limited by D1, the LED will not be fully illuminated. This is the basic operating principle of this measurement circuit.

If you rotate P1 from its maximum setting towards the minimum until the LED is fully illuminated, you know that the voltage is limited by D1 at this setting, and that the voltage across D1 is exactly 2.5 V. The voltage across R1 is then equal to ($U_{D1} - U_{BE1}$) = (2.5 – 0.5) V = 2 V. This means that you have used P1 to set the voltage across R1 to 2 V. Based on this, you can determine the input voltage as follows. The voltage across R1 is given by:

$$U_{R1} = \frac{R1}{R1 + \alpha \cdot P1} \cdot U_{in}$$

From this, it follows that the input voltage is:

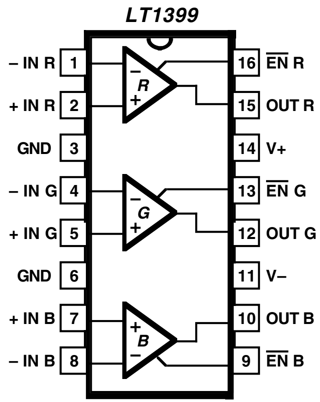
$$U_{in} = \left(1 + \frac{\alpha \cdot P1}{R1}\right) U_{R1}$$

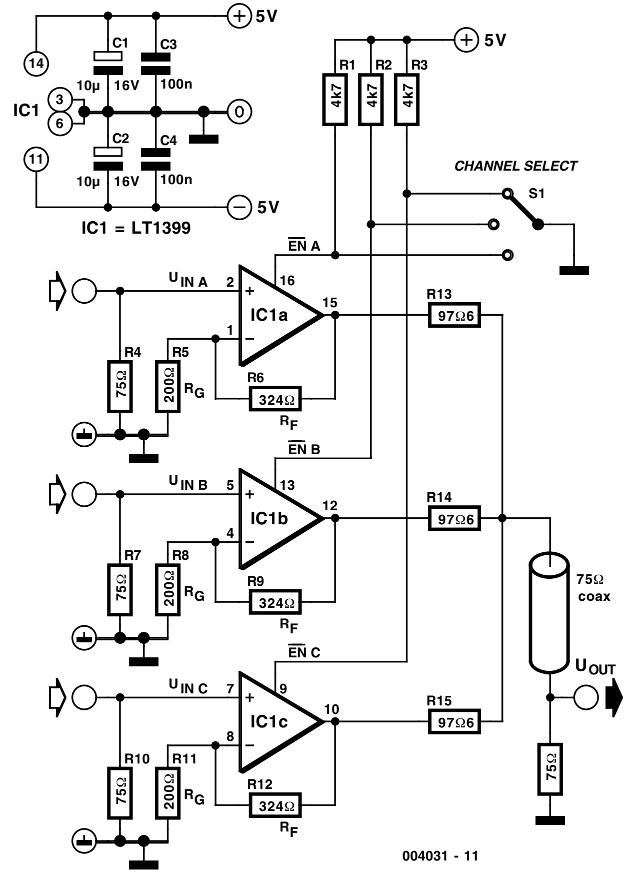
You can see that U_{in} is nearly equivalent to the amount of rotation of P1. All you have to do to complete the voltmeter is to provide a scale for P1. You can calibrate the scale by connecting a real voltmeter in parallel with the circuit and varying the input voltage.

The current consumption of this circuit is approximately 8 mA. The voltage source to be measured must be able to easily supply this current, since otherwise the voltage will drop too much during the measurement.

(004078-1)

014 3-Input Video MUX Cable Driver





Based on a Linear Technology Application Note

The circuit diagram shows a low-cost 3-input video MUX cable driver. In this circuit, the amplifier is loaded by the sum of R_F and R_G of each disabled amplifier. Resistor values have been chosen to keep the total back termination at 75 Ω while maintaining a gain of 1 at the 75-Ω load. The switching time between any two channels is approximately 32 ns when both enable pins are driven.

When designing a circuit board for this cable driver, care should be taken to minimize trace lengths at the inverting input. The ground plane should also be pulled away from R_F and R_G on both sides of the board to minimize stray capacitance.

Current consumption of the cable driver is a modest 8 mA.

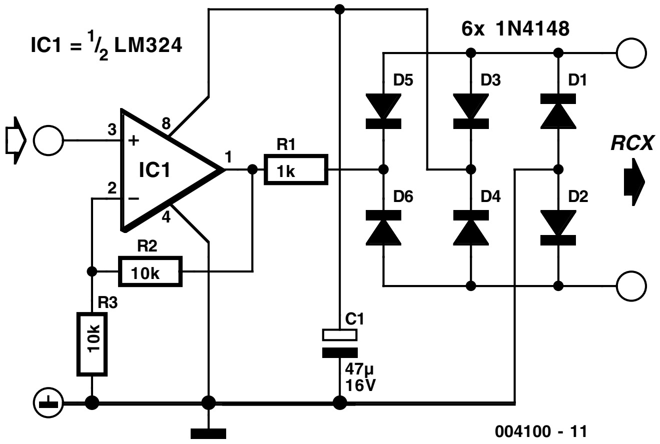
(004031-1)

015 Analogue Input for Lego RCX Module

H. Steeman

Up to now, there has not been an analogue input for the RCX block. The circuit presented here allows the RCX block to detect analogue voltages. The approach used here even allows this function to be used with existing Lego software, although this naturally amounts to a certain 'misuse' of the interface.

If an RCX input is set to the 'light sensor' mode, the internal logic senses the resistance between the two terminals of the selected input. The circuit presented here converts an analogue input voltage in the range of 0 to 2.5 V into something like a resistance. It's easy to explain how this is done. The analogue input voltage, amplified by a factor of two, appears at the out-



put of the opamp. Resistor R1 is located between the output of the RCX block and the buffered input voltage. If the input voltage is low, there is a relatively large voltage across R1. The RCX block measures a relatively high output current and thus thinks that the sensor has a low resistance. If the buffered input voltage is nearly 2.5 V, there is almost no voltage across R1. Little current flows, and the RCX block sees a high resistance.

Based on experiments, a value of 1 kΩ is a good choice for R1, but you may want to try other values. The value of R1 may be increased up to a maximum of 3.3 kΩ. Above this value, the characteristic of the adapter deviates too much from that of the light sensor, so that it is not possible to use the full measurement range. With the suggested value for R1, the measurement value ranges from 95 (with an input of 0 V) to approximately 5 (with an input of 2.5 V). The curve between these two values is reasonably linear in the region between 0.5 V and 2.5 V.

The rest of the circuit is self-explanatory. Diodes D1 through D4 form a full-wave bridge rectifier, so that the polarity of the connecting wires for the RCX input does not matter. The rectified voltage is buffered by capacitor C1 and then used to power the opamp. Measurement resistor R1 is also connected to the RCX input by two diodes, in order to avoid polarity problems with the connections.

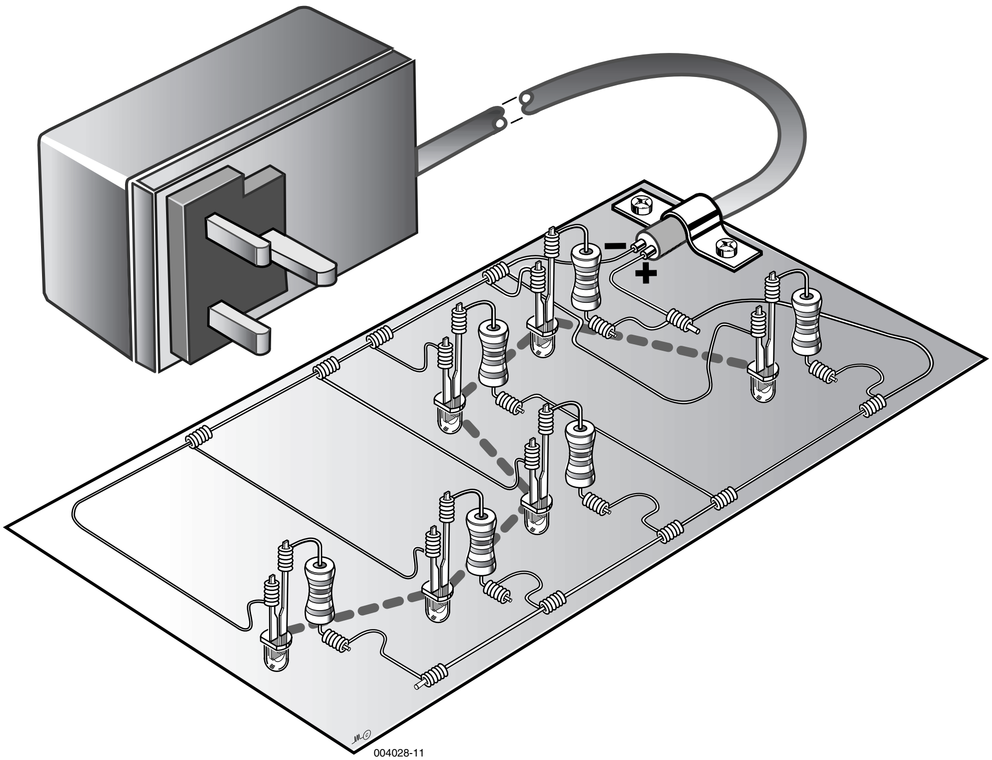
(004100-1)

016 Night Sky Model

K. Walraven

Now that white LEDs are available at reasonable prices, one nice application is an imitation of the night sky. The LEDs emit a rather cold bluish light (6500 to 8000 Kelvin), which gives a good 'star' effect. The magnitudes (relative brightnesses) can be reproduced quite well by using different values of series resistors.

There are various ways to build such a model. The simplest is to use a piece of black cardboard, with holes for the LEDs punched in the proper locations. The LEDs can then be taped to the cardboard. 'Umbrella' models are also very popular. Here a (large) black umbrella provides a darkened background, and the LEDs are glued into small holes that are carefully made in the umbrella fabric.



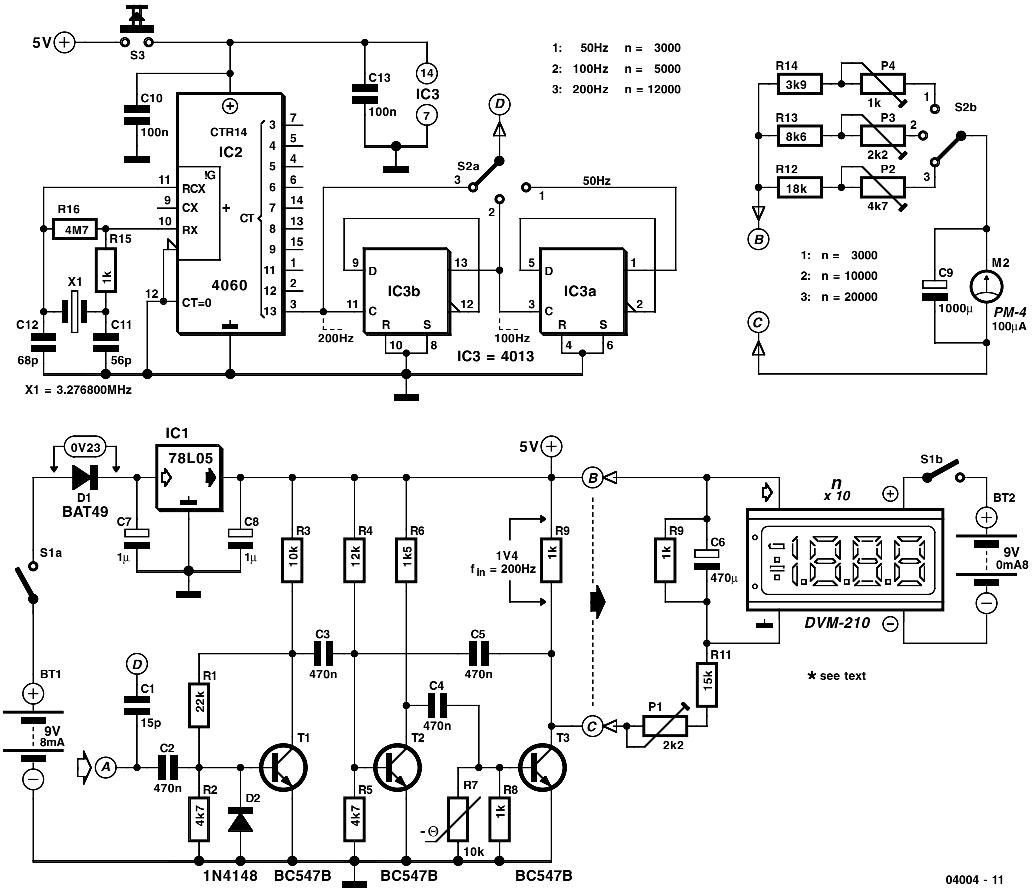
The circuit is based on a normal mains adapter, which delivers a safe voltage. A standard adapter can deliver 300 mA, which is enough to power 15 LEDs at 20 mA each. For a larger night sky model, you will need a more powerful adapter (500 or 1000 mA), or you will have to divide the sky into several sections, each with its own mains adapter. We want to have at most 20 mA flow through each LED. If the adapter output is set to 6 V, the value of the series resistor can be easily calculated:

$$R = (6 - 3.5)\ \mathrm{V} \div 20\ \mathrm{mA} = 125\ \Omega$$

We can use a standard value of 120 Ω or 150 Ω. Larger values will cause the LED to be less bright.

(004028-1)

017 Universal Rev Counter



K.-H. Lorenz

This rev counter is unusual in that it has no direct electrical connection to the ignition system of the engine. It is also unnecessary to add a winding to an ignition lead. If the ignition system on the engine is electronic then all that is necessary for the spark impulse signal pickup is a length of wire positioned 10-30 cm from the ignition lead. For an engine with magneto ignition this distance will be less.

The original idea for this design was to make a hand held rev counter or tachometer with its own battery supply that could be used to check engine speed or to calibrate another built-in rev counter. The good resolution of this design and its convenience in operation mean that it is also ideal for use when adjusting the engine idle speed.

The display for this circuit is a commercially available DVM panel meter and will show engine speeds up to 19,999 r.p.m. Alternatively a 100 µA-moving coil meter can also be used to give an analogue display.

The circuit consists of three transistors. T1 amplifies the incoming signal while T2 and T3 perform pulse shaping. The resulting pulses are fed to an integrating network formed by preset P1, R11 and C6; the DVM measures the DC voltage across C6.

An optional built-in generator is also shown. This is used to calibrate the rev counter and also test the display. If a DVM panel meter is used for the display, then the 200 Hz output of the 4060 (IC2) at pin 3 can be connected directly to the input of the rev counter (point D in the diagram). 200 Hz corresponds to an engine speed of 12,000 r.p.m. for a 2-stroke, single cylinder engine, so adjust preset P1 until the display corresponds to this.

If a moving coil meter is used for the display, the speed measurement is divided into three ranges selected by S2. The second set of contacts of S2 (S2a), are used to switch an appropriate calibration frequency from the generator. So, with S2 in position 1 preset P4 is adjusted until 3000 r.p.m. is shown on the meter (i.e. full scale deflection). The same process is repeated with S2 in position 2 and 3 this time adjusting P3 and P2 for 6000 and 12000 r.p.m. respectively.

The supply to the rev counter is provided by a 9 V battery to a low-drop 5 V voltage regulator. Schottky diode D1 gives protection against accidental battery reversal.

The DVM will also require its own supply (not necessary if you are using a moving coil meter). The current consumption of the rev counter is approximately 8 mA, while the calibration circuit consumes substantially less and is only active during calibration when S3 is pressed.

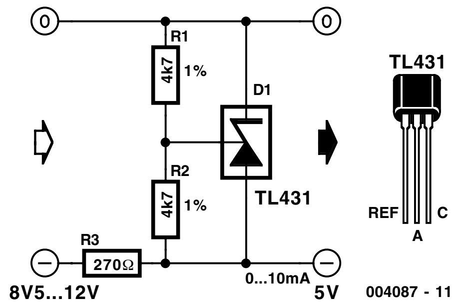
Note that the r.p.m. referred to above is valid for a single cylinder 2 stroke engine and will need to be converted for the type of engine configuration which is being measured. The relationship between the impulse frequency and the r.p.m. of an engine is given by the formula:

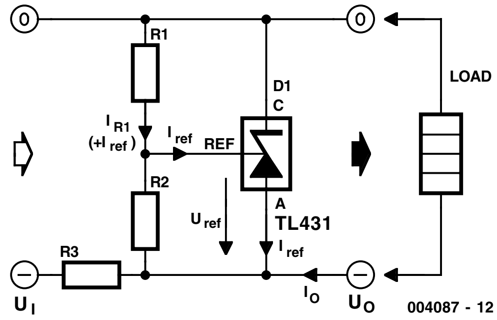
$$f = 2nc/60t$$

Where f = Impulse frequency, n = r.p.m. c = No of cylinders, t = type of engine (2 = 2-stroke, 4 = 4-stroke).

004004

018 Negative Shunt Regulator





K. Thiesler

Using just a few components and the circuit shown (**Figure 1**) it is possible to build a regulated negative supply with an output voltage of –2.5 V to –36.0 V with an output current in the order of 100 mA. The TL431 shunt regulator from Fairchild Semiconductor has good accuracy and only requires an unregulated negative voltage source. The Drop-out-voltage (Vin – Vout) of the regulator appears across resistor R3. The shunt regulator must always have a load connected at its output.

The negative output voltage is set by the voltage divider formed by R1/R2

$$U_O = (1 + R1/R2)\cdot U_{REF} + R1\cdot I_{REF}$$

where
$U_{REF} = -2.495$ V (nominal)
$I_{REF} = -2\ \mu A$

This gives the relationship between the output voltage, the voltage divider network and the internal voltage reference of the TL431.

To ensure correct operation, the current through the divider network (I_{R1}) must be many times higher than the reference current (I_{REF}) the value of R1 is given by:

$$R1 = (U_O - U_{REF}) / I_{R1}$$

So choosing an output voltage of –5 V and using an I_{R1} of –100 µA i.e. 50 times I_{REF} gives a value for R1 of:

$$R1 = (-5\ V - 2.495\ V) / -10^{-4}\ A \approx 25.05\ k\Omega$$

The second resistor is given by:

$$R2 = U_{REF} / [(U_O - U_{REF})/R1 - I_{REF}] \approx 25.46\ k\Omega$$

For correct operation of the regulator the value of cathode current must be between –1 mA and –100 mA, and the load current should be constant. Both currents flow through R3 and are used here to estimate its value:

$$(U_{Imin} - U_O)/(I_{Omax} + I_{Cmin}) \geq R3 \geq (U_{Imax} - U_O)/(I_{Omin} + I_{Cmax})$$

Using the minimum resistor values will obviously increase the power dissipation in the circuit. The power dissipated in resistor R3 is given by:

$$P_{R3} = (U_{Imax} - U_O)^2 / R3$$

The power dissipated in the regulator is given by:

$$P_{TL431} = U_O \cdot I_{Cmax}$$

If an output voltage of –5 V is required then calculations becomes a bit simpler, R1 and R2 will be of equal value and the formula reduces to:

$$U_O = -2\cdot U_{REF} - R1\cdot I_{REF}$$

$$R1 = R2 = (U_O + 2\cdot U_{REF}) / I_{REF}$$

With a nominal reference voltage (U_{REF}) of –2.495 V this gives a resistor value of 5 kΩ. Changing this value appreciably will alter the reference current and therefore the output voltage. The diagram in **Figure 2** gives the complete circuit with component values for a –5 V voltage regulator.

There are three versions of this chip from Fairchild that improve on the U_{REF} specification. The standard TL431 with no suffix has a U_{REF} tolerance of ±2 %, while the A suffix has ±1 %, and the **L** suffix has ±0.4 %. A data sheet (albeit with a few typos) can be downloaded from the Fairchild website:
http://www.fairchildsemi.com/ds/TL/TL431.pdf
Maxim also produce a similar device with fixed output voltages. The MAX6330 and MAX6331 is available in 5.0 V, 3.3 V and 3.0 V versions, with an output current of 100 µA to 50 mA. The voltage divider network is integrated on the chip.

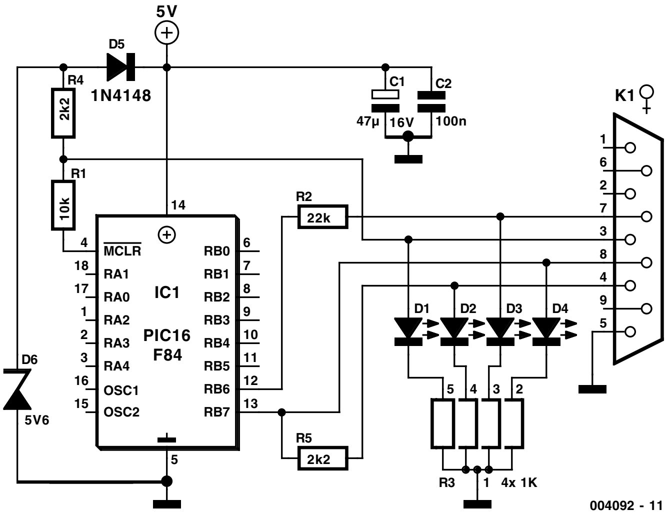
(004087)

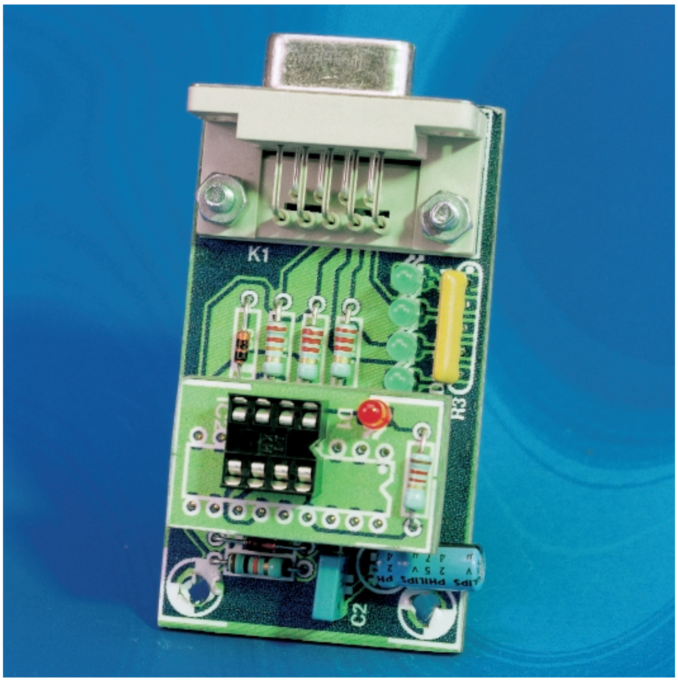
019 PIC16F84/16C84 Mini Programmer

J. Klein

As new microcontroller chips appear on the market, today's microcontroller programming tools are becoming more and more 'universal' to cope with different programming conventions. It is also sadly the case that the more 'universal' the programmer, the more you need to pay. In practice, most people will only use a fraction of the capabilities of such a programmer, making it difficult to justify such an expense.

The project here describes a minimal solution to the programming problem for one of the most popular types of controller. The PIC16F84 (1k-Flash-memory) and the PIC16C84 (1k-ROM) with 13 I/O-lines. Using a PC together with this relatively simple interface and some software it is possible to build a low-cost programmer





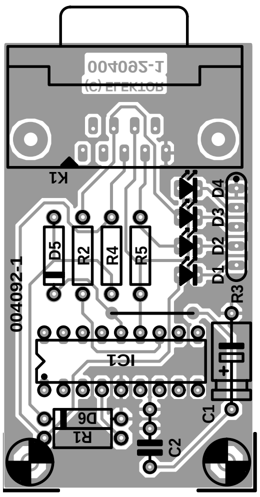
The design for the programmer is described on the author's website. The programmer connects to the serial port of a PC. Pin 3 of the port supplies the power and zener diode D6 along with D5 regulates the supply to the chip at 5 V. C1 and C2 smooth the regulated supply. The unregulated supply is fed to pin MCLR of the PIC to configure it in programming mode. R1 limits current into this pin and an internal regulator ensures the correct programming voltage on chip. A high on this pin switches the PIC into programming mode. Data exchange between the PC and the PIC occurs over the lines TxD (Pin 3), DTR (Pin 4) and CTS (Pin 8) and can be viewed on the LEDs D2, D3 and D4.

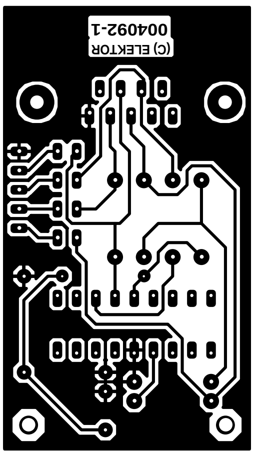
A control software package comprising NTPICPROG, PIX and Euro13 for Windows and DOS (altogether 198 kB) can be downloaded free from the 'Elektro' page of the authors website at http://jump.to/gate

Also available from the website is the Eagle and PDF data for the author's circuit board, along with the circuit diagram and some pictures. The circuit board shown is an *Elektor Electronics* design, the layout can also be downloaded from the Free Downloads section on the *Elektor Electronics* web site: http://www.elektor-electronics.co.uk

The board is unfortunately not available ready-made through the Publishers' Readers Services.

(004092-1)





Elsewhere in this edition is an EEPROM Adapter that can be used with this PIC programmer.

COMPONENTS LIST

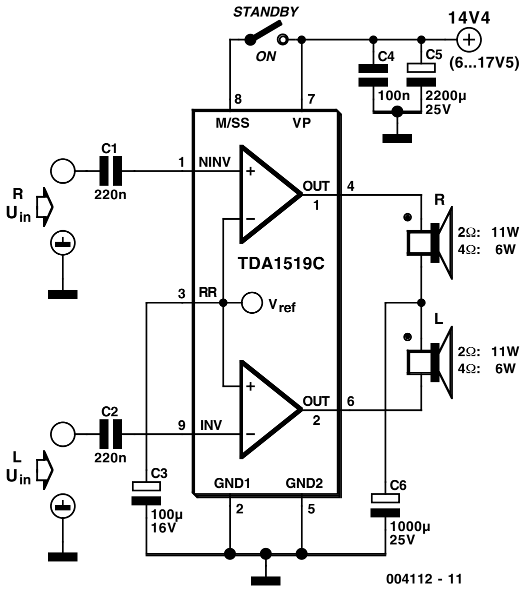
Resistors:
R1 = 10kΩ
R2 = 22kΩ
R3 = 4-way SIL array 1kΩ
R4,R5 = 2kΩ2

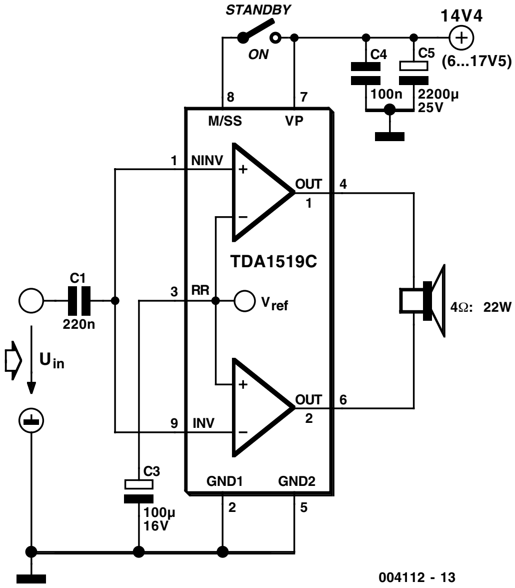
Capacitors:
C1 = 47µF 16V
C2 = 100nF

Semiconductors:
D1-D4 = LED
D5 = 1N4148
D6 = zener diode 5V6, 100 mW
IC1 = PIC16F84

Miscellaneous:
9-pin sub-D socket (female), angled pins, PCB-mount version

020 11 W Stereo or 22 W Mono Power Amp





G. Kleine

Integrated AF power amps have seen great improvements in recent years offering improved power and easier use. The TDA1519C from Philips contains two power amplifiers providing 11 W per channel stereo or 22 W mono when the two channels are connected in a bridge configuration. The special in-line SIL9P package outline allows the chip to be conveniently bolted to a suitable heatsink. The TDA1519CSP is the SMD version, in this case the heat sink is mounted over, and in contact with, the top surface of the chip.

The operating voltage of this device is from +6 V to +17.5 V. The two channels of the amplifier are different in that one channel, between pins 1 and 4, is a non-inverting amplifier, while the other between pins 9 and 6 is an inverting amplifier. It is therefore necessary in stereo operation, to wire the speakers so that one of them has its polarity reversed. Each amplifier has an input impedance of 60 kΩ and a voltage gain of 40 dB, i.e. 100 times. When both amplifiers are used in a bridge configuration, the inputs are in parallel so that the input impedance will be 30 kΩ.

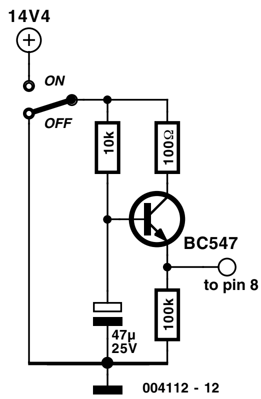
A combined mute/standby function is provided on pin 8. In its simplest form this can be connected to the positive rail via a switch. When the switch is open the amplifier will be in standby mode and current consumption is less than 100µA. When the switch is closed, the amplifier will be operational.

A circuit is also shown that uses the mute input to prevent the annoying switch-on plop heard when power amps are first switched on This is caused by the rush of current to charge capacitors C1 and C2. The circuit shown generates a ramp voltage, which is applied to pin 8. At switch on, as the voltage rises from 3.3 V to 6.4 V, the amplifier will switch out of standby mode and into mute mode allowing C1 and C2 to charge. Only when the ramp voltage on pin 8 reaches 8.5 V will the amplifier switch into active mode.

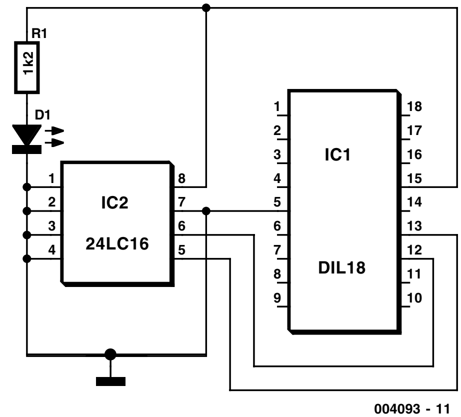
Protection built into the TDA1519C would seem to make it almost foolproof. The two outputs can be shorted to either of the supply rails and to each other. A thermal shutdown will prevent overloading and the power supply input is protected against accidental reversal of the supply leads up to 6 V.

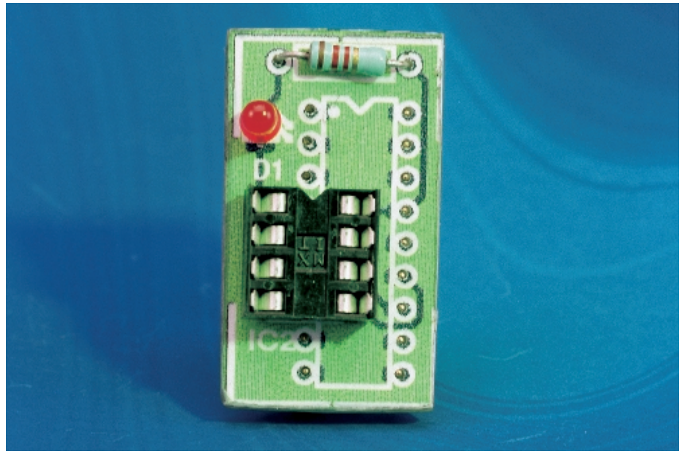
(004112)

Internet Address: www.semiconductors.philips.com



021 EEPROM Adapter





J. Klein

Using this adapter, the PIC16F84/16C84 Mini Programmer described in this issue of *Elektor Electronics* can also be used to program serial EEPROMs such as the 24LC16 and similar. The adapter itself consists of a small PCB onto which is mounted an 8-pin DIL socket to take the EEPROM. Underneath the PCB is an 18-pin DIL header (a socket with pins on both sides) which plugs into the PIC socket of the Mini Programmer.

Construction could not be simpler. Firstly solder the 8-pin IC socket, LED and resistor onto the postage stamp sized PCB. Next, position the 18-pin DIL header on the track side of the PCB, spaced so that you will be able to reach each pin with the soldering iron bit and solder it to the PCB.

To program the EEPROMs you will need the PC control software "NT Pic Programmer", this can be downloaded, free from the author's website at *http://jump.to/gate*

The PCB shown here is unfortunately not available ready-made through the Pubishers' Readers Services.

(004093-1)

COMPONENTS LIST

Resistors:
R1 = 1kΩ2

Semiconductors:
D1 = LED
IC1 = 24LC16

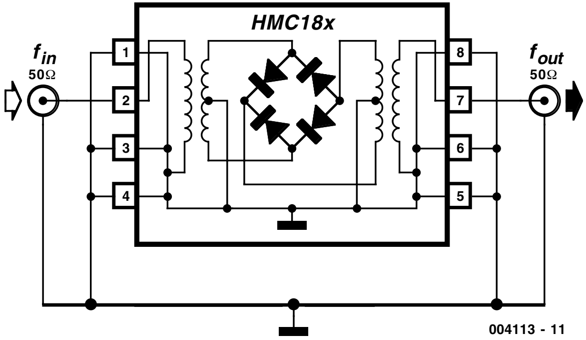
Miscellaneous:
18-way DIL header with pins at both sides
8-way IC socket

022 Frequency Doubler

G. Kleine

If you are working at frequencies of the order of 850 MHz to 4 GHz and find that a frequency multiplier is required, the HMC 187, HMC 188 and HMC 189 (see table) frequency doubler may be just the solution you are looking for. The isolation performance of these devices ensures that the input frequency (f_{in}) and its harmonics $3f_{in}$ and $4f_{in}$ are attenuated by 35 dB relative to the wanted output frequency $2f_{in}$. This excellent isolation specification reduces the need for additional output filtering and is also an advantage where several doublers are connected in series to produce four or eight times the input frequency.

The tiny outline of the HMC18x- series device occupies a board area of 3 mm by 4.8 mm and measures just 1.07 mm high.



Internally the device contains balanced to unbalanced transformers (baluns) to match the doubler circuit with the output and input. The doubler circuit itself is passive and comprises a full wave Schottky diode bridge rectifier. The monolithic baluns which are integrated on-chip give the device a relatively high low-frequency roll-off at 850 MHz. Lower frequencies can also be multiplied but the conversion loss factor (given as typically 15 dB) will increase. The input and output are matched for 50 Ohm operation and the input signal level should be of the order of +15 dBm which will give a output level of approximately 0 dBm.

The main characteristics of the three versions of this device are summarised in the table below

Type	f_{in}	f_{out}	Conversion Loss	Isolation at output		
				f_{in}	3 f_{in}	4 f_{in}
HMC 187	0,85 - 2 GHz	1,7 - 4 GHz	15 dB	45 dB	52 dB	40 dB
HMC 188	1,5 - 2,5 GHz	3 - 5 GHz	15 dB	45 dB	50 dB	45 dB
HMC 189	2 - 4 GHz	4 - 8 GHz	13 dB	34 dB	40 dB	40 dB

(004113)

Internet Address: www.hittite.com

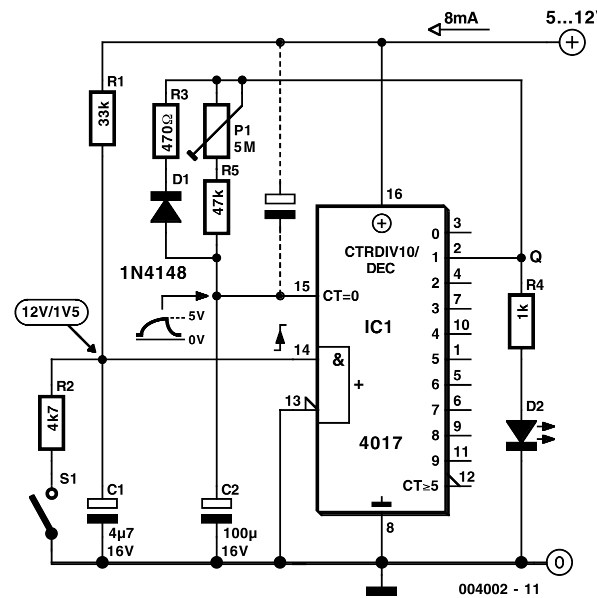
023 Flipflop Timer

J. Graßmann

This circuit shows how a 4017 CMOS decade counter can be used to build a timer circuit. Pushbutton S1 will discharge capacitor C1 through resistor R2. When S1 is released, C1 will charge up through R1 causing a rising edge at the clock input of IC1. This causes the output Q1 to go high (to the supply voltage). Current will flow through R4 and LED D2 will light. At the same time C2 will begin charging through preset P1 and R6. When the voltage on C2 reaches approximately half the supply voltage it will reset IC1. Q1 will go low, the LED will go off and C2 will discharge through D1 and R3. The circuit will now remain stable in this reset condition until S1 is pressed again. Preset P1 allows the ON time of the circuit to be adjusted between 5 seconds and 7 minutes.

The current consumption of this circuit in its reset state is only a few micro-amps, rising to approximately 8 mA mainly due to the LED current, when S1 is pressed. When power is applied to the circuit IC1, can be in an indeterminate state and the LED may be on. Pressing S1 until the LED goes off clears this condition. Alternatively C2 may be connected to the supply rail (as shown dotted in the diagram) this will ensure that IC1 will always power up in a reset state. A disadvantage of this configuration is that any noise on the supply rail will be coupled through to the reset pin of IC1 and may affect the timing period.

(004002-1)



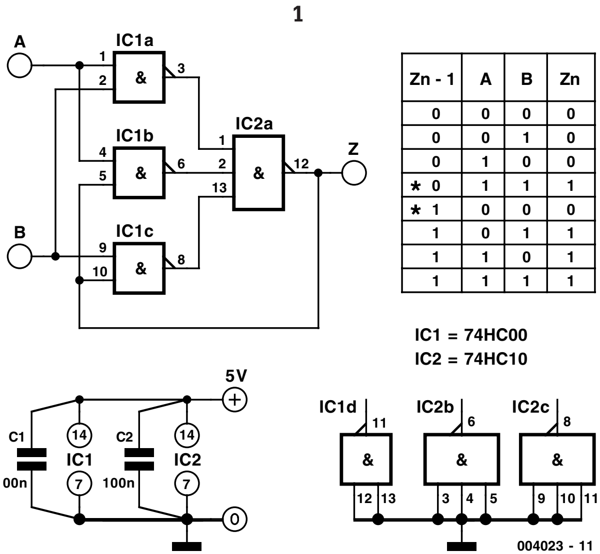
024

The Muller-C Element

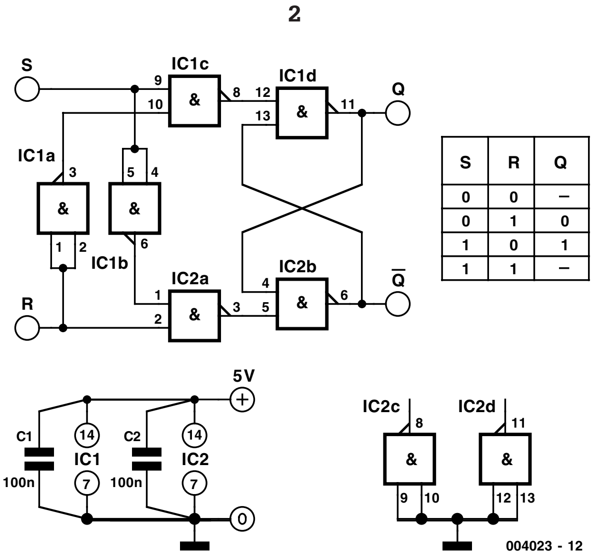
K.-J. Thiesler

The Muller-C element is a special type of flipflop. Its function is that it will only SET its output when all its inputs change to a 1 (on the rising edge). Likewise it will only RESET its output when all its inputs change to a 0 (on the falling edge). All other input combinations will result in no change to its output. **Figure 1** shows an implementation of the circuit using NAND gates. The memory function of the Muller-C element has some distinct advantages over other flipflop configurations:

1. There are no forbidden input combinations as with some other flipflops.
2. The propagation delay through the flipflop will be equal to the delay through only two gates.
3. The Muller-C element does not require an external clock. Its output changes state only when both inputs have a new state. It is self-synchronising.



Zn - 1	A	B	Zn
0	0	0	0
0	0	1	0
0	1	0	0
* 0	1	1	1
* 1	0	0	0
1	0	1	1
1	1	0	1
1	1	1	1



S	R	Q
0	0	–
0	1	0
1	0	1
1	1	–

4. In a control circuit using many coupled C elements there will be no clock i.e. in a synchronous system, the clock edges cause many circuit elements to switch at the same time. Careful design is necessary to comply with EMC requirements

Figure 2 shows a Set/Reset flipflop with additional NAND gates to prevent the forbidden input condition. As you can see, it uses more gates than the Muller-C element.

(004023)

025 Bit Calculator

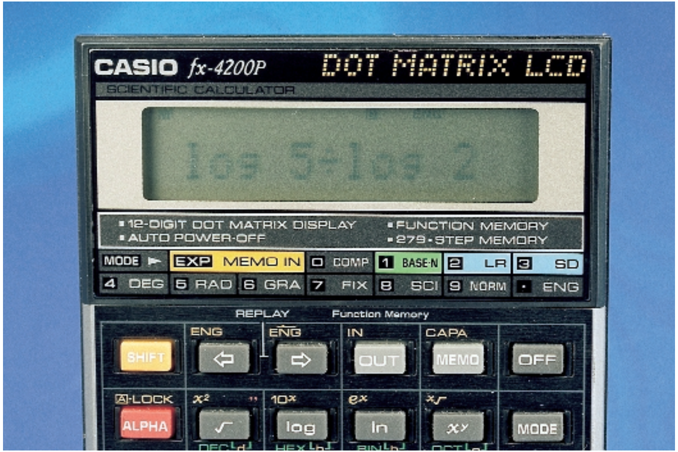
H. Bonekamp

Suppose you want to make a counter that can count from 0 to a certain decimal number M. If you want to implement this as a binary counter, you can calculate how many bits you need with a simple pocket calculator, using the following formula:

$$N = \frac{{}^{10}\log(M)}{{}^{10}\log(2)}$$

In principle, you can use any logarithm, including ln( )! Don't forget that you must round off the result!

(004080-1)

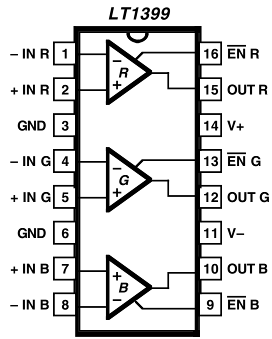


026 RGB-to-Colour difference converter

Based on a Linear Technology Application Note

The circuit diagram shows two LT1398's from Linear Technology used to create buffered colour-difference signals from RGB (red-green-blue) inputs. In this application, the R input arrives via 75-Ω coax. It is routed to the non-inverting input of amplifier IC1a and to 1.07-kΩ resistor, R8. There is also an 80.6-Ω termination resistor R11, which yields a 75-Ω input impedance at the R input when considered in parallel with R8. R8 connects to the inverting input of a second LT1398 amplifier (IC1b), which also sums the weighted G and B inputs to create a –0.5Y output. Yet another LT1398 amplifier, IC2a, then takes the –0.5Y output and amplifies it by a gain of –2, resulting in the +Y output. Amplifier IC1a is configured for a non-inverting gain of 2 with the bottom of the gain resistor R2 tied to the Y output. The output IC1a thus results in the colour-difference output R–Y.

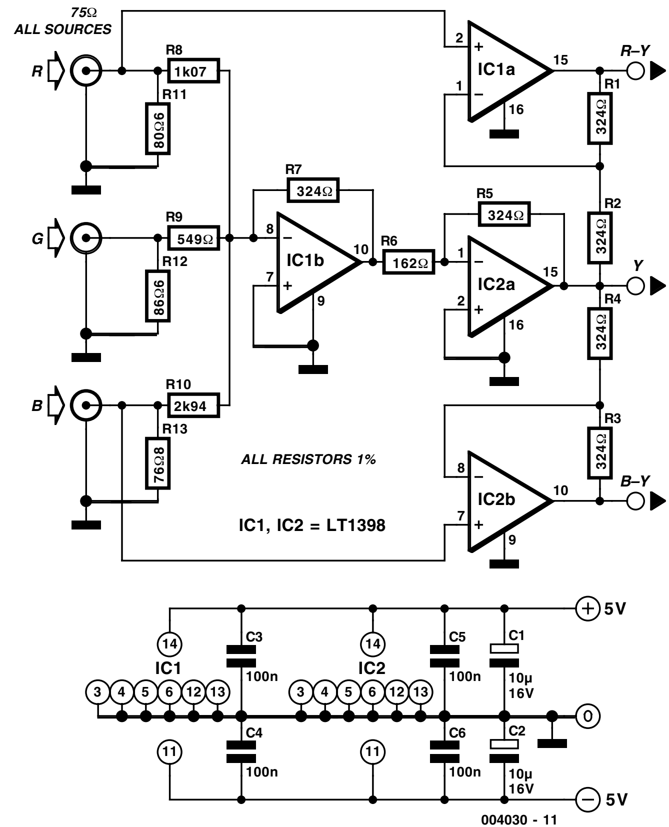
The B input is similar to the R input. Here, R13 when considered in parallel with R10 yields a 75-Ω input impedance. R10 also connects to the inverting input of amplifier IC1b, adding the B contribution to the Y signal as discussed above. Amplifier IC2b is configured to supply a non-inverting gain of 2 with the bottom of the gain resistor R4 tied to the Y output. The output of IC2b thus results in the colour-difference output B–Y.



The G input also arrives via 75-Ω coax and adds its contribution to the Y signal via resistor R9, which is tied to the inverting input of amplifier IC1b. Here, R12 and R9 provide the 75-Ω termination impedance. Using superposition, it is straightforward to determine the output of IC1b. Although inverted, it sums the R, G and B signals to the standard proportions of 0.3R, 0.59G and 0.11B that are used to create the Y signal. Amplifier IC2a then inverts and amplifies the signal by 2, resulting in the Y output.

The converter draws a current of about 30 mA from a symmetrical 5-volt supply.

(004030-1)

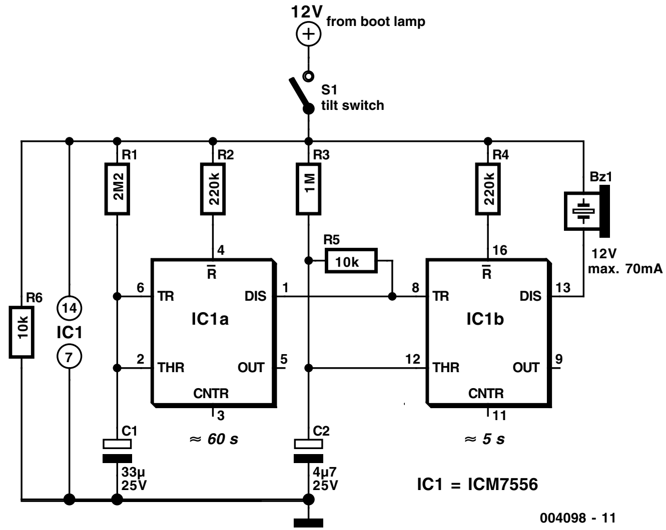


027 Car Boot Lamp Warning

B. C. Zschocke

On many cars, the boot light will not go out until the lid is properly closed. It is all too easy when unloading the car, to leave the lid ajar. If you are unlucky and the car remains unused for some time, the next time you try to start it, the lamp will have drained the battery and you will no doubt utter a few appropriate words.

The circuit described here will give a warning of just such a situation. A mercury tilt switch is mounted in the boot so that as the lid is closed, its contacts close before the lid is completely shut. The supply for the circuit comes from the switched 12 V to the boot lamp and through the mercury switch. When the lid is properly closed, the boot lamp will go out and the supply to the circuit will go to zero. If however the lid is left ajar, the lamp will be on and the mercury switch will close the circuit. After 5 seconds, the alarm will start to sound, and unless the lid is shut, it will continue for 1 minute to remind you to close the boot properly. The 1-minute operating period will ensure that the alarm does not sound continuously if you are, for example, transporting bulky items and the boot will not fully close.



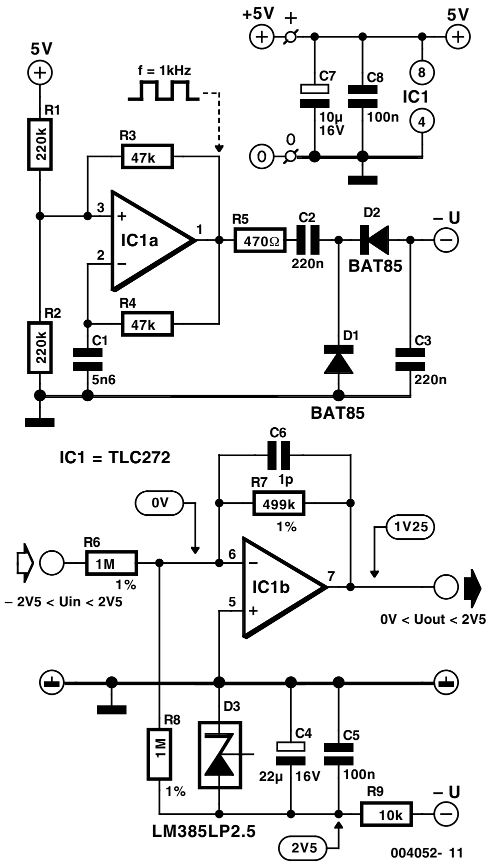
The circuit consists of a dual CMOS timer type 7556 (the bipolar 556 version is unsuitable for this application). When power is applied to the circuit (i.e. the boot lid is ajar) tantalum capacitors C1 and C2 will ensure that the outputs of the timers are high. After approximately 5 seconds, when the voltage across C2 rises to 2/3 of the supply voltage, timer IC1b will be triggered and its output will go low thereby causing the alarm to sound. Meanwhile the voltage across C1 is rising much more slowly and after approximately 1 minute, it will have reached 2/3 of the supply voltage. IC1a will now trigger and this will reset IC1b. The alarm will be turned off. IC1a will remain in this state until the boot lid is either closed or opened wider at which point C1 and C2 will be discharged through R6 and the circuit will be ready to start again. To calculate the period of the timers use the formula:

$t = 1.1RC$

Please note that the capacitor type used in the circuit should be tantalum or electrolytic with a solid electrolyte. The buzzer must be a type suitable for use at D.C. (i.e. one with a built in driver).

004098

028 Single-Supply Measurement Amplifier



H. Bonekamp

This circuit is eminently suitable for adding bipolar input capability to the unipolar A/D converter of a microprocessor circuit (in a data logger, for example), without having to use a dual power supply.

A bipolar input stage normally takes the form of two amplifier stages powered from a symmetrical supply. The first stage functions as an input buffer, with an input impedance of 1 M?. The second stage adds an offset to the buffered input signal, to make it suitable for the following unipolar ADC.

This circuit takes a different approach. Here, a buffer with an asymmetric power supply is used, and a negative offset current provides the conversion from bipolar to unipolar. The current is derived from a negative offset voltage, which is generated using IC1a. This opamp operates as a Schmitt trigger, due to the positive feedback network formed by R1, R2 and R3. The RC time constant of R4 and C1 in the negative feedback loop provides a timing element, which causes the Schmitt trigger to work as a square-wave oscillator. C2, C3, D1 and D2 convert the output voltage of the opamp into a negative voltage. To minimise interference, a relatively low frequency (1 kHz) has been chosen for the generator. R5 minimises the load current, and thus the amount of interference. Since the maximum output level of the TLC272 is only 3.5 V, diodes D1 and D2 must be Schottky diodes in order to obtain a sufficiently large output voltage.

Finally, R9, D3, C4 and C5 convert the output voltage into a stable –2.5-V reference voltage. The offset current for IC1b is drawn from this voltage via a 1-MΩ resistor (R8).

The brief specifications of this circuit are:

Input impedance :	1 MΩ
U_{in} :	–2,5 to +2,5 V
U_{out} :	0 to +2,5 V
Bandwith :	250 kHz
Current drain :	<2 mA

(004052-1)

029 Three Switches on One RCX Input

H. Steeman

One of the main limitations of the RCX module of the Lego Robotics Invention System is that only one switch can be connected to a given input. Even though each input has a measurement range of 0 to 1023, using an internal 8-bit A/D converter, the software works with only two states for detecting a switch: open and closed.

If you use the standard Lego software, you will have to accept this limitation. However, if you use a different programming language, such as Visual BASIC, it is not that difficult to find a way around it. In fact, the RCX block measures with a resolution of eight bits, even when detecting the state of a switch. If you connect several resistors with different values to the switch input, it is possible to read out several switches using a single input.

The schematic diagram shows that only three resistors are

needed for connecting three switches. With the proper choice of resistances, it is possible to detect eight different combinations of switch states. RAW will have the values listed in the table, according the states of the three switches. For a programmer with a bit of experience, it is no problem to cause certain actions to take place according to on these values.
The table lists the (theoretical) values that RAW will have if a single switch or a combination of switches is closed. The more precisely the resistances correspond to the given values, the closer the measured values of RAW will be to the theoretical values. However, with the illustrated arrangement, you can reliably use standard resistors with a tolerance of 5 percent.

(004101-1)

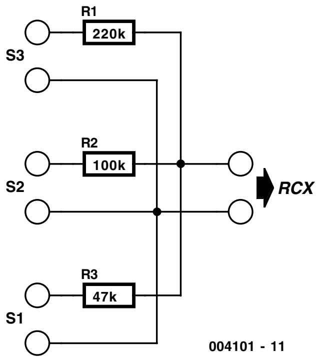


Table 1. Theoretical value of RAW for 8 switch states.

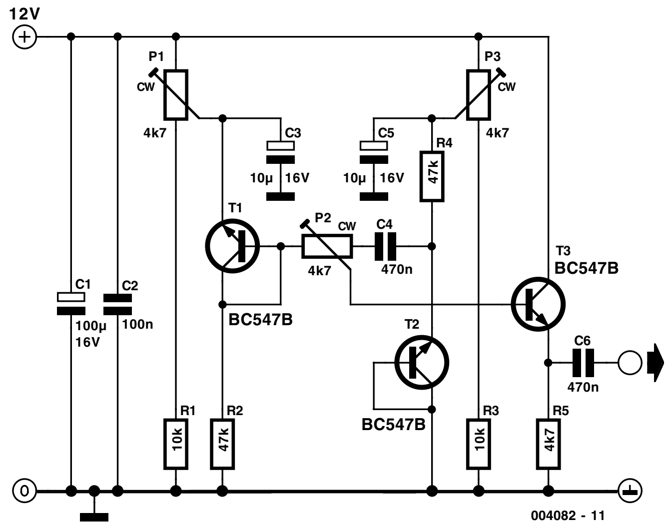
Switch 1	Switch 2	Switch 3	measured value of RAW
0	0	0	1 023 – 1 001
0	0	1	1 000 – 955
0	1	0	954 – 912
0	1	1	911 – 869
1	0	0	868 – 830
1	0	1	829 – 798
1	1	0	797 – 768
1	1	1	767 – 0

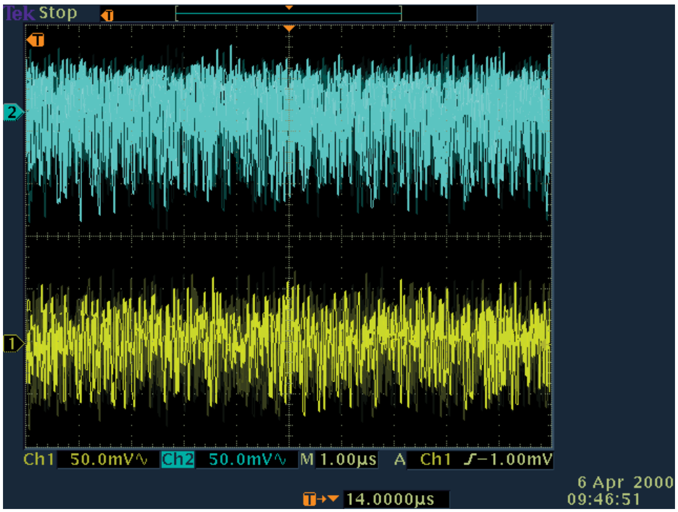
030

Symmetric Noise Source

H. Bonekamp

If a transistor junction operating in Zener breakdown is used as a noise source, the amplitude of the noise signal is asymmetric. This problem can be solved by using two transistors as two independent noise sources. One of these has a series resistor to earth, and the other has a series resistor to the supply line. Each of these noise sources produces an asymmetric noise voltage, with opposite asymmetry. If these two voltages are combined, the amplitude of the result will be symmetric.
In the circuit diagram, T1 and T2 are the noise sources. The series resistors are R2 (to earth) and R4 (to the positive supply line).
The supply voltage for the noise sources has been made adjustable, to allow the noise generation of the transistors to be optimised. This is because the amount of noise produced depends on the power supply voltage. P1 and R1 provide an adjustable supply voltage between 8 and 12 V for the noise stage around T1, while P3 and R3 perform the same function for T2. C3 and C5 smooth these voltages.
Since the amplitudes of the two noise sources will never be the same, it is necessary to take a weighted sum of the two signals. Consequently, P2 is included between the outputs of the noise sources as a sort of balance control. Since the DC levels of the two noise sources are not the same, C4 is also included in the balance network. The weighted sum of the two signals is present on the wiper of P2, superimposed on the DC signal of noise source T1. This DC level is also used for the DC bias of the buffer stage T3. The buffer isolates the noise sources from whatever circuit is connected to the output.
To adjust the circuit, connect an oscilloscope to the output. First, turn P2 all the way to the left. Now rotate P1 until a maximum noise signal is seen on the oscilloscope. Next, turn P2 all the way to the right, and then adjust P3 for the best noise signal. Finally, adjust P3 so that the noise signal looks symmetric.
The circuit provides an output voltage of approximately 150 mV_{pp}. The current consumption is 2 mA.
The oscillogram shows the asymmetric noise signal on channel 2, and the symmetric noise signal on channel 1.





(004082-1)

031 555 Signal Generator Kit

H. Baggen (text)
design © 2000 Velleman

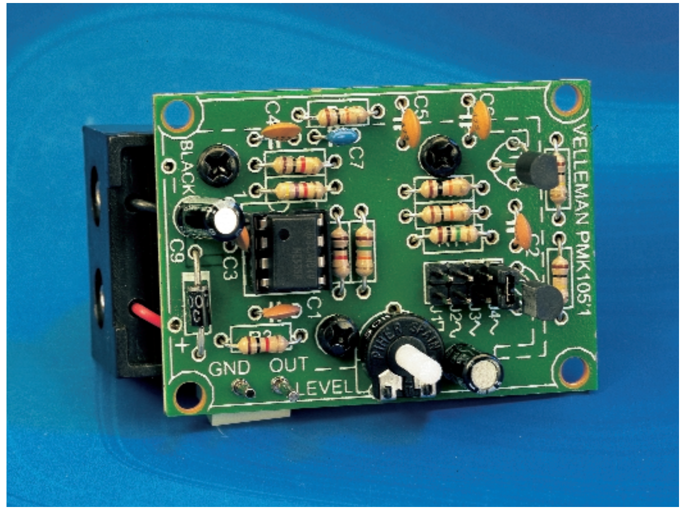
The 555 is a timer IC that has so many applications that it is not possible to name all of them. Here, a 555 forms the heart of a simple signal generator, which can produce square wave, rounded square wave, triangular and sinusoidal signals. A wire bridge or jumper can be used to connect one of these signals to an output buffer. The circuit produces a fixed output frequency of 1 kHz, and the output level can be adjusted between 0 and approximately 200 mV_{eff} using a potentiometer.

The 555 is wired as a squarewave generator, with the output frequency determined by the values of R1, R7 and C3. The symmetry of the squarewave depends on the ratio of R1 and R7. The output signal goes via the voltage divider R8 and R2 and jumper J1 to transistor T2, which is configured as an emitter follower. A low-impedance signal is available at the emitter of T2. This is finally brought to the output via the coupling capacitor C8 and potentiometer RV1.

The squarewave is integrated by the RC time constant of R4 and C5, and the resulting rounded squarewave is available at J2, from which it can be buffered by T2. A second RC stage (R5 and C6) makes the signal transitions even more gradual, so that something resembling a triangular waveform is obtained. This signal is available at J3. Finally, the tops of the triangular waveform are rounded off by the amplifier stage around T1, due to the presence of C2. This signal is available at J4.

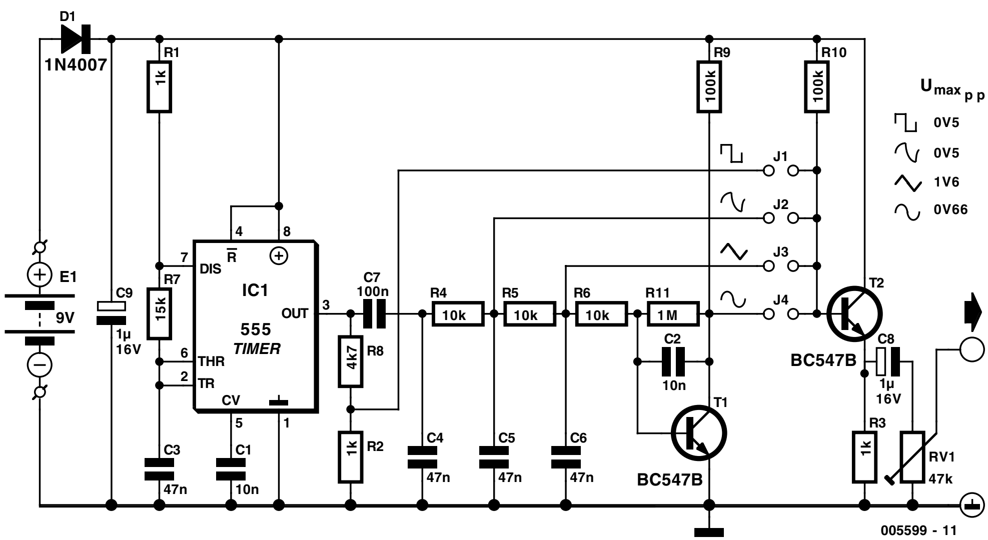
The current consumption of the circuit is approximately 10 mA, so it can easily be powered by a 9-V battery. Diode D1 provides reverse-polarity protection, so that the 555 and the transistors will not go up in smoke if you accidentally connect the battery the wrong way around.

(005099-1)



The order code of this Velleman kit is MK105.

For further information, visit http://www.velleman.be

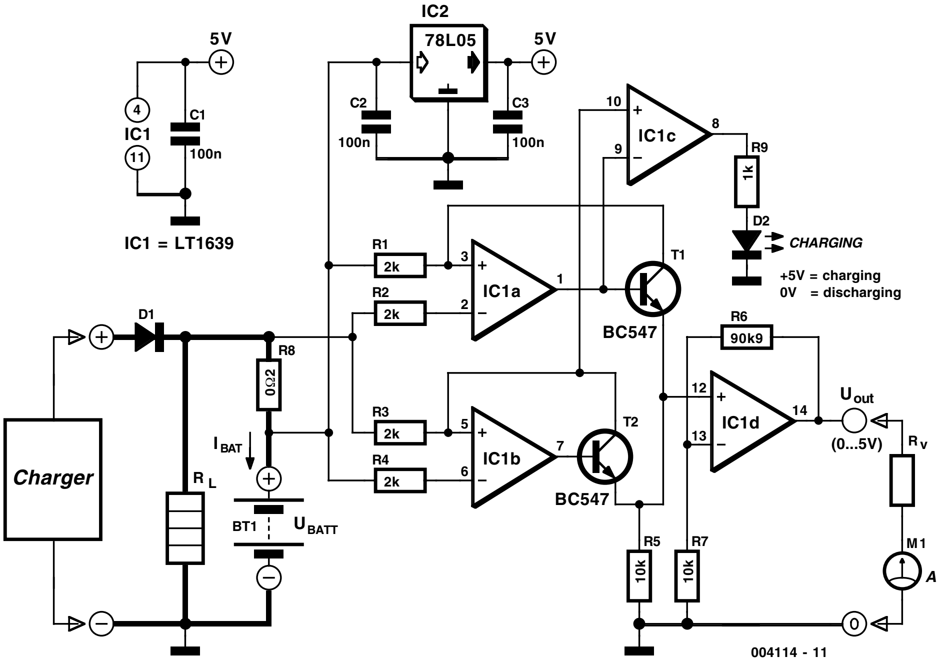


032 Battery Charger Display

By G. Kleine

The Over-the-Top type of operational amplifier is ideal for use as a current sense for battery charger applications. The design described here can be used with chargers for rechargeable batteries (Lead/acid or NiCd etc). The 5 V operating supply for the circuit is derived from the battery on charge. The circuit uses a sense resistor R8 to determine the value of current flowing in or out of the battery. An LED output shows whether the battery is charging or discharging and an analogue output displays the battery charge or discharge current. The circuit can also be altered to shown different ranges of charging current to cater for higher capacity cells.

IC1a and IC1b together with T1 and T2 form two current sources, which produce a voltage across R5. The voltage across R5 is proportional to the current through resistors R8 and R1 (for IC1a) or R8-R3 (for IC1b). The current source formed by IC1a and T1 is active when the batteries are discharging and IC1b and T2 is active when the batteries are being charged. In each case the inactive opamp will have 0 V at its output and the corresponding transistor will be switched off. IC1d amplifies the voltage across R5, which is proportional to the sense current. The component values given in the diagram produce an amplification factor or 10. A sense current of 0.1 A will produce an output voltage of +1 V. The supply voltage to the circuit is +5 V so this will be the maximum value that the output can achieve. This corresponds to a maximum charge/discharge current of 0.5 A To display currents from 0 to 5.0 A, resistor R7 can be omitted to give IC1d a voltage gain of 1. Higher currents can be displayed by using a lower value of sense resistor R8. A DVM or analogue meter can be used at Vout to give a display of the charge/discharge current.



The constant current sources can only function correctly when the supply to the voltage regulator circuit ($U_{Batt.}$ e.g. 6 V or 12 V) is greater than the operating voltage of the opamps (+5 V). The supply voltage to the LT1639 can be in the range of +3 V and +44 V and voltages up to 40 V over the supply voltage are acceptable at the inputs to the opamp.

IC1c controls the charging/discharging LED output. The inputs to this opamp are connected to the outputs of the current source opamps and its output goes high when the battery is being charged and low when it is discharging.

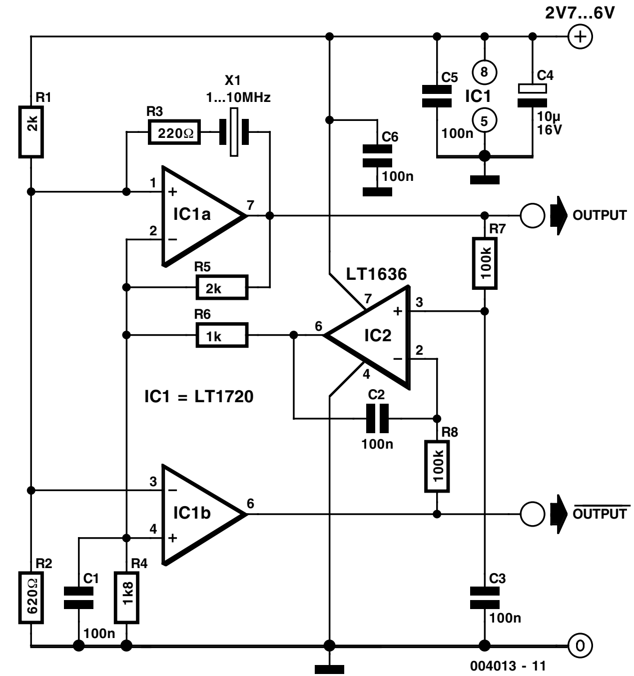
(004114e)

Internet Address: www.linear-tech.com

033 Compator-based Crystal Oscillator

Based on A Linear Technology Application

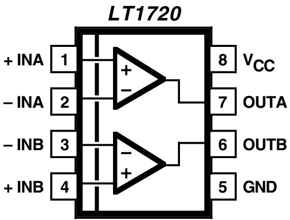
Although a simple crystal oscillator may be built from one comparator of an LT1720/LT1721, this will suffer from a number of inherent shortcomings and design problems. Although the LT1720/LT1721 will give the correct logic output when one input is outside the common mode range, additional delays may occur when it is so operated, opening the possibility of spurious operating modes. Therefore, the DC bias voltages at the inputs have to be set near the centre of the LT1720/LT1721's common mode range and a resistor is required to attenuate the feedback to the non-inverting input. Unfortunately, although the output duty cycle for this circuit is roughly 50%, it is affected by resistor tolerances and, to a lesser extent, by comparator offsets and timings. If a 50% duty cycle is required, the circuit shown here creates a pair of complementary outputs with a forced 50% duty cycle. Crystals are narrow-band elements, so the feedback to the non-inverting input is a filtered analogue version of the squarewave output. The crystal's path provides resonant positive feedback and stable oscillation occurs. Changing the non-inverting reference level can vary the duty cycle. The 2k-680Ω resistor pair sets a bias point at the comparator + (Comparator IC1a) and – (Comparator IC1b) input. At the complementary input of each comparator, the 2k-1.8k-0.1µF path sets up an appropriate DC average level based on the output. IC1b creates a complementary

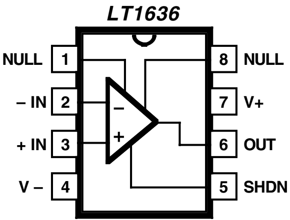


output to IC1a by comparing the same two nodes with the opposite input.

IC2 compares band-limited versions of the outputs and biases IC1a's negative input. IC1a's only degree of freedom to respond is variation of pulse width; hence the outputs are forced to 50% duty cycle. The circuit operates from 2.7 V to 6 V. When 'scoping the oscillator output signal, a slight dependence on comparator loading, will be noted, so equal and resistive loading should be used in critical applications. The circuit works well because of the two matched delays and rail-to-rail outputs of the LT1720.

(004013-1)





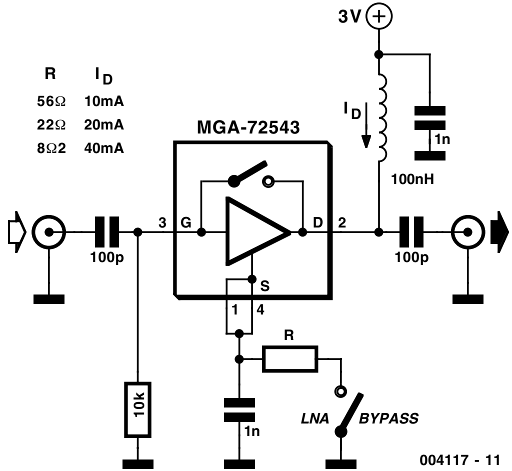
034 MMIC Antenna Amplifier with Bypass Switch

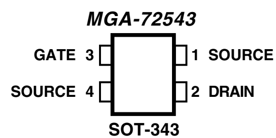
G. Kleine

Specifications for receiver frontends, converters and detectors are generally very demanding. On the one hand they need to have high sensitivity and low noise to amplify small input signals while on the other, they must be able to handle large input signals without distortion or overdriving subsequent stages. A solution to this problem is provided by the MGA72543 GaAs-RFIC. This IC contains a low noise RF preamp that, at high signal levels, can be bypassed altogether with internal switches.

This IC provides an amplification factor of 14 dB over the frequency range of 100 MHz to 6 GHz with a noise figure of less than 2 dB. With the amplifier bypassed the insertion loss is 2.5 dB. The operating voltage is between +2.7 V and +4.2 V and the input and output are matched to 50 Ω. The value of supply current to the device controls the input signal handling characteristics. At operating currents of 40 to 50 mA, the output power at 1 dB gain compression (P_{1dB}) is +16 dBm, while at 10 mA operating current this figure is +8 dBm.

Biasing the MGA72543 is similar to using a discrete GaAs FET, the DC levels at the input (gate) and output (drain) are calculated to produce the desired operating current (10 - 50 mA). A gate bias method can be used but needs a negative voltage, this is only convenient if a negative voltage is available elsewhere in the circuit. The more usual method and the one described here is to use a resistor in the source lead to set the operating point. Negative feedback in this configuration controls the drain current. The DC path to the gate is relatively high impedance and an RF choke is used to isolate the RF signal from the DC supply. The 1 nF capacitor connected in parallel witho bias resistor R effectively acts as a short circuit at RF, ensuring that R is seen as a short circuit. The switch is used to put the device into bypass mode by open circuiting the source resistor. The operating current then drops to near zero.





(004117)

Internet Address: www.agilent.com

035 Precision Voltage to Current Converter

H. Bonekamp

In measurement and control systems a common method of sending control or measurement data over long distances is by means of a ‘current loop’ using two wires. This system converts the measurement values into corresponding current levels on the two wire loop.

In such a system a precision converter is required to translate the measured values (typically in the form of a voltage level) into a current on the loop i.e. a current source controlled by a voltage.

IC1b together with T1, R5–R8 and P2 form an instrumentation amplifier, this amplifier controls the voltage across resistor R9 to ensure that it exactly follows the input voltage. Because R9 has a fixed resistance, the current through it will be exactly proportional to the voltage across it (according to Mr Ohm). The current through R9 is therefore exactly proportional to the input voltage. IC1b therefore forms the voltage controlled current source.

P2 is used to optimise the common mode rejection of this instrumentation amplifier by compensating for unequal resistances and imbalance caused by R9.

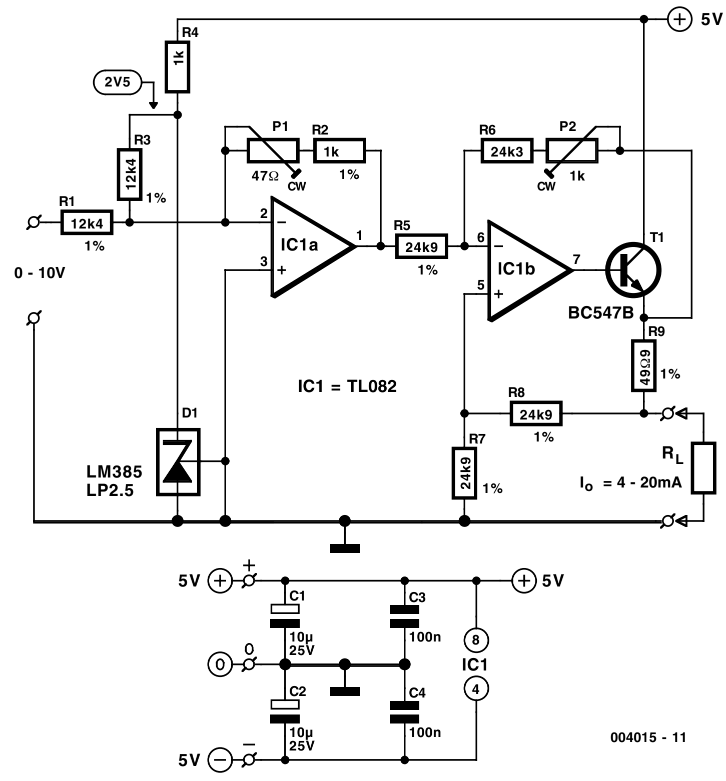
The input amplifier formed by IC1a ensures that the voltage into the instrumentation amplifier is limited and within its control range. D1, R3 and R4 add an offset to this input voltage to ensure that with 0 V input the output will be, not 0 mA but 4 mA. The maximum output current is 20 mA. This circuit is therefore a '20 mA current loop' interface, often used in industrial control applications.

The amplification factor of the complete circuit is performed by IC1a and can be adjusted by preset P1. To calibrate the circuit:

1. Common mode rejection optimisation using P2:
Connect a DMM in series with a resistor of 47 Ω. Connect this to the output (in place of RL), measure the output current. Now short circuit the 47Ω resistor and measure the output current again. If there is no difference in the output current then P2 is correctly adjusted. Otherwise, adjust P2.

2. Voltage range adjustment:
With the output short-circuited, adjust P1 until 200 mV is measured across R9.

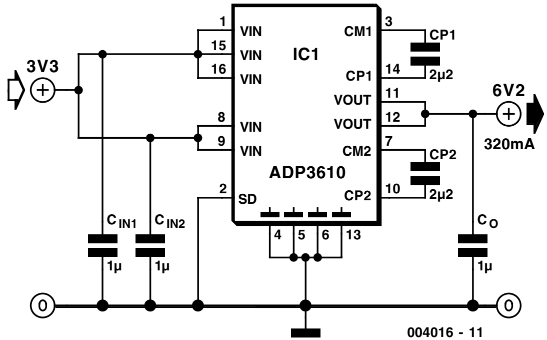
(004015-1)

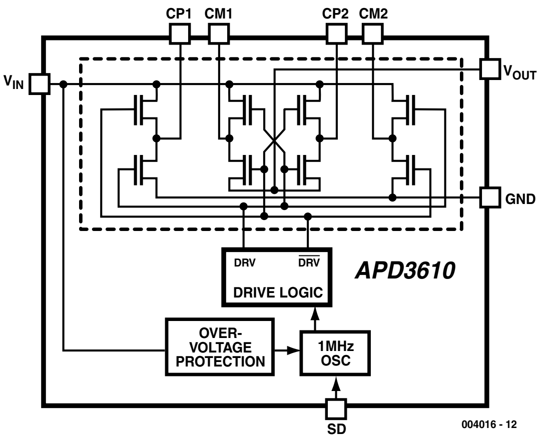


036 Switching Voltage Regulator

H. Baggen
(from an Analogue Devices application note)

The Analog Devices ACP3610 is a voltage doubler that works with a switched-capacitor converter, using the push-pull principle. The switching frequency at the output is approximately 550 kHz. The term ‘push-pull’ refers to the two charge pumps, which work in parallel but in opposite directions in order to deliver the output voltage and current. Whenever one capacitor is supplying current to the output, the other one is being charged. This technique minimises voltages losses and output ripple. The converter works with input voltages between 3 and 3.6 V. It provides an output voltage of around 6 V at a maximum current of 320 mA, if 2.2-μF switched capacitors with low ESR (equivalent series resistance) are used. A shutdown input is provided to allow the voltage doubler to be enabled or disabled by a logic-level signal. The IC is enclosed in a special package, which can dissipate up to 980 mW at room temperature.





The schematic diagram shows a typical application for the ADP3610. Here it works as a non-regulated voltage doubler. In theory, a voltage doubler can provide exactly twice the input voltage at its output, but in practice the combination of internal losses in the electronic switches and the internal resistances of the capacitors always causes the output voltage to be somewhat lower. The output voltage drops from a no-load value of 6 V to 5.4 V with a 320-mA load, with a nearly linear characteristic.

A small capacitor is connected across the two supply pins at the input of the IC. It suppresses noise, brief voltage fluctuations, and current peaks when the ADP3610 switches. This capacitor (C_{IN}) must have a low internal resistance (ESR). A larger capacitance value is necessary if long supply leads to the ADP3610 are present.

The 1-μF output capacitor (C_O) is alternately charged by the two capacitors of the charge pump, CP1 and CP2. The internal resistance is an important factor here as well. It largely determines the amount that the voltage drops under load, and the amount of ripple in the output voltage. Ceramic or tantalum capacitors are recommended. The ESR can also be reduced by connecting several smaller-value capacitors in parallel. With small loads, the value of C_O may be reduced.

(004016-1)

037 0 – 2 μF Variable Capacitor

H. Bonekamp

This circuit makes it possible to simulate a variable capacitance, with the value set by a potentiometer. The amplification stage around IC1b has an adjustable gain, which can be set between +1 and –1 using P1. IC1a is an impedance buffer, which makes the ultimate converted capacitance as close as possible to an ideal capacitance.

This circuit is described by the following formula:

$$U1 = (2 \cdot \alpha - 1) \cdot U_{in}$$

in which α represents the setting of P1 (with a range of 0 to 1).

The input current I_{in} is given by:

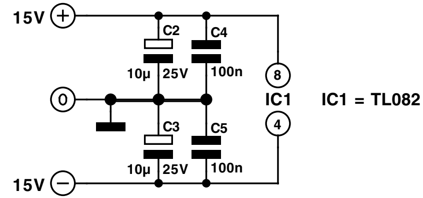
$$I_{in} = \frac{U_{in} - U_1}{\frac{1}{j \cdot \omega \cdot C1}} = \left(U_{in} - (2 \cdot \alpha - 1) \cdot U_{in}\right) \cdot j \cdot \omega \cdot C1 = j \cdot \omega \cdot (1 - \alpha) \cdot 2 \cdot C1 \cdot U_{in}$$

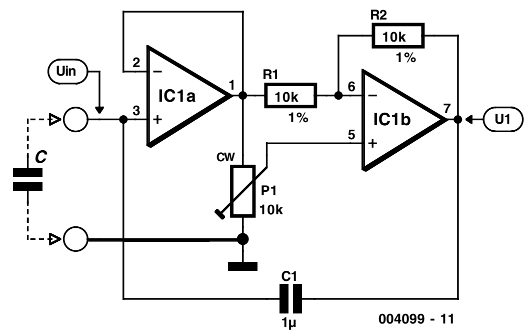
The input impedance is thus:

$$Z_{in} = \frac{U_{in}}{I_{in}} = \frac{1}{j \cdot \omega \cdot (1 - \alpha) \cdot 2 \cdot C1}$$

From the last formula, we can see that the input capacitance is:

$$C_{in} = (1 - \alpha) \cdot 2 \cdot C1$$





The maximum allowable voltage across C_{in} is 10 V_{pk}. The current consumption of the circuit is around 5 mA.

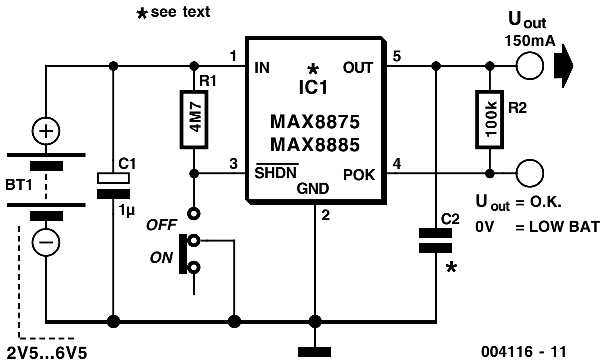
(004099-1)

038

150 mA LDO Regulator with Power OK O/P

By G. Kleine

To get maximum battery life from portable equipment, Low Drop Out (LDO) regulators are often used. These devices can supply a well regulated output voltage while having a drop-out voltage (Vin – Vout) of only 100 mV. Battery powered equipment always has a problem when the battery voltage falls and the regulator can no longer maintain its designed output voltage. Operation of the circuit becomes unreliable and data can be lost. A solution to this problem is provided by the new LDO regulator from Maxim, the MAX 8875/8885. This chip has a power-OK output; it indicates when the output voltage has fallen to below (typically) 5% of the regulated voltage. This is useful, for example in processor applications, so that data can be safely stored in a non-volatile environment before the voltage drops further and the system dies completely.



The MAX 8875/8885 is a fixed voltage regulator with outputs of +2.5 V, +2.7 V, +3.0 V, +3.3 V or +5.0 V. Each have a guaranteed output current of 150 mA and a typical drop out voltage of 110 mV at a supply current of 100 mA. The input voltage range is from +2.5 V to +6.5 V. The chip has built-in protection against output short circuit and over temperature. Also useful for portable equipment is the accidental battery reversal (up to 7 V) protection. A shutdown pin is also provided and grounding this will turn off the regulator and its supply current will fall to less than 1 μA.

The stability of the regulator is largely dependent on the type of output capacitor (C2) used. The 8875 is designed to be used with a ceramic output capacitor (optimally 1μF), while the 8885 is suitable for more economical capacitors with a higher ESR (Effective Series Resistance) such as tantalum. The recommended value of output capacitor here is from 1 – 4.7 μF.

Device	V_{out}	Identification MAX8875	Identification MAX8885
MAX 88x5EUK25	2.5 V	ADKZ	ADLE
MAX 88x5EUK27	2.7 V	ADLA	ADLF
MAX 88x5EUK30	3.0 V	ADLB	ADLG
MAX 88x5EUK33	3.3 V	ADLC	ADLH
MAX 88x5EUK50	5.0 V	ADLD	ADLJ

(004116)

Internet Address: www.maxim-ic.com

039 Optimised Semiconductor Noise Source

H. Bonekamp

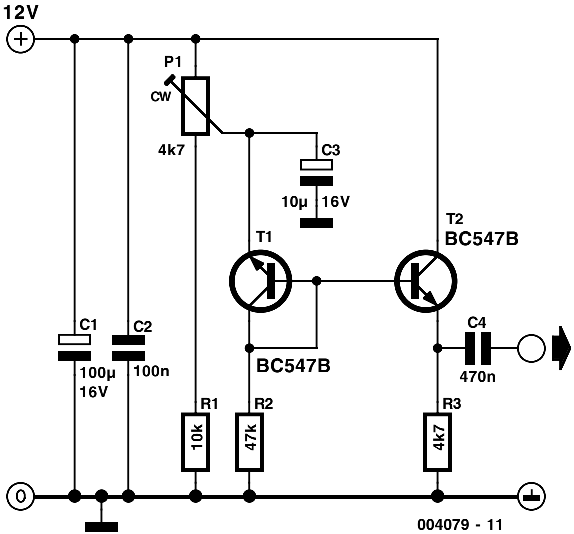
We have already published designs that use a transistor junction operating in Zener breakdown as a noise source. Anyone who has experimented with a reverse-biased transistor knows that the amplitude of the noise voltage generated in this manner is strongly dependent on the supply voltage. The variation between individual transistors is also rather large. An obvious solution is to use an adjustable supply voltage for the noise generator stage.

A BC547B starts to break down at around 8 V. Using P1 and R1, you can adjust the voltage across T1 and R2 between 8 and 12 V. C3 decouples the reduced supply voltage. An impedance buffer in the form of T2 and R3 is added to the circuit, to prevent the connected load from affecting the noise source. This buffer is powered directly from the 12-V supply.

To adjust this circuit, connect the output to an oscilloscope. Then adjust P1 to obtain the highest signal amplitude, combined with the best 'shape' of the noise signal.

The output voltage is approximately 300 mV_{pp}, and the current consumption is around 2 mA.

(004079-1)



040 External RF Power Control for 2m/70cm Handhelds

N.S. Harisankar, VU3NSH

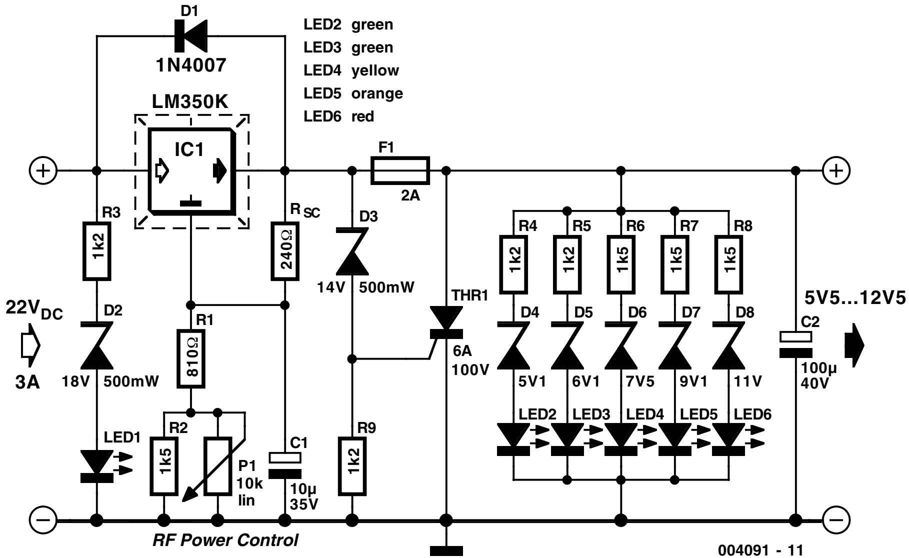
Most modern handheld transceivers for 2 m and 70 cms ham radio incorporate MOSFET RF power modules and are capable of operating at supply voltages between 4.5 V and 13.8 V. When using four NiCd batteries (4.8 V), RF power on high power mode (H) is typically 1.8 watts. When using a 9.6 V battery pack, the H power level is typically 5 watts, while 500 mW or 50 mW is produced in L (low power) and EL (economical low) mode, respectively.

Sometimes low RF power is not sufficient, while high power seems wasteful. The circuit shown here enables transceivers like the Yaesu F11/41R, Kenwood TH22AT, Icom ICT22A and others to supply any RF power level between 2 and 5 watts while also providing a low power mode.

The LM350K voltage regulator used in this circuit is capable of supplying 1.25 V to 33 V at up to 3 A. The output voltage is defined by R_{SC}, R1, R2 and P1. The latter acts as the RF power control.

A 'crowbar' thyristor is used on the supply output line to protect the expensive radio from possibly disastrous supply voltage levels. In the unlikely case of the output voltage rising above 14 V (as defined by zener diode D3), the thyristor (a 6A/100V type) is triggered and it will faithfully destroy fuse F1. The action is similar to throwing a crowbar across the output terminals!

Each of the LEDs at the output will light when the output



voltage exceeds the threshold set up by the associated zener diode. In this way you get a visible and colourful RF power indicator. The actual span of the output voltage will be 5.5 to 12.5 V. The voltage regulator should be fitted with a suitably sized heatsink.

(004091-1)

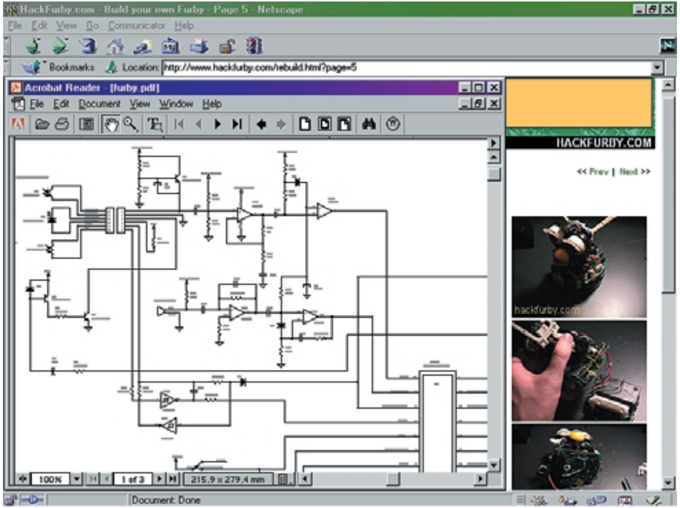
041 Furby Revealed

H. Baggen

The unbridled popularity of the toy animal ‘Furby’ is largely due to the clever design of the electronics in this furball, which acts like it actually communicates with children. However, in time their parents may grow tired of listening to the constant babble from this little beast.

Anyone who would like to experiment with a Furby can find a lot of useful information regarding its construction and operation on the web site *www.hackfurby.com*. You can find all sorts of information here. Undocumented features are reported under ‘Secrets’. An extensive discussion under ‘Rebuild’, complete with photos, provides step-by-step instructions for taking apart a Furby, following which you can use its skeleton and electronics for making your own doll. For electronics types, however, the most interesting item is naturally the schematic diagram of the electronics module. Someone has taken the trouble to completely puzzle this out, and you can download the result as a PDF document. In addition, you can find a lot of information and news about Furbys at this site. It can be recommended for both Furby-lovers and Furby-haters!

(004109-1)



042 Simple Sump Pump Controller

S. van Rooij (text) and
H. Gulikers (design)

After one too many flooded basements, the author of this design was fed up. The mechanical float switch for the sump pump, which pumps the water from the washing machine up to the sewer level, was increasingly often jammed by caked-on detergent. It seemed like a good idea to replace this switch by an electronic version with two sensors. In order to avoid unnecessary effort, an existing *Elektor Electronics* circuit was taken as a model, namely the 'simple moisture detector' in the December 1998 issue. This was equipped with two sensors: one for the lower level of the water in the sump in which the pump is located, and one for the upper level.

In the original circuit, input 1 of IC1a is used for a manual reset switch. In the new version, this switch is replaced by a second sensor (S1), which detects the *low* water level for the sump pump. Sensor S2 is naturally used to detect the high level.

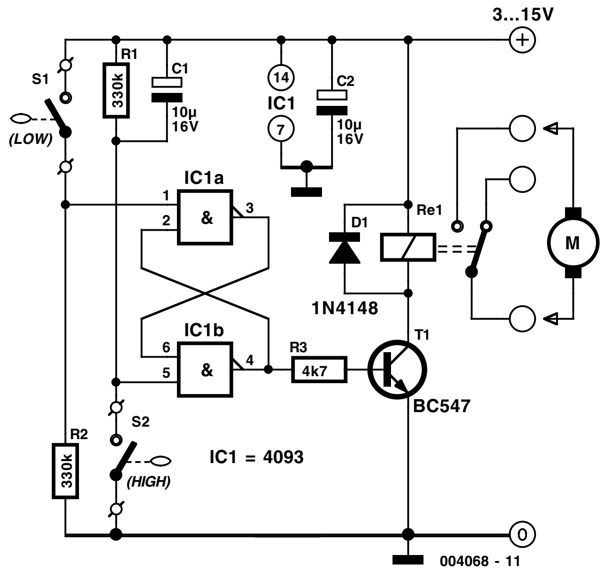
It all works very simply. If the water rises above the level of the high-level sensor S2, input 5 goes to '0', which causes output 4 to go to '1' and the relay to be energised. This in turn activates the pump, which will cause the water level to drop. If the water level drops to somewhere between S1 and S2, nothing changes and the pump continues to run. The flip-flop holds the relay engaged. However, as soon as the water drops below the low-level sensor S1, so that S1 no longer conducts, input 1 will no longer pulled 'high' and will drop to zero. The flip-flop will be reset, and the relay will drop out.

The relay contacts are connected in series with the main power cable of the sump pump. The original (defective) float switch is set to a permanently closed position, so that the pump always runs whenever voltage is applied to the supply cable.

Each of the sensors is made from two pieces of mains distribution wire mounted approximately 1 cm apart, with the insulation stripped from the lower ends.

Based on experiments, the series resistor for the two sensors (R1) was reduced from 1 MΩ to 330 kΩ. This is because the sensors do not have to detect moisture, as in the original circuit, but come fully in contact with the water.

The prototype is powered by a standard 6-V mains adapter. The relay is a normal 6-V DC type with double-pole contacts rated at 250 V.



The circuit is so simple that it can be very quickly put together on a piece of prototyping board. Naturally, you must be careful with the connections between the relay and the 240-V mains cable for the pump. Thoroughly isolate the connections, and make sure that there is adequate separation between the individual AC connectors and between the connectors and the rest of the circuit.

This controller has worked several months in the author's house with no problems. Flooded basements appear to be definitely a thing of the past.

(004068-1)

043 Instrumentation Amplifier with Improved CMRR

H. Bonekamp

This circuit is a relatively simple instrumentation amplifier that can be powered from a single 5-V supply. The output voltage is given by the formula:

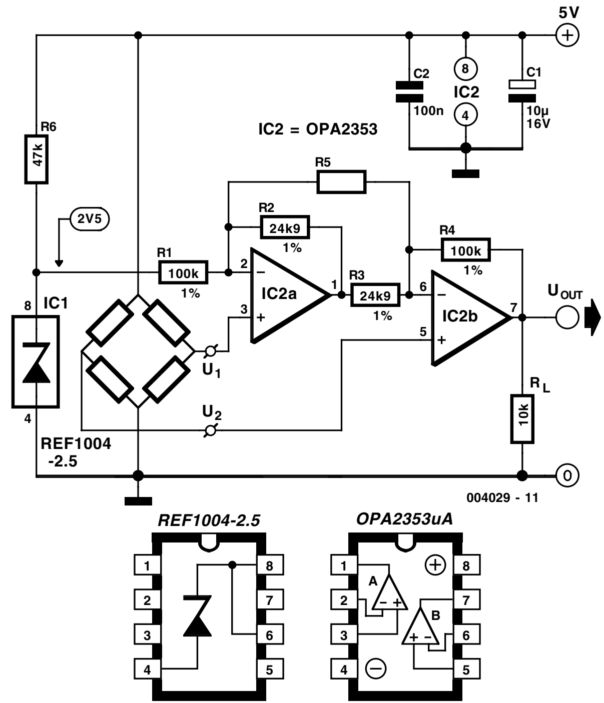
$$U_O = 2.5V + \frac{R4}{R5} \cdot \left(1 + \frac{R2}{R3}\right) \cdot (U1 - U2) + \left(1 + \frac{R4}{R3}\right) \cdot U1 - \frac{R4}{R3} \cdot \left(1 + \frac{R2}{R1}\right) \cdot U2$$

For the optimum CMRR, the ratio of R1 to R2 should be the same as that of R3 to R4. If we let R1 = R4 and R2 = R3, the formula becomes:

$$U_O = 2.5V + \left(1 + \frac{R1}{R2} + 2 \cdot \frac{R1}{R5}\right) \cdot (U1 - U2)$$

Thanks to the 2.5-V reference voltage provided by IC1, this circuit can work with a single supply voltage. With regard to common-mode voltage range, it is important that R2/R1 < 1, since otherwise IC2a will be driven to its limit too quickly by the common-mode voltage! The current consumption of this circuit is approximately 10 mA.

(004029-1)



044 Pushbutton-Controlled Trimpot ICs

G. Kleine

The DS1809 and DS1869 ICs are electronically controlled trimpots that can be adjusted in 64 linear steps using up and down pushbuttons. Since they hold their settings via built-in EEPROMs when the power is switched off, they can be used in place of mechanical trimpots. Values of 10 kΩ, 50 kΩ and 100 kΩ are available. The supply voltage may lie between +3 V and +5 V or +8 V, respectively. The voltages on the three connections RH, RL and RM may not be more than 0.5 V higher than the supply voltage, nor more than 0.5 V negative.

The up and down pushbuttons are debounced inside the ICs, in order to avoid jumping more than one step at a time. Internal 100-kΩ pull-up resistors simplify the external circuitry for the UC and DC pins. Both button pulses must be applied for at least 1 ms. With the DS1809, the delay between successive button pulses must be at least 0.5 s, while with the DS1869 the delay must be at least 1 s. If a pushbutton is held pressed for longer than the delay time, the IC goes into autorepeat mode, in which the wiper setting moves up or down at the rate of 10 steps per second. This means that it takes at most 7 s to go from one end of the range to the other.

Figure 1 shows the DS1809 in autostore operation. The Schottky diode (BAR42, BAR43, BAT45 or similar) and the 10-µF electrolytic capacitor provide sufficient reserve energy when the power is switched off to allow the current wiper setting to be written to the EEPROM. The manufacturer guarantees 50,000 error-free write cycles, and the EEPROM is only written if the setting is not the same as what is already stored in the EEPROM.

If the STR pin of the DS1809 is held to earth while the supply voltage is switched on, the IC enters the command-initiated storage mode. In this mode, a High pulse at the STR input with a duration of at least 1 µs triggers the storage process. In this case, the wiper setting is not stored when the power is switched off.

The DS1869 automatically stores the wiper setting each time it is changed. Instead of a Store input (STR), it has a digital input (D) on the same pin. This input can be driven by a microcontroller or similar device, and it operates immediately, without any debouncing delay. If it is not used, it may simply be left open.

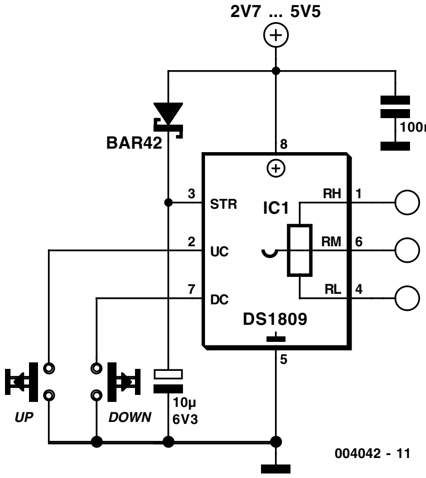
In addition to the two-pushbutton operating mode already described and illustrated in Figure 2a, the DS1869 can also be operated using only one pushbutton, as illustrated in **Figure 2b**. In this case, the 'DC' pin must always be tied to V+, particularly during the switch-on interval. The direction of adjustment reverses after the pushbutton has not been actuated for longer than one second. A series of button pushes separated by less than one second causes the wiper to steadily move in the same direction. If it reaches the end of the adjustment range, it automatically reverses direction.

In this configuration, the autorepeat mode described above is also activated if the pushbutton is held pressed for longer than 1 s, so that the wiper moves automatically at the rate of 10 steps per second. With the DS1869, in contrast to the DS1809, the wiper automatically reverses direction and continues moving in the opposite direction if it reaches the end of the adjustment range.

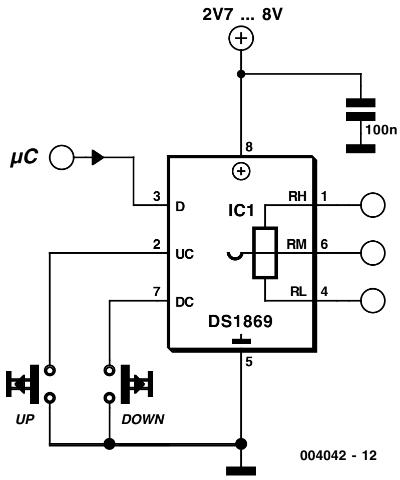
You can download data sheets for these 'Dallastat' ICs from the Internet site *www.dalsemi.com*.

(004042-1)

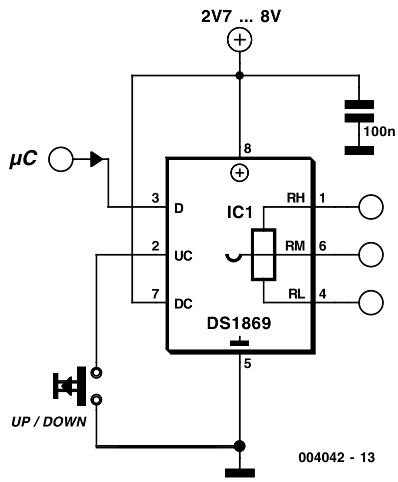
1



2a



2b



045 Stepper Motor Microstep Driver

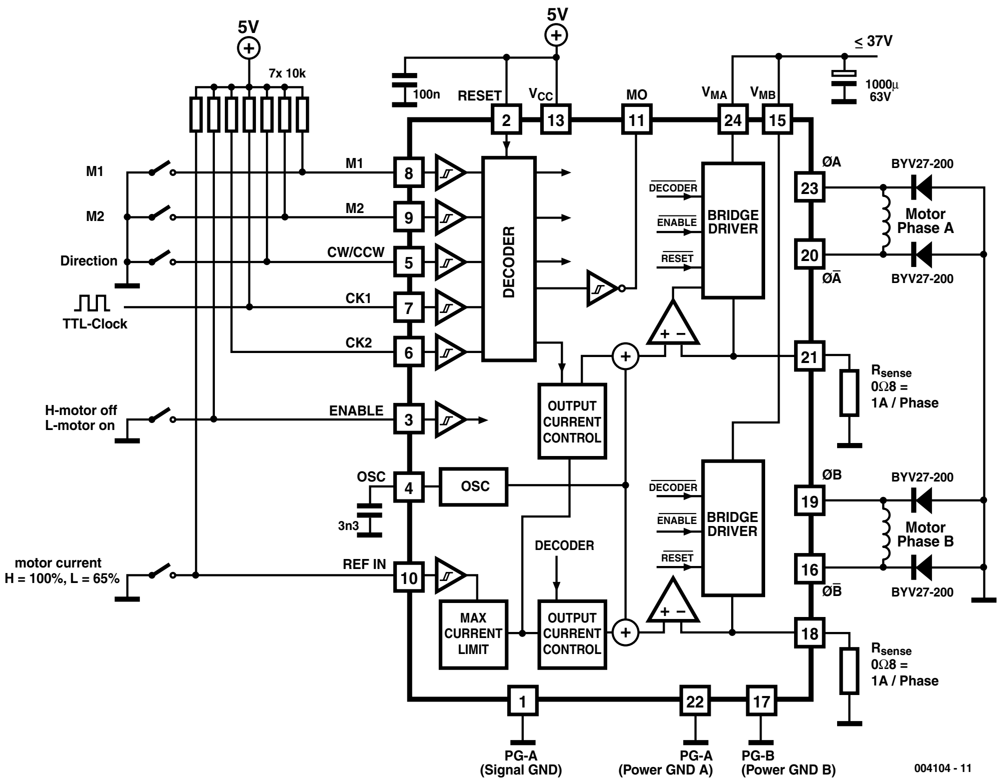


Table 1. Step settings

Input		Mode
M1	M2	
L	L	1/1 step
H	L	1/2 step
L	H	1/4 step
H	H	1/8 step

K. Walraven (text)
from a Nanotec application note

The IMT 901 IC from the German firm Nanotec makes it easy to build a stepper motor controller. You can find a data sheet with complete technical specifications at the Nanotec site (*http://nanotec.de*).

The circuit diagram can be kept very simple. In the application shown here, the IC is controlled manually, but it can naturally also be controlled by a microcontroller or PC. Signals M1 and M2 select the number of current steps for each mechanical step (see **Table 1**). Bear in mind that the more current steps there are, the more clock pulses are needed to move the motor by one step. In other words, with more current steps the motor will run more slowly at the same clock rate.

The motor current is set by the resistor R_{sense}. The desired current level naturally depends on the motor used. The value of R is given by

$$R = 0.8 \div (\text{desired current in ampère})$$

Use a 5-W resistor. The IC will limit the motor current to this value using pulse-width modulation. The IC runs from a 5-V supply. The two supply voltages on V_{MA} and V_{MB} must be equal to the specified voltage for the motor used (in practice, at most 37 V).

(004104-1)

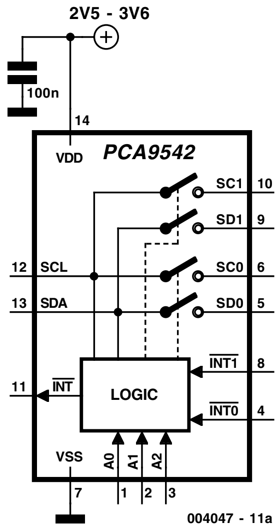
IMT 901 specifications

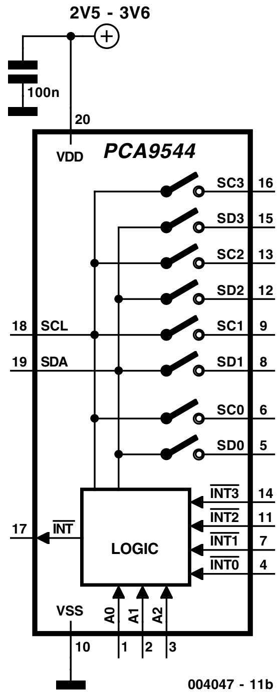
Supply voltage V_{CC}	5.5 V
Supply voltage V_M	40 V
Output current I_{out}	1.5 A (average) 2.5 A (peak)
Power dissipation P_d	4 W (no heat sink) 40 W (with heat sink) T_c = 85 °C
Operating temperature	–40 to +85 °C
Storage temperature	–55 to +150 °C

INPUT					MODE
CK1	CK2	CW/CCW	Enable	Reset	
↑	H	L	L	H	CW
⎍	L	L	L	H	INHIBIT
H	↑	L	L	H	CCW
L	⎍	L	L	L	INHIBIT
↑	H	H	L	H	CCW
⎍	L	H	L	H	INHIBIT
H	↑	H	L	H	CW
L	⎍	H	L	H	INHIBIT
X	X	X	H	H	Z
X	X	X	X	L	Z

004104- 12

046 I²C Multiplexer





G. Kleine

Anyone who wants to build large I²C bus structures sometimes has the problem that the four or eight addresses of a particular type of IC are not sufficient, or that the capacitive loads on the SDA and SCL lines are very high, so that a low clock frequency must be chosen for the bus. The multiplexer ICs described here allow a large I²C bus system to be partitioned into several zones. These ICs connect the I²C bus controller to the zone that is to be addressed at any given time.

It can also happen that parts of the I²C bus must be operated at 3 V, while other parts must be operated at 5 V. The multiplexers are also helpful here, since their inputs and outputs can both be loaded with 5-V signals, even though the multiplexer operates with only +3 V.

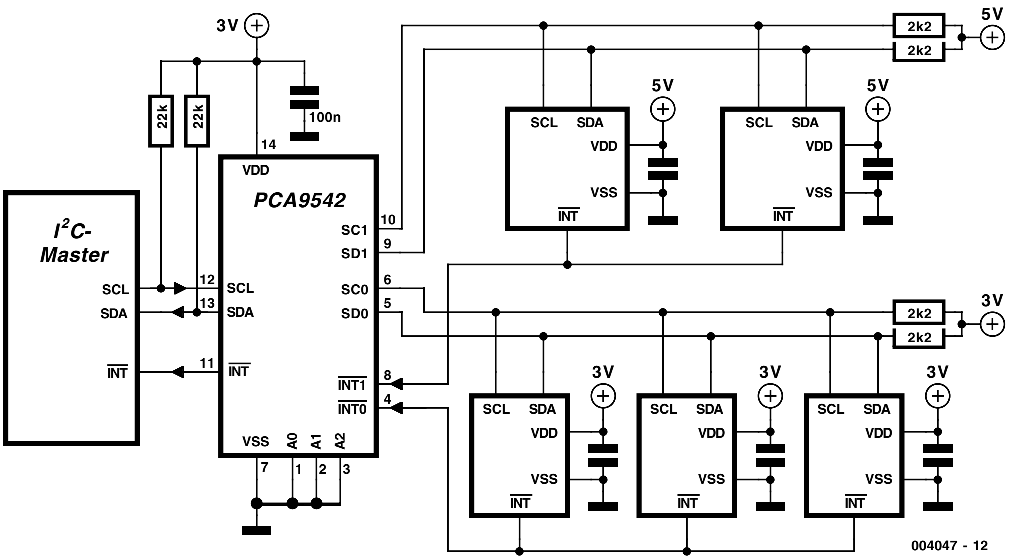
The PCA9542 and PCA9544 ICs (**Figure 1**) contain 1-to-2 and 1-to-4 multiplexers, respectively, for both of the bidirectional SDA and SCL lines. These ICs can be operated from supply voltages between 2.5 and 3.6 V (3 V nominal). When operating power is first applied to the multiplexer IC, the attached I²C 'islands' are initially isolated, since the multiplexer is in a high-impedance state. The IC must first be accessed using its own I²C address and a control byte, in order to allow one of the two or four attached I²C line pairs to be connected through to the main I²C bus. In order to avoid interference to data transfers on the I²C bus, the connection is made only after the following Stop state, when the bus is again free.

Each of the multiplexer ICs has three address pins (A0, A1 and A2), so up to eight PCA954x ICs can be operated in parallel. Provisions are also made for handling interrupts. Each of the I²C zones has its own interrupt line, and all of the lines for the various zones are coupled in an AND arrangement. The I²C bus of the zone that is the source of the interrupt does not have to be actively connected to the main bus. The summed interrupt signal arrives at the interrupt input of the controller via an open-drain output. The controller detects that an interrupt has been generated somewhere in the overall system, and it must then determine the source zone of the interrupt by polling the multiplexers (by reading out their control bytes). After this, it can poll the ICs in the identified zone. The interrupt line of the I²C bus is released only after the IC that triggered the interrupt has been accessed and the interrupt has been acknowledged. If other interrupts have been generated in the meantime, the interrupt line will remain Low, and the controller must again search for the IC(s) that triggered the interrupt(s).

Figure 2 shows an example of an I²C bus system with one part that works at +3 V and a second part that is wired with +5-V pull-up resistors. Note that the pull-up resistors between the controller and the PCA9542 multiplexer are necessary to allow the multiplexer to be initially accessed after the 3-V supply voltage comes up, when it is still in the isolating state. If these resistors have larger values than the pull-up resistors used in the 5-V zones (by a factor of ten), the voltage in the +5-V zones will be able to rise to an adequate level.

The PCA9542 comes in a TSSOP-14 package, while the PCA9544 comes in a TSSOP-20 package. Additional information can be found at www.semiconductor.philips.com.

(004047-1)



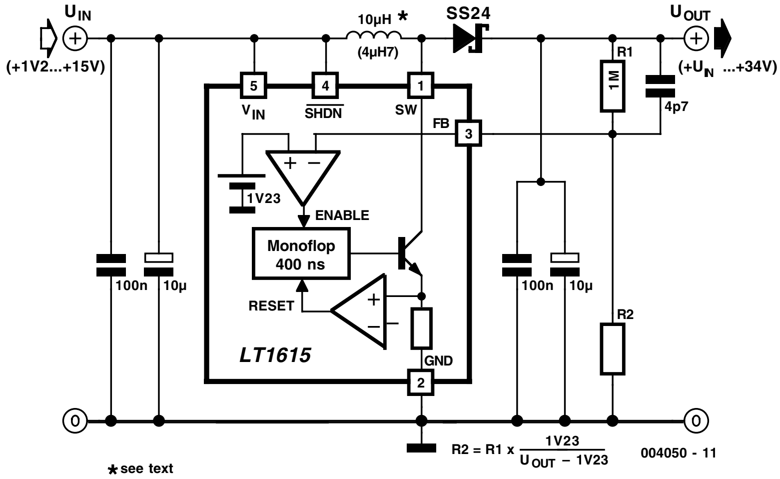
047 DC/DC Converter from 1.5 V to +34 V

G. Kleine

An interesting DC/DC converter IC is available from Linear Technology. The LT1615 step-up switching voltage regulator can generate an output voltage of up to +34 V from a +1.2 to +15-V supply, using only a few external components. The tiny 5-pin SOT23 package makes for very compact construction. This IC can for example be used to generate the high voltage needed for an LCD screen, the tuning voltage for a varicap diode and so on.

The internal circuit diagram of the LT1615 is shown in **Figure 1**. It contains a monostable with a pulse time of 400 ns, which determines the off time of the transistor switch. If the voltage sampled at the feedback input drops below the reference threshold level of 1.23 V, the transistor switches on and the current in the coil starts to increase. This builds up energy in the magnetic field of the coil. When the current through the coil reaches 350 mA, the monostable is triggered and switches the transistor off for the following 400 ns. Since the energy stored in the coil must go somewhere, current continues to flow through the coil, but it decreases linearly. This current charges the output capacitor via the Schottky diode (SS24, 40 V/2 A). As long as the voltage at FB remains higher than 1.23 V, nothing else happens. As soon as it drops below this level, however, the whole cycle is repeated. The hysteresis at the FB input is 8 mV. The output voltage can be calculated using the formula

1



$$V_{out} = 1.23\ V \cdot (R1+R2) / R2$$

The value of R1 can be selected in the megohm range, since the current into the FB input is only a few tens of nanoampères.

When the supply voltage is switched on, or if the output is short-circuited, the IC enters the power-up mode. As long as the voltage at FB is less than 0.6 V, the LT1615 output current is limited to 250 mA instead of 350 mA, and the monostable time is increased to 1.5 μs. These measures reduce the power dissipation in the coil and diode while the output voltage is rising.

In order to minimise the noise voltages produced when the coil is switched, the IC must be properly decoupled by capacitors at the input and output. The series resistance of these capacitors should be as low as possible, so that they can short noise voltages to earth. They should be located as close to the IC as possible, and connected directly to the earth plane. The area of the track at the switch output (SW) should be as small as possible. Connecting a 4.7-μF capacitor across the upper feedback capacitor helps to reduce the level of the output ripple voltage.

The selection of the coil inductance is described in detail in the LT1615 data sheet at *www.linear-tech.com*. Normally, a 4.7-μH filter choke is satisfactory for output voltages less than 7 V. For higher voltages, a 10-μH choke should be used. In the data sheet, the Coilcraft DO1608-472 (4.7 μF) and DO1608-100 (10 μF) are recommended. The Schottky diode must naturally have a reverse blocking voltage that is significantly greater than the value of the output voltage. The types MBR0530 and SS24 are recommended.

The shutdown input (/SHDN) can be used to disable the step-up regulator by applying a voltage that is less than +0.25 V. If the voltage at this pin is +0.9 V or higher, the LT1615 is active. You must bear in mind that even when the IC is disabled, the input voltage still can reach the output via the coil and the diode, reduced only by the forward voltage drop of the diode.

The second circuit diagram for the LT1615 (**Figure 2**) shows how you can make a symmetric power supply using this switching regulator. Here the switch output of the IC is tapped off and rectified using a symmetrical rectifier. The voltage divider at the positive output of the rectifier determines the output voltage

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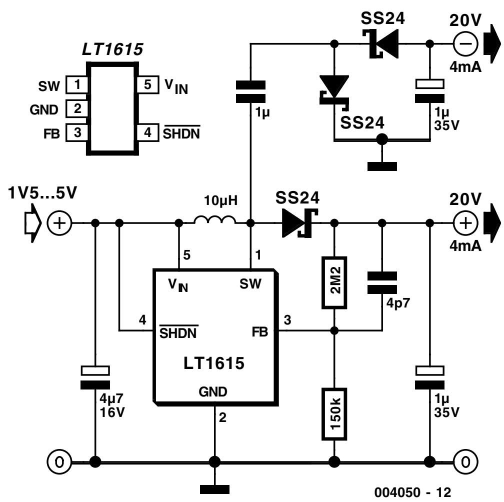
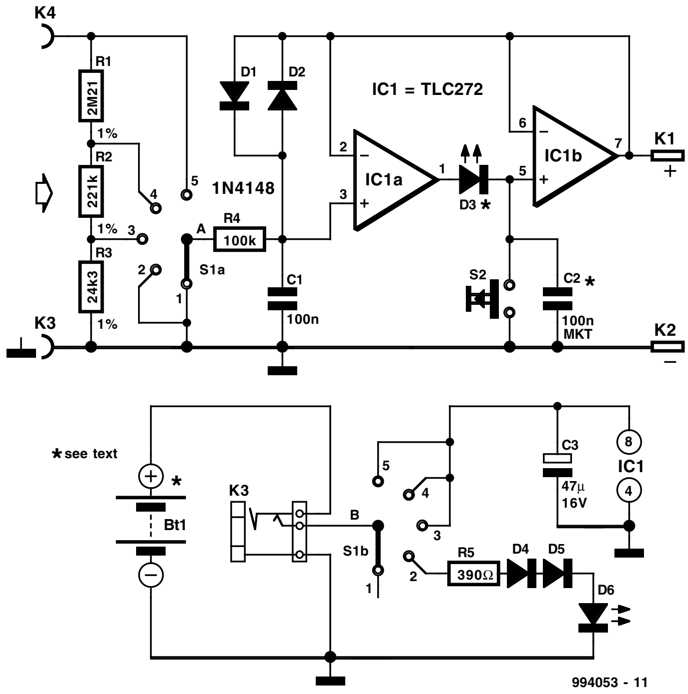


048 Hold Adaptor for Voltmeters

F. Hueber



Modern, good-quality multimeters have a hold function that enables a measurand to be read even after the test prods have been removed. The present adaptor is intended to add this facility to multimeters and voltmeters that are not so equipped.

In meters with a hold function, the measurand is quantized by an analogue-to-digital converter (ADC) and held in memory. This type of hold function is fairly sophisticated, but for can be attained by a rather less expensive and simpler analogue circuit.

A hold function is invariably based on a capacitor that is charged to the full value of the measurand. The resulting voltage across the capacitor is used to drive the display. In the case of analogue memories a variety of phenomena may affect the reading and lead to errors. Also, the charge on the capacitor does not remain constant owing to self-discharge, leakage currents on the board, the input current to a measurement amplifier, and so on. These deficiencies can, however, be negated by the use of a capacitor with a very high insulation resistance, a customized board layout, a current operational amplifier (op amp) with an input impedance in the region of teraohms.

The discharge current of a capacitor cannot be controlled properly by a silicon diode: it is far better to use a light-emitting diode (LED) that is totally screened, so that no light can escape. The photo current is then reduced from a few nA to some pA. This property of LEDs has been known for may years, but is hardly ever used for practical purposes. As an illustration, in the prototype the output voltage had dropped by only 1%, from 1.000 V to 0.990 V, thirty minutes after the original measurement was taken.

The circuit is simplicity itself as is evident from the diagram. In this type of circuit, the potential divider at the input is unavoidable. Here, the divider is arranged to measure input voltages of 2 V, 20 V, and 200 V: the voltmeter connected to the output remains set to the 2 V range.

The divider is followed by low-pass filter $R_{4\text{-}C}1$ which decouples the op amps from any noise and interference on the input signal. Resistor R_4, in conjunction with diodes D_1 and D_2, provides overvoltage protection.

The two op amps are arranged as voltage followers. Diode D_3 is the earlier mentioned screened LED, while C_2 is the reservoir capacitor with a very high insulation resistance. The potential across C_2 is applied to the non-inverting input of IC_{2b}. The low-impedance output of this voltage follower allows the use of a analogue pointer voltmeter or multimeter. Note that C_2 is shunted by miniature push-button switch S_2, which enables the capacitor to be discharged instantly when required.

Switch S_1 serves not only as input selector, but also as on/off switch (position 1), and enables the supply voltage to be monitored (position 2). In conjunction with diodes D_4 and D_5, light-emitting diode D_6 lights only when the supply voltage $<$2.8 V.

The supply voltage may be derived from a 3.6 V Li-ion battery, a 9 V size PP3 dry or rechargeable battery, or a mains adaptor. The current drawn by the hold adaptor is small: in the quiescent mode $<$1 mA, and in the absence of a hold voltage about 0.2 mA. The specified value of R_5 is right for supply voltages up to 5 V; when a 9 V battery or mains adaptor is used, the value should be increased to 1.2 kΩ.

When set to positions 3–5, switch S_1 functions as input selector. Diode D_6 is then not in operation.

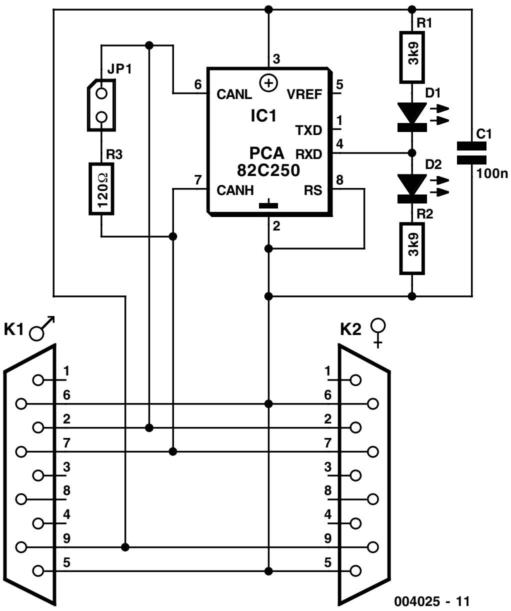
Only good-quality components should be used. Diode D_3 is a standard (not high-efficiency) red LED, which has been dipped a couple of times in black lacquer, but it can also be made light-tight with good-quality black insulating tape. Its terminals should also be well insulated to prevent leakage currents. The IC should be placed in a good-quality suitable socket. If a board is used, make sure that it is very clean (use pure alcohol) and free of any smudges such as fingerprints or handprints. When the construction work has been completed, it is beneficial to spray the track side with a good insulating lacquer. This will reduce the risk of leakage current sand also keep the tracks clean and free from corrosion.

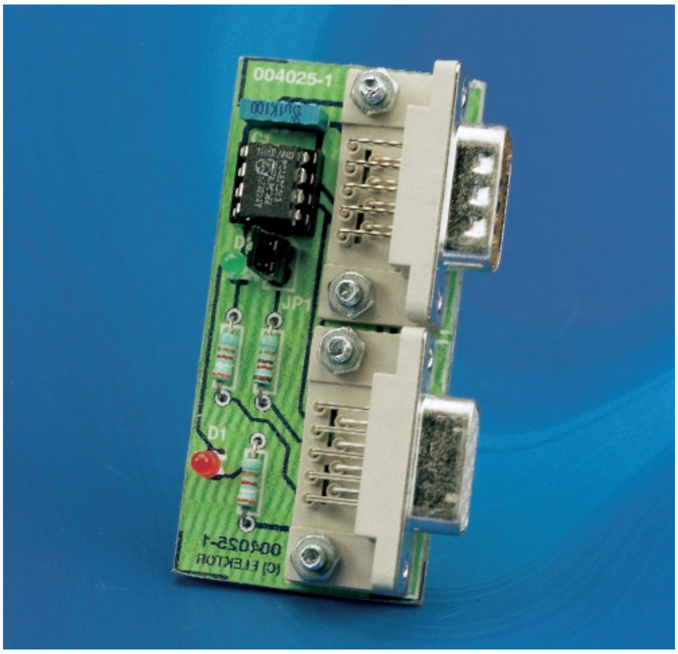
The adaptor need not be calibrated. After it has been switched on and after every measurement, press S_2 for about a second to make sure that capacitor C_2 is thoroughly discharged. Even then, an offset voltage of 2–3 mV may be measured. However, this is of no consequence, as long as the measurand is higher than this potential.

In principle, the circuit may also be used for measuring alternating voltages: the value of C_1 must then be reduced to about 0.001 μF, which gives an upper limit of 1000 Hz. Owing to the high input resistance and to prevent stray voltages from interfering with the measurement, the input sockets should be replaced by a BNC socket. Owing to the asymmetric supply voltage, the IC, in conjunction with D_3, forms a peak-voltage rectifier. This means that with a sinusoidal input signal, the output of the hold adaptor is $\times$1.414 higher than the r.m.s. value of the input signal.

[994053]

049 CANopener





A. Grace

This project was stimulated by a recent series of CAN bus articles in *Elektor Electronics*. One of the articles in the series included a constructional project (pages 20–24 in the November 1999 edition), which described a CAN bus interface. This construction project is based on that article.

At the fastest data rate of 1 Mbits/second the maximum bus length is 40 metres, whilst at the slowest speed of 50 kbits/second, the bus length can extend up to 1000 metres (1 km). If you imagine a CAN bus based on the slowest speed, the data cable could stretch for up to a maximum of 1 km — this is long enough to pass through several buildings! Take an industrial complex that uses a CAN bus to transmit data from one plant room to another (which could be on a different floor, or even in a separate building), and it develops a cable fault. You could spend hours wandering back and forth between the various bits of equipment trying to find the fault. What you need is a small hand held, loop-powered device that you can insert into the network to see if there is data on that node.

This circuit uses the Philips PCA82C250 transceiver chip. However, as the CANopener is being used to monitor CAN data only, the unused transmitter input is left floating — if grounded this could become a DOMINANT transmitter. Data enters the circuit through K2 and leaves unmodified through K1. IC1 converts the CAN bus data stream into a digital logic signal which is indicated by the two LEDs. If the CANopener is to replace a terminated piece of equipment, a terminating resistor can be switched into place through JP1, but in normal use JP1 is left open.

The PCB for this project is available ready-made from the Publishers.

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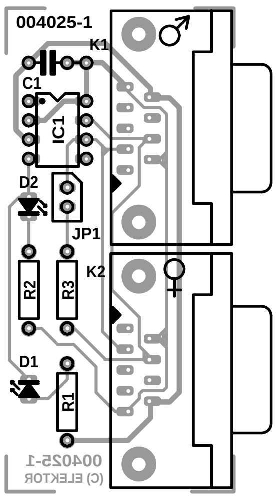
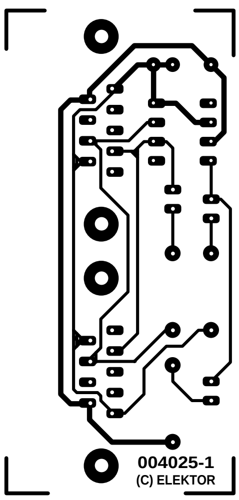




COMPONENTS LIST

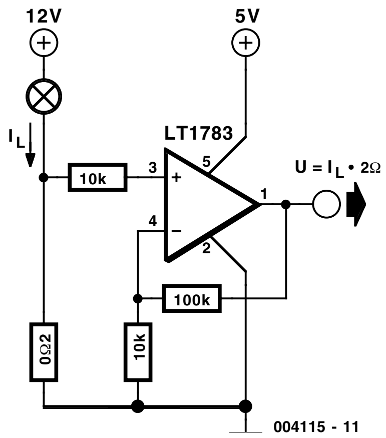
Resistors:
R1,2 = 3kΩ9
R3= 120Ω

Semiconductors:
D1 = LED, 3mm, high-efficiency, red
D2 = LED, 3mm, high-efficiency, green
IC1 = PCA82C250

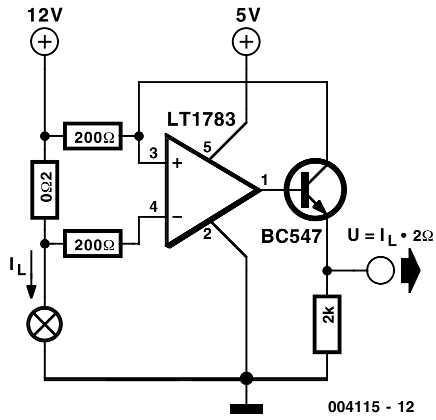
Miscellaneous:
K1 = 9 way sub-D male connector (plug), PCB mount version
J2 = 9 way sub-D female connector (socket), PCB mount version
JP1 = 2-way 2.54 mm pin strip and jumper

050 Rail-to-Rail and Over-the-Top Opamp

a



b



By G. Kleine

The LT1783 rail to rail opamp is an exceedingly robust device. It has reverse battery protection up to 18 V, a common mode input voltage of up to 18 V and protection on the input allowing voltages of –10 V to +24 V relative to the negative supply pin 2 (V–, for most applications at earth potential). With these protective features and its rail-to-rail input and output capability and over-the-top i.e. with one or both its inputs above the positive rail, this device is ideal for current sense applications with the sense resistor either in the supply side or earth side of the current path

The first diagram (**a**) shows an earth side current sensor circuit with linear output voltage. The opamp has a fixed gain of 10 so that the output voltage will be linearly proportional to the current through the 0.2 Ω sense resistor.

The second diagram (**b**) shows a positive supply rail current sense circuit. The opamp together with the transistor form a current source. The opamp controls the current through the upper 200 Ω resistor so that it has the same voltage drop as across the 0.2 Ω sense resistor. This current will therefore be 1000 times smaller than the current I_L through the lamp. This current through the 2 KΩ emitter resistor will produce a voltage at the output proportional to the load current. The relationship between the output voltage and load current is given in the diagram. You will notice that the lamp voltage in this application is +12 V while the supply voltage to the opamp is only +5 V.

The final diagram (**c**) shows a circuit that detects a blown lamp. The current sense resistor is again in the high side of the supply. When the lamp is switched off, both inputs to the opamp will be close to earth potential and when it is switched on they will both be close to the lamp supply voltage which, again, is much higher than the 3 V opamp supply. An output of +3 V from the opamp indicates that the lamp is burnt out. When the lamp is good and switched on, current through the 0.5 Ω sense resistor will produce 50 mV. When the lamp is switched off, a current through the 100 kΩ and 5 kΩ resistors will generate 15 mV across the inputs of the opamp. In both cases this will cause the output of the opamp to go low indicating that the lamp is OK.

When the lamp filament has blown the only path to earth will be through the internal 500 kΩ input resistors of the opamp. This time, the 5 kΩ resistor will generate a voltage between the inputs of the opamp of about 10 mV but of opposite polarity than before thereby producing a +3 V positive output from the opamp, indicating that the lamp is defective, the diode here will be reverse biased.

(004115)

Internet Address: www.linear-tech.com

c