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ELEKTOR ELECTRONICS

THE INTERNATIONAL ELECTRONICS MAGAZINE

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DIO & HI-FI • COMPUTERS & MICROPROCESSORS • DESIGN IDEAS • RADIO, TELEVISION & COMMUNICATIONS • SCIENCE & TECHNOLOGY • TEST & MEASUREMENT

PANDA-WEGENER STEREO DECODER

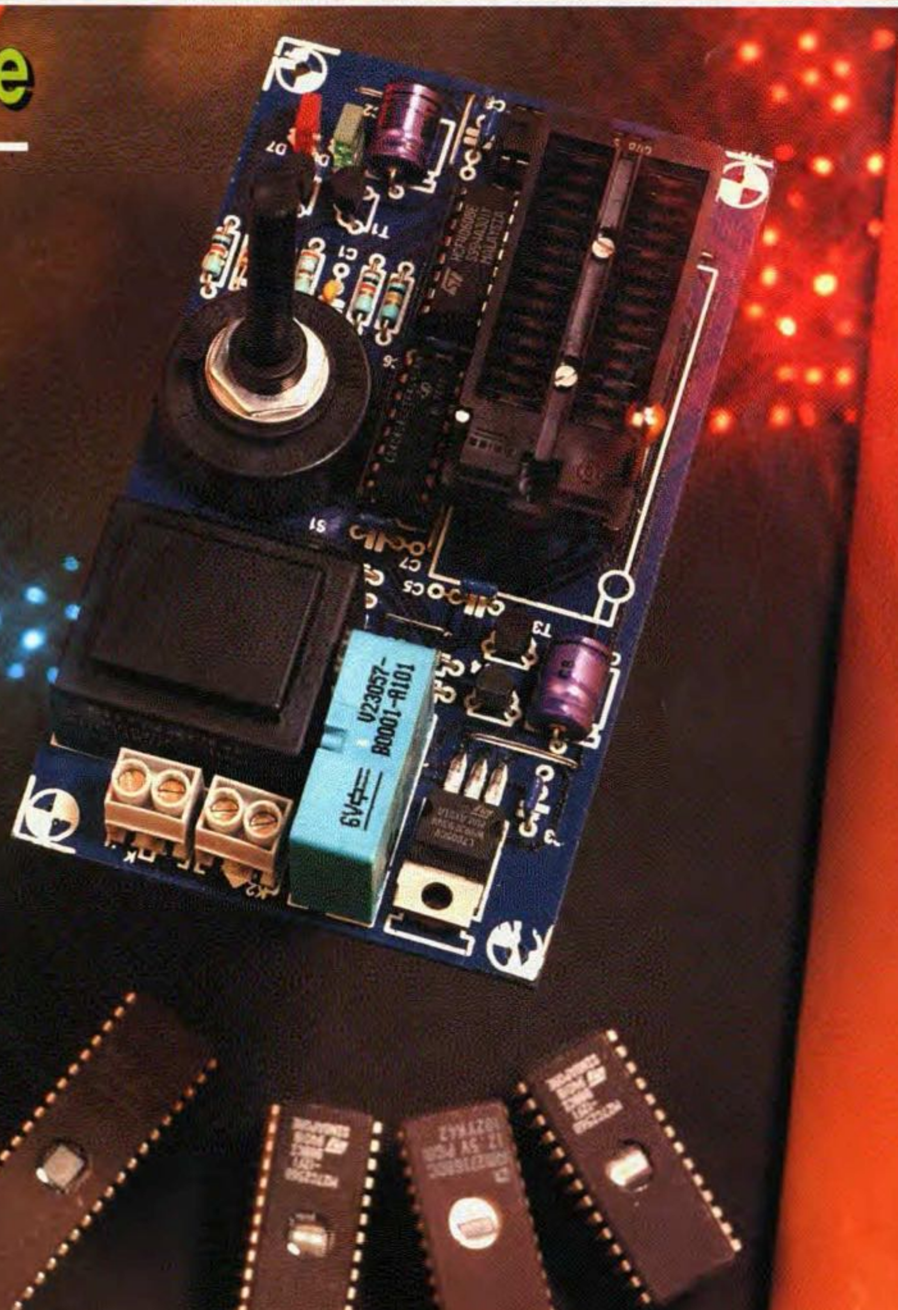
LED oscilloscope

4½ digit frequency meter

Analogue rev counter

Anemometer and flow meter

Switchable a.c. supply



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ELEKTOR ELECTRONICS

CONTENTS

September 1994
Volume 20
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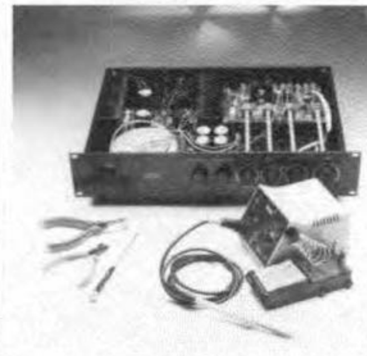
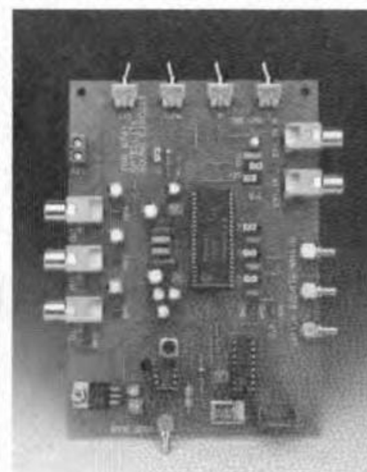
In next month's issue

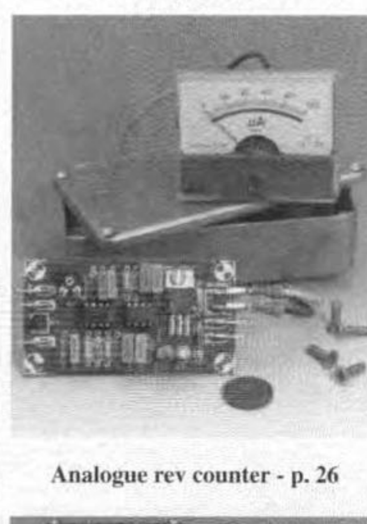
- TV line monitor
- Car audio amplifier
- Stable DC-DC converter
- Car battery charger
- Infra-red remote-control tester
- Checking power supplies
- Application note: Voice scrambler
- and others for your continued interest.



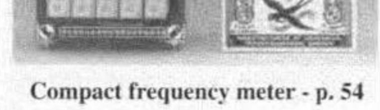
Integrated a.f. amplifier - p. 5



Panda-Wegener stereo decoder p. 10



Analogue rev counter - p. 26



Compact frequency meter - p. 54

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INTEGRATED A.F. AMPLIFIER - PART 1

Design by T. Giesberts

It sometimes seems as if the design of an audio amplifier is a competition to achieve the lowest possible distortion. However, reducing distortion below a certain (low) level merely increases the cost of the amplifier and does not further improve the sound quality. It is for this reason that the design of the integrated amplifier has been kept simple, although the maximum distortion is well below the discernible (audible) level.

To the normal, healthy human ear, the sound produced by an amplifier with a total harmonic distortion of, say, 0.001% sounds exactly the same as that produced by one which has a distortion figure of 0.2%, like the present integrated audio amplifier. Of course, the latter could not, in the world of the audio buff, be called 'hi-fi'. Neverthe-less, it sounds excellent, produces 85 W into 4 Ω or 45 W into 8 Ω and covers a frequency range of 10 Hz to 70 kHz.

Circuit description

The circuit of the preamplifier is shown in **Fig. 1**. The integrated input selector enables switching from one input to another simply by a direct voltage, which eliminates any likelihood of interference between the various inputs. The recording output uses an identical IC, so that this can also be switched without any interference arising and irrespective of the position of the input selector.

The voltage amplifier and tone control are based on Type NE5532 integrated operational amplifiers. This type of IC offers a virtually unrivalled price/quality ratio. The (arguably) slightly improved performance of 'better' devices bears no comparison to the far higher cost of these ICs.

The input signals are applied to the circuit via audio sockets K_1–K_8. There are four inputs: TAPE, TUNER, CD and AUX(iliary). All four are terminated into the standard 47 kΩ (R_1–R_8).

Four-channel analogue switches IC_1 and IC_2 are connected in parallel. IC_1 functions, in combination with S_1, as the RECORD SELECT control, while IC_1 and S_2 form the SOURCE SELECT control. These ICs contain buffer stages which can be actuated in pairs by the application of 12 V to their

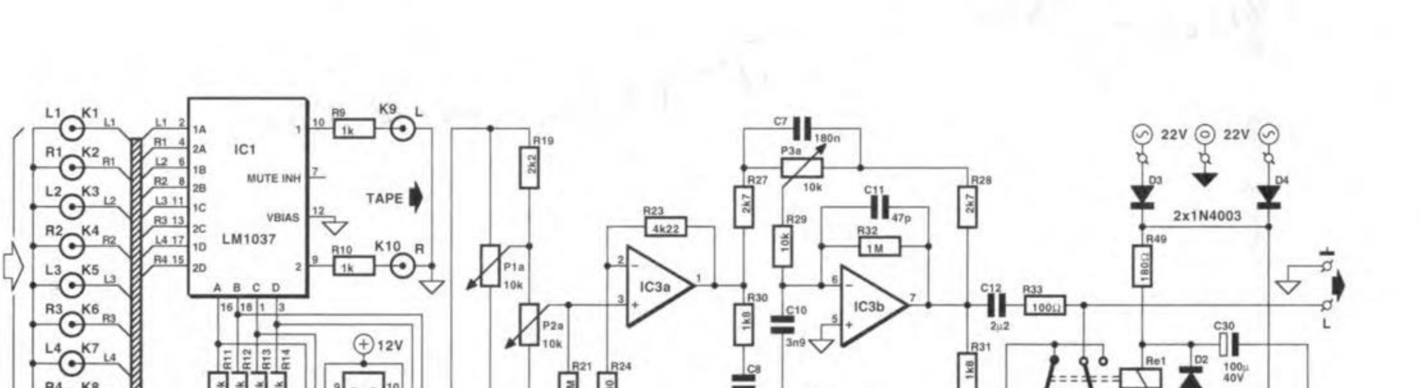

Fig. 1. Cicuit diagram of the preamplifier.

control inputs (A, B, C, D). They can operate from 5–28 V, have low distortion (0.04%) and a low noise level (5 µV).

During quiescent operation, the control inputs are held low by pull-down resistors R_{11}–R_{18}. They are actuated when 12 V is applied to them by S_1 or S_2. The ICs and the switches are connected via box header K_{11}.

The two outputs of IC_1 are applied to TAPE out sockets K_9 and K_{10}. Note that these signals are entirely independent of the SOURCE signal selected with IC_2.

The outputs of IC_2 are fed to the preamplifier proper via BALANCE control P_1. Resistors R_9 and R_{10} ensure that the balance control from zero is gradual. Logarithmic potentiometer P_2 functions as VOLUME control.

Voltage amplification is provided by IC_{3a} and IC_{4a}. The amplification is determined by the ratio R_{23}:R_{24} and R_{25}:R_{26}. In the present design, it is arranged at ×5 to give a nominal output voltage of 1 V for an input of about 250–300 mV. Bear in mind that the balance control introduces some attenuation.

Circuits IC_{3b} and IC_{4b} are active elements in the tone control, which is a standard Bandaxall design. Its cross-over point is at 1 kHz and it provides up to 12 dB attenuation or amplification. Potentiometer P_3 is the BASS control and P_4 the TREBLE control.

The output signal is 'freed' of annoying on-off switching plops by relay Re_1, which is controlled by T_1.

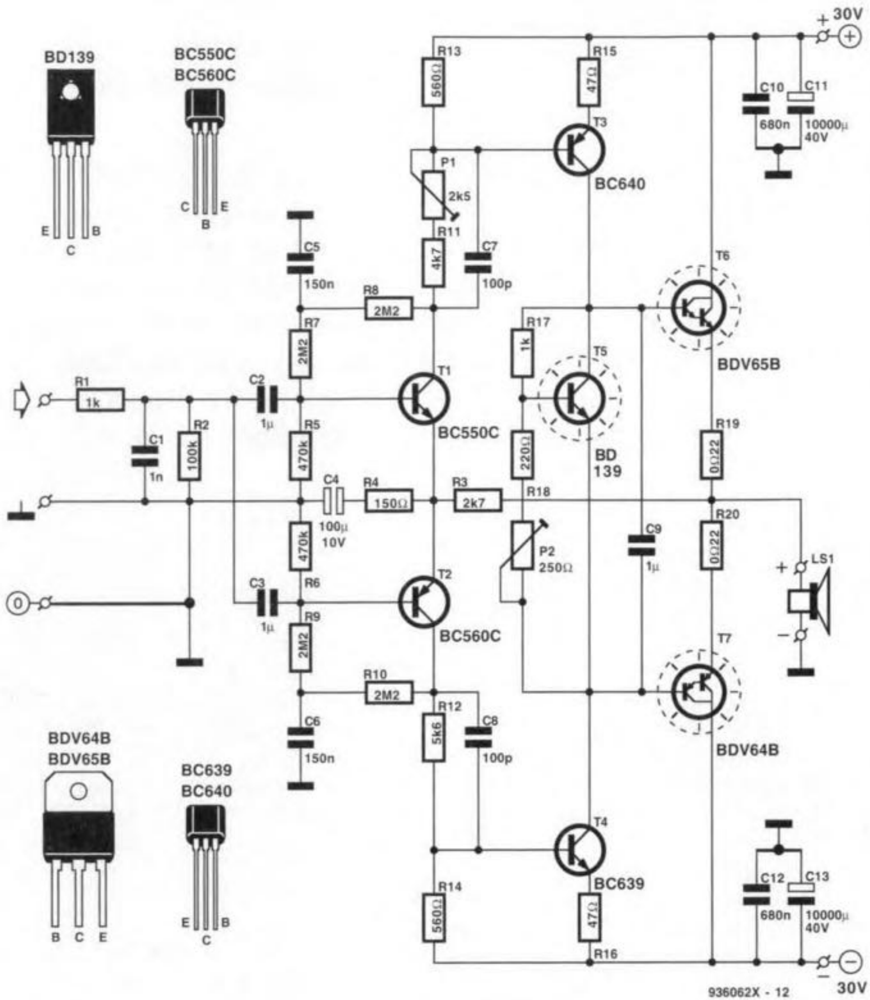

Fig. 2. Circuit diagram of the power amplifier.

Immediately the supply voltage is switched off, the relay is denergized, so that its contacts short-circuit all inputs to earth. When the supply is switched on, C_{14} must be charged, via R_{47}, to at least 1.4 V before T_1 comes on to actuate the relay. This means that the outputs are not enabled for a few seconds after the supply is switched on.

The circuit of the power amplifier is given in **Fig. 2**. Transistors T_1 and T_2 form the input amplifier and T_3 and T_4 the drivers for output stages T_6 and T_7. Transistor T_5 serves to set the quiescent current through the output transistors.

The input impedance of the power amplifier is determined by R_2 and parallel-connected R_5 and R_6. The input signal is applied to T_1 and T_2 via C_2 and C_3. The operating point of the transistors is set with the aid of R_5–R_{12}.

The amplified signal is then applied to driver stages T_3 and T_4. The operating point of these transistors is set by R_{11} and R_{12} respectively. Preset P_1 serves to ensure that the two operating points are identical, so that in the absence of an input the output is exactly zero.

Power darlington transistors T_6 and T_7 form the output stages, which provide the necessary current for driving the loudspeakers. Resistors R_{19} and R_{20} further stabilize the operating point of the transistors.

Transistor T_5 serves to set the temperature-independent quiescent current setting of the output stages. The current through these transistors is set to a given value with P_2. The collector-emitter voltage of T_5 is set to cause a tiny current (≈100 mA) through T_6 and T_7 to obviate cross-over distortion in the output stages. This kind of distortion arises when the transistors do not react instantly when a small signal is applied to their bases. This happens because the transistors do not start to conduct until the signal level exceeds the level of the base-emitter potential. When a small bias is present on the bases, the transistors react immediately even when the applied signal is small.

Capacitor C_9 decouples the (direct voltage) operating point.

Since T_5–T_7 have virtually the same temperature coefficient and, moreover, are thermally coupled, the operating points of T_6 and T_7 are automatically compensated for changes in the ambient temperature.

Like all transistors, T_1 and T_2 have a collector-base capacitance, which causes high frequencies to be amplified to a lesser degree than low and middle frequencies. This inequality is compensated by C_7 and C_8, which ensure that amplification is more or less even over the entire frequency range.

The overall amplification is determined by the feedback provided by R_3 and R_4. Capacitor C_4 ensures that theis feedback has no effect on the direct voltages. The feedback not only determines the amplification, but it also lessens distortion of the signal.

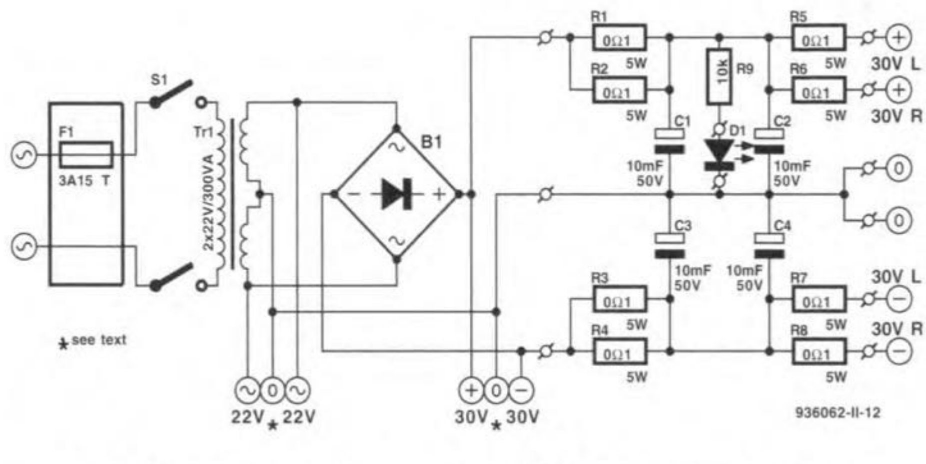

Fig. 3. Circuit diagram of the power supply for the integrated amplifier.

The circuit diagram of the power supply is shown in **Fig. 3**. The supply is simple, but robust. It consists of a sturdy toroidal mains transformer, a 35 A bridge rectifier and four large, heavy-duty electrolytic capacitors. Resistors R_1–R_4 protect these capacitors against too high charging currents. Resistors R_5–R_8 prevent the charging currrent to the power amplifier board rising unduly. Diode D_1 provides mains on indication.

The 30 V output lines are taken direct to the power amplifier (**Fig. 2**). and to the preamplifier (**Fig. 1**). In the preamplifier, they are brought down to 2×12 V by regulators IC_5 and IC_6. Series resistors R_{45} and R_{46} limit the voltage drop across these regulators, which, nevertheless, need to be mounted on a heatsink.

The relay delay circuit has its own power supply, which is derived directly from the secondary of mains transformer Tr_1. The 2×22 V alternating voltage is rectified by D_3 and D_4 and applied to T_1. The supply voltage for Re_1 is reduced by the few volts drop across R_{49} and buffered by C_{30}.

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BARNSLEY & DISTRICT AMATEUR RADIO CLUB

The Barnsley & District Amateur Radio Club will be holding its fourth Amateur Radio Rally on 13 November, 1994, at the Metrodome Complex in Barnsley town centre, which is less than 2 miles from Junction 37 of the M1 in South Yorkshire.

This is a **new** venue, all on one level with excellent facilities for the disabled, a licensed bar/restaurant and a separate cafeteria.

The Rally will have all the usual amateur radio and computer dealers, radio clubs, specialist groups and 'bring and buy'. There is ample car parking at the Metrodome, as well easy access from the bus and railway stations.

For further information contact Ernie, G4LUE QTHR, on (0226) 716 339 between 6 p.m. and 8 p.m. (Mondays 6–7 p.m.).

ROBBING THIEVES OF THE OPPORTUNITY TO STEAL

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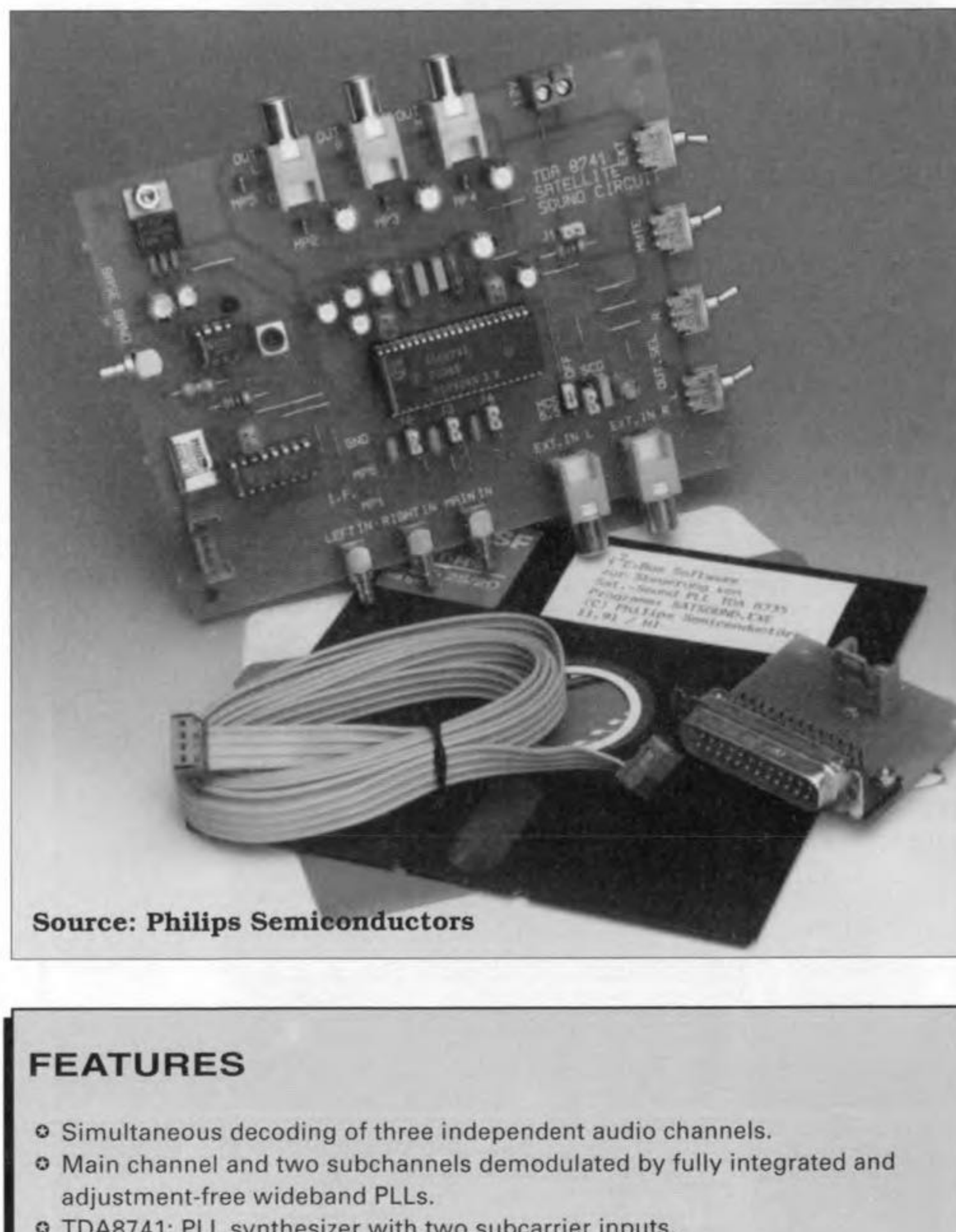
PANDA-WEGENER STEREO SOUND DECODER

Many channels on the Astra 1A/1B/1C TV satellites use the Panda-Wegener analogue companding system for their audio subcarriers. Unfortunately many older and low-end satellite TV receivers lack an appropriate decoder to bring out the excellent stereo sound quality offered by these subcarriers. This article describes a Panda-1 compatible decoder based on ICs from Philips Semiconductors.



Source: Philips Semiconductors

FEATURES

- Simultaneous decoding of three independent audio channels.
- Main channel and two subchannels demodulated by fully integrated and adjustment-free wideband PLLs.
- TDA8741: PLL synthesizer with two subcarrier inputs.
- Noise cancellation to Astra specifications.
- Audio switching between stereo, channel 1, channel 2, mono, external.
- Automatic mute during tuning (optional).
- Three SCART compatible line outputs.
- Tuning via I^2C bus.

ON a satellite TV receiver without a Panda-Wegener sound decoder, the companded sound channels received on the Astra channels sound 'flat', hissing, and sometimes even distorted. Obviously, some kind of decoder is in order!

Apart from the video signal, a baseband spectrum transmitted by a satellite TV transponder also contains one main sound carrier and up to eight sub-carriers for auxiliary sound channels. The main sound channel is monaural, and usually found at 6.5 MHz. It has a fixed pre-emphasis, and uses wideband FM modulation. The auxiliary subcarriers are also frequency modulated, and appear between 7.02 MHz and 8.28 MHz in a 180 kHz channel raster. Adaptive pre-emphasis is used at the transmitter side to improve the signal-to-noise ratio within the limits of the available channel bandwidth. The 'narrow-band' subcarriers in the baseband are used to convey hi-fi stereo music programmes or multi-language comment.

The Panda-1™ companding (compression/expanding) system developed by Wegener Communications has become the *de facto* standard for high-quality analogue stereo sound with satellite TV broadcasts. The basics of this interesting system are discussed below.

Panda-1 companding system

The aim of the Panda-1 companding system is to obtain a high signal-to-noise ratio for FM audio channels, whilst keeping the bandwidth used within limits. A good trade-off between these conflicting requirements is achieved by using adaptive modulation of the audio signal with signal frequency and signal level as the parameters. The bandwidth of a frequency modulated carrier depends on the modulation index (i.e., the gradient of the modulator characteristic), and the level of the modulation signal. To stay within reasonable limits as regards the bandwidth, the deviation is reduced to about 50 kHz. Also, the power of the Panda-1 subcarrier is reduced by about 6 dB relative to the main audio subcarrier. Consequently, the bandwidth is reduced to about 130 kHz, and the power to a quarter of the main

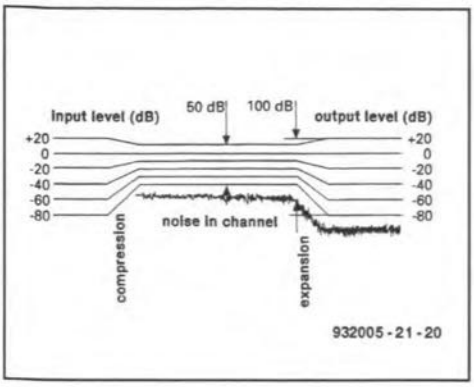


Fig. 1. Basic operation of a compander/expander system.

audio subcarrier (which is located at 6.5 MHz in the baseband spectrum). Although these measures allow the subcarriers to be transmitted with a spacing of 180 kHz in the baseband spectrum, they reduce the attainable signal-to-noise ratio, at least, theoretically. This reduction is counter-acted by a compander system based on adaptive pre-emphasis. The principle of operation is illustrated in **Fig. 1**.

Transmitter side

Pre-emphasis means that high-frequency components in the modulation signal are 'lifted' before they are applied to the modulator (which forms part of the transmitter). At the receiver side, the reverse process ('de-emphasis') is applied, so that the net result is nil (theoretically). If a fixed time constant is used in the pre-emphasis, the modulator is easily overdriven by high-frequency components in the audio input signal. At the same time, signal components with a high frequency but a relatively low level are, under certain conditions, insufficiently amplified and consequently, noisy. This may be avoided by using a pre-emphasis which adapts itself to the instantaneous signal level and signal frequency, hence the term 'adaptive' pre-emphasis. If the modulation signal presents strong high-frequency components, it is even given a small de-emphasis to prevent the modulator from being overdriven. By contrast, high frequencies with a relatively low level are given a strong pre-emphasis to lift them above the noise inherent to the channel. The modulation signal so obtained has a limited dynamic range, and its spectrum contains mainly high-frequency components. Certainly not great stuff to listen to!

The designation 'Panda-1' is a registered trademark of Wegener Communications, and refers to a specific set of companding characteristics only, not to hardware (although Wegener-licensed decoders are available). The present decoder is not officially approved by Wegener and, therefore, only *compatible* with the Panda-1 system.

Receiver side

The signal processing characteristic at the receiver side is the 'mirror image' of that at the transmitter side. The upshot is that the noise content of the channel is considerably reduced, resulting in a clear improvement of the signal-to-noise ratio (S/N).

Philips Semiconductors supply a series of satellite TV sound concepts based on the TDA8740 and the TDA8741/42 integrated circuits. Both offer you a complete audio signal processing path from baseband right up to the audio pins on the SCART socket. The best and most advanced concept is shown in **Fig. 2**. Three (mostly) analogue ICs are used in combination with the TDA8741 as the central building block. The baseband signal is applied to the familiar NE612A mixer/oscillator via a high-pass filter which separates the sound subcarrier from the video signal. The mixer/oscillator serves to shift the main audio carrier and the subcarriers such that they fit into three bandpass filter characteristics. The centre frequencies of these characteristics are 10.7 MHz (wide-band for the main audio carrier and narrow-band for the first subcarrier), and 10.52 MHz (narrow-band) for the second subcarrier.

For the frequency shift operation the VCO in the NE612A needs a control voltage. That is supplied by a PLL (phase-locked loop) type TDA8735. This IC is selected and controlled via the I²C bus. The software to do so is available from Philips Semiconductors, and is written for IBM PCs and compatibles.

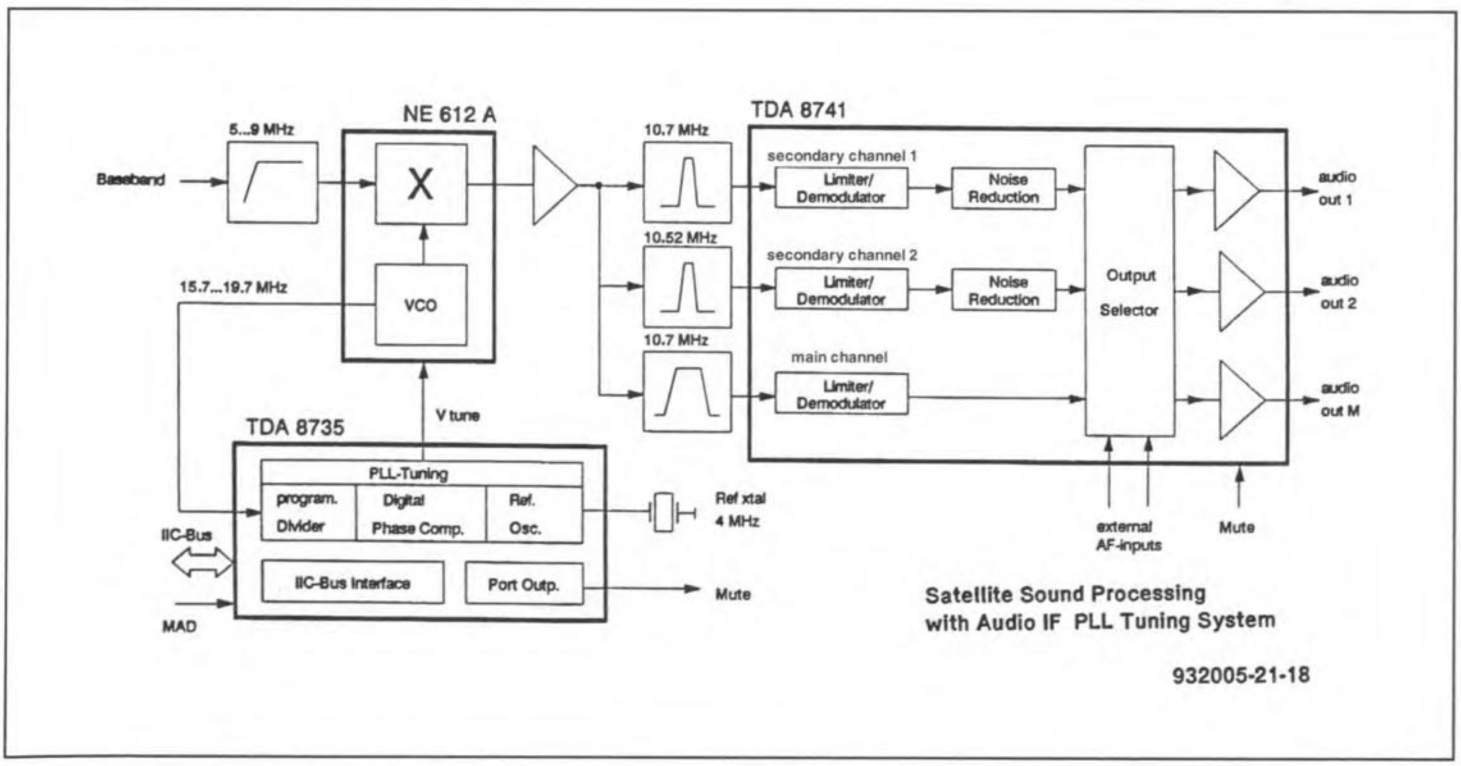


Fig. 2. Block diagram of the Wegener Panda-1 compatible stereo sound decoder. A Philips Semiconductors concept is used based on control via the I²C bus.

'Sat sound' ICs

The baseband signal, which is freed from video information below about 5 MHz, contains one main sound channel and up to eight subcarriers in a 180-kHz raster.

The baseband signal is applied to one input of the NE612A (single-ended use of the mixer input, see **Fig. 3**). The mixer is a multiplier which generates two output spectra, one below and one above the oscillator frequency. The upper spectrum, $f_{osc+fbb}$, appears in the normal position, and is not required here. Instead, the 'mirror image' spectrum, $f_{osc-fbb}$, is used, which also appears on the unbalanced mixer output, pin 5. The function of the mixer is illustrated in **Fig. 4**: the desired sound carrier spectra are fitted into the passband of the subsequent ceramic filters. The main (mono) sound carrier used on virtually all Astra channels is moved from 6.5 MHz to 10.7 MHz. Alternatively, two adjacent subcarriers are moved to 10.52 MHz and 10.7 MHz to recover a stereo or two-language programme. Three ceramic filters with bandwidths of 140 kHz (10.7 MHz) and 50 kHz (10.52/10.7 MHz) are used for the respective frequency bands. In this application, the oscillator in the NE612A is voltage-controlled, and also forms part of a phase control loop to ensure the highest possible tuning accuracy.

The other part of the PLL is contained in the TDA8735, of which the internal circuit is shown in **Fig. 5**. At the left, the VCO signal (pin 7) arrives at a prescaler and a programmable 13-bit divider. The reference oscillator, which uses an external 4-MHz quartz crystal on pins 1 and 4, is drawn at the right. In between these circuits sits a digital phase detector whose output signal controls a 5 µA or 500 µA current source. This, in turn, controls an external variable-capacitance diode (varicap) which determines the control voltage for the oscillator.

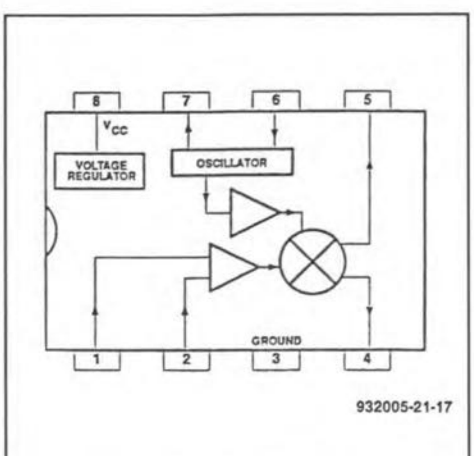


Fig. 3. Internal circuit of the NE612A.

Name	Byte	Bits	Function		
CP	DB0	D0	Charge pump, 0=low; 1=high		
S0-S14	DB0	D1-D7	Input frequency scaler		
	DB1	D0-D7	$(S0{\cdot}2^0+S1{\cdot}2^1+...+S13{\cdot}2^{14})\ f_{ref}$		
	DB2	D5	Enable prescaler 1: 0=on; 1=off		
REF1	DB2	D7	REF1	REF2	Frequency (kHz)
REF2	DB2	D6	0	0	1
			0	1	10
			1	0	25
			1	1	none
OPAMP	DB2	D4	Tuning voltage amplifier: 1=on; 0=off		
BS	DB2	D2	Open-collector switching output (Mute): 1=current to ground; 0=open-circuit		

databyte DB3 must be 0, and is reserved for test purposes.

Table 1. Databyte functions in I²C transmission.

The TDA8735 is controlled via the I²C bus. The main function of the control bus is programming the 13-bit counter which tunes the entire circuit. Auxiliary, but equally essential, functions controlled via the I²C bus are the charge pump current selection, the reference frequency selection, the mute control, and the control of the opamp in the control voltage output circuit.

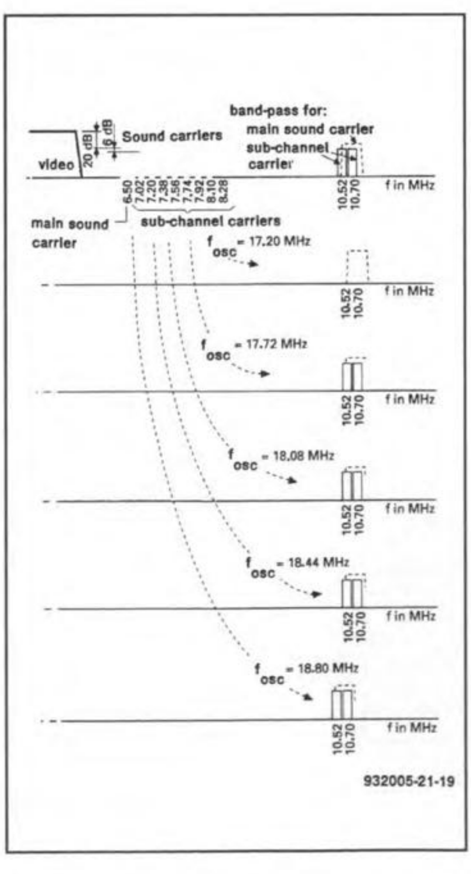


Fig. 4. Illustrating the up-conversion of the sound carriers to the frequency ranges determined by the ceramic filters.

Figure 6 shows the protocol for data and commands conveyed via the I²C bus. After the start condition comes the device address. Note that A1 must be '0' because the AS pin of the TDA8735 is permanently tied to ground. This addressing variant allows two TDA8735s to be used in the system. Since the TDA8735 is a 'slave device', the next bit, R/$\overline{W}$ (A0) must be '0'. The acknowledge bit of the PLL synthesizer is followed by the sub-address byte. This byte indicates which databyte is transmitted first. Next come the databytes in numerical order, for as long as they are requested after the acknowledge condition. The data transport is terminated by a 'stop' condition. The structure of the four databytes is given in **Table 1**.

Satellite Sound Circuit with Noise Reduction TDA8741

The heart of the circuit is formed by the TDA8741, whose internal organisation is shown in **Fig. 7**. The up-converted baseband signal is applied to three ceramic filters via a transistor amplifier stage. The narrow-band subcarriers are applied to pins 8 and 16 of the TDA8741, and the main audio subcarrier, at pin 18.

Inside the TDA8741, each of the high-frequency signals is applied to a limiter, and then to an adjustment-free PLL with demodulator. The audio signals are found at the outputs of these stages. The audio signal extracted from the main audio carrier ('ex-6.5 MHz') is available straight away. The stereo subcarriers, however, are separately demodulated, and each of the resulting audio signals is taken

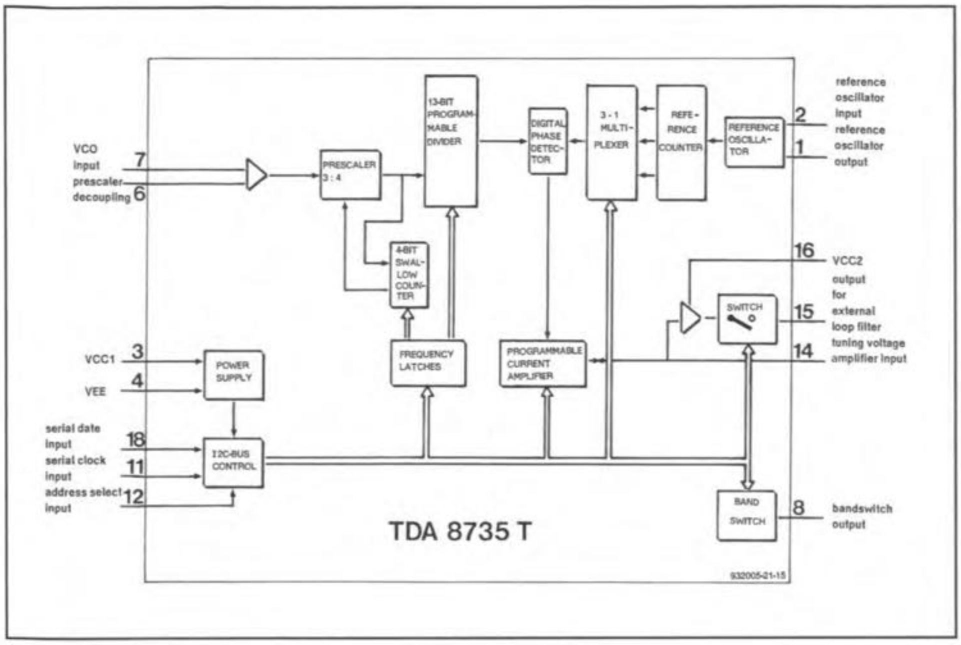


Fig. 5. Internal structure of the TDA8735 frequency synthesizer with I^{2}C bus control.

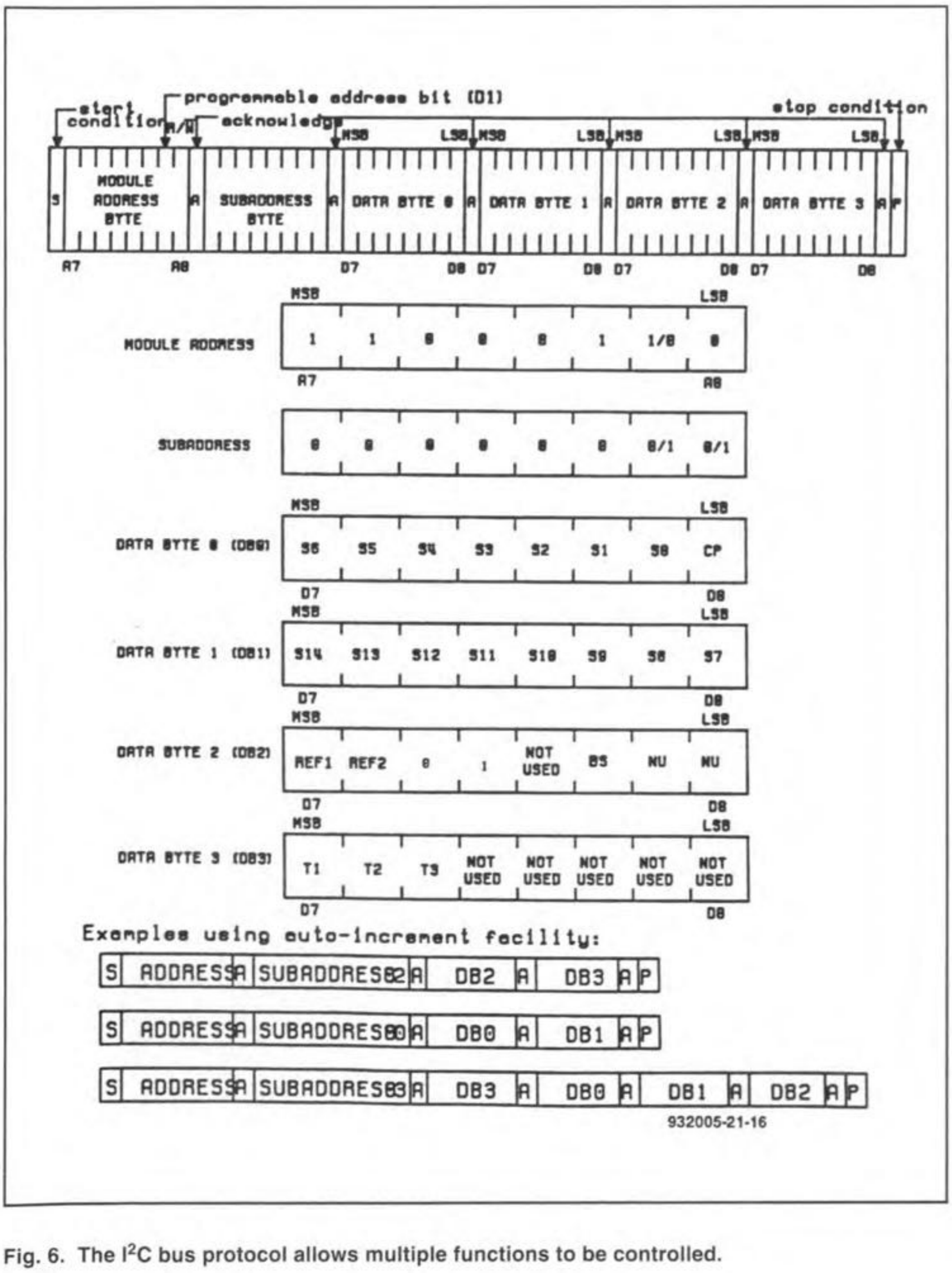


Fig. 6. The I^{2}C bus protocol allows multiple functions to be controlled.

through a direct-coupled low-pass filter and a noise reduction stage which is responsible for the Panda-1 compatible expansion ('decompanding'). Via a switching network, the audio signals recovered from the three subcarriers (or from one subcarrier and two external sources) are routed to +10-dB output amplifiers and from there to the IC outputs.

The level applied to the MCS pin of the TDA8741 determines the capture range of the PLL:

Level	**capture range**
0 V	5.5 to 7.5 MHz
1 V	10.0 to 11.5 MHz
1.8 to 2.8 V	off

The 0-V level may be used when a fixed sound IF is used and not, as in the present application, a tuneable IF (controlled by a PLL). For fixed sound channel applications, up-conversion is not necessary, and the centre frequency of the IF filter should be 6.5 MHz. For tuneable sound, the IF is set at 10.7 MHz (centre frequency).

The d.c. decoupling after the PLL is necessary to eliminate the effects of d.c. errors on demodulator output. These errors occur inevitably at high gain factors. The relevant coupling capacitor is connected to pin 41. The audio signal arrives at pin 20 via a buffer and a 1.5-kΩ internal resistor. Together with capacitor CDM, this resistor determines the de-emphasis on the main audio channel. The time constant equals 1500 C_{DM}, i.e. 50 µs with C_{DM}=33 nF, and 75 µs with C_{DM}=47 nF. For certain French and Italian transponders, a J17 de-emphasis is needed (Ref. 1). From pin 21, the signal travels to the switching subcircuit, and from there to an output buffer which supplies an output level of about 500 mV.

Like the main sound carrier, the subcarriers at 10.52 MHz and 10.7 MHz are first limited and then demodulated. The gain of the buffers at the demodulator outputs is determined by resistors R_{S1} and R_{S2}. A.c. coupling is used (capacitors C_{DCL} and C_{DCR}) and the signal is taken through a fourth-order Butterworth filter to eliminate the effects of spurious high-frequency products on the noise suppression circuit.

The noise suppression circuit may be looked upon as an input-level controlled low-pass filter (for adaptive de-emphasis) followed by a fixed de-emphasis. At the maximum input level (0 dB), the transfer characteristic of the adaptive de-emphasis is virtually linear. The lower the input level, the higher the attenuation of the higher frequencies with respect to the lower frequencies ('1:2 expansion').

Switch position	MUTE status	Pin 15, S1 OUTSEL L	pin 17, S2 OUTSEL R	pin 26, S3 EXT/INT	pin 13, S4
1	stereo	1	1	0	0
2	left	1	0	0	0
3	right	0	1	0	0
4	mono	0	0	0	0
5	external	*	*	1	0
6	MUTESEC	*	*	0	1
7	MUTEALL	*	*	1	1

* = don't care

Table 2. Truth table for the output signal selector.

TDA8741 MAIN SPECS

Supply voltage:	12 V
Main sound carrier	
Input sensitivity:	1 mV
PLL capture range:	10.0-11.5 MHz or 5.5-7.5 MHz
Output level:	500 mV
Noise distance:	65 dB(A)
Aux. sound carriers	
Input sensitivity:	0.5 mV
PLL capture range:	10.0-11.5 MHz
Output level:	500 mV
Noise distance:	74 dB(A)
Crosstalk	
Aux. to main carrier:	64 dB
Main to aux. carriers:	74 dB
Between aux. carriers:	74 dB

The behaviour of the noise suppression circuit at different input levels is shown in **Fig. 8**. The proper expansion characteristic is only achieved if the *RC* combinations responsible for the attack and decay times of the expander have the right values. The combinations are: C_{ATT} R_{EC}, R_{ECT} and $C_{NR\ D}$ (identical parts in the left and right channel). The time constant of the fixed de-emphasis is determined by the internal 2.3-kΩ resistor and capacitor CD.

For high-end applications, the input level of the noise suppression circuit must be adapted for the best possible transfer characteristic. The maximum level, 0 dB, corresponds to a maximum frequency deviation of 50 kHz for the companded sub-carriers. If the input level is too low (or too high), the high-frequency attenuation (or amplification) of the noise suppression circuit is greater than absolutely necessary. Consequently, an input level error of 1 dB causes a difference of 1 dB be-

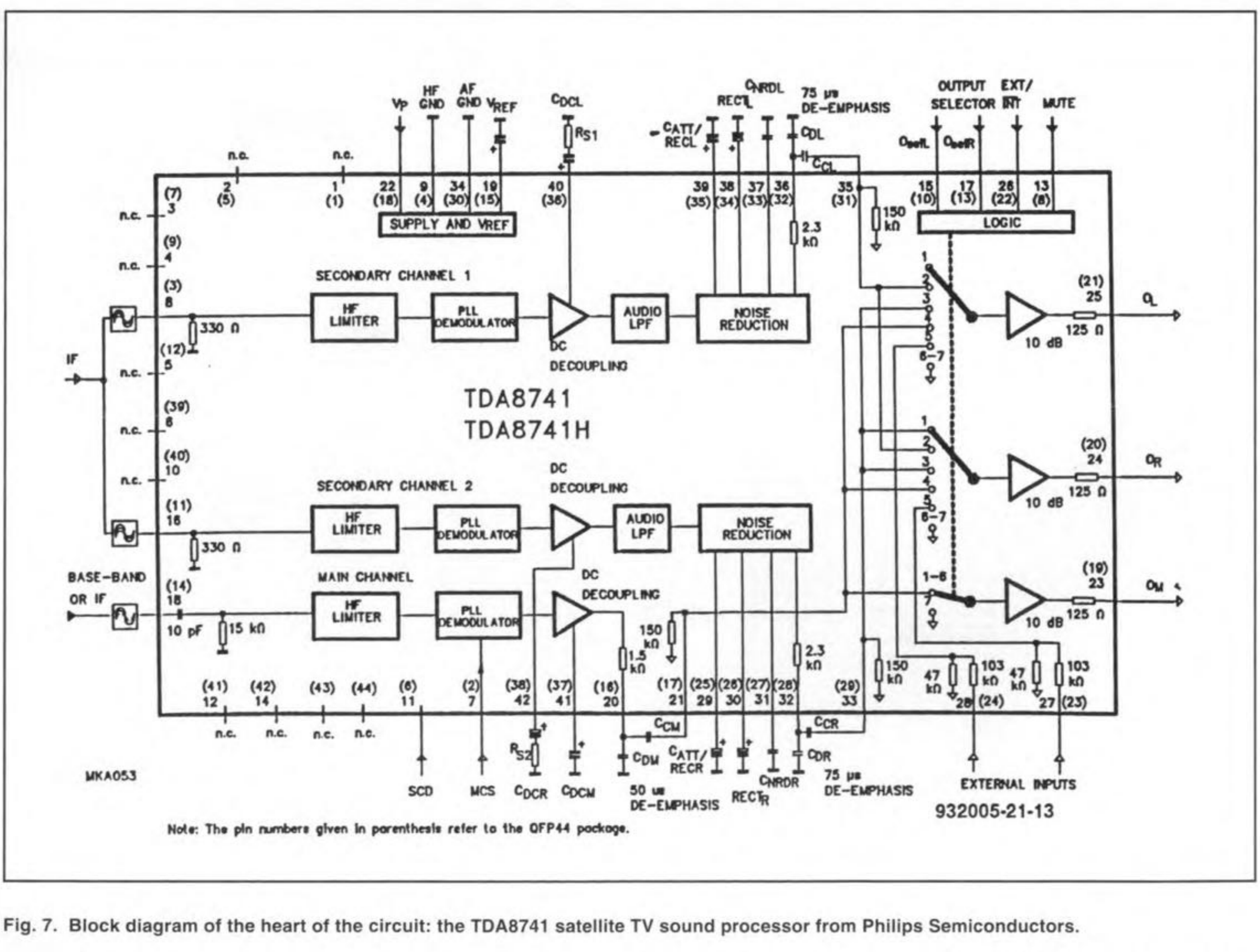


Fig. 7. Block diagram of the heart of the circuit: the TDA8741 satellite TV sound processor from Philips Semiconductors.

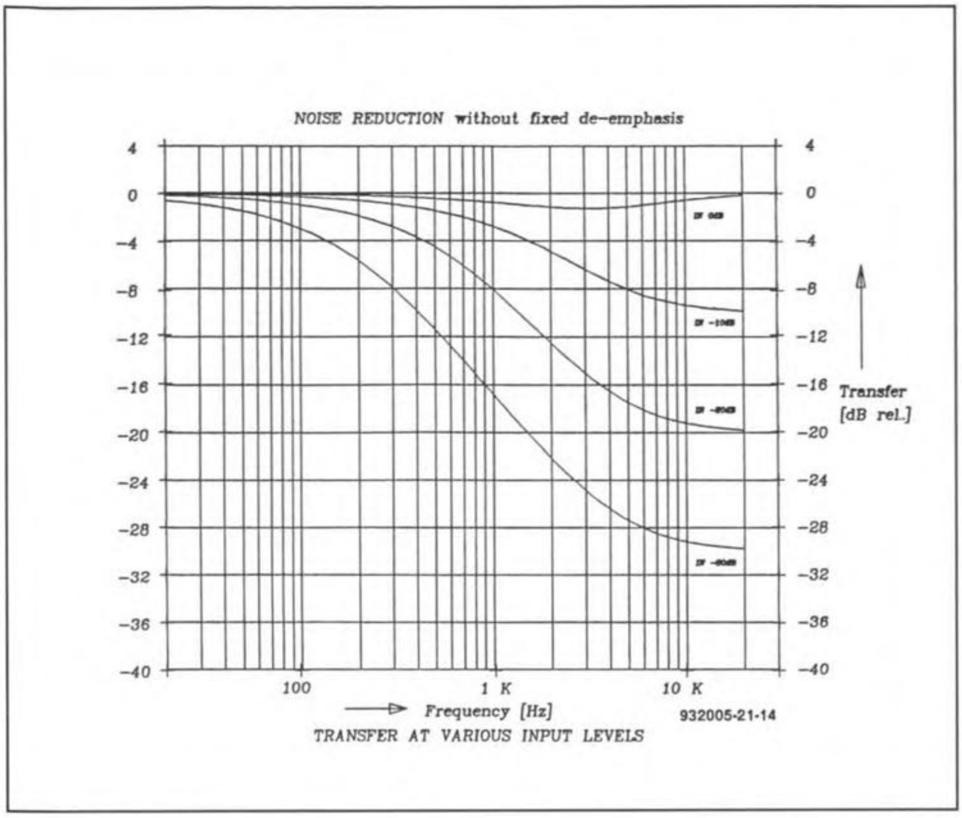


Fig. 8. The adaptive noise suppression operates virtually independently of signal level and frequency.

tween the lowest (50 Hz) and the highest (15 kHz) audio frequency. External resistors R_{s1} and R_{s2} allow the signal level applied to the noise suppression circuits to be set such that the output level at 15 kHz is exactly 0.25 dB lower than at 50 Hz, both at an input level of 0 dB.

Coupling capacitor C_C feeds the signal to the switching sub-circuit. Depending on the audio mode selected via pins 13, 15, 17 and 26, the signal(s) are routed to three output amplifiers. The switching options are summarized in **Table 2**.

Philips application board

The circuit diagram of the Panda-1 compatible stereo sound decoder is shown in **Fig. 9**. Only a handful of parts are used in addition to those which belong to the standard application circuits of the three ICs discussed above.

A large number of jumpers is used on the board to allow experiments with the many different functions offered by the decompander. Sockets K_3, K_4 and K_5 enable signal generators to be connected for measurement and test pur-

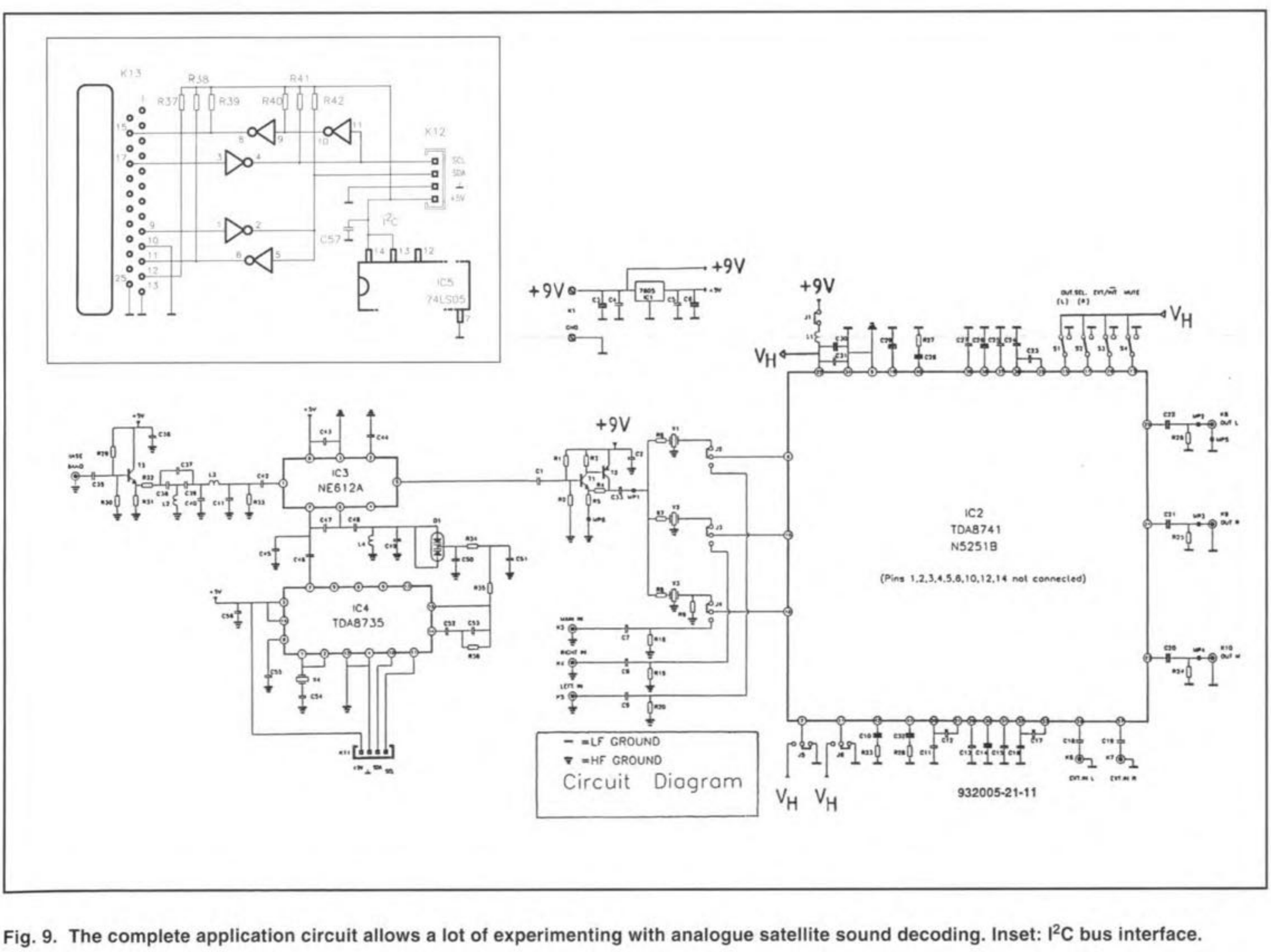


Fig. 9. The complete application circuit allows a lot of experimenting with analogue satellite sound decoding. Inset: I²C bus interface.

COMPONENTS LIST

Resistors:
R1 = 3kΩ9
R2 = 1kΩ8
R3;R4;R28 = 470Ω
R5 = 270Ω
R6-R9 = 330Ω
R18= 51Ω
R19;R20 = 56Ω
R23;R27 = 680Ω
R24-R28 = 100kΩ
R29 = 5kΩ6
R30 = 2kΩ2
R31 = 220Ω
R32 = 150Ω
R33;R35 = 1kΩ
R34 = 56kΩ
R36 = 8kΩ2

Capacitors:
C1;C33;C42;C48 = 1nF
C2;C4;C5;C31;C36;C43 = 22nF
C3 = 47µF 16V radial
C7;C8;C9 = 2nF2
C10;C14;C20;C21;C22;C26;C28;C29;
C30;C30;C32 = 10µF 16V radial
C11;C16;C24 = 33nF
C12;C17;C18;C19;C23 = 220nF
C13;C27;C52 = 1µF MKT
C15;C25 = 4nF7
C35 = 470pF
C37;C38 = 68pF
C39;C40;C41 = 100pF
C44;C51;C56 = 100nF
C45;C46 = 10pF
C47;C49 = 18pF
C50 = 560pF
C53;C55 = 10nF
C54 = 27pF
C34 = not fitted

Inductors:
L1 = 22µH
L2 = 4µH7
L3 = 10µH
L4 = 0µH82

Semiconductors:
D1 = BB204
T1;T3 = BC848B
T2 = BC858B
IC1 = 7805
IC2 = TDA8741
IC2 = NE612A
IC3 = TDA8735

Miscellaneous:
X1 = SFE10.7MJA10
X2 = SFE10.52MJA10
X3 = SFE10.7MS2
K6-K10 = RCA (audio line) socket for PCBmounting
42-pin SDIL socket
8-pin DIL socket
16-pin DIL socket

I^2C INTERFACE:

Resistors:
R37-R42 = 4kΩ7

Capacitor:
C57 = 100nF

Semiconductor:
IC5 = 74LS05 (SMD)

Miscellaneous:
Satsound software for I^2C bus (Philips Semiconductors).

poses. These inputs are terminated into 50 Ω. Jumpers J_2, J_3 and J_4 must, of course, be fitted accordingly. Jumper JP_5 sets the MCS mode discussed above. The default is a lock range of 10.0 MHz to 11.5 MHz. Jumper JP_6 (SCD) disables the PLLs for the subcarriers.

Switch position '1' (stereo) causes the left channel to be output via socket K_8, and the right channel, via socket K_9. In position '2', K_8 and K_9 supply the left channel. In position '3', K_8 and K_9 supply the right channel. In position '4', the monaural main subcarrier is fed to K_8 and K_9. In position '5', external signals from K_7 and K_6 are routed to K_8 and K_9. Position '6' causes the two subcarriers to be muted, to which the main subcarrier is added if the switch is turned to position '7'. In positions '1' through '6' the demodulated signal of the main sound subcarrier is output via socket K_{10}.

The circuit requires two supply voltages. IC_2 and the transistor amplifier around T_1 and T_2 are supplied with a regulated voltage between 8 V and 9 V. IC_3, IC_4 and input buffer T_3 are operated at 5 V, which is derived from the 9-V rail by IC_1. The I^2C bus is connected to the circuit via connector K_{11}.

The introductory photograph shows the application board designed for the circuit by Philips Semiconductors. Unfortunately this board is available to design laboratories and equipment manufacturers only. Permission to reproduce the artwork for the board could not be obtained from Philips Seniconductors. However, the decoder should not be too difficult to build on stripboard, since the board layout is not particularly critical. High-frequency connections should be kept as sort as possible, and the inputs and outputs should be kept as far apart as possible. Also, separate ground tracks should be used for RF and AF.

Software

The control program for the sound decoder is called SATSOUND.EXE. It runs on all IBM PCs and compatibles, and uses the Centronics parallel printer port for the I^2C bus connection. The program may be launched from floppy or hard disk by typing SATSOUND <ENTER>. The program saves its configuration in a file called SAT-

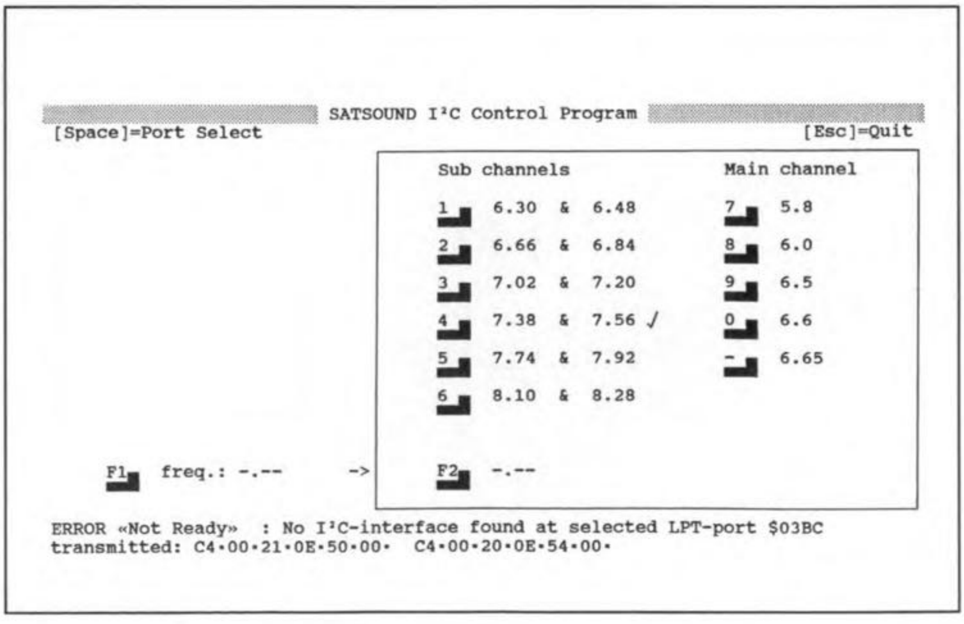


Fig. 10. Main menu of SATSOUND.EXE, the control software that belongs with the TDA8741 development board.

SOUND.INI. The Centronics port selection is either manual or automatic using the 'S' key. The selection does not require the interface (inset in the circuit diagram) to be powered or even connected to the decoder. The port selection is indicated by a 'tick' (√) behind the port address. The selection is saved by pressing the <space> bar. Selection without saving is accomplished by pressing the <ESC> key. The port selection may be called up at any time in the main program by pressing the <space> bar.

The main menu of the program is shown in **Fig. 10**. A total of five main sound subcarriers and six stereo subcarrier pairs (Astra-compatible) is available to choose from. The relevant frequencies are selected by pressing the numerical key indicated in tube menu. Other frequencies up to 9.99 MHz may be entered after pressing the <F1> key. Pressing <ENTER> causes these frequencies to be saved in the SATSOUND.INI file, while <F2> actually enables them. The stereo subcarrier pair selection is identical to that of the main sound subcarrier. Frequency pairs may be selected with a 'spacing' of 180 kHz.

All actuated functions are marked in **bold** characters on the screen, as well as by a √. Pressing any key actuates or changes the current status, and causes the corresponding set of data to be conveyed to the PLL via the I^2C bus. An error alert may be produced when an error occurs in the data transport. The most recent function and a list of the associated databytes is displayed in the bottom two lines of the screen. Note that the program does not receive information from the TDA8735, although it does process the acknowledge messages received from this IC. If you are not sure of the current status, call it up it by pressing the <ENTER> key.

The program also supports automatic muting wile the PLL is being tuned. To enable this function, remove jumper JP7, and fit a wire between pin 8 of the TDA8735 and pin 13 of the TDA8741. Also connect a 5-kΩ pull-up resistor to +5 V on this line. This modification disables the 'mute' switch, S_4. The default mute time is 100 ms, and is read as a parameter from the SATSOUND.INI file. Consequently, it may be given a different value simply by editing SATSOUND.INI using a word processor, and then re-launching SATSOUND proper. (932005-21)

Reference:

1. J17 equalizing network for satellite TV receivers, *Elektor Electronics* November 1991.

APPLICATION NOTE

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

HOME BUS MODEM IC ST7537

The ST7537 modem IC from SGS Thomson Microelectronics enables a home bus to be implemented which uses the mains wiring to carry information around in buildings. The half-duplex modem achieves a speed of 1200 bits per second.

MANUFACTURED in CMOS technology, the ST7537 is an asynchronous half-duplex modem (modulator/demodulator) specially designed for automation applications in buildings. Commands issued by a master control system are transmitted to slave devices via the existing mains wiring. Slave devices can communicate information back to the master via the same lines. In other words, the mains wiring is used as a low-speed local area network (LAN). The communication protocol complies with the European EN 50061-1 Cenelec standard. The ST7537 was developed by SGS-Thomson as a contribution to a part of the Esprit project aimed at setting new standards for automation in buildings. Consequently, it can safely be assumed that this technology will soon spread from its original application area, industrial buildings, via the office, to the consumer market. An essential condition for this to happen is that the components required become available in large numbers and at low prices.

The maximum data speed offered by the ST7537 is 1200 bits per second, using a carrier frequency of 132.45 kHz. According to the EN 50061-1 Cenelec specification, the carrier frequency may lie between 125 kHz and 140 kHz (North American FCC rules indicate an upper limit of 450 kHz). Assuming a crystal frequency of 11.0592 MHz, a '0' at the modulator input causes a frequency of 133.05 kHz, and a '1' a frequency of 131.85 kHz. Both frequencies are well within the Cenelec specifications.

The designers of the ST7537 have opted for frequency shift keying (FSK) because that is a good trade-off between reliable data exchange between devices on the one hand, and low cost on the other (the latter aspect is important if the device is to be successful in a large market). Thanks to FSK, the link is relatively immune to noise and other interfering signals which are unfortunately present on the mains. The modem IC integrates all functions needed to transmit and receive data. The only external components needed to realise a complete modem are a line transformer with a driver stage, and some 'intelligence', for instance, in the form of a microprocessor. As regards the application proposed here, the line transformer forms part of the basic circuit, while the digital intelligence is vested in a PC which communicates with the modem.

Transmitter section

The block diagram in **Fig. 1** shows the modem's functions burned into silicon. The circuit enters 'transmit' mode when pin 20 (RX/$\overline{\text{TX}}$) is made logic low. If the level remains 'low' longer than one second, the IC reverts to 'receive' mode. To select 'transmit' mode again, the pin has to be 'high' for a minimum of 2 µs before another '0' is applied. In other word, to permanently select 'transmit' mode, pin 20 has to be driven with a pulsed signal.

Data to be transmitted reach the asynchronous FSK modulator at a nominal speed of 1200 bits per second. The modulated data which emerge from the FSK modulator are filtered by a switched-capacitor bandfilter, which suppresses the harmonics contained in the transmitted spectrum. The last section of the internal chain is a buffer which for proper biasing requires a feedback signal from the external power amplifier. The latter serves to drive the line transformer. Pin 23 (RxD) is logic 'high' while data is being transmitted.

Receiver section

This part of the IC comes to life when a high level is applied to pin 20 (RX/$\overline{\text{TX}}$). Data received via the line interface is applied to pin 3 (RAI), after which it is cleaned. The bandfilter has a bandwidth of about 12 kHz around the centre frequency. The filter output signal is amplified 20 dB, which causes larger signals are limited symmetrically. The resulting signal is mixed down. The mixer uses a carrier frequency supplied by the FSK modulator. An intermediate frequency (IF) bandfilter with a central frequency of 5.4 kHz improves the signal-to-noise ratio of the modulated signal before it is applied to the FSK modulator. An external 100-nF capacitor ensures the coupling between the output of the IF section (IFO) and the demodulator input (DEMI). The capacitor also removes the d.c. component in the signal path.

The RxD output supplies the demodulated signal if output 'CD' is low. If this output is not active, the RxD output is continuously high.

The ST7537 has two separate 10-V supply rails (A_{VDD} and D_{VDD}) with associated ground connections (A_{VSS} and D_{VSS}). The analogue and digital ground connections must be joined at the ground terminal of the power supply. In addition to the 10-V rails a 5-V supply is required for the digital buffers. This supply is indicated as D_{VCC}.

Source: SGS-Thomson Microelectronics

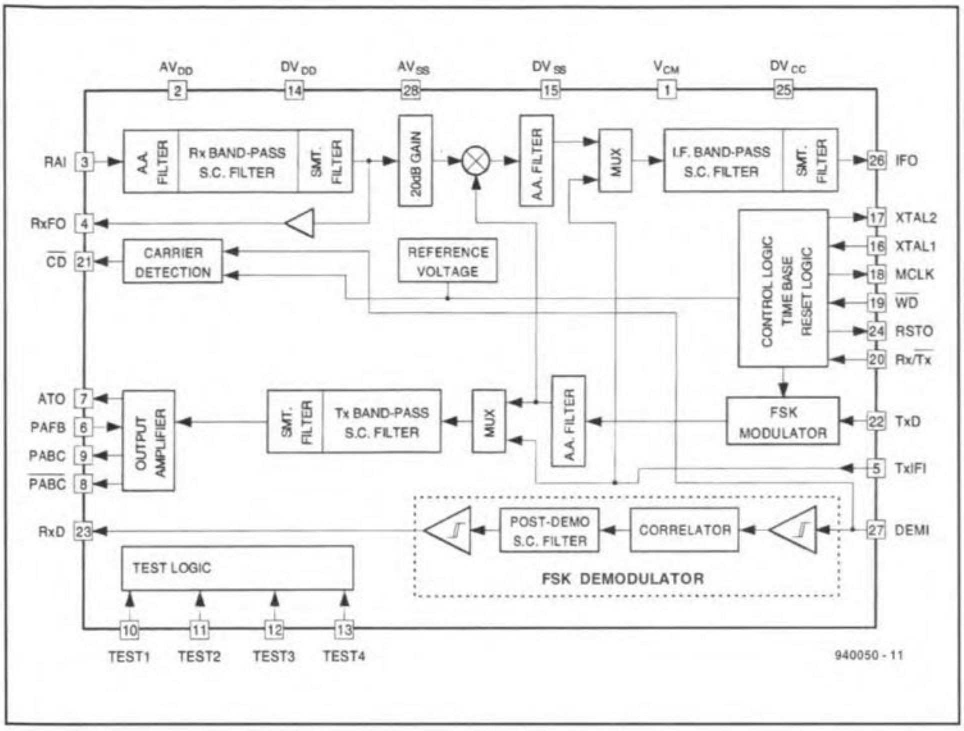


Fig. 1. Block diagram of the ST7537 modem IC from SGS-Thomson.

Application circuit

The standard application circuit for the ST7537 as developed by SGS-Thomson Microelectronics is shown in **Fig. 2**. Apart from a standard RS232 interface for the connection to the PC, the circuit contains a line transformer (from Toko) to establish the link to the mains network. Because the ST7537 operates with TTL logic levels only (0 V and 5 V), the familiar MAX232 voltage converter/line interface has been added. This IC converts the positive and negative voltage swings on the RS232 line into 0 V/5 V swing, and vice versa. The MAX232 has an on-board symmetrical d.c./d.c. step-up converter, and can make do with a single 5-V supply voltage.

Switches SW_1 through SW_8 with associated indicators LD_1-LD_4 are only functional for test purposes, and may be omitted for normal use.

Unfortunately the electrical isolation afforded by the T1002 transformer is insufficient to fully satisfy safety regulations as regards mains coupled apparatus. Full compliance can be achieved however by adding optically coupled inputs and outputs to the MAX232.

The line transformer is driven by a push-pull circuit consisting of transistors Q_5 and Q_2. The output driver is enabled via pin 7 of IC_1, and transistors Q_3 and Q_4. In this arrangement transistors Q_6 and Q_1 supply the base current for Q_5 and Q_2. In 'receive' mode the entire driver stage is switched off by cutting off Q_6 and Q_1.

Diode D_1 is a so-called 'transit' which protects the circuit against voltage surges which could enter the circuit via the mains. The transit reduces such surges to a harmless level of 6.8 V within 5 ns. If difficult to obtain, the transit may be replaced by an ordinary protection diode such as the

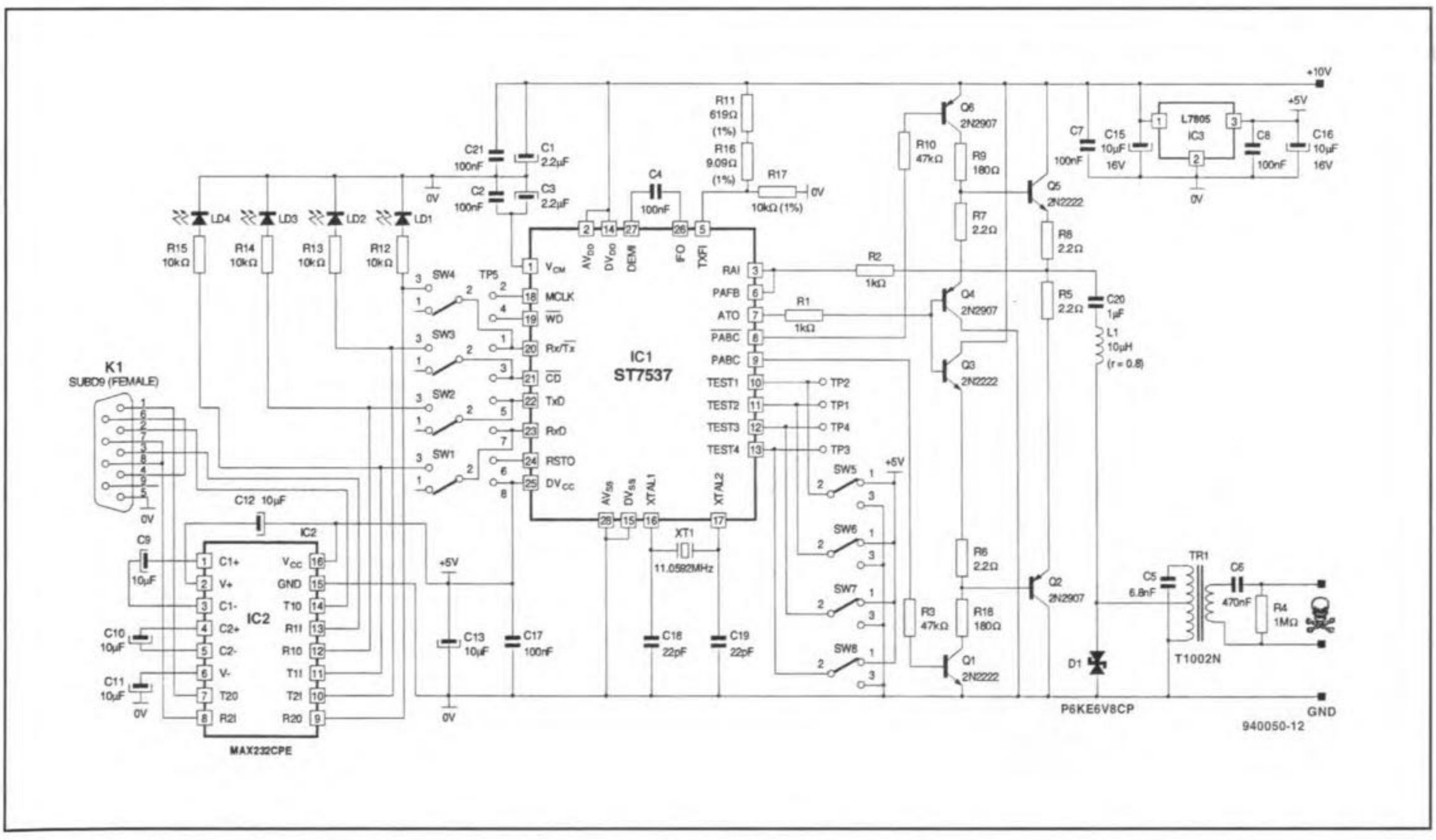


Fig. 2. Application circuit of the ST7537: a low-speed modem which communicates via the mains.

BZT03/C15 from Philips Semiconductors. Two of these devices should be connected in series with reverse polarity.

A 5-V voltage regulator, IC_3, is used to power IC_2 as well as the D_{VCC} and 'test' pins of the modem chip.

A number of these circuits may be used to set up a small network in which data is conveyed via the mains at a speed of 1200 bits per second.

The circuit shown in Fig. 2 is supplied as a demoboard by SGS Thomson Microelectronics distributors. Unfortunately the demoboard kit is fairly expensive, and therefore probably within reach of laboratories and colleges only. The application note published by SGS-Thomson in relation to the ST7537 demoboard contains the control program which is run on the PC.

(940050)

For further reading:

ST7537 Power line Modem - Application Note & Data Sheet, SGS Thomson Microelectronics.

SGS-Thomson Microelectronics UK head office:

Planar House, Parkway, Globe park, Marlow SL7 1YL. Tel. (0628) 890800. Fax: (0628) 890391.

SGS-Thomson Microelectronics distributors in the UK:

Abacus Electronics Ltd., tel. (0635) 33311, fax: (0635) 38670.

Access Electronic Comp. Ltd., tel. (0462) 480888.

Anzac Components Ltd., tel. (0628) 604411, fax: (0753) 661915.

Farnell Electronic Components Ltd., tel. (0532) 636311, fax: (0532) 633411.

HB Electronics Ltd., tel. (0204) 25544, fax: (0204) 384911.

Impulse Electronics Ltd., tel. (0833) 346433, fax: (0833) 346061.

ITT Multicomponents, tel. (0753) 824131, fax: (0753) 824160.

Polar Electronics PLC, tel. (0525) 377093.

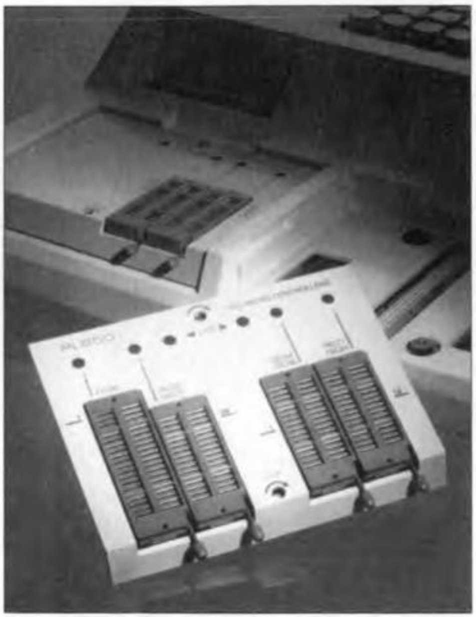
MMD Microprocessor & Memory Distribution, tel. (0734) 313232, fax: (0734) 313255.

NEW PRODUCTS

Lloyd Research L9000 programmer now supports PICs

Lloyd Research has now added a PIC module to support Microchip Technology's range of low-cost microcontrollers such as the 16C5x, 16C71/84 and 17C42 with the PL650 socket.

The L9000 is ideal for production programming because it can either operate in stand-alone mode, or be operated from a PC or from a batch file on a PC.



The PL650 was developed using the latest algorithms from Microchip. Easy device selection and the ability to automatically secure devices are key features.

Great care has been taken to optimise programming times. Most devices can be programmed in about three seconds. Devices which are typically programmed in serial mode on development programmers are programmed in parallel mode to increase speed.

The L9000 programmer is unique in that it offers the user two mehtods of operation. In production mode, all operations are simplified, and two verify operations are performed to ensure devices work at both high and low V_{cc}.

The L9000 main frame has positions for two modules. If a PIC module is fitted in one position, another module can be fited in the other position for other microcontroillers or EPROMs. The facility to cater for two or more different families makes the L9000 a cost-effective and versatile programmer.

LLoyd Research Ltd., 7/7A Brook Lane, Warsash, Southampton SO3 9FH. Tel. (0489) 574040 or 885515. Fax: (0489) 885853.

The Electronics Software Compendium on CD ROM

The Electronics Software Compendium is a collection of programs and data files that pertain to electronics, broadcasting, amateur radio and SWL activity. Over 15,000 files in total, with automatic unzipping to your hard drive. The disc is updated in April and October. Over 200 megabytes of PC material is resident on the CD ROM, and over 20 megabytes for the MAC. The price of the Electronics Software Compendium is £21.99 incl. VAT and P&P.

The Disc Trader, 85 Curzon Street, Derby DE1 1LN. Tel. (0332) 362770. Fax: (0332) 294486.

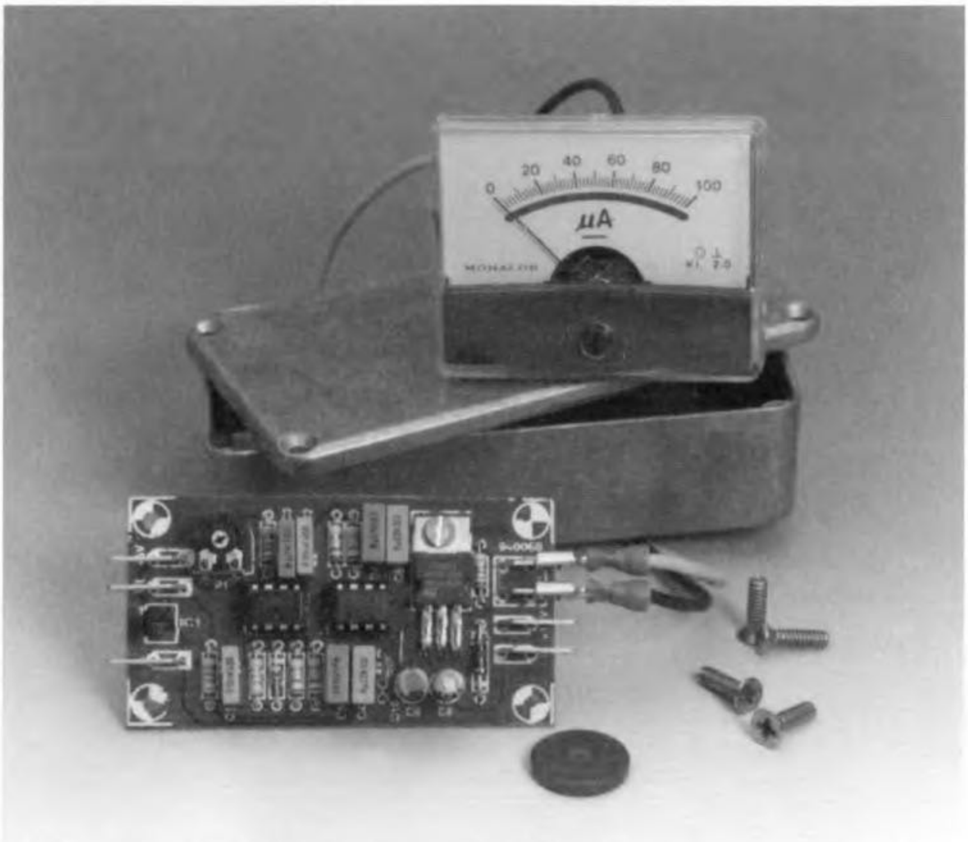
Catalogue for Elektor parts hunters and project builders

Now available from C-I Electronics is their 12-page mini-catalogue containing an overview of kits and components for projects published in *Elektor Electronics*, from May 1992 up to this September 1994 issue. The catalogue may be obtained against one IRC.

C-I Electronics, P.O. Box 22089, NL-6360 AB, Nuth, Holland. Fax: (+31) 45 241877.

REVOLUTION COUNTER

R
Design by F. Giamachi



Many small and older cars are not provided with a rev counter, which is a pity, because this instrument is very useful for driving with fuel economy in mind. Fortunately, it is not too difficult to add one, as this article shows.

The first item that is required for an accurate rev counter is a sensor. There is a variety of sensors to choose from, but in the present counter a Hall-effect type is used. This registers the movements of a small magnet which is fitted to the pulley that drives the generator, and produces corresponding voltage pulses. It is followed by a pulse shaper that converts the signal to a series of neat, single pulses; incidental variations in the width and amplitude of the pulses do not matter in this application. Any spurious signal variations are removed by a low-pass filter. The average d.c. component of the pulse train is then used to drive a moving coil meter.

Circuit description

In **Fig. 1**, IC_1 contains the Hall sensor just discussed. It produces a voltage at its output when it passes through a magnetic field. Thus, if it is fitted in close proximity to a rotating magnet, a pulse will appear at its open-collector output (pin 3) every time the magnet passes by. Since the type of sensor used reacts only to the south pole of the magnet, only one pulse per revolution is generated. The device also contains a voltage regulator, a temperature stabilizer, an amplifier, and a Schmitt trigger. These stages ensure that the output has one of two levels, corresponding to the presence or absence of a magnetic field respectively.

The pulse rate at the output of IC_1 accords with the number of engine revolutions. However, since the width of this signal varies with the number of revolutions, IC_1 is followed by a retriggerable monostable multivibrator, IC_2, which converts the output of IC_1 into pulses of identical width. The width is preset with P_1. To obviate IC_2 being retriggered when the input pulses are very wide, the output of IC_1 is first converted into narrow trigger pulses by differentiating network R_2-C_1.

To obtain an accurate average level of the pulses at the output (pin 3) of IC_2, the pulses are passed through a fourth-order low-pass filter. This is formed by two active second-order sections based on IC_{3a} and IC_{3b}. The cut-off point is arranged at 4 Hz (about $^1/_{60}$ of the number of revolutions), which ensures that as many as possible of the spurious signal variations are rejected. To ensure rapid reaction

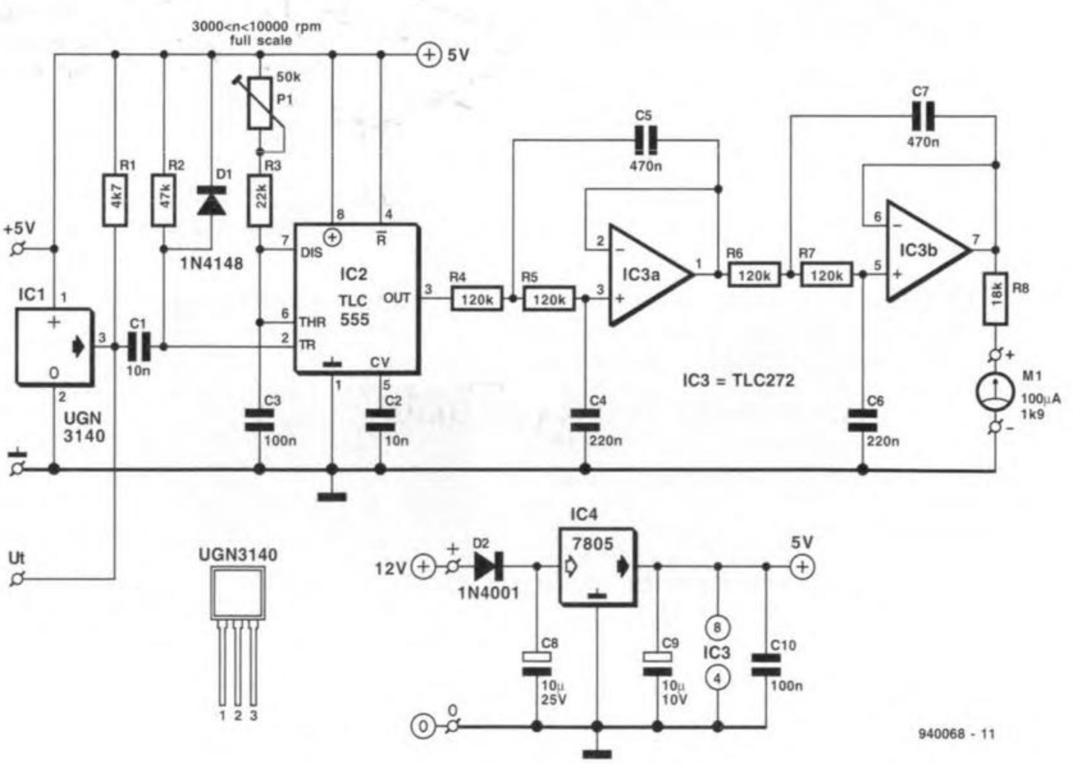


Fig. 1. Circuit diagram of the revolution counter.

Fig. 2. Printed circuit board for the revolution counter.

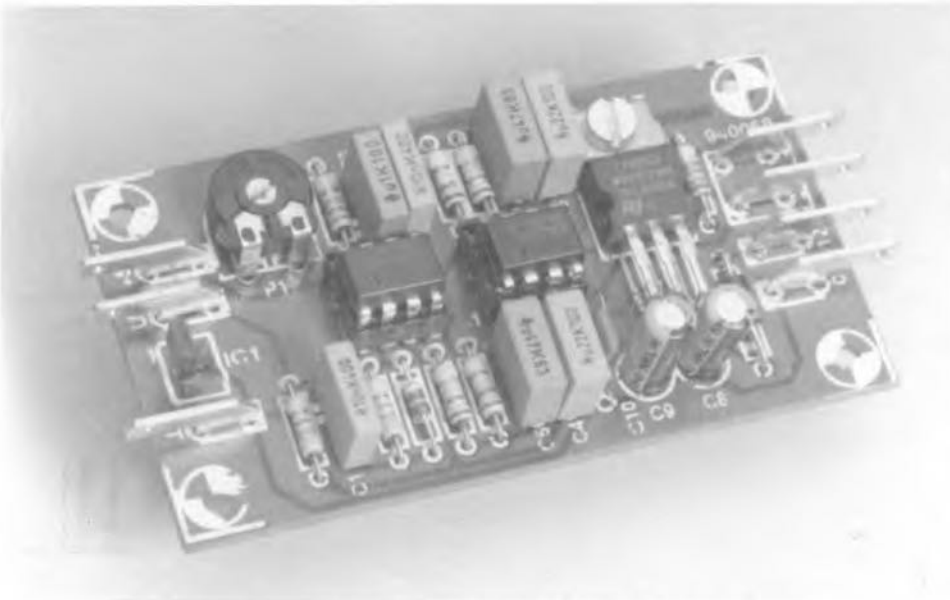


Fig. 3. The completed printed circuit board of the prototype.

of the circuit to changes in the number of revolutions, the filter has a sub-critical damping factor of 0.7.

The direct voltage so obtained is applied via R_8 to a moving coil meter that has a sensitivity of 100 μA.

Diode D_2 in the regulator section protects the circuit against wrong polarity of the connection to the 12 V car battery.

Construction

The rev counter is best built on the printed-circuit board shown in **Fig. 2**. Completing the board is straightforward; **Fig. 3** shows a completed board.

The connecting points for the meter, the battery and the sensor are clearly marked on the board. Note that use of right-angle connectors is made for these connections—see **Fig. 3**.

In many cases, it may be possible to fit IC_1 on the PCB. More about this a little later.

The small magnet must be glued (preferably with epoxy resin) to the pulley, next to its axle, that drives the generator. The PCB is fitted in a compact, watertight case (aluminium or plastic), which is then mounted close in front of the magnet. The actual mounting will depend on the type of car, so that no further guidelines can be given here. It all entails a little experimentation.

The voltmeter must, of course, be mounted at or near the dashboard. It is connected to the enclosed board by heat resistant wire.

Calibration

Before the unit can be mounted in a car, it needs to be calibrated. This may be done with the use of another rev counter as reference, but it can also be carried out as follows. A sufficiently strong 50 Hz field is ideal for calibrating the unit. Such a field is conveniently provided by the erase coil of a tape recorder. The Hall sensor of the prototype picked it up readily. A frequency of 50 Hz corresponds to 3000 rev/min. Adjust P1 so that this speed gives a meter reading at about mid-scale. Mark this position clearly.

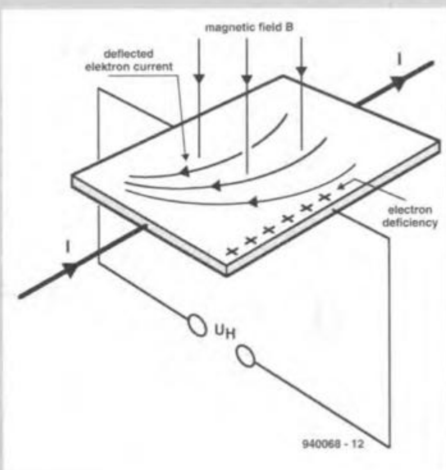
The values of R_3, P_1 and C_3 enable

Hall effect

The Hall effect is named after the American physicist, Edwin Hall, who first observed it. When a current-carrying conductor is placed in a magnetic field such that the field is at right angles to the direction of the current, an electric field (sometimes called Hall field), E_H, is produced in the conductor at right angles to both the current and the magnetic field. The field produced is related to the current, I, and the magnetic flux density, B, by:

$E_H = -R_H(I \times B)$,

in which R_H is the **Hall coefficient**. The electric field results in a potential difference, the Hall voltage, U_H, across the material.



Measurement of the Hall coefficients of various materials shows that in some cases the direction of the Hall field is reversed; that is, such materials have a positive Hall coefficient. This effect, the Suhl effect, indicates that the current is carried by positively charged carriers or holes in these materials. Measurement of the Hall coefficient therefore determines whether the charge carriers are electrons or holes and the value of the carrier concentration.

A practical Hall sensor consists of a thin wafer of metal or semiconductor material. When a current, I, is made to flow through the material, it is 'bent' when the wafer enters a magnetic field, B. This causes an electron deficiency at one side of the material, which results in a potential difference (Hall voltage, U_H) across the wafer.

The magnitude of the Hall voltage depends on the nature of the material and the dimensions of the wafer. If these are assumed to be constant, U_H appears to be directly proportional (up to about 2 V) on I and B. It is **not** dependent on the speed at which the magnetic field varies.

the instrument to indicate a maximum engine speed of 3000–10 000 rev/min. In most cases, this will be more than adequate; if not, however, the value of P_1 or R_3 may be altered to suit. As a rule of thumb, their combined value should be 220 MΩ divided by the engine speed in rev/min. For example, if a maximum speed of 15 000 rev/min is required, the combined value of the two elements should be 14.7 kΩ: make R_3 4.7 kΩ and P_1 10 kΩ.

In exceptional cases, it may be necessary to alter the values of the filter components. Calculation of the elements in the prototype assumes a stationary engine speed of 500 rev/min. If the actual speed is appreciably different, the values of R_4–R_7 may be altered to suit. Here, the rule of thumb is that the value of these resistors is 60 MΩ divided by the stationary engine speed.

Other types of sensor

If it proves difficult to get good results with the Hall sensor, or if this type of sensor is not readily available, the magnetoresistive sensor described in 'Fuel consumption meter'[1] may be used. The circuit of this is shown in **Fig. 4**. There is a ready-made PCB available (Ref. 940045) for this. The board has three connecting points, U_t, +5 V and ground (⊥), which are to be connected directly to the corresponding points on the present board.

The magnetoresistive circuit is so sensitive that it will always be able to pick up a usable pulse signal. It may be necessary to alter the values of P_1, C_3 and R_3 in the revolution counter by trial and error.

It is, of course, possible to use a different type of sensor that reacts to another location in the engine electrics. In some cars, the distributor generates an alternating magnetic field that may be used.

Parts list

Resistors:
R_1 = 4.7 kΩ
R_2 = 47 kΩ
R_3 = 22 kΩ
R_4–R_7 = 120 kΩ
R_8 = 18 kΩ
P_1 = 50 kΩ (47 kΩ) preset

Capacitors:
C_1, C_2 = 10 nF
C_3, C_{10} = 100 nF
C_4, C_6 = 220 nF
C_5, C_7 = 470 nF
C_8 = 10 μF, 25 V, radial
C_9 = 10 μF, 10 V, radial

Semiconductors:
D_1 = 1N4148
D_2 = 1N4001

Integrated circuits:
IC_1 = UGN3140 (Sprague)
IC_2 = TLC555CP
IC_3 = TLC272CP
IC_4 = 7805

Miscellaneous:
7 off right-angle PCB connectors
M_1 = 100 μA f.s.d., 1.9 kΩ
1 off small magnet
Enclosure (aluminium or plastic)
PCB Ref. 940068
PCB Ref. 940045 (if required - see text)

[940068]

[1] *Elektor Electronics*, June 1994, p. 14

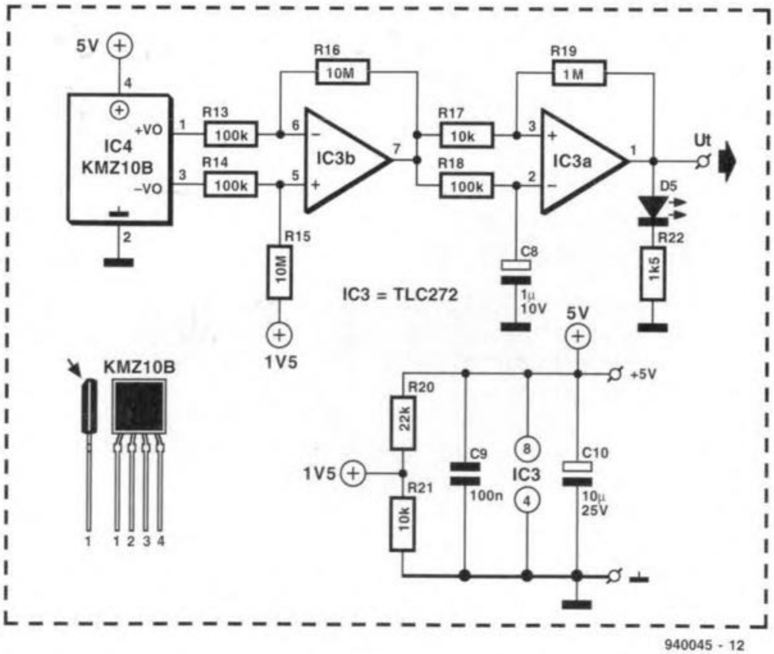
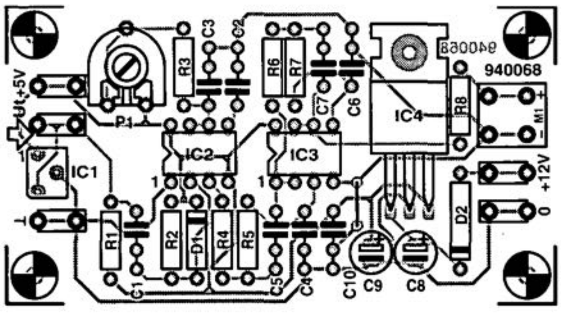
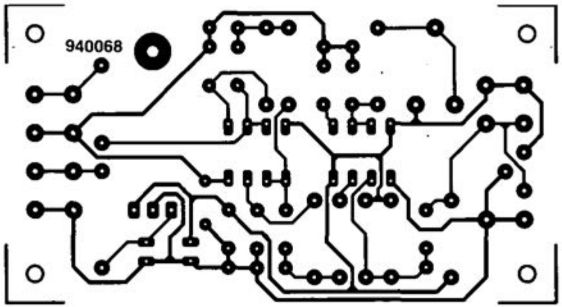


Fig. 4. Circuit diagram of the magnetoresistive sensor.


940068
Ut+5V
P1
R3
C3
C2
R6
R7
C7
C6
IC4
R8
M1
IC1
IC2
IC3
+12V
0
R1
R2
D1
R4
R5
D2
C1
C5
C4
C10
C9
C8


940068

FIGURING IT OUT

PART 19 – BEGINNINGS AND ENDINGS

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas. We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

This month we take the Laplace transform a step further by considering how to use it to find the initial and final states of a circuit. But, before we do this, we will look at circuits containing capacitive elements.

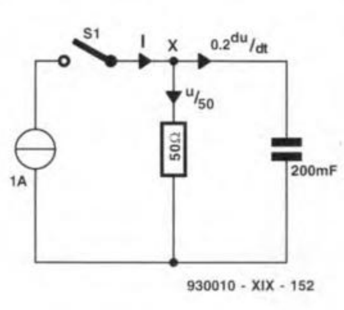


Fig. 152

A simple example is the capacitor of **Fig. 152**, which initially has no charge and is charged when S_1 is closed. The current being supplied to the circuit rises instantly from zero to 1 A. Last month, we described this as a **unit step function**. The currents arriving at and leaving node X at any instant are shown in the figure. The current through the resistor is obtained by applying Ohm's law. The current that is charging the capacitor is obtained from the current equation for a capacitor (Eq. 27, Part 5). Using Kirchhoff's current law, we can write:

$u(t)/50 + 0.2\mathrm{d}u/\mathrm{d}t = 1.$

The differential equation describes how voltage varies in the circuit from the moment S_1 is closed. Transforming this equation, using the tables of transforms given in boxes 1 and 2 of last month's article:

$U(s)/50 + 0.2[(sU(s) - u(0^+)]$
$= 1/s.$

As always, the use of the transform means that instead of **differentiating** a function of u, we simply have to **multiply** $U(s)$ by s, which is a much easier operation. Assuming that the capacitor is uncharged to begin with, so that $u(0^+) = 0$, the equation simplifies to:

$U(s)(1/50 + 0.2s) = 1/s$

$U(s) = 50/s(1 + 10s).$

Converting the right-hand side into partial fractions (see last month, Box 3):

$U(s) = 50/s - 500/(1 + 10s).$ [Eq. 138]

The second expression on the right will obviously transform back into the decay function, but we must first reduce the coefficient of s to unity. Divide top and bottom of this expression by 10:

$50/(0.1 + s).$

The transform becomes

$U(s) = 50[1/s - 1/(0.1 + s)].$ [Eq. 139]

The inverse transform is:

$u(t) = 50(1 - e^{-0.1}).$ [Eq. 140]

Figure 153 shows the graph of **Eq. 140**. The voltage across the resistor and capacitor starts at 0 V, increasing at an exponentially decreasing rate until it reaches 50 V.

Capacitor current

To model the currents in a capacitive circuit, we need to begin with a voltage equation. In Part 5 we gave the voltage equation of a capacitor in the form:

$$U = \frac{1}{C}\int I\,\mathrm{d}t + \frac{c}{C}.$$ [Eq. 21]

In this equation, u is the voltage across the capacitor at any given time, C is the capacitance, I is the current entering or leaving the capacitor at any given time, and c is the initial charge (if any) on the capacitor. Rewriting the equation with symbols to remind us that voltage and current are both functions of time, we obtain:

$$u(t) = \frac{1}{C}\int_0^t i(t)\mathrm{d}t + q(0^+)/C$$

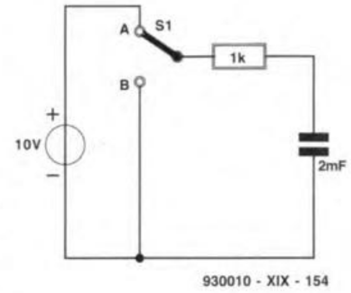


Fig. 154

[Eq.141]

The initial charge is now represented by $q(0^+)$. To work with equations which contain integral expressions, we use the transform given in Box 2 of Part 18. The transform of

$\int_0^t i(t)\,\mathrm{d}t$ is $I(s)/s + i(0^+)/s$.

Note that to integrate an expression, we **divide** its transform by s. This is easier than integrating.

As a practical example, consider the circuit of **Fig. 154**, in which a capacitor is charged to 10 V with the switch in position A. When $t = 0$, the switch is moved to position B. If the current in the circuit at any given time is $i(t)$, the voltage across the resistor is $Ri(t) = 1000i(t)$. The voltage across the capacitor is given by

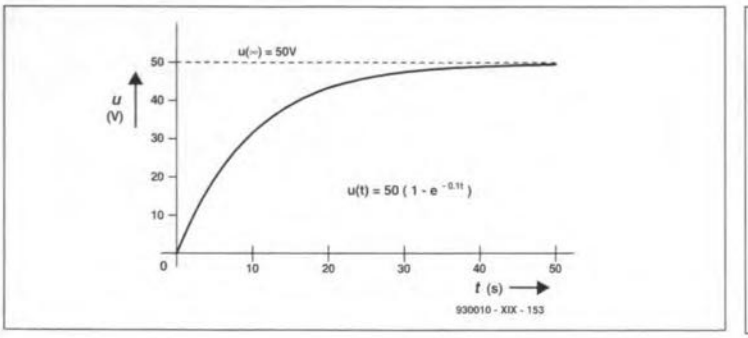


Fig. 153.

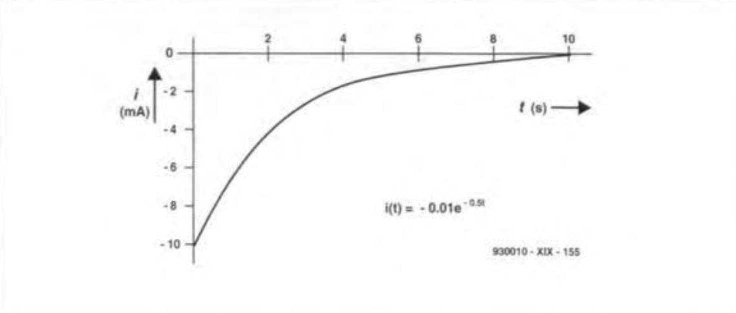


Fig. 155

Eq. 141 as:

$$\frac{1}{C}\int_0^t i(t)\mathrm{d}t+q(0^+)/C$$

Substituting C = 0.002 F and $q(0^+)/C = u(0^+)$ = 10 V, the initial voltage across the capacitor, into the equation for Kirchhoff's voltage law:

$$1000i(t)+500\int_0^t i(t)\mathrm{d}t+10=0$$

As usual, the equation contains an unknown function, in this case $i(t)$. We want to find it and we do not want to have to integrate it. So we transform the equation as in Boxes 1 and 2, Part 18:

$1000I(s)$
$+ 500[I(s)/s+I(0^+)/s]$
$+ 10/s = 0$.

The initital current, $I(0^+) = 0$. Substituting this in the equation:

$1000I(s)+500I(s)/s+10/s = 0$
$I(s)(1000+500/s)+10/s = 0$
$I(s)(1000s+500)/s = -10s$
$I(s) = -10/(1000s+500)$.

Dividing top and bottom of the fraction by 1000:

$I(s) = -0.01/(s+0.5)$. [Eq. 142]

The reverse transform is

$i(t) = -0.01e^{-0.5t}$. [Eq. 143]

Figure 155 shows the current from the moment the switch is changed. The current is −0.01 A (−10 mA) to begin with, as might be expected when the capacitor has 10 V across it and is discharging through a 1 kΩ resistor. The negative sign indicates the reverse direction of the current. The current decays exponentially to zero.

Initial value

The initial value of a current or voltage may be easy to calculate, as in the example above. Putting t = 0 in **Eq. 142**, we find:

$i(0) = -0.01e^0 = -0.01$ A.

There is another way to arrive at this result, using the Laplace transform. We apply the **initial value theorem**, which states that:

$$f(0)=\lim_{s\to\infty} s\mathrm{F}(s)$$

We perform the Laplace transform, multiply it by s, and find the limit of the product as s tends to infinity. In the example above, from **Eq. 141**, and taking limits as s approaches infinity:

Element	Voltage equation	Transformed voltage
Step voltage, $u(0^+)$	$u(t) = u(0^+)$	$U(s) = u(0^+)\,/\,s$
Resistance, R	$u(t) = i(t)\,r$	$U(s) = I(s)\,r$
Capacitance, C	$u(t) = (1/C)\int i(t)\mathrm{d}t$	$U(s) = I(s)\,/\,Cs$
Inductance, L	$u(t) = L\mathrm{d}i(t)\,/\,\mathrm{d}t$	$U(s) = Ls\,I\,(s)$

Table 1. Circuit transforms (voltages).

$i(0) = \lim -0.01s/(s+0.5)$.

Substituting 'infinity' directly into this equation leads to a meaningless indeterminate result. We overcome this problem by looking at the equation to see what is the highest power of s. The highest power is s^1, which is s. Divide top and bottom by s:

$$i(0)=\lim\frac{-0.01s/s}{s/s+0.5/s}$$
$$=\lim\frac{0.01}{1+0.5/s}$$

As s becomes very large, $0.5s$ becomes very small. Eventually, it is virtually zero and:

$i(0) = -0.01/1 = -0.01$.

This is the same result as obtained from our knowledge of electric circuits. The initial value theorem has its uses when the behaviour of the circuit is not so obvious, as is shown by a later example.

Final value

The **final value theorem** tells us the final value of the current, voltage or other quantity. This theorem states that:

$$f(\infty)=\lim_{s\to 0} s\mathrm{F}(s)$$

Using this theorem to find the final value of the current in the previous example:

$i(\infty) = \lim -0.01s/(s+0.5)$.

Substituting s = 0 in the above:

$i(\infty) = 0$.

As the capacitor discharges, the current eventually falls to zero.

If we apply these theorems to the transform of **Eq. 139**:

$$u(0)$$
$$=\lim_{s\to\infty} 50s[1/s-1/(0.004+s)]$$
$$=\lim_{s\to\infty} [50-50s/(0.004+s)]$$
$$=\lim_{s\to\infty} [50-50]=0$$

The initial voltage is zero. Next, for the final value:

$$u(\infty)$$
$$=\lim_{s\to 0} 50s[1/s-1/(0.004+s)]$$
$$=\lim_{s\to 0} [50-50s/(0.004+s)]$$
$$=\lim_{s\to 0} [50-0]=50$$

The final voltage is 50 V as originally deduced.

LC circuit

Now to look at something a little more complicated: a circuit that contains both capacitive and inductive elements (**Fig. 156a**). The capacitor is already charged to 6 V when the switch is closed. It helps the analysis to draw a tranformed circuit, as in **Fig. 156b**. The voltage across the capacitor is represented by the voltage source. Closing the switch is equivalent to introducing a step voltage into the circuit; the transform of this is $6/s$. The voltage equation for the capacitor is given in**Eq. 141**. Since we have already allowed for the initial charge on the capacitor, the term $q(0^+)/C$ is zero. Thus, the transformed voltage across the capacitor is $I(s)/Cs$. In **Fig. 156b** we have written the transformed impedance, which is the transformed voltage divided by $I(s)$.

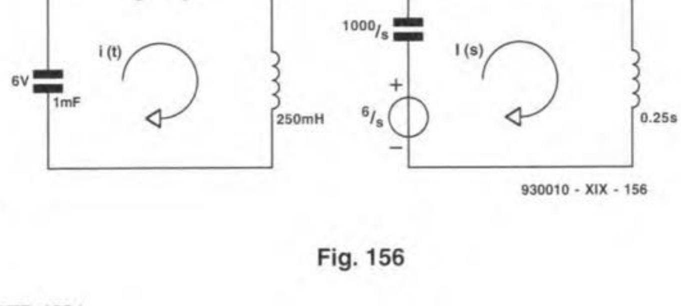


Fig. 156

This is $1/Cs = 1000/s$.

The voltage equation of an inductor, given in Part 5, is:

$$u_L=L\frac{\mathrm{d}i(t)}{\mathrm{d}t}$$

which transforms to $Ls/(s)$. The impedance is $Ls = 0.25\ s$. These transforms are summarised in **Table 1**. This is the table to use when working with voltages according to KVL, in order to calculate currents. In this example, we sum the transformed voltages around the circuit, according to KVL:

$I(s)1000/s+0.25sI(s) = 6s$
$I(s)[1000/s+0.25s] = 6/s$
$I(s)[(1000+0.25s^2)/s] = 6/s$
$I(s) = 6/(1000+0.25s^2)$. [Eq. 144]

This is the first time that we have wanted to reverse-transform an equation with s^2 in the denominator. Looking at Box 1, Part 18, the obvious candiate for the function is $\sin\omega t$, which transforms to $\omega/(s^2+\omega^2)$. First, we multiply the fraction of **Eq. 144**, top and bottom, by 4 to obtain s^2 in the denominator:

$I(s) = 24/4000+s^2$.

This makes $\omega^2 = 4000$, so that $\omega = 63.246$. The numerator is 24, which is ω multiplied by 24/63.246 = 0.3795. With the calculated value for ω, the equation becomes:

$I(s) = 0.3795\omega\,/\,(\omega^2+s^2)$.

The reverse-transformed equation is:

$i(t) = 0.3795 \sin 63.246t$. [Eq. 145]

The current oscillates in the circuit with an amplitude of nearly 4 V and with a frequency $f = w/2\pi$ = 10.07 Hz. The equation states that the oscillations continue indefinitely, but resistive and other losses in the circuit will eventually damp them out.

The circuit is a tuned circuit with a resonant frequency given by the equation

$f = 1/2\pi\sqrt{(LC)}$ = 10.07 Hz.

As in the other examples, we have been able to check the Laplace transform method by independent means.

Current transforms

If you are working with currents and wish to calculate voltages by applying KCL, **Table 2** supplies the required transforms. This table could have been used with the example of **Fig. 152**. Note that the capacitive and inductive transforms are the **duals** (see Part 5) of those in **Table 1**.

LCR circuit

The behaviour of the circuit in **Fig. 157a** is not readily predictable. **Figure 157b** is its transformed equivalent. Applying KVL:

$500I(s)+0.1sI(s)+10^8I(s)/s = 2/s$

$I(s)\,(500+0.1s+10^8/s) = 2s.$

Multiplying both sides by s:

$I(s)\,(500s+0.1s^2+10^8 = 2$

$I(s) = 2/(0.1s^2+500+10^8).$

Multiply top and bottom by 10 to get s^2 in the denominator:

$I(s) = 20/(s^2+5000s+10^9)$ [Eq. 146]

If we try to factorise the denominator, with the aim of finding a pair of fractions, we find that the denominator has imaginary roots. We must try another approach. One of the standard functions not included in last month's Box 1 is $e^{-at}\sin\omega t$, which has the transform:

$\omega / [(s+a)^2+\omega^2].$

Our plan is to remould the expression in **Eq. 146** into this form. The expansion of $(s+a)^2$ is $(s^2+2as+a^2)$. So the 5000 in **Eq. 146** corresponds to 2a in the expansion. In other words, $a = 2500$. Now $a^2 = 625\times10^4$, so the 10^9 in **Eq. 146** is made up of $625\times10^4+\omega^2$. This makes $\omega^2 = 10^9-625\times10^4 = 99375\times10^4$. From this we find that $\omega = 31524$.

The 20 in the numerator consists of ω multiplied by a factor. This factor is $20/\omega = 20/31524 = 6.34\times10^4$. We can now rewrite **Eq. 146** in the form of a standard transform:

$$I(s)=6.34\times10^4\times \times\frac{31524}{(s+2500)^2+31524^2}$$

The reverse transform yields:

$i(t) = e^{-2500t}\sin 31524t.$ [Eq. 147]

The graph of this equation (**Fig. 158**) is a sine wave with exponentially decaying amplitude. In comparison with the circuit of **Fig. 156**, the presence of the resistor means that the oscillations are damped in a few milliseconds. A similar circuit was analysed in Part 13 (**Fig. 115**). This is an underdamped circuit; with different component values, the circuit might be overdamped or critically damped. The Laplace transform will help us find the degree of damping.

Element	Current equation	Transformed current
Step current, $i(0^+)$	$i(t) = i(0^+)$	$I(s) = i(0^+)/s$
Resistance, R	$i(t) = u(t)/r$	$I(s) = U(s)/r$
Capacitance, C	$i(t) = C\mathrm{d}u(t)/\mathrm{d}t$	$I(s) = CsU(s)$
Inductance, L	$i(t) = (1/L)\int u(t)\,\mathrm{d}t$	$I(s) = U(s)/Ls$

Table 2. Circuit transforms (currents).

Initial and final states

If we are interested only in the initial and final currents or voltages, there is no need for the reverse transform. The initial and final value theorems require only the transformed equation. In the previous example:

$I(s) = 20/(s^2+5000s+10^9).$

$$\text{Initial value is } \lim_{s\to\infty} sI(s)$$

$$=\lim_{s\to\infty} 20s/(s^2+5000s+10^9).$$

The highest power of s is s^2, so divide throughout by s^2:

Initial value is

$$=\lim_{s\to\infty} 20s/(s^2+5000s+10^9)$$

$$=\lim_{s\to\infty} (20/s)/(1+5000/s+10^9/s^2).$$

As s becomes very large, $20/s$, $5000/s$ and $10^9/s^2$ approach zero: initial value is $0/(1+0+0) = 0$.

$$\text{Final value is } \lim_{s\to 0} sI(s)$$

$$=\lim_{s\to 0} 20s/(s^2+5000s+10^9)$$

$$=(20\times0)/(0^2+5000\times0+10^9)$$

$$=0/(0+0+10^9)=0.$$

The current is zero to begin with and finally dies away to zero.

Help at hand

We have tried to illustrate the various modelling techniques with examples in which the mathematics is not too involved. With circuits that are more complicated than those we have analysed here, the principles of analysis remain the same, but the calculations soon begin to spread over many pages. There is little new in such calculations; one just has to keep a clear head and avoid arithmetical errors. For those who do not relish too much maths, there are several circuit modelling or circuit simulation programs for personal computers. Next month, we conclude this series by looking at the subject of circuit simulations and using one of these programs to investigate the behaviour of circuits of varying degrees of complexity.

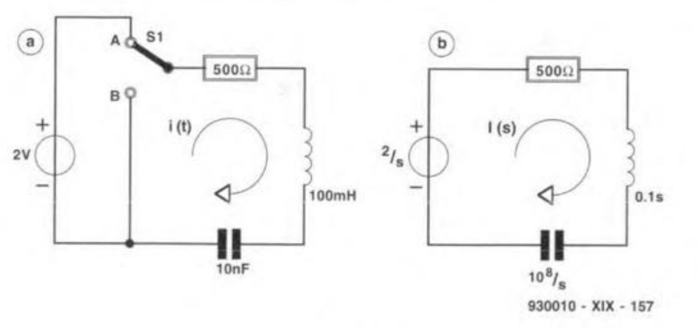


Fig. 157

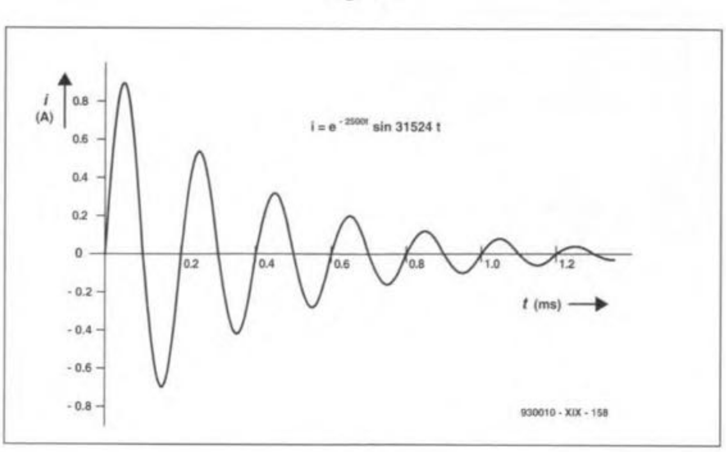


Fig. 158

Test yourself

1. Draw the transform of a circuit like **Fig. 152**, except that the resistance is 220 Ω and the capacitance is 180 µF. Find equations for $U(s)$ and $u(t)$, and calculate the initial and final voltages.

2. Given the circuit of **Fig. 157a**, but with the capacitor replaced by 10 µF, calculate the equation for current against time. Is this circuit underdamped, critically damped or overdamped? Hint: the transform used previously with this circuit is **not** used now.

Answers to Test yourself (Part 18)

1a $2/(s+6)$

1b $3s/(s^2+16)$

2a $1-e^{3t}$

2b $4\sin 7t$

3 $I(s) = 1/(s-2)-1/(s+3)$
$i(t) = e^{2t}-e^{-3t}$

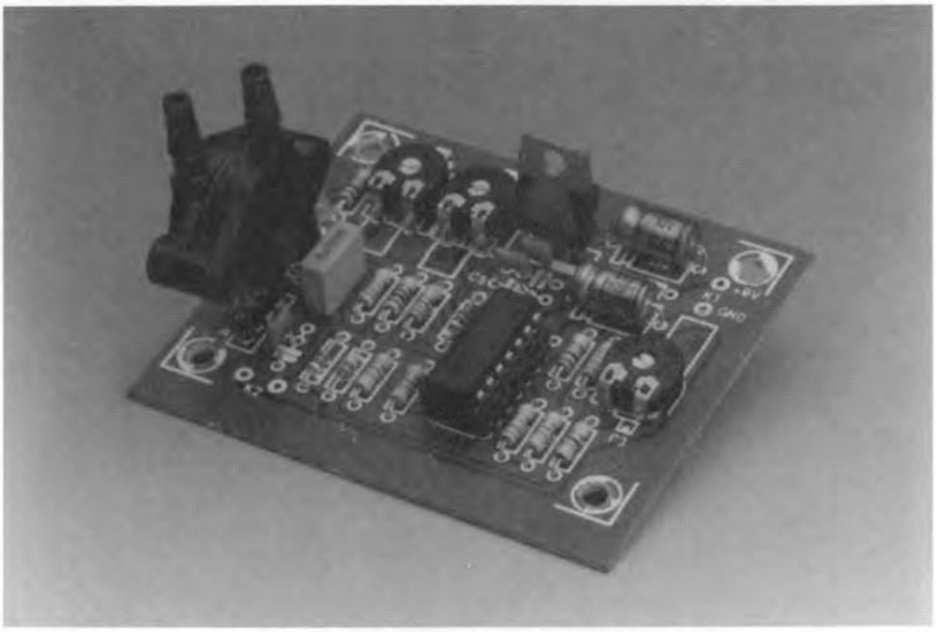
4a $2.5\mathrm{d}i(t)/\mathrm{d}t+20i(t) = 10-200t$

4b $I(s)=\dfrac{10s-200+1.25s^2}{s^2(2.5s+20)}$

4c $i(t) = 1.75-10t-1.25e^{-8t}$.

[930010-XIX]

ANEMOMETER AND FLOW METER



The meter described in this article is suitable as an anemometer for use by yachtsmen and surfers, and also as a flow meter for use in air, water or other liquids

Design by K. Bachun

Motorola's Type MPX10DP sensor enables the design and construction of a unit that is suitable for use not only as a small wind gauge without moving parts, but also a variety of applications that range from pressure monitor in an air conditioning system and measuring pollution (air resistance) in filters to level measurements in liquid containers. Since the unit can be fed from either a 9 V battery or a 12 V battery, application in motor vehicles is also possible.

Circuit description

Pressure sensor R_3 in **Fig. 1** is the heart of the circuit. Its operation is based on the piezoelectric effect of four resistors mounted on a silicon membrane and connected in a bridge network. When pressure is exerted on the membrane, a proportional voltage is developed across the bridge. **Figure 3** shows the construction of the sensor.

The supply to the sensor consists of a constant current provided by T_1. The base voltage of this transistor is held constant by D_5. The resulting emitter current, and thus collector current, is about 3 mA, so that the collector voltage is 2.5–3.5 V.

The temperature sensor has a negative temperature coefficient of $-19\%\ K^{-1}$, so that the output voltage drops with rising temperature. However, the current source has a positive temperature coefficient of about $+17\%\ K^{-1}$, which virtually negates the temperature dependence of the sensor. The remaining deviation of the sensor with temperature can be ignored for most applications.

The remainder of the circuit is formed by an instrument amplifier and an analogue display.

The inevitable offset error of the bridge is compensated by IC_{1a}, R_4, P_1 (fine control) and P_2 (coarse control). This arrangement ensures (provided the two preset potentiometers have been adjusted correctly) that, in the absence of any pressure difference, the voltage between the non-inverting inputs of IC_{1b} and IC_{1c} is zero.

To reduce the load on the sensor as much as possible, the input resistance of the instrument amplifier is as high as practicable. The voltage amplification is determined by R_5, R_6, R_{12} and P_3. The amplification (that is, the sensitivity of the amplifier) is set with P_3.

The symmetrical measurand (quantity measured) is converted into an asymmetric potential by IC_{1d}. The current flowing through moving-coil meter M_1 is thus directly proportional to the measured pressure difference.

Since the offset compensation works in only one direction, M_1 is given a bias voltage of about 0.1 V, which should be borne in mind if a different display is used.

Construction

The circuit is best built on the printed-circuit board shown in **Fig. 2**. Populating the board should be straightfor-

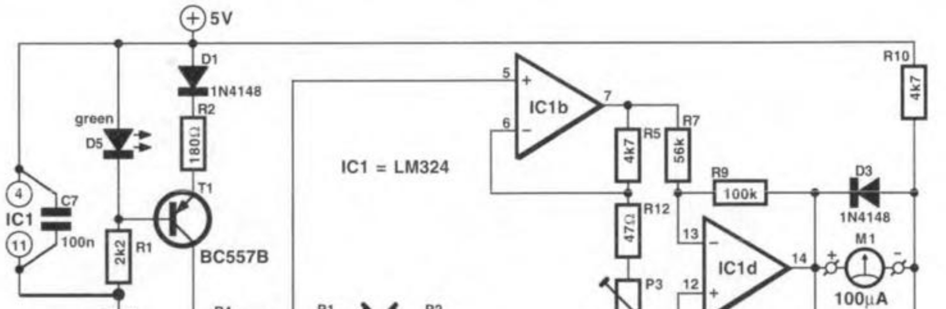


Fig. 1. Circuit diagram of the wind gauge and flow meter.

ward. Although normal types of preset may be used, it is advisable to use Cermet multiturn types. The board allows either type to be used.

The sensor is mounted with the aid of a 4-way PCB connector; the notch in the ground pin prevents its being inserted incorrectly.

The more sensitive Type MPX11DT (0.50 mV mbar^{-1}) sensor may also be used.

Since the potential of batteries drops relatively quickly in use, it is advisable, if a battery power supply is used, to replace the Type 7805 voltage regulator by low-drop Type 4805. This will ensure a longer battery life.

Each of the two nipples of the sensor may be inserted into a half of a table tennis ball cut into two as shown in **Fig. 5**. This is the most inexpensive way of providing dishes for intensifying and rarefying the pressure. Such dishes, therefore, also provide a means of determining the wind direction.

Figure 4 shows a number of other ways of using the pressure sensor: as flow meter in a tube (**4a** and **4b**); to measure filter performance (**4c**); and as level meter in a liquid container (**4d**).

Calibration

- Switch on the meter and let it warm up for about 1 minute.
- Set P_1 and P_2 to the centre of their travel.
- Adjust P_2 for approximate zero reading on M_1.
- Adjust P_1 for precise zero reading on M_1.

The sensitivity depends on the type of sensor used and on the amplification set with P_3. To obviate problems with the stability, it is best not to set the amplification too high. At the same time, the sensitivity depends also on the type of application. If the instrument is to be used only as an anemometer, expose the dishes in

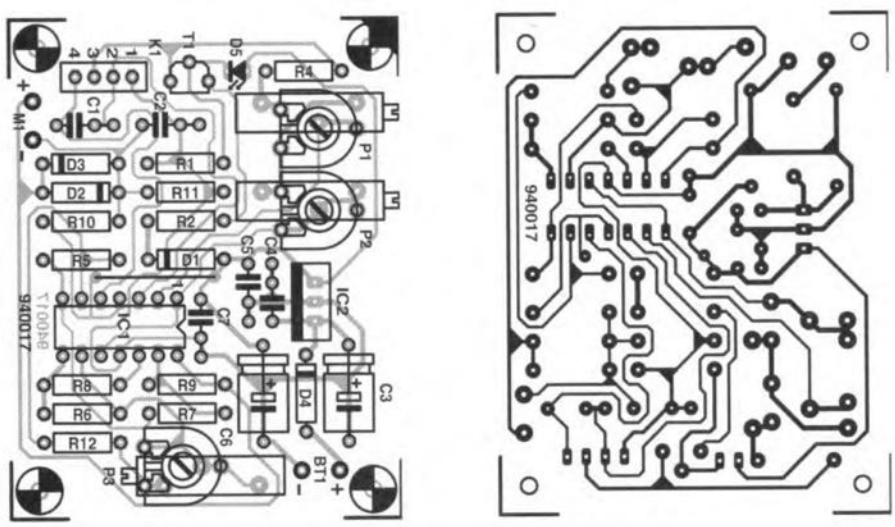
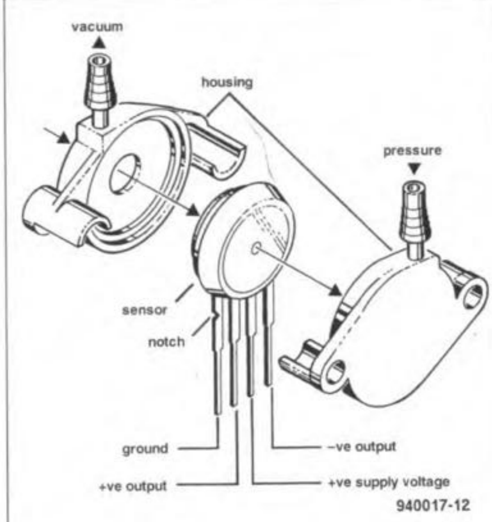


Fig. 2. Printed-circuit board for the wind gauge.



Measuring range	Max 100 mbar
Sensitivity	0.35 mV mbar^{-1}
Max overpressure	1000 mbar
Linearity	±1%
Offset	Min 0 mV
	Max 35 mV
Temperature range	–40 °C to
	+125 °C

Fig. 3. Construction of the Motorola Type MPX10DP sensor and some of its main parameters.

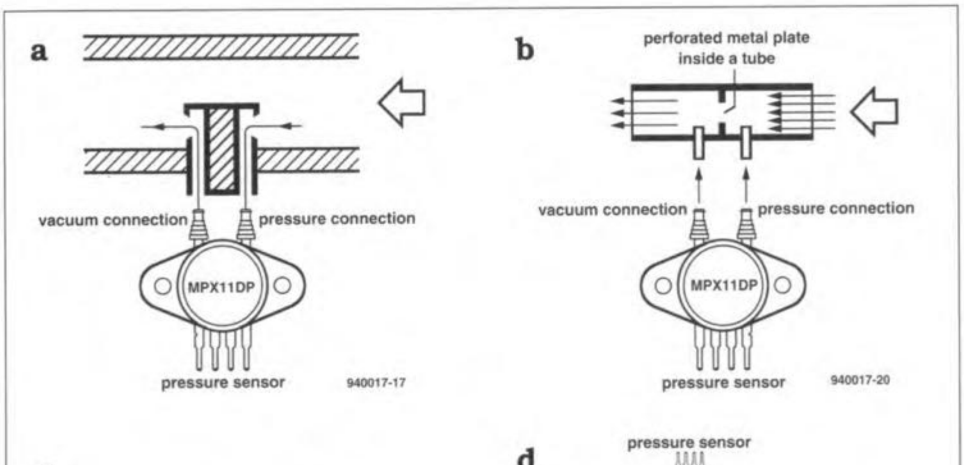


Fig. 4. Some other applications of the sensor.

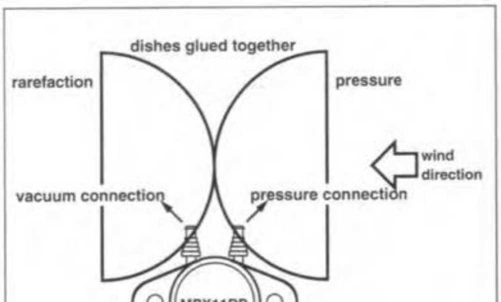


Fig. 5. Suitable dishes can be made from a table tennis ball cut into two.

Fig. 5 to a wind speed of 100 km h^{-1} and adjust P_3 for full-scale deflection (f.s.d.) of M_1. The wind speed can be simulated by driving a car at exactly 100 km h^{-1}, and have a passenger stick the instrument out of the window and adjust P_3 for f.s.d.

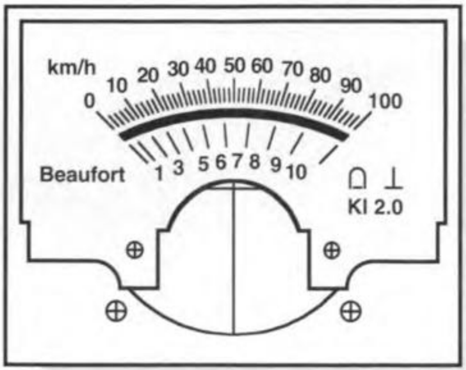


Fig. 6. This new scale for the meter correlates the Beaufort wind force and the corresponding speed in km h^{-1}.

Figure 6 shows a new scale for the meter correlating Beaufort force (up to 10) and wind speed in km h^{-1}. If a scale in miles per hour (knots) is desired, a correspondingly correlating scale can be made with the aid of the formula

$$V = (1.52B)^{3/2},$$

where V is the mean wind speed in miles per hour and B is the Beaufort wind force.

For other applications, it is, of course, necessary to produce an appropriate different scale.

Parts list

Resistors:
R_1 = 2.2 kΩ
R_2 = 180 Ω
R_3 = pressure sensor Type MPX10DP or MPX11DP (Motorola)
R_4, R_9 = 100 kΩ
R_5, R_6, R_{10} = 4.7 kΩ
R_7, R_8 = 56 kΩ
R_{11} = 100 Ω
R_{12} = 47 Ω

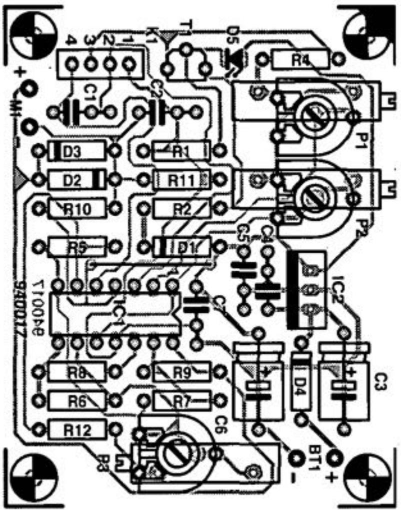
Capacitors:
C_1, C_4, C_5, C_7 = 100 nF
C_2 = 220 nF
C_3 = 100 μF, 10 V
C_6 = 47 μF, 10 V

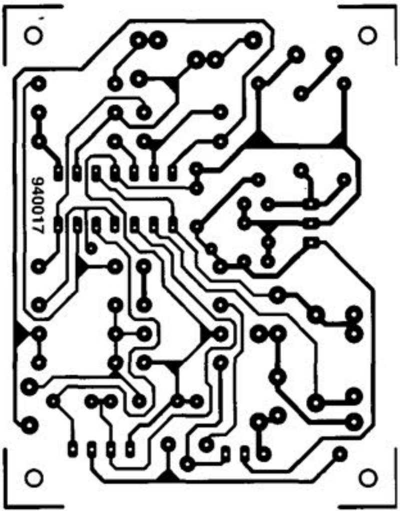
Semiconductors:
D_1–D_3 = 1N4148
D_4 = BAT85
D_5 = LED, green
T_1 = BC557B

Integrated circuits:
IC_1 = LM324
IC_2 = 7805 or 4805 – see text

Miscellaneous:
K_1 = 4-way socket for PCB mounting
M_1 = 100 μA moving-coil meter
Bt_1 = 9 V battery with clip
S_1 = single-pole, single-throw switch
PCB Ref. 940017 (see p.79)

[940017]


K1
T1
D5
R4
1
2
3
4
+
M1
C1
C2
P1
P2
D3
R1
D2
R11
R10
R2
R5
D1
C5
C4
IC2
IC1
940017
C7
R8
R9
C3
D4
R6
R7
C6
R12
R3
BT1
+
−


940017

PIC® PROGRAMMING COURSE

PART 2: IMPORTANT REGISTERS

The main subject of the this second part of the course is the processor hardware and the way it interacts with the software. In addition to the stack, a number of special registers including the prescaler, the watchdog timer and a real-time clock/counter are scrutinized.

By our editorial staff.

Source: Microchip Data Book, Microchip Technology Inc.

Stack registers

The devices in the PIC16C5x family have two-level hardware push/pop stack registers. A CALL instruction pushes the current program counter value, incremented by 1, into stack level 1. Next, stack level 1 is pushed to level 2. That allows the program to jump out of its main flow up to two times. If more calls are nested, the return addresses are lost. A RETLW instruction loads the contents of stack level 1 back into the program counter, while the contents of stack level 2 are copied into stack level 1. Stack level 2 retains its value.

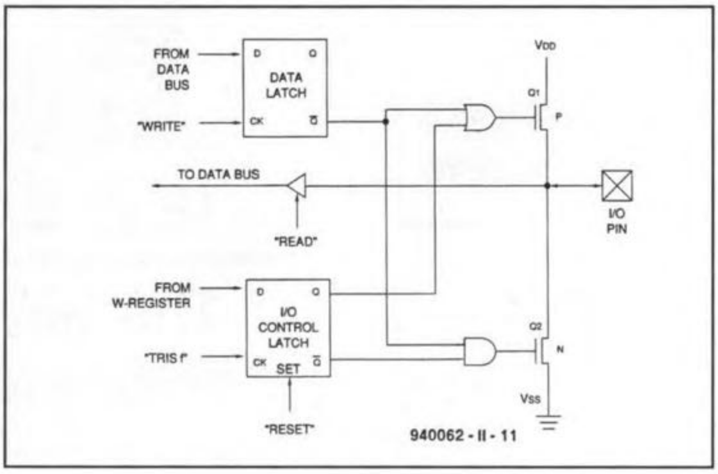


Fig. 1. Equivalent circuit for a single I/O pin.

I/O registers

The I/O registers (ports) are the interface between the software and external hardware devices. These registers can be written and read. A read instruction always reads the state of a pin, regardless of it being programmed as an input or an output. After a reset, all pins are defined as input (high-impedance mode). Next, the TRIS instruction may be used to set them up as outputs. A '0' in the TRIS register is necessary to assign the 'output' function to a particular I/O pin. As soon as a pin is defined as an output, it is switched to the level indicated by the associated register bit. Consequently, you are well advised to ensure that these bits are at a 'safe' level before they are switched to the 'output' function. The equivalent circuit of an I/O pin is shown in **Fig. 1**.

Port A is programmed via register f5. Only the four low-order bits, RA0-RA3, are used. Bits RA4-RA7 are not implemented, and read as zeros. To ensure that application software remains compatible with later, more advanced versions of the controller, these bits must be treated as undefined.

Ports B and C are programmed via registers f6 and f7 respectively. Because the PIC16C54 and PIC16C56 do not have a port C, f7 may be used as a general purpose register in those controllers.

General purpose registers

The organisation of the addresses in the data registers is shown in **Table 1**.

The desired register bank is selected using bits 5 and 6 in the FSR register. Since the instructions have up to five bits available for addressing, the register address has to be converted. The information needed to do so may be obtained from **Table 1**. For instance, to address register f5D, bit 6 in the FSR has to be set. At the same time, bit 5 is de-actuated. Next, address 1DH is included with the instruction.

Special purpose registers

W register
The working register, W, may be compared (functionally) with the accumulator as used in most microprocessors. A special feature of the W register in PIC16C5x processors is that it need not always be the 'target' destination of an operation in the ALU.

TRIS registers
The TRIS ('TRI-State™') registers are used to configure the I/O pins. Each pin is represented by a configuration bit in the respective TRIS register. A '1' in the relevant bit position causes the pin to function as an output. The TRIS instruction has a write function only. After a reset, all bits are '1' so that all port pins behave as inputs.

OPTION register
This register is used to program the prescaler of the WDT (watch dog timer) or that of the RTCC (real time clock/counter). The option register has a width of 6 bits, and is 'write-only' using the OPTION instruction. The functions of the individual bits may be found in **Table 2**. Here, too, all bits are at '1' after a reset.

Watchdog timer

The watchdog timer (WDT) is a special counter which is programmed to reset a program when an overflow occurs in the counter ('watchdog time-out'). As long as a program runs normally, the software resets the counter before it can produce an overflow. If the program crashes, the counter is no longer reset. Consequently, the WDT produces an overflow, and resets the program, causing it to start from the beginning. The WDT is very useful in automated control systems where faults can create criti-

Table 1. Register addresses and organisation

PIC 16C54/C55/C56	
$f08_{HEX}$ to $f1F_{HEX}$	general-purpose register files
PIC 16C57	
$f08_{HEX}$ to $f0F_{HEX}$	general-purpose register files
$f10_{HEX}$ to $f1F_{HEX}$	general-purpose register files in memory bank 0
$f20_{HEX}$ to $f2F_{HEX}$	identical to $f00_{HEX}$ to $f0F_{HEX}$
$f30_{HEX}$ to $f3F_{HEX}$	general-purpose register files in memory bank 1
$f40_{HEX}$ to $f4F_{HEX}$	identical to $f00_{HEX}$ to $f0F_{HEX}$
$f50_{HEX}$ to $f5F_{HEX}$	general-purpose register files in memory bank 2
$f60_{HEX}$ to $f6F_{HEX}$	identical to $f00_{HEX}$ to $f0F_{HEX}$
$f70_{HEX}$ to $f7F_{HEX}$	general-purpose register files in memory bank 3

Register	FSR		Address
	bit 6	bit 7	
$f10_{HEX}$ to $f1F_{HEX}$	0	0	10_{HEX} to $1F_{HEX}$
$f30_{HEX}$ to $f3F_{HEX}$	0	1	10_{HEX} to $1F_{HEX}$
$f50_{HEX}$ to $f5F_{HEX}$	1	0	10_{HEX} to $1F_{HEX}$
$f70_{HEX}$ to $f7F_{HEX}$	1	1	10_{HEX} to $1F_{HEX}$

cal situations. Obviously, the programmer must ensure that the WDT is reset at a fixed location in the program. The WDT may not be reset in an interrupt routine.

The time which elapses before the counter produces an overflow lies between 9 and 30 ms. The counter clock pulses are supplied by an internal *RC* oscillator. If a longer time is required, the internal prescaler may be used. The prescale value may be programmed via the OPTION register.

The WDT is enabled via a bit in the OPTION register. Once enabled, the WDT can not be disabled by any instruction. The only way to stop the timer is to remove the system clock. The WDT and the associated prescaler can only be reset by the CLRWDT and SLEEP instructions.

Dual-purpose prescaler

The PIC processor has an 8-bit counter which functions as a prescaler that can be assigned to the WDT or the RTCC by programming the PSA bit in the OPTION register. In fact, the prescaler is a post-scaler for the watchdog timer. For the sake of simplicity, however, it will be referred to as 'prescaler' in the discussion below.

The prescale value is programmed via bits PS0, PS1 and PS2 which are also found in the OPTION register. If the prescaler is assigned to the real-time clock, it is cleared at each write action to the clock. If the prescaler is coupled to the watchdog timer, it is cleared by the CLRWDT instruction. Since the function of the counter is determined by software, the configuration may be changed at any time during program execution. The WDT/RTCC assignment is mutually exclusive. To prevent an unintended device reset, the instruction sequence listed in **Program 1** is suggested by Microchip when changing the prescaler assignment from RTCC to WDT.

The first two steps are required only if an external RTCC clock source is used. The last two steps are necessary only if the desired prescale value is 000 or 001.

To change the prescaler from WDT to RTCC, the sequence in **Program 2** is recommended.

RTCC

The real-time clock/counter (RTCC) in the PIC controller is an 8-bit wide counter whose contents ('state') can be read and modified at any time. The RTCC is basically an ordinary memory location. As soon as the maximum state, FFH, is reached, the counter 'wraps around' to 00H. The counter has two possible clock sources: (1) an internal signal, which is a quarter of the crystal frequency (fosc/4), or (2) an external clock signal which is applied to the RTCC input. When external clocking is used, the signal edge which increments the clock is selectable via software.

The prescaler can be 'connected' ahead of the counter to prevent the latter from wrapping around after 256 clock pulses already. Remember, the prescaler is assigned either to the WDT **or** the RTCC, **not** to both counters at the same time. The prescale value is programmed via the OPTION register.

When the RTCC is used, programmers should take into account that clock pulses are delayed two machine cycles by the synchronisation unit. Consequently, a write instruction to the

Program 1
Changing prescaler from clock/counter (RTCC) to watchdog timer (WDT) in real time.

```
MOVLW xx0x0xxxB    ;Select internal clock and new prescale value.
                   ;
OPTION             ;If new value is 000 or 001, then select
                   ;any other prescale factor temporarily.
CLRF 1             ;Clear RTCC and prescaler.
MOVLW xxxx1xxxB    ;Select WDT, do not change prescale value.
OPTION             ;
CLRWDT             ;Clears prescaler and WDT.
MOVLW xxxx1xxxB    ;Select new prescale factor.
OPTION             ;
CLRWDT             ;Clears WDT and prescaler.
MOVLW xxxx1xxxB    ;Select new prescale value
OPTION             ;
```

Program 2
Changing prescaler from WDT to RTCC.

```
CLRWDT             ;Clears WDT and prescaler.
MOVLW xxxx0xxxB    ;Select RTCC, new prescale value and clock
                   ;source
OPTION             ;
```

PIC16C5x PROGRAMMING

This short course is aimed at providing an introduction into programming and hardware aspects of the PIC16C5x family of microcontrollers manufactured by Microchip Technology Inc.

An **assembler** will be offered on disk later in the course. This assembler is distributed with the permission of Microchip tecnology Inc., and supports the PIC16C5x and PIC16Cxx series of controllers. It offers a full featured macro and conditional assembly capacity. It can also generate various object code formats including several hex formats to support Microchip's proprietary development tools as well as third party tools. Also supported are hex (default), decimal and octal source and listing formats. An assembler users manual is available from Microchip Technology distributors for detailed support.
The disk, which will be distributed via the *Elektor Electronics* Readers' Services, will also contain a software **simulator**.

The files produced by the assembler can be downloaded to the **PIC programmer** described in *Elektor Electronics* March 1994.

JOIN THE COURSE!

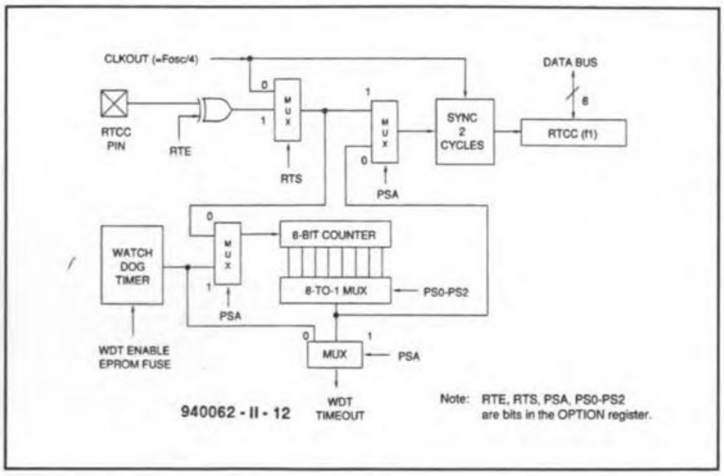


Fig. 2. Block diagram of the prescaler (an 8-bit counter) which can be connected 'ahead of' the RTCC, or 'behind' the watchdog timer.

Table 2. OPTION register organisation.

OPTION Register (reset value: 111111b)					
bit 5	**bit 4**	**bit 3**	**bit 2**	**bit 1**	**bit 0**
RTS	RTE	PSA	PS2	PS1	PS0
PS2	**PS1**	**PS0**	**RTCC**	**WDT**	
0	0	0	1:2	1:1	
0	0	1	1:4	1:2	
0	1	0	1:8	1:4	
0	1	1	1:16	1:8	
1	0	0	1:32	1:16	
1	0	1	1:64	1:32	
1	0	0	1:128	1:64	
1	1	1	1:256	1:128	
PSA	**prescaler assignment bit**				
0	real-time clock/counter (RTCC)				
1	watchdog timer (WDT)				
RTE	**RTTC signal edge**				
0	increment on rising (low-to-high) edge at RTCC pin				
1	increment on falling (high-to-low) edge at RTCC pin				
RTS	**RTCC signal source**				
0	Instruction Cycle Clock (CLKOUT= $f_{osc}/4$)				
1	transition at RTCC pin				

RTCC register is not actually recognized by the program until after two instruction cycles plus one clock pulse. If the prescaler is used, it is essential to know that the synchronisation unit is 'connected' behind the prescaler. If the RTCC is clocked internally, it is still necessary to keep the RTCC input pin at a fixed level.

A clock signal applied to the RTCC pin should comply with the following conditions:

Prescaler not used:
RTCC high: $\geq 2t_{osc} + 20$ ns
RTCC low: $\geq 2t_{osc} + 20$ ns

Prescaler used:
RTCC period: $\geq (4t_{osc} + 40 \text{ ns})/N$
RTCC high: ≥ 10 ns
RTCC low: ≥ 10 ns

where t_{osc} is the period of the oscillator signal, and N the prescale value.

The combined block diagram of the watchdog timer, the prescaler and the real-time clock/counter is shown in **Fig. 2**. The different options which may be programmed for these units (using the OPTION register) are easily recognized.

Power Down Mode (SLEEP)

The power-down mode is entered by executing a SLEEP instruction. The instruction resets the watchdog timer (if it was enabled), clears the PD bit in the status register f3, actuates the TO bit, and stops the clock oscillator. The I/O ports maintain the status they had before the SLEEP command was executed. For lowest current consumption in this mode, all I/O pins, and the RTCC pin, should be at a fixed logic level, i.e., at V_{dd} or V_{ss}. When the clock oscillator is stopped, the RTCC is stopped also.

The device can be awakened by a watchdog time-out, or by a positive-going edge at the end of a 'low' level applied to the $\overline{\text{MCLR}}$ input. In both cases, the PIC processor will go through one oscillator start-up timer period before the program is started at from the reset address.

The PD bit in the status register may be used to determine if the processor was powered up, or awakened from power-down mode. Similarly, the TO bit tells you if the 'wake-up' was caused by an external $\overline{\text{MCLR}}$ signal or by a watchdog time-out. More information about these features may be found in **Tables 1** and **2**.

Reset

There are three ways to reset the PIC processor: (1) a low level at the $\overline{\text{MCLR}}$ input; (2) a power-up; or (3) a watchdog time-out. The controller remains in the reset state during the oscillator startup time (OST), and/or as long as the $\overline{\text{MCLR}}$ input is logic low. The OST starts on the positive-going edge of the $\overline{\text{MCLR}}$ signal. If the $\overline{\text{MCLR}}$ input is connected to the positive supply line, the OST starts the instant the supply voltage is switched on. The OST has a duration of 9 ms to 30 ms.

During a reset condition, the state of the PIC controller is defined as:

- the oscillator is running, or being started (power-up or wake-up from SLEEP);
- all I/O port pins are switched to high-impedance mode by setting the TRIS registers to 'all ones' (= input mode);
- the program counter is set to 'all ones', i.e., 01FFH in PIC16C54/55, 03FFH in PIC16C56, 07FFH in PIC16C57;
- the OPTION register is set to 'all ones';
- the watchdog timer and its prescaler are cleared;
- the upper three bits of the status register (f3) are cleared to zero;
- for RC type devices only: the CLKOUT signal on the OSC2 pin is made low.

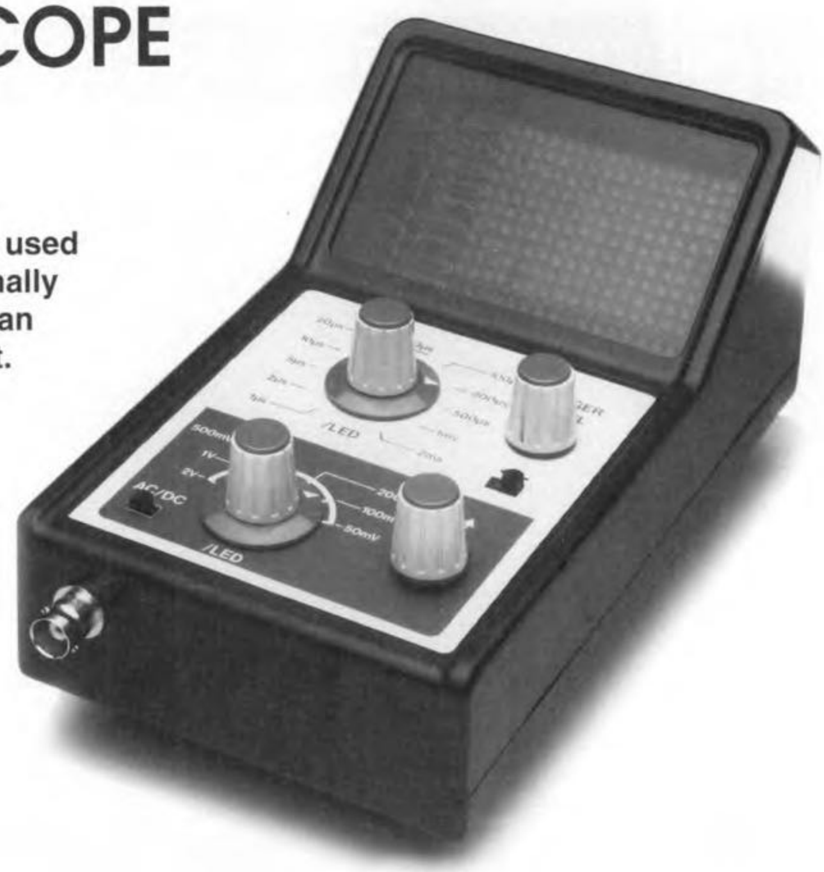
(940062-2)

Continued in the October 1994 issue.

LED OSCILLOSCOPE

Design by A. Rietjens

This inexpensive oscilloscope can be used at frequencies up to 20–30 kHz (optionally up to 1 MHz). It works from 12 V and can thus be used as a portable instrument.



The cost of the oscilloscope is kept down by the use of 3-mm LEDs instead of a cathode ray tube (CRT). These standard LEDs can be bought (in quantity) for between 5p and 8p each (in UK).

The LED matrix that produces the image is shown in **Fig. 1**. The horizontal grid lines represent the *X*-axis (or time base) of the oscilloscope and the vertical ones, the *Y*-axis. A given waveform, say, a sine wave, is shown by making a number of LEDs light in succession. The inertia of the human eye makes it appear as if all the relevant LEDs light.

The lighting of the LEDs is controlled by transistors T_1–T_{10}, buffers (line drivers) IC_1 and IC_2, and the circuit shown in **Fig. 2**.

If D_{151} is required to light, T_1 must conduct and pin 11 of IC_2 must go low. The anode of the diode is then supplied with +10 V via the transistor, while its cathode is earthed via IC_2. Series resistor R_1 limits the current through the diode. The circuits in **Fig. 2** ensure that only one transistor can conduct, and that only one of the outputs of IC_1 and IC_2 can become low, at any one time. It is thus impossible for more than one LED to light at any one time.

One of transistors T_1–T_{10} is made to conduct by making the relevant base low, while one of the outputs of IC_1, IC_2 is made low by making the relevant input high. The control voltages for this are pro-

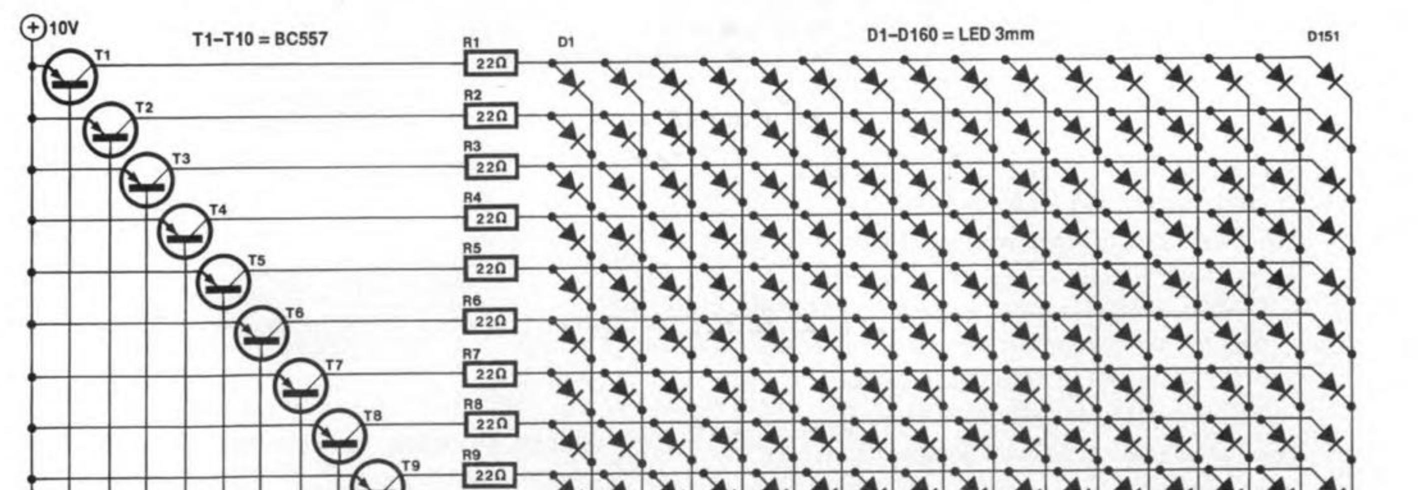


Fig. 1. The trace of the oscilloscope is produced by a matrix of LEDs of which only one is on at any one time.

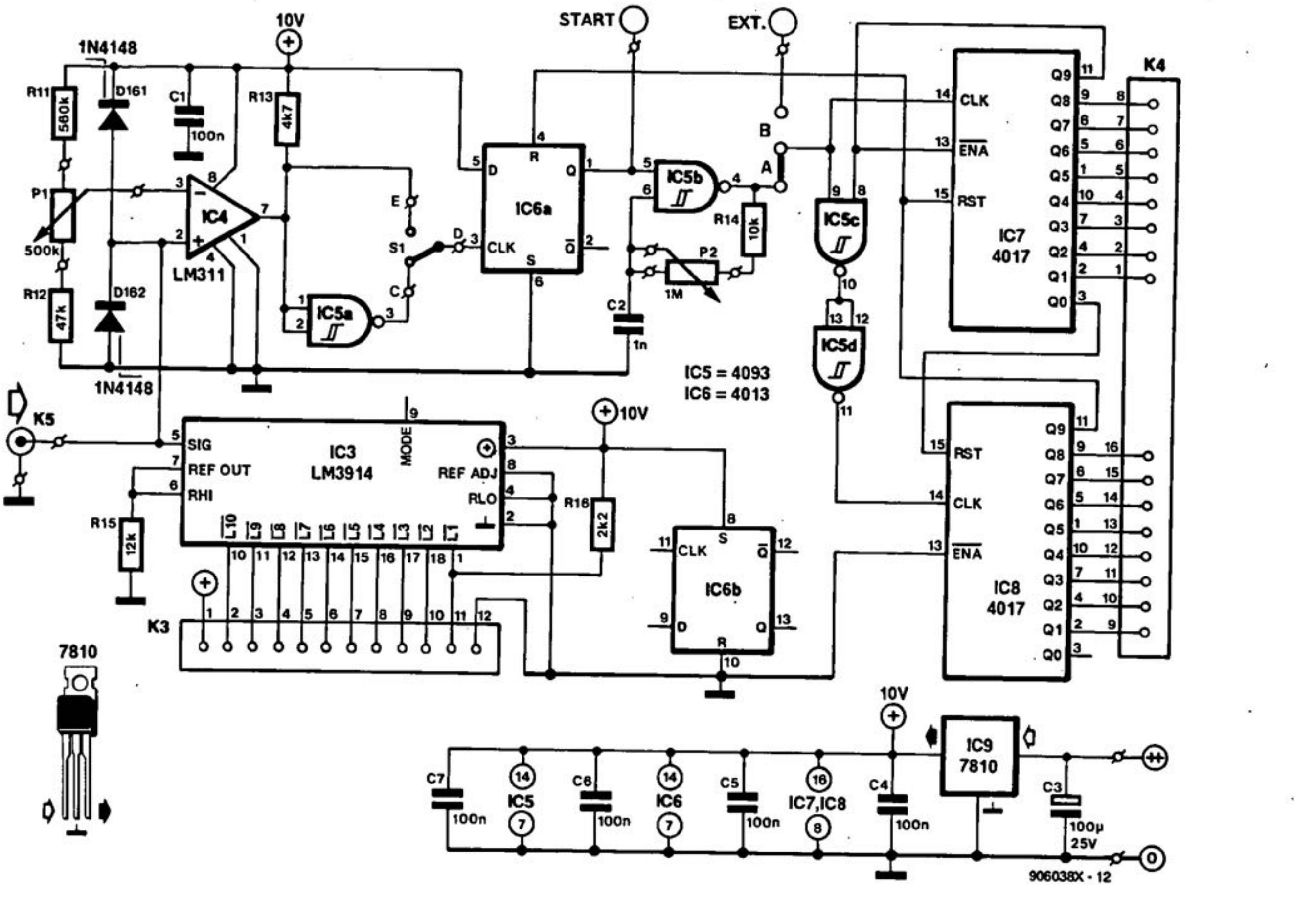


Fig. 2. Drive circuit for the LED matrix.

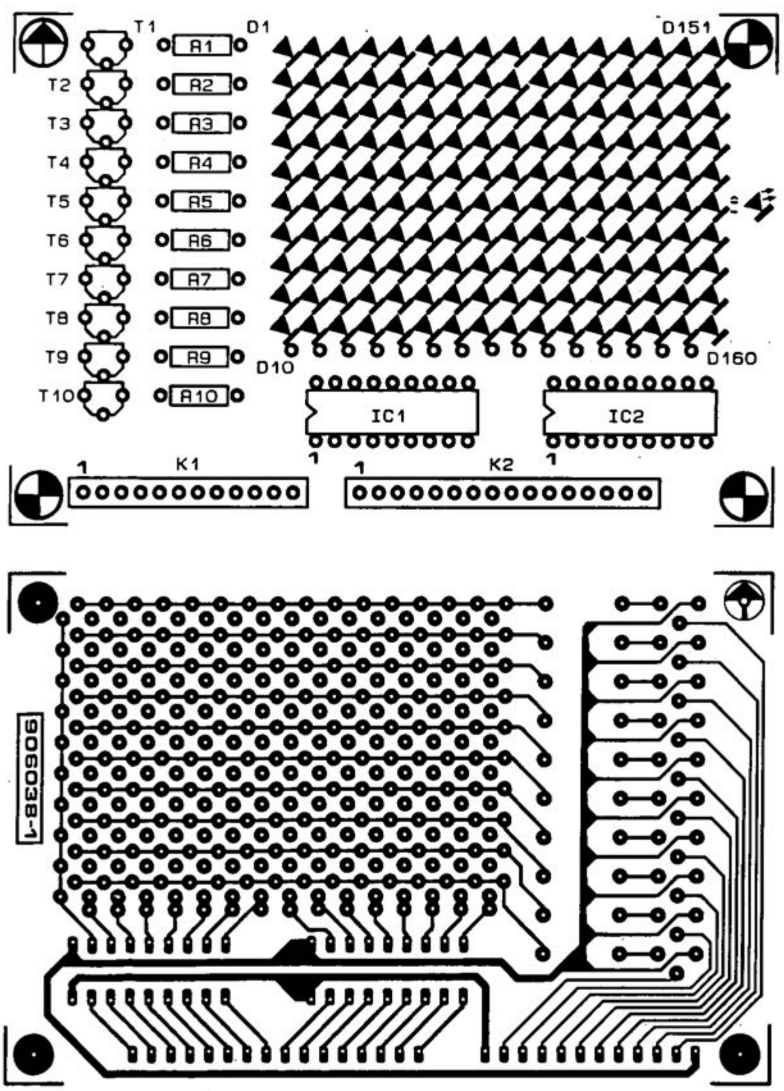


Fig. 3. Printed circuit for the display board (not available ready made).

vided by the circuit in **Fig. 2** via connectors K_1 and K_2.

Y-axis

As already stated, the amplitude of the input signal (the measurand) is represented by a vertical 'deflection'. Thus, for a large measurand, one of the upper row of LEDs must light, for a smaller signal, one of a lower row of diodes. This is arranged by analogue-to-digital (A-D) converter IC_3 in **Fig. 2**.

When a slowly increasing signal is applied to the SIG input (pin 5) of IC_3, outputs L_1–L_{10} of this IC go low sequentially, one at a time. This means that when the signal level at pin 5 is low, L_1 is low, when the input is high (1.25 V), L_{10} is low, and when the level is somewhere between these values, say, 0.6 V, L_5 or L_6 is low.

Since the input level must be 0–1.25 V, a signal amplifier/attenuator is required prior to K_5 and this will be described a little later on in the article.

X-axis

The horizontal trace is provided by (decade) counters IC_7 and IC_8 in **Fig. 2**. Control of the matrix LEDs is from outputs $\overline{Q_1}$–$\overline{Q_8}$ of these ICs via IC_1 and IC_2. The counters are in series and connected, with the aid of IC_{5c} and IC_{5d}, to form a 16-counter.

To make the Q-outputs high, one by one, starting with Q_1 of IC_7 and ending with Q_8 of IC_8, clock pulses must be applied to the CLK inputs of both ICs. To ensure that the process proceeds as indicated, the clock to IC_8 is interrupted by IC_{5c}-IC_{5d} when one of the Q-outputs of IC_7 is high. Only when pin 8 of IC_{5c} is high, are clock pulses applied to IC_8; they are applied to IC_7 at all times.

Because Q_9 of IC_7 is connected to its own ENA(ble) input and to pin 8 of IC_{5c}, as soon as this output is high, IC_7 is disabled and clock pulses are applied to IC_8. When Q_8 of IC_8 becomes high, IC_7 is reset, whereupon its Q_0 goes high, which causes IC_8 to be reset also. This is shown on the matrix by the LEDs lighting, one by one, from left to right.

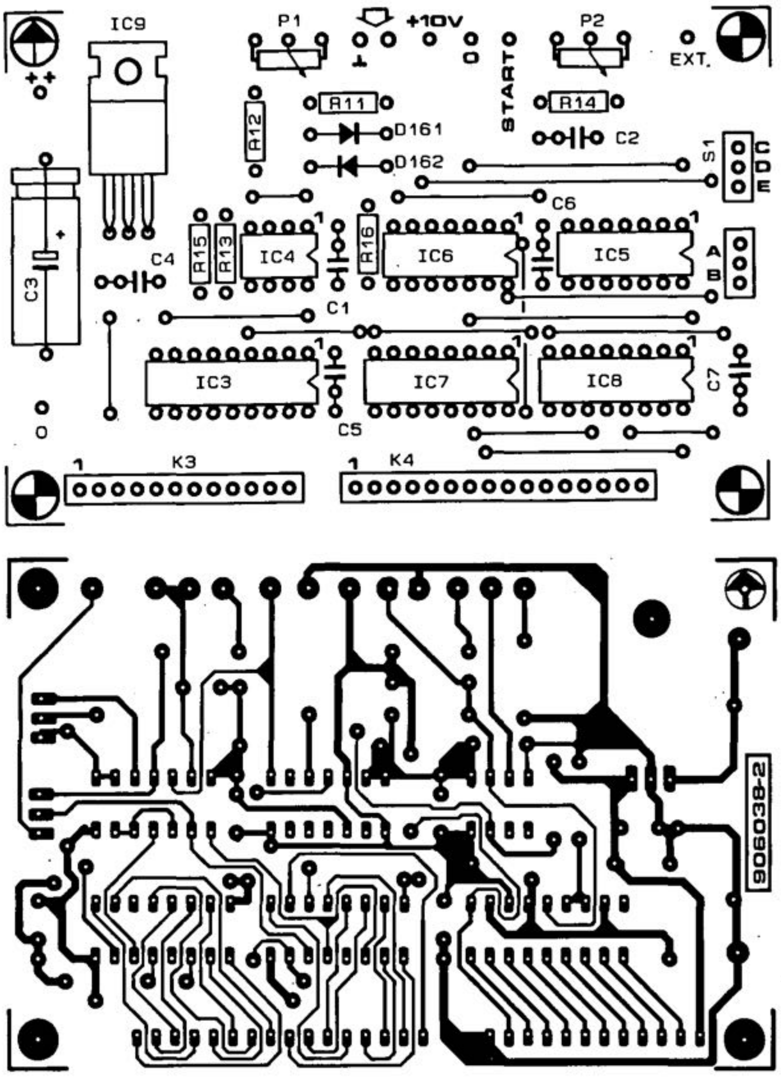


Fig. 4. Printed circuit for the driver board (not available ready made).

Clock

The clock pulses for IC_7 and IC_8 are provided by rectangular-wave generator IC_{5b}. The generator is enabled only when its pin 5 is high. Its operation depends on C_2 being charged (when pin 4 is high) and discharged (when pin 4 is low) alternately via R_{14}-P_2. Since the gate is arranged as an inverter, pin 4 is high when pin 6 is low and low when pin 6 is high.

When the level of the potential across C_2 approaches that at pin 4, the output of the generator instantly changes state. This means that if the capacitor was being charged (discharged) just prior to the change of state, it will be discharged (charged) after the change. Since this is a continuous action, the output of the generator consists of a train of rectangular pulses whose repetition rate (frequency) depends on the setting of P_2.

Trigger circuit

When the clock generator runs continuously, a still trace is obtained only if the frequency of the input signal is the same as that at which horizontal deflection takes place, or a whole multiple of it.

Since the input frequency is better not, or cannot be, altered, the horizontal deflection frequency must be variable. In **Fig. 2**, this is made possible by P_2. None the less, because of changes in temperature and other factors, one or both of the frequencies will shift to some extent. This will result in a trace that constantly moves across the screen.

It is thus necessary to ensure that the input frequency and the deflection frequency are synchronized at all times. As is normal, this is achieved by triggering the time base with the input signal. To that end, part of the input signal is applied to pin 2 of IC_4 (the trigger circuit consists of IC_4, IC_5, and IC_{6a}).

To ensure that the input levels of IC_3 and IC_4 are not exceeded, the signal input is provided with a voltage limiter consisting of series-connected diodes D_{161} and D_{162}. The supply voltage exists across this series network, but no current flows through this, because the diodes are inverse-biased. When the level at the input exceeds 10.6 V, D_{161} begins to conduct; when the input drops below –0.6 V, D_{162} begins to conduct. This means that the diodes limit the range of input levels to between 0.6 V below earth potential and 0.6 V above the supply voltage.

Circuit IC_4 is arranged as a comparator, that is, there is no feedback resistor between its output (pin 7) and its inverting input (pin 3). This means that, in theory, the amplification of the opamp is infinite, so that the level of its output voltage can only be 0 V (earth potential) or 10 V (supply voltage). The output cannot be set to any intermediate value. In other words, if the level of the signal at pin 2 rises above a certain value, the output instantly becomes 10 V; if the level drops below that certain value, the output immediately becomes 0 V. The value at which this happens, the change-over or toggle point, is set with P_2.

Since the output of IC_4 is thus either 10 V or 0 V, even for very small input signals, any input is converted into a rectangular voltage, whose frequency is identical to that of the input signal. This means that the output is synchronous with the input signal.

The output of IC_4 is used to trigger clock generator IC_{5b} at exactly the right moment via D-bistable IC_{6a}. When IC_4 applies a clock pulse to IC_{6a}, the Q-output of this stage goes high, whereupon IC_{5b} is enabled. Switch S_1, in conjunction with inverter IC_{5a}, determines whether the clock is enabled at the leading or at the trailing edge of the input signal. The exact instant of onset of the trace is set with preset P_1.

When decade counters IC_7 and IC_8 reach their maximum counter state, they are reset to their starting state via Q_9 of IC_8; IC_6 is then also reset, so that the clock generator is disabled. The clock is reenabled at the next output of IC_4. If, therefore, there is no input signal, the LEDs are out, because there is no trigger.

Construction

The oscilloscope is constructed on two printed-circuit boards, one for the display (**Fig. 3** and **5**) and the other for the driver (**Fig. 4**, **6**, and **7**).

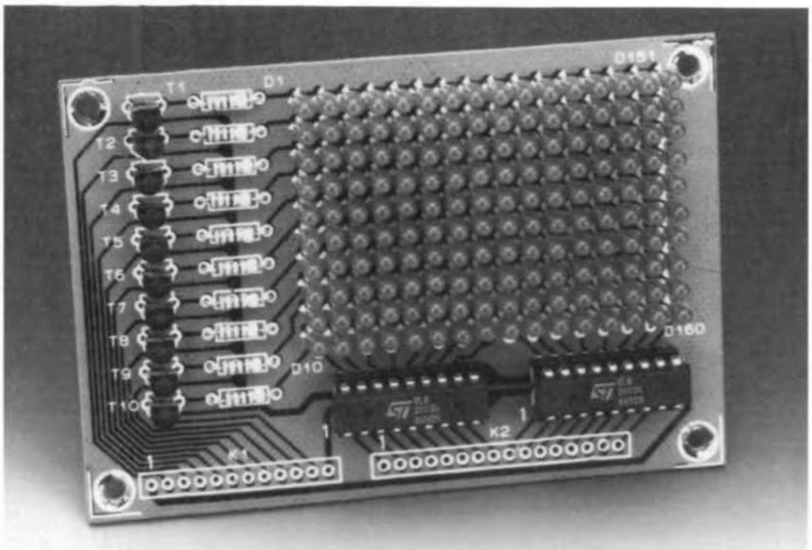


Fig. 5. Completed display board (component side).

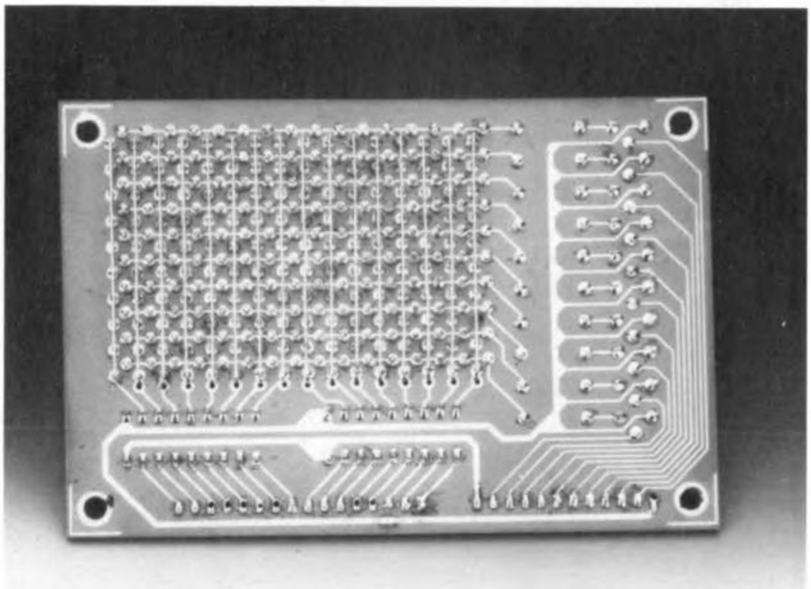


Fig. 6. Completed display board (track side).

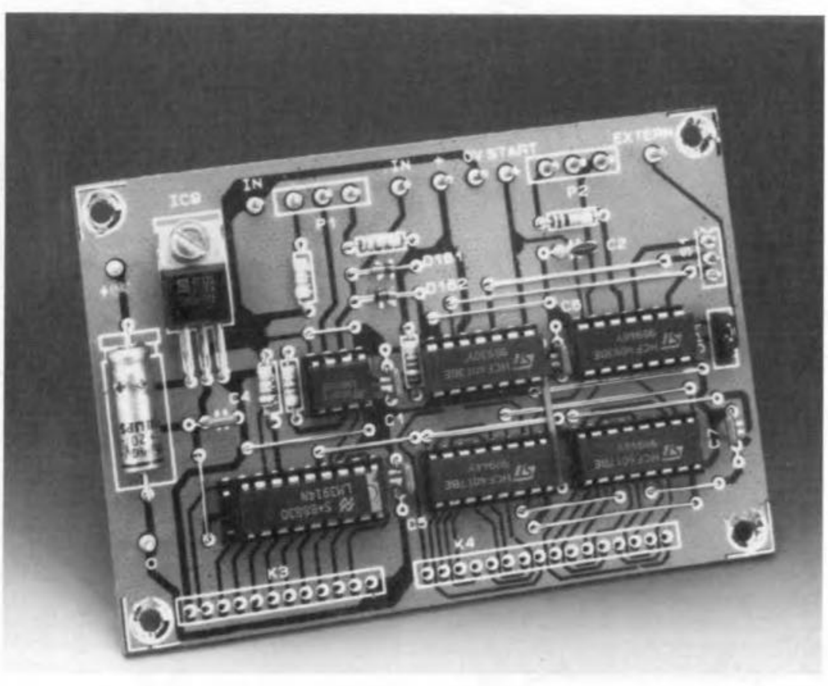


Fig. 7. Completed driver board (component side).

The cathodes of each row of LEDs must be linked via a suitable length of wire which remains suspended above the track side of the board and is connected to the relevant pin of IC_1 and IC_2—see **Fig. 1**. The anodes are soldered directly to the board.

Finishing the boards is straightforward. Since the two boards are the same size, they may be assembled into a 'sandwich' with the aid of suitable spacers.

Optional attenuator/amplifier

The single input voltage range of the oscilloscope may be extended to six ranges with an optional attenuator/amplifier. Both the attenuation and amplification factors are well defined to ensure accurate measurements. Furthermore, both are as nearly frequency-independent as possible. The unit described is usable at frequencies up to 40 kHz.

The circuit of the unit is shown in **Fig. 8**: the (variable) attenuator is formed by switch S_3 and resistors R_{17}–R_{24}. The remainder of the circuit functions as a differential amplifier.

With S_2 open, C_8 prevents any direct voltage at the input entering the circuit. This is useful if, for instance, the ripple on a supply voltage is to be measured.

The values of resistors R_{17}–R_{24} have been chosen to provide six ranges of sensitivity as shown in **Table 1**.

Operational amplifiers IC_{10} and IC_{11} form a differential ×4 amplifier. Since they are supplied with a positive voltage only and the input signal can be negative, the input levels are shifted. This is achieved by linking the earthy input connection not to supply earth, but to a point halfway between the supply earth and +10 V, that is, to +5 V (since R_{25} and R_{26} have identical values). This means that the input voltage alternates around +5 V.

Operational amplifiers IC_{10a} and IC_{10b} are protected against too high inputs by diodes D_{163}–D_{166}. If the level at the 'earthy' input pin rises above +10 V, D_{163} conducts, whereas if it drops below 0 V, D_{164} conducts. Diodes D_{165} and D_{166} function similarly when the level at the upper input terminal rises above 10 V or drops below 0 V respectively.

The potential difference across the input pins is amplified at the outputs of

Position S_3	Sensitivity (volt per LED)
1	2
2	1
3	0.5
4	0.2
5	0.1
6	0.05

Table 1.

IC_{10a} and IC_{10b}, from where it is applied to the –ve and +ve inputs of IC_{11}. This circuit sums the two voltages which are then applied to the input of the oscilloscope.

The level of the d.c. component at the output (pin 6) of IC_{11} is set with P_3, which means that the waveform on the oscilloscope can be shifted vertically with P_3.

Since the level at the input of the oscilloscope must not exceed +1.2 V, diodes D_{167} and D_{168} conduct when the output of IC_{11} rises above 1.2 V. Note that the output cannot become negative, because the supply of the opamp is 0 (earth) to +10 V.

Construction. The unit is best built on the printed-circuit board shown in **Fig. 9** and **10**: note, however, that this is not available ready made. Switch S_3 and potentiometer P_3 must be soldered directly to the board.

Since the 'apparent earth' of the input terminals floats 5 V above real earth, only the input terminals are accessible from outside, the real earth is not.

The unit is best tested in conjunction with the oscilloscope. Apply a variable alternating voltage to the input of the attenuator/amplifier and check that six different sensitivities can be selected with S_3. Rotating P_3 should shift the trace on the oscilloscope vertically.

More accurate time base

The on-board time base of the oscilloscope is intended for applications where the accent is more on viewing waveforms than on accurate measurements. If, however, not only the waveform, but also the frequency (or period) of a certain signal is to be assessed, a more accurate time base is required. In an oscilloscope, an unknown frequency is measured by comparing it with a known frequency: that of the time base generator. It is clear that the time base frequency must be accurate if it is to function as a reference. In the present time base a fixed 1 MHz generator is used, which is followed by a variable divider, so that not only 1 μs pulses, but also longer ones, are available.

The time base circuit is shown in **Fig. 11**. The 1 MHz clock is formed by

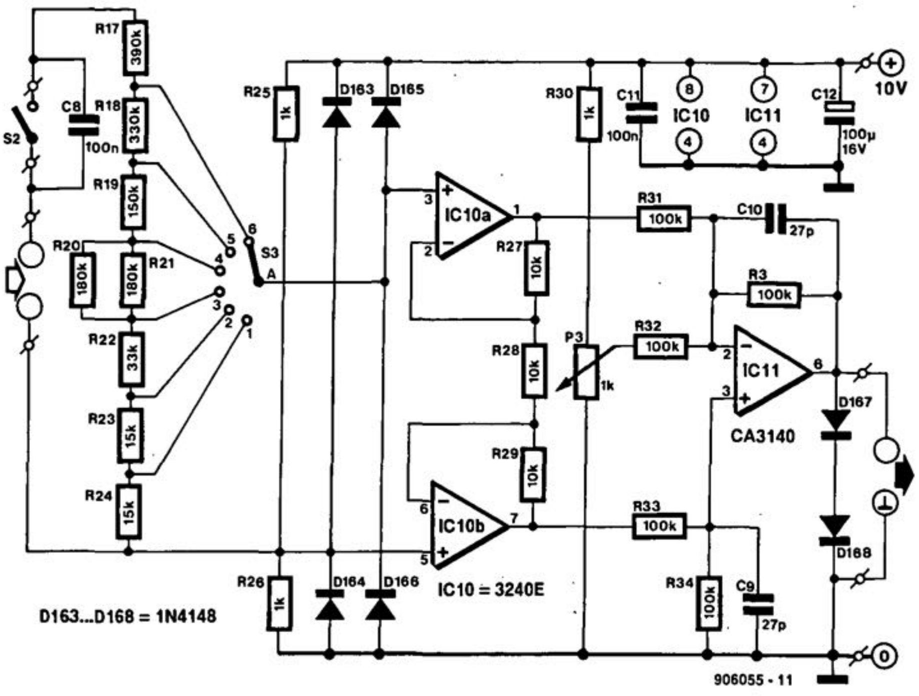


Fig. 8. Circuit diagram of the (optional) attenuator/amplifier.

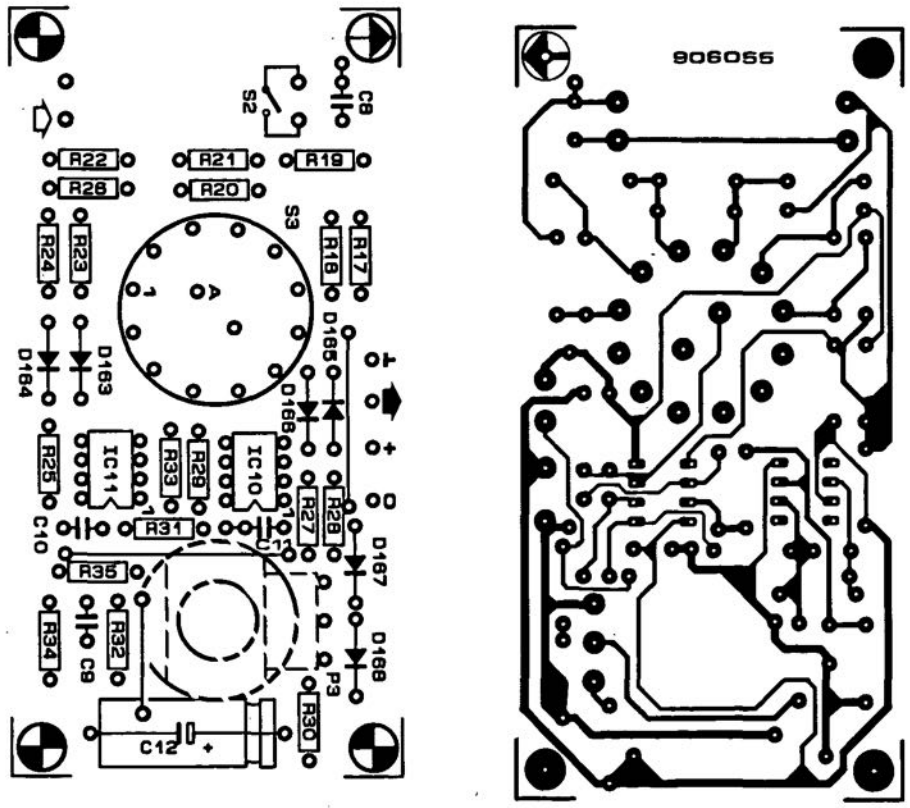


Fig. 9. Printed-circuit board for the attenuator/amplifier (not available ready made).

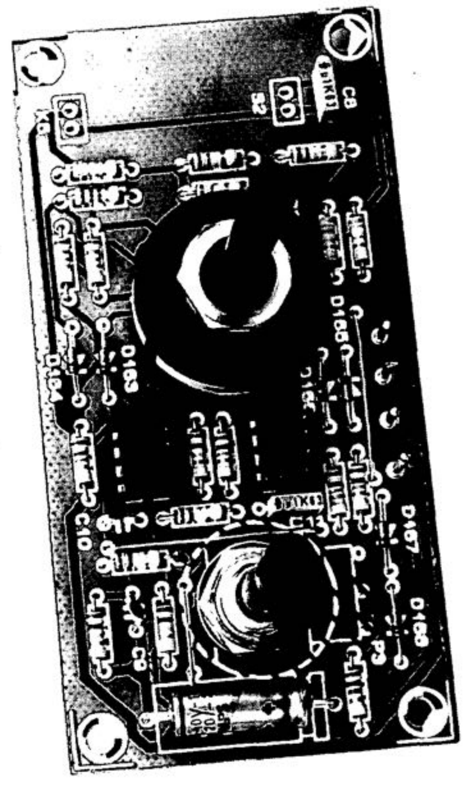


Fig. 10. Completed attenuator/amplifier board.

Schmitt trigger NAND gate IC_{13a}. When C_{14} is charged to a certain level, IC_{13a} changes state, whereupon C_{14} is discharged. When it is discharged to a certain level, IC_{13a} again changes state, and the capacitor is charged again. This results in a triangular voltage across the capacitor.

The output of IC_{13a} is a stream of rectangular pulses with a PRR (pulse repetition rate), f_{ck}, of 1 MHz, and this is used as the clock for the time base. When the START input of the time base circuit is made high by the oscilloscope, the pulse train is applied to the clock input of divider IC_{12} via IC_{13d}.

The 11 dividers (bistables) contained in IC_{12} are cascaded. The construction of each bistable is such that upon each high-low transition at its input, its output changes state. That is, at the first high, the divider is set, at the second, it is reset, at the third, it is set again, and so on. This means that a rectangular signal is divided by two, so that the frequency at the output of the bistable is half that at its input. If IC_{12} were considered on its own, the frequency at Q_0 would be $f_{ck}/2$, that at Q_1, $f_{ck}/4$, and that at Q_{10} $f_{ck}/2048$. Thus, as long as there is a clock signal at pin 10, all binary combinations between 000 0000 0000 and 111 1111 1111 would be available.

However, although the bistables in IC_{12} are set at a high-low transition at their input, they are reset immediately after being set (upon their output going high). Therefore, at the next high-low transition, the bistable is already reset, so that it is set again. With S_4 in position 1, Q_0 is set at every high-low transition of the clock pulses, which means that the frequency at Q_0 is equal to f_{ck}.

The circuit which ensures that a bistable is set and almost immediately reset is the delay network formed by IC_{13b} and IC_{13c}. The reset pulses are provided by the selected Q output, relevant diodes and S_4. For example, when the first divider is set, Q_0 goes high and this high level is transferred immediately via S_4, IC_{13b} and IC_{13c} to the reset input of IC_{12}, whereupon Q_0 becomes low again. This process results in a very short pulse at the EXT output, which is used to set the decoders on the mother board to their next state.

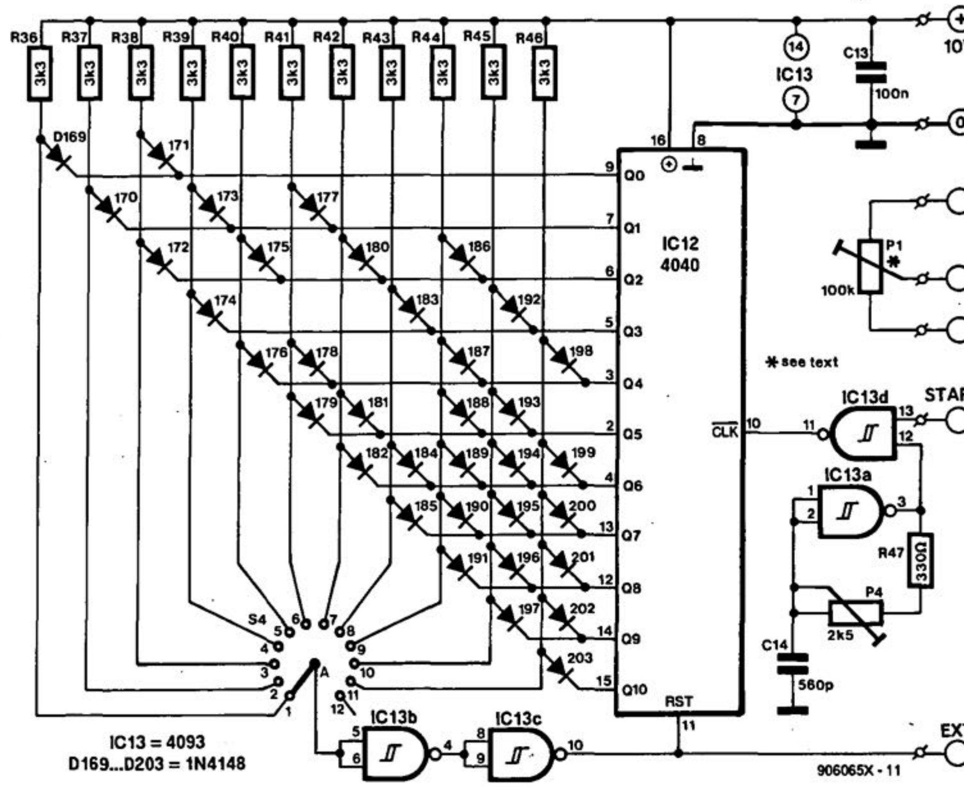


Fig. 11. Circuit diagram of the time base.

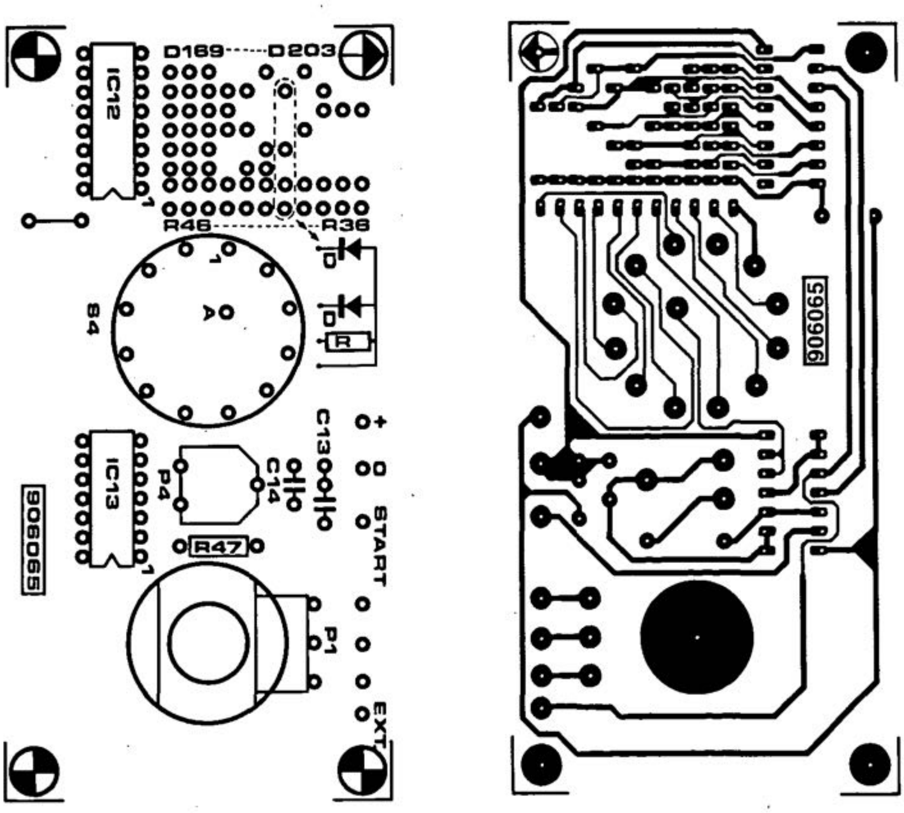


Fig. 12. Printed-circuit board for the time base (not available ready made).

Fig. 13. Completed printed-circuit board for the time base.

Diode matrix. The input of the delay network is connected to the Q outputs of IC_{12} via a diode matrix. This matrix ensures that IC_{12} does not always have to go through all its counter states, but is reset, that is taken back to zero, after a given number of states. As mentioned before, IC_{12} is reset by the pulse provided by the relevant Q output and diodes, and S_4. Thus, the diode matrix, in conjunction with S_4, ensures that resetting takes place only when certain binary codes (numbers) exist on the Q outputs.

For example, with S_4 in position 1, the divider can count only to 1: in **Fig. 11**, the input of the delay network is connected only to Q_0 (via S_4 and D_{169}). This means that IC_{12} is reset almost immediately: the binary number at its Q outputs is then 000 0000 0001. When S_4 is in position 2, IC_{12} is reset when Q_1 goes high: counter position 000 0000 0010. With S_4 in position 3, Q_0 and Q_2 must be high simultaneously before a reset pulse is generated.

Since the position of S_4 determines when a reset of IC_{12} takes place, it also determines at what intervals the time base generates short pulses at the EXT output. The diode matrix has been designed to provide intervals in the ratio 1:2:5, that is, 1 µs, 2 µs, 5 µs, 10 µs, 20 µs, 50 µs, 100 µs, 200 µs, 500 µs, 1 ms and 2 ms.

Preset P_1 is the trigger control already discussed in the section 'Trigger circuit'.

Construction. The time base circuit is best built on the printed-circuit board in **Fig. 12**, which is, however, not available ready made. The finished board is shown in **Fig. 13**.

Test the circuit by applying a 10 V supply to it and holding it near a MW receiver: the 1 MHz clock should be heard clearly.

Assembly. The four boards and a suitable enclosure for them are shown in **Fig. 14**. A suitable (not mandatory) way of fitting the boards in the case is illustrated in **Fig. 17**. Other types of assembly are, of course, possible. A suggested front panel is shown in **Fig. 15**.

The interconnections between the boards are shown in the diagram in **Fig. 16**. Do not forget to place the wire bridge at the EXT connection on the driver board to position B.

If the original clock on the driver board is not intended to be used, P_2 may be omitted.

All wires carrying analogue signals should be single screened cables with the screen earthed at one end only. This applies particularly to the signal input, the attenuator/amplifier and the circuit around IC_4 on the driver board.

All earth wires between the boards should be separate wires.

Before the case can be closed, set the clock generator to exactly 1 MHz with P_4, which is best done with the aid of a frequency counter.

Parts list

Display board

Resistors:
R_1–R_{10} = 22 Ω

Semiconductors:
T_1–T_{10} = BC557
D_1–D_{160} = LED, 3 mm, red

Integrated circuits:
IC_1, IC_2 = ULN2803A

Driver board

Resistors:
R_{11} = 560 kΩ
R_{12} = 47 kΩ
R_{13} = 4.7 kΩ
R_{14} = 10 kΩ
R_{15} = 12 kΩ
R_{16} = 2.2 kΩ
P_1 = 500 kΩ linear potentiometer
P_2 = 1 MΩ logarithmic potentiometer

Capacitors:
C_1, C_4–C_7 = 100 nF
C_2 = 1 nF
C_3 = 100 µF, 25 V

Semiconductors:
D_{161}, D_{162} = 1N4148

Integrated circuits:
IC_3 = LM3914
IC_4 = LM311

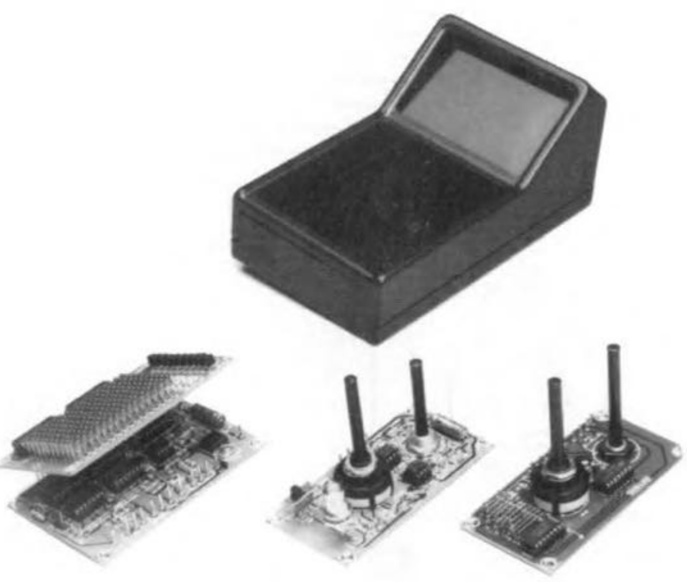


Fig. 14. The constituents of the oscilloscope.

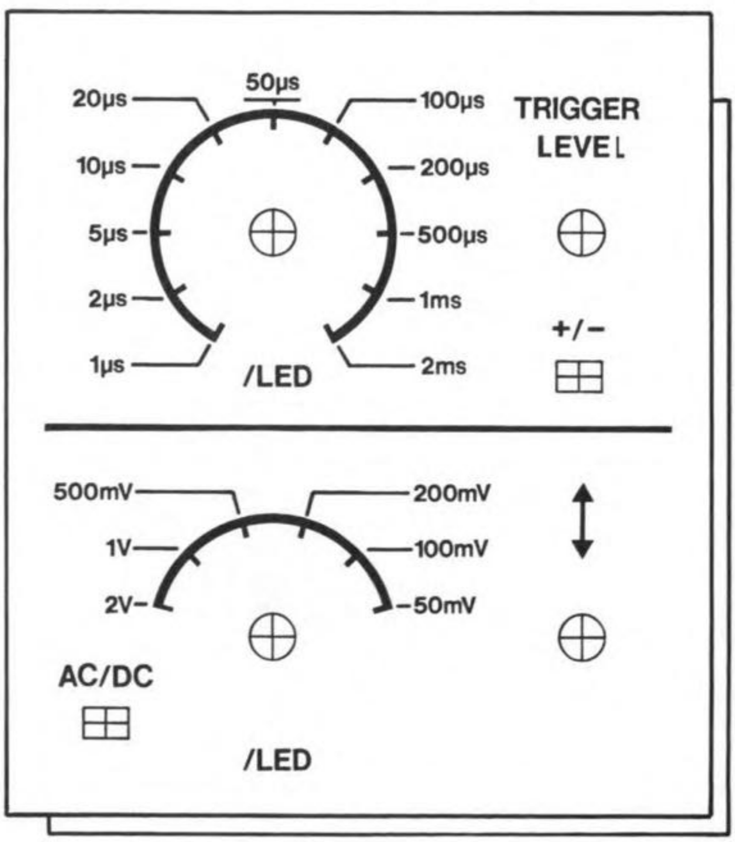


Fig. 15. Suggested front panel layout for the oscilloscope.

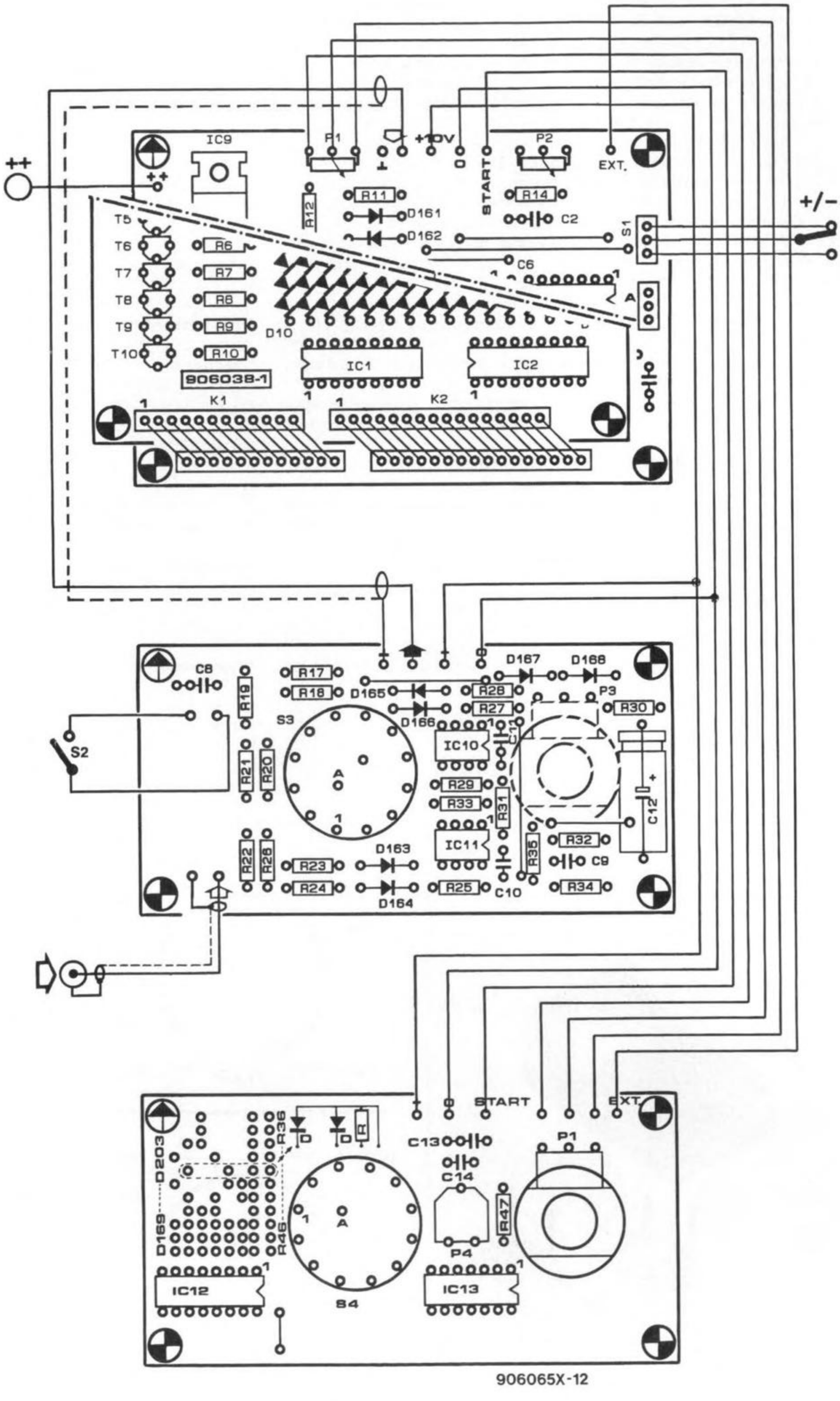


Fig. 16. Interwiring diagram for the complete oscilloscope.

IC_5 = 4093
IC_6 = 4013
IC_7; IC_8 = 4017
IC_9 = 7810

Miscellaneous:
S_1 = single-pole change-over switch

Attenuator/amplifier

Resistors:
R_{17} = 390 kΩ
R_{18} = 330 kΩ
R_{19} = 150 kΩ
R_{20}, R_{21} = 180 kΩ
R_{22} = 33 kΩ
R_{23}, R_{24} = 15 kΩ
R_{25}, R_{26}, R_{30} = 1 kΩ
R_{27}–R_{29} = 10 kΩ
R_{31}–R_{35} = 100 kΩ
P_3 = 1 kΩ linear potentiometer

Capacitors:
C_8; C_{11} = 100 nF
C_9; C_{10} = 27 pF
C_{12} = 100 μF, 16 V

Semiconductors:
D_{163}–D_{168} = 1N4148

Integrated circuits:
IC_{10} = CA3240E
IC_{11} = CA3140

Miscellaneous:
K_5 = BNC connector
S_2 = single-pole on-off switch
S_3 = 2-pole, 6-position rotary switch

Time base

Resistors:
R_{36}–R_{46} = 3.3 kΩ
R_{47} = 330 Ω
P_4 = 2.5 kΩ preset

Capacitors:
C_{13} = 100 nF
C_{14} = 560 pF

Semiconductors:
D_{169}–D_{203} = 1N4148

Integrated circuits:
IC_{12} = 4040
IC_{13} = 4093

Miscellaneous:
S_4 = 1-pole, 12-position rotary switch
Enclosure: Suggested Bopla BP 680 (available from Phoenix Mecano, phone (0296) 398 855; fax (0296) 398 866

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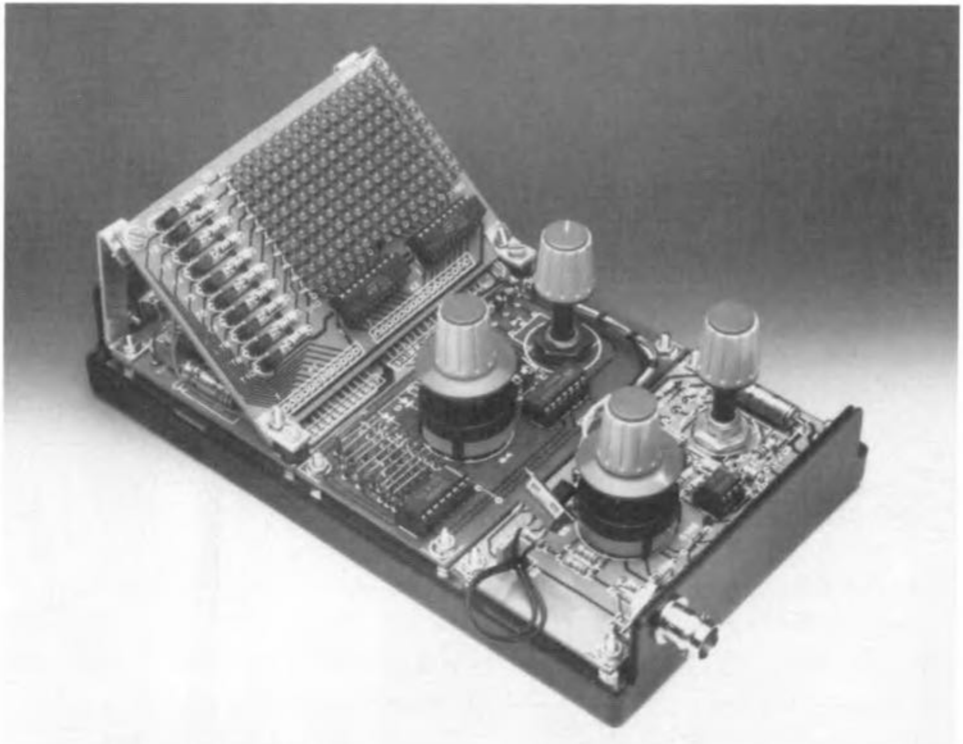

Fig. 17. Suggested assembly of the complete oscilloscope.

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COMPACT FREQUENCY METER

Modern integration techniques make possible the design of a digital frequency meter with a $4^1/_2$-digit LED display not much larger than a matchbox as shown in the photograph. With the (optional) prescaler, it enables frequencies up to about 1 GHz to be measured. Apart from use as a stand alone instrument, the meter can also be integrated in an existing apparatus.

Design by C. Wolff

Test and measurement are such indispensible aspects of electronics that there is always a demand for accurate instruments that can measure all sorts of electrical quantity precisely. The frequency meter described in this article can measure signals with a frequency of up to 199.99 kHz. Adding the optional prescaler to it increases the frequency range to over 1 GHz. The meter can be integrated in, say, a function generator based on an XR2206, and also be used as a stand alone instrument.

Circuit description

A look at the circuit in **Fig. 1** shows that it consists of three distinct sections, indicated by the dashed lines.

The measurand (signal to be measured) is applied to IC_2 via IC_{5c} and IC_{5d}. Counter IC_2 drives four of the five 7-segment, common cathode displays, LD_2–LD_5. The fifth display is controlled by an auxiliary circuit, which will be reverted to later.

Circuit IC_1 generates the gate time during which clock pulses are applied to the counter. With a crystal frequency of 5.24288 MHz, the gate time is 100 ms or 1 s. Which of these times is available depends on the position of jumper JP_1. If pin 11 of IC_1 is high, the gate time is 100 ms, which is long enough for measurands up to 200 kHz. When pin 11 is low, signals of up to 20 kHz can be counted. In either case, one measurement is carried out every two seconds.

Circuit IC_4 is an overflow detector which controls display LD_1. When IC_2 passes the maximum counter position (9999), a pulse appears at its C/B (carry) output, pin 1. Circuit IC_4 counts these pulses. Since the meter is reset at the onset of each gate time, it begins each measurement in position 0. When the carry pulse reaches IC_4, its output Q_1 goes high to indicate that 10000 pulses have been counted. Circuit IC_6 is then triggered via IC_{5a}, whereupon LD_1 is actuated minimally (with the specified value of C_4) until the start of the next measurement cycle. To obviate visible flickering, IC_6 is made retriggerable by T_3. This is necessary, because during the measurement period the data at the output of IC_2 are stored in a buffer. Such a buffer is not available for the data appearing on LD_1, which is why IC_6 fulfils this function during the measurement (which lasts not more than 1 s). If during the measurement a second carry pulse were generated, output Q_2 of IC_4 would go high and an unused segment in LD_1 would be actuated via T_2. This is the warning signal that an overflow situation exists. The frequency at which the segment flickers corresponds to the repeat time of the measurement, that is, $^1/_2$ second.

It is imperative that the 'A' version of the ICM7217 (IC_2) be used, because this can count pulses up to 9999. Since the IC contains all the necessary drivers, including multiplexers, and associated resistors for driving the LED displays, the interface can be kept fairly simple. Note that the display control input (pin 20) is linked to earth to ensure that the suppression of the leading zeros is disabled. This means that when no signal is applied to the frequency meter, the display is not blank, but reads 0000.

The measurement cycle is started by a short pulse at the store input (pin 9) of IC_2, whereupon the present counter position is stored in the output register. Then, a reset pulse at pin 14 returns the IC to 0 (if necessary, IC_6 is enabled for LD_1). Next, the pulses to be counted are applied to the CNT input (pin 8). The cycle ends after no more than 1 s, and a new cycle is started.

Circuit IC_{12} in the prescaler divides signals up to 1 GHz by 64. It is manufactured in ECL (emitter-coupled logic) technology and performs to specification at 1 GHz even at signal levels as low as 5 mV. The overall prescaler divides the measurand by 1000 and converts the ECL levels to TTL levels.

The input signal is applied to IC_{12} via protection network, D_2–D_3, which limits the input level to ±300 mV.

The output (pin 7) is converted to TTL level by gates IC_{8a}–IC_{8c} and associated components.

The :64 signal at the output of IC_{8c} is divided by 1.5625 in IC_9 and IC_{10}, so that the signal at the output of IC_{10a} is the measurand divided by 100. This is then applied to decade scaler IC_{11}, so that the output of the prescaler circuit (pin 10 of IC_{11}) is the measurand divided by 1000. This signal is applied to the input of the main circuit (pins 8 and 9 of IC_{5c}).

Parameters

Display	$4^1/_2$-digit LED
Measurement rate	1 every 2 seconds
Measurement range	
Counter (gate time 0.1 s)	199.99 kHz
(gate time 1 s)	19.999 kHz
with prescaler	1 GHz
Sensitivity	2–5 mV (with prescaler)
	TTL level (without prescaler)
Overflow	detection with display – see text

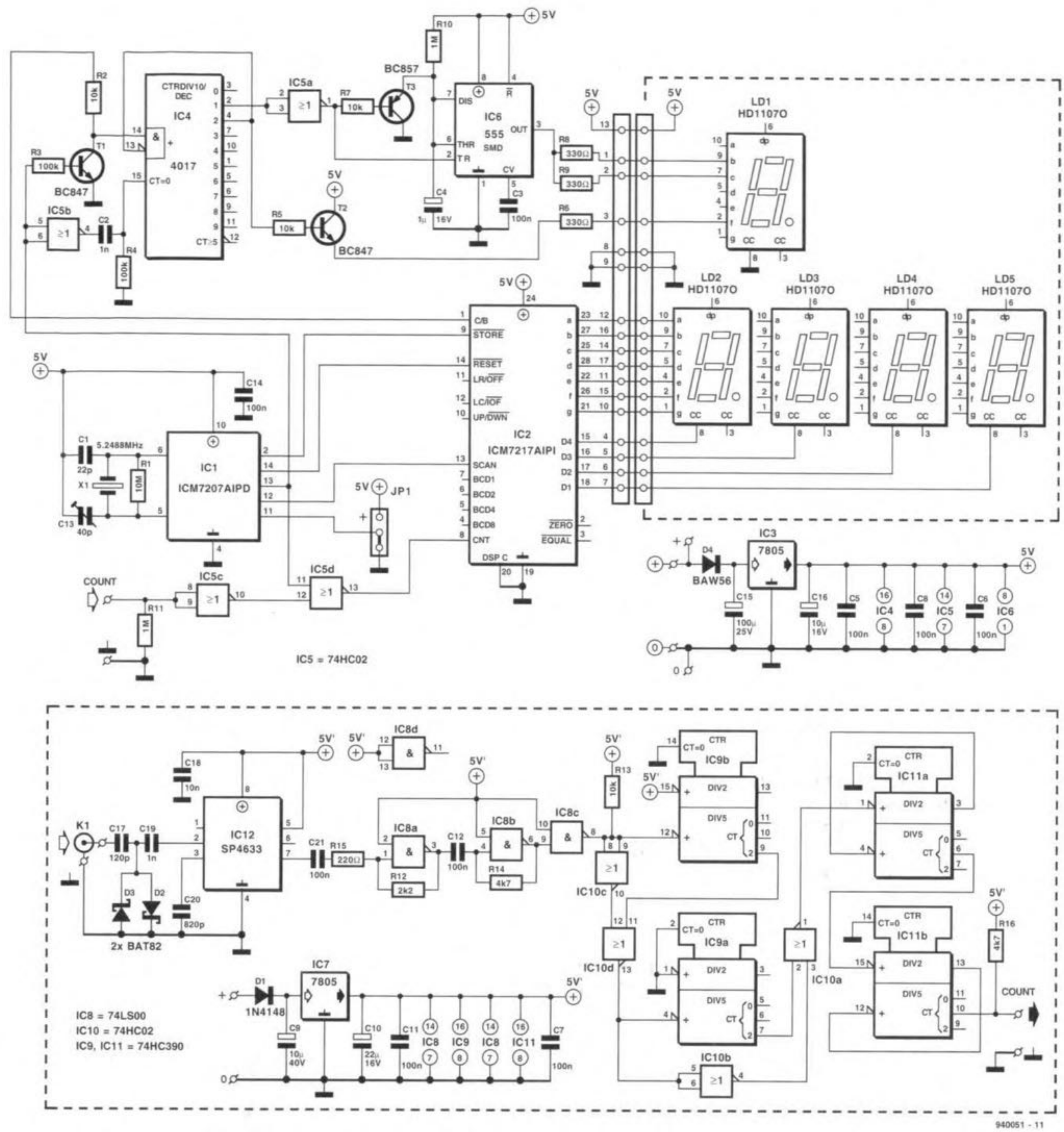


Fig. 1. Circuit diagram of the frequency meter, incl. the prescaler.

Construction

The frequency meter is best built on the printed-circuit board shown in **Fig. 2**. Before any work is done, this PCB must be snapped into three along the separation grooves.

Start with the board for the displays. First solder the wire bridges and then the displays into place.

The mother board is populated with conventional components as well as surface mount devices (SMDs). Start by soldering the SMDs at the track side of the board. Then, solder the wire bridges, followed by the conventional components, at the component side.

The mother board is linked to the display board by soldering a right-angle header to them. This results in a fairly solid entity.

Trimmer C_{13} on the mother board must be adjusted with the aid of an accurate frequency meter connected to pin 14 of IC_1 or IC_2. Depending on the position of JP_1, adjust the trimmer for a meter reading of 0.1 s or 1 s. If a frequency meter is not available, set C_{13} to the centre of its travel.

The prescaler, which is required for measuring high frequencies, is built on the third board. Like the mother board, it uses SMDs as well as conventional components. First, solder the SMDs to the track side and then the wire bridges and the conventional components to the component side.

Both the mother board and the prescaler board contain a 7805 voltage regulator. Although not strictly necessary, it is advisable to fix these regulators to the respective boards with some glue or double-sided sticky tape. This prevents these devices from vibrating and thus extends their reliability.

The reason that both boards have a regulator is that this makes the prescaler a general purpose unit which may be used with other circuits as well as with the present one.

If the prescaler is used, link its board and the mother board in a 'sandwich' construction as shown in **Fig. 3**. Use M3 screws, nuts, washers and spacers. Note that the track side of the prescaler board points upwards and that of the mother board downwards, that is, the component sides of the boards face one another.

Link the output of the prescaler to the input of the meter proper and interconnect the supply lines of the boards. Finally, connect a suitable mains adaptor (9 V, 200 mA) to the supply connections.

Parts list

Resistors:
R_1 = 10 MΩ*
R_2, R_5, R_7, R_{13} = 10 kΩ*
R_3, R_4 = 100 kΩ*
R_6, R_8, R_9 = 330 Ω*
R_{10}, R_{11} = 1 MΩ*
R_{12} = 2.2 kΩ*
R_{14}, R_{16} = 4.7 kΩ*
R_{15} = 220 Ω*

Capacitors:
C_1 = 22 pF*
C_2, C_{19} = 1 nF*
C_3, C_5–C_8, C_{11}, C_{12}, C_{14}, C_{21} = 100 nF*
C_4 = 1 μF, 16 V, radial
C_9 = 10 μF, 40 V, radial
C_{10} = 22 μF,16 V, radial
C_{13} = 40 pF trimmer
C_{15} = 100 μF, 25 V, radial
C_{16} = 10 μF, 16 V, radial
C_{17} = 120 pF*
C_{18} = 10 nF*
C_{20} = 820 pF*
* = SMD

Semiconductors:
D_1, D_4 = PRLL4001 (=1N4001 SMD)
D_2, D_3 = BAS82 (=BAT82 SMD)
LD_1–LD_5 = HD11070
T_1, T_2 = BC847 (=BC547 SMD)
T_3 = BC857 (=BC557 SMD)

Integrated circuits:
IC_1 = ICM7207AIPD
IC_2 = ICM7217AIPI
IC_3, IC_7 = 7805
IC_4 = 4017*
IC_5, IC_{10} = 74HC02*
IC_6 = 555*
IC_8 = 74LS00*
IC_9, IC_{11} = 74HC390*
IC_{12} = SP4633

Miscellaneous:
K_1 = BNC connector
X_1 = quartz crystal, 5.24288 MHz
JP_1 = 3-way jumper
PCB Ref. 940051-1 (see p. 78)

* SMD

[940051]

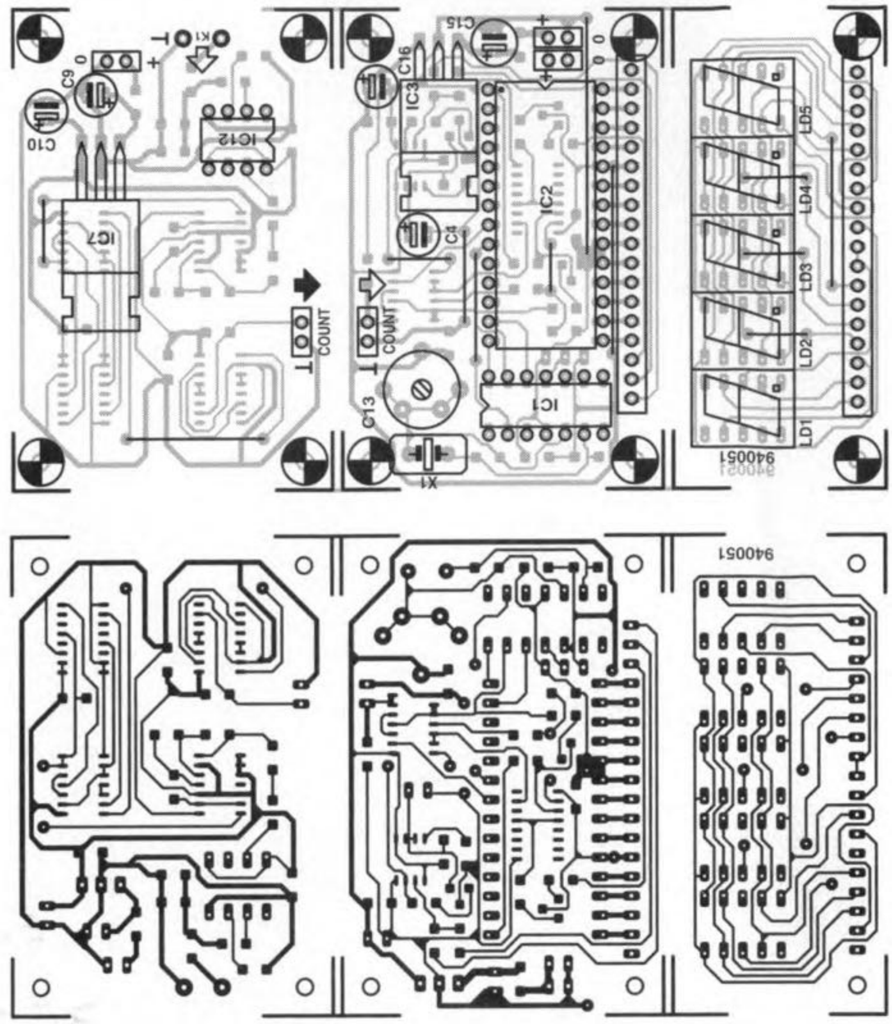


Fig. 2. The PCB for the frequency meter must be snapped into three.

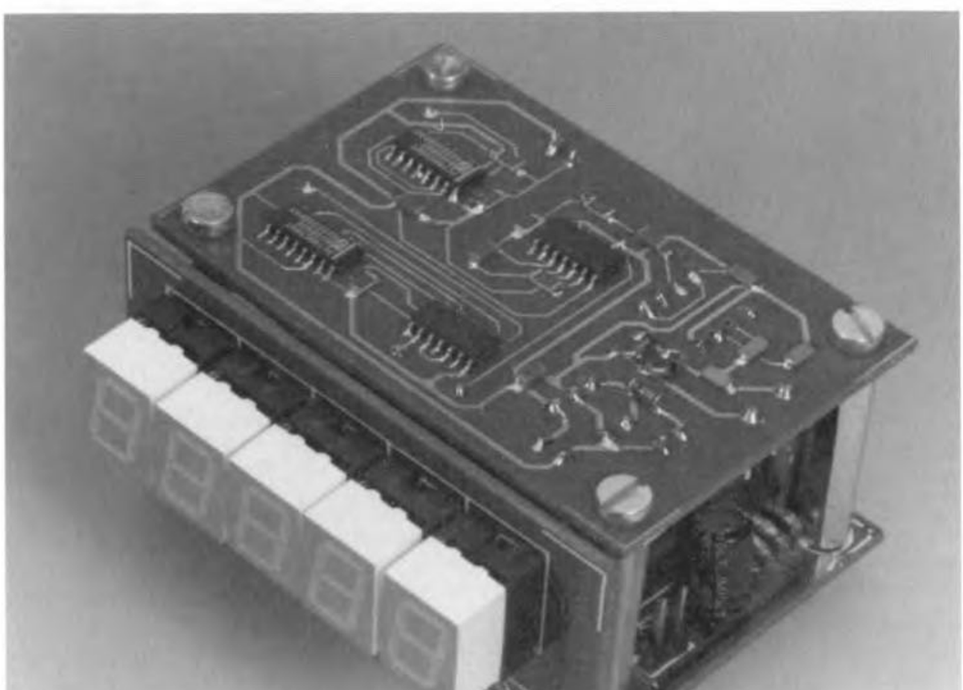
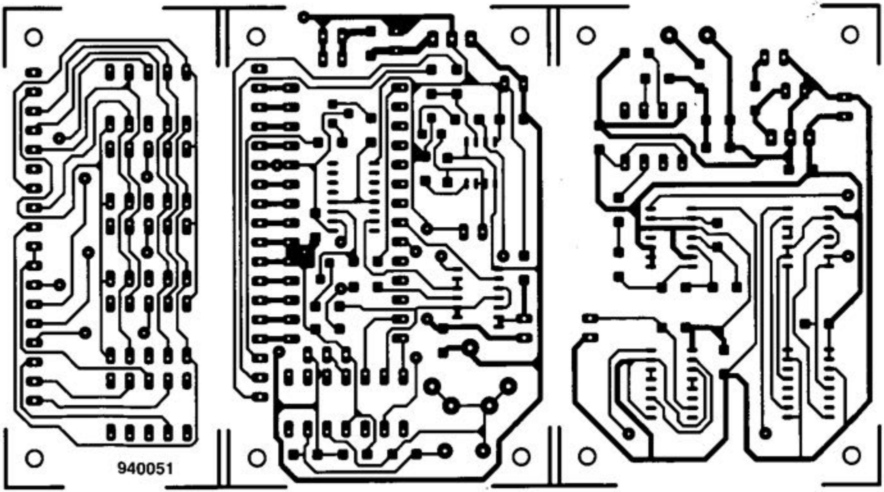


Fig. 3. Assembly of the completed boards: the display board is at right angles to the mother board, and the prescaler board is mounted above the mother board on suitable spacers.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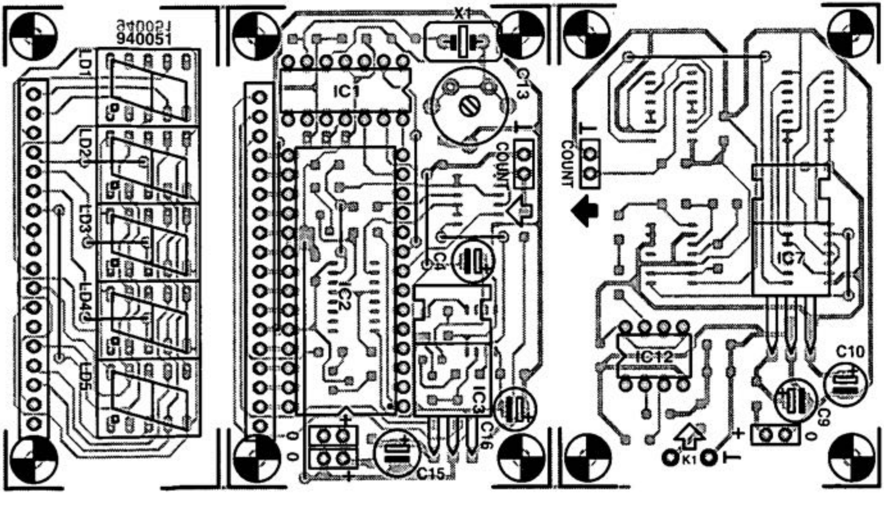
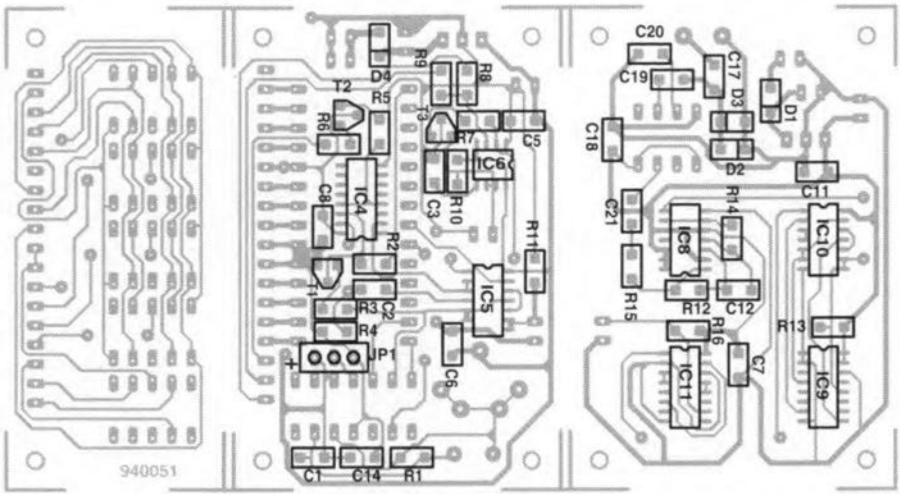

940051
LD1
LD2
LD3
LD4
LD5
X1
IC1
C13
COUNT
IC2
C4
IC3
C16
C15
IC7
IC12
C10
C9
K1

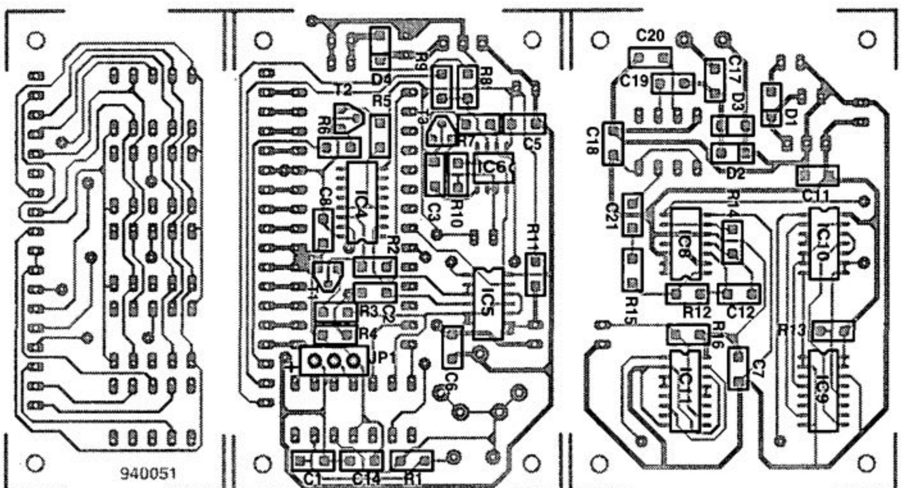
COMPACT FREQUENCY METER (SEPTEMBER 1994)



We apologise for the inconvenience caused to readers by the omission from our September 1994 issue of the third part of the double-sided through-plated printed-circuit board for the 'Compact frequency meter', which is now reproduced on the left.

The order forms are on page 71 this month.

COMPACT FREQUENCY METER (SEPTEMBER 1994)



We apologise for the inconvenience caused to readers by the omission from our September 1994 issue of the third part of the double-sided through-plated printed-circuit board for the 'Compact frequency meter', which is now reproduced on the left.

940080 - I

SURVEY COMMENTS

...I would like to see more technical books (Radio, Test & Measurement, etc.) published by Elektor Electronics.

...I would appreciate more on software design of your microcontroller articles. While I can understand that selling preblown EPROMS has a commercial bias, I would be most interested in blow diagrams and/or code on how the application was programmed.

...When you publish application notes for ICs, could you also give name and address of the IC's distributors or possible suppliers. This also applies to ICs used in your projects.

...How about publishing a master index yearly? Often, I need potential circuits on a particular topic and am then obliged to research more than 10 years' work.

...When publishing a project with an MCV (i.e., microcontroller or microprocessor), please do include source code listing for the MCV (if the listing is reasonable in size).

...I would like to commend you on the quality of design in your construction projects. I have built the GAL programmer and am very pleased with the hardware and software. Your magazine is better than any other electronic hobbyist magazine published in the USA.

...Please keep your project-oriented focus – yours is the only magazine of that kind I know of!

...More articles on Amateur Radio would be greatly appreciated, in particular any articles pertaining to (advanced) receiver design.

...Your magazine gives me hours of entertainment and hobby work. I have followed the 8051 course and built many of the related projects with great pleasure.

...Your series 'Figuring it Out' is great.

...Even with CAD I cannot get my circuit diagrams as good as yours.

...I would be interested in a list of suppliers for audio grade components, software programs (BASIC) for the calculation of loudspeaker cross-over networks and loudspeaker enclosures.

...More application notes, please! Good high-quality PCB construction teach-in would benefit many readers with limited resources...

...I would like to see more audio and visual linked projects, as well as automotive projects.

...Please could you start a beginners course on computing technology.

...Some components used in your projects are not easy to get hold of. Some of the recent microcontroller articles are tedious. But...your board layouts and the quality of the magazine (both print and text) are extremely good.

...Having used the Readers service on a few occasions, I have found this to be an efficient service – even when I make a mistake with the order numbers!

...I would like to see more projects covering Satellite TV and decoders, also information covering items as found in *What Satellite TV*, but with more technical detail.

...Your magazine has been a benchmark by which I have measured many others, all of which have failed to come up to your high standards. The breadth of the articles and the depth of the projects are unequalled.

...I would like to see more small-scale projects (AF to RF).

...Your microcontroller articles/courses are very good. I would like to see more music related articles.

...I would appreciate more British sources listed for specialized components.

...Seeing yet another audio amplifier does not amuse me!

...I would like to see a project for a Dolby Pro Logic Decoder and one for an 8751 to decode the Rugby clock fully.

...I find your magazine provides a very good means of learning about electronics.

...I would like to see more on ways of how electronics can help those less able than us lucky ones: the blind, the deaf, the old, the disabled, and so on. Surely our speciality should not only be fun, but also strive to improve the lot of as many people as possible?

...Please, please, let us have the construction of more mundane items (but not the trivia found in some other magazines). When I read articles like the 68HC11 processor board and the 80C535 hardware/assembler, I just shrug my shoulder and say "so what?"

...I would like to see more projects with microcontrollers (es. 8051) – they don't have to be very advanced. Just small tips and tricks...Also, HOME AUTOMATION would be very interesting.

...I express my compliments about your courses, because a periodical should help people to be able to build their own projects rather than just build circuits whose behaviour they don't (fully) understand.

...You have something of interest each month – and most of them work!

COMPONENT RATINGS

In resistor and capacitor values, decimal points and large numbers of zeros are avoided wherever possible. Small and large values are usually abbreviated as follows:

p (pico-) $= 10^{-12}$
n (nano-) $= 10^{-9}$
μ (micro-) $= 10^{-6}$
m (milli-) $= 10^{-3}$
k (kilo-) $= 10^{3}$
M (mega-) $= 10^{6}$
G (giga-) $= 10^{9}$

Note that nano-farad (nF) is the international way of writing 1000 pF or 0.001 μF. Resistors are $^1/_3$ watt, 5% metal film types unless otherwise specified.

The direct working voltage of capacitors (other than electrolytic or tantalum types) is assumed to be ≥60 V. As a rule of thumb, a safe value is about 2× direct supply voltage.

Direct test voltages are measured with a 20 kΩ/V meter unless otherwise specified.

Mains (power line) voltages are not listed in the articles. It is assumed that our readers know what voltage is standard in their part of the world.

Readers in countries that use 60 Hz supplies, should note that our circuits are usually designed for 50 Hz. This will not normally cause problems, although if the mains frequency is used for synchronization, some modification may be required.

The international letter symbol '*U*' is used for voltage instead of the ambiguous '*V*'. The letter V is reserved for 'volts'.

The size of a metric bolt or screw is defined by the letter M followed by a number corresponding to the overall diameter of the thread in mm, the × sign and the length of the bolt or screw, also in mm. For instance, an M4×6 bolt has a thread diameter of 4 mm and a length of 6 mm. The overall diameter of the thread in the BA sizes is: 0 BA = 6.12 mm; 2 BA = 4.78 mm; 4 BA = 3.68 mm; 6 BA = 2.85 mm; 8 BA = 2.25 mm.

EIB: EUROPEAN INSTALLATION BUS

Practically from the (very primitive) start, mains systems in private homes have had switches to enable loads to be connected to or disconnected from the electricity supply. With relatively few apparatus and lamps in the 'old' days, the number of mains connections in each home was also quite small. That has changed dramatically in the course of time. Not only the number of connections has gone up considerably; ease of use has also become the prime requirement for electrical appliances. Consequently, the need of an intelligent bus system for automation purposes in buildings in on the rise.

By Michael Rose

THE desire for more and more user comfort is the main cause of the increasing number of electrical energy users. The number of electrical apparatus in itself also rises as a result of regulations covering the environment and energy saving technology. Apparatus that come to mind in this respect are solar cells, systems for water control, and intelligent systems for heating and air conditioning.

Obviously, the complexity of an electrical system increases with the number of apparatus connected. Further problems arise from the fact that most apparatus want connections to a fair number of peripheral devices in addition to their connection to the mains. This is because sensors, actuators and control devices are applied in a growing number of cases. Consequently, a system is called for which enables vast numbers of different components in an electrical system to exchange information as regards their status.

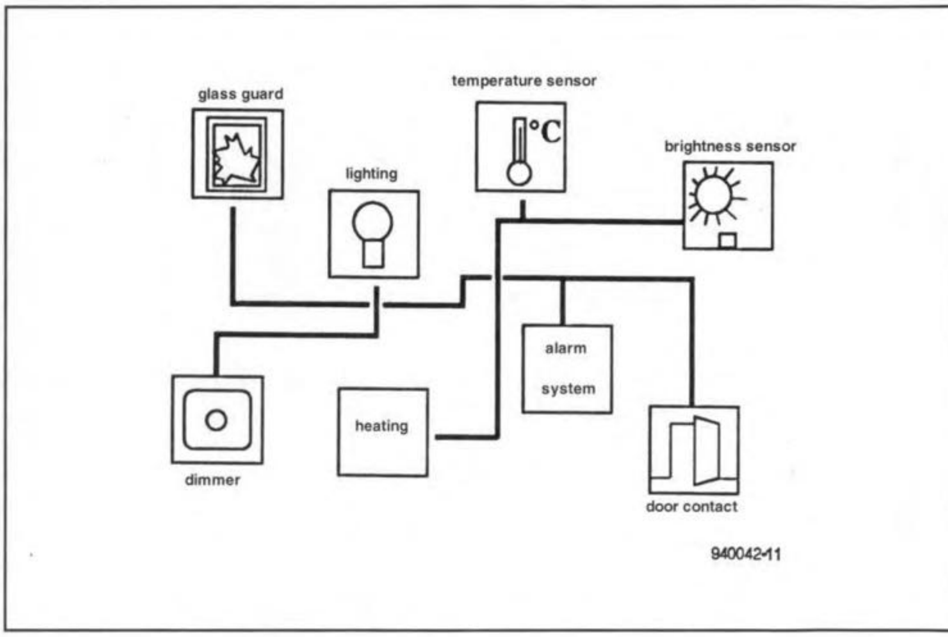


Fig. 1. Conventional electrical systems usually consist of 'islands'.

The limitations of conventional electrical systems and installations soon become evident in the light of the demands mentioned above. Also evident is the need of measures against fire hazards caused by ever more complex electrical systems.

Traditional electrical installation methods and materials are hardly up to the large changes and requirements brought about by the 'market'. To complicate the problems further, traditional solutions lead to small 'islands' in an electrical system — integrated systems being few and far between. The drawing in **Fig. 1** shows how three of these 'islands' function independently in a building.

Typical system components causing 'islands' in a system are:
- air conditioning;
- heating system;
- weather-dependent control of blinds;
- alarm system.

In some 'conventional' electrical systems, an added network is already used to exchange information. This network interconnects control and operating elements, such as sensors and actuators. This approach is fairly universal and therefore found in different 'surroundings', such as the office, the industry and the private home. These applications only differ in respect of the complexity of the system. Meanwhile, in utility buildings the number of cables and their total length increases more than proportionally with the number of control signals.

In the near future, the demand for systems capable of exchanging information between different components will increase as a result of more and more 'electrification', modernised systems, the introduction of measures to protect the environment, and the use of energy-saving technologies. A modern, standardised bus system has no problems in coping with these changes and the needs of the users. In that respect, it adds functionality to the existing electrical installations. **Figure 2** shows the structure of such a system, in this case, the European Installation Bus (EIB), originally developed by Siemens, and now firmly proposed for standardization in Europe.

The bus system has the following advantages:
- All users (i.e., loads) are connected

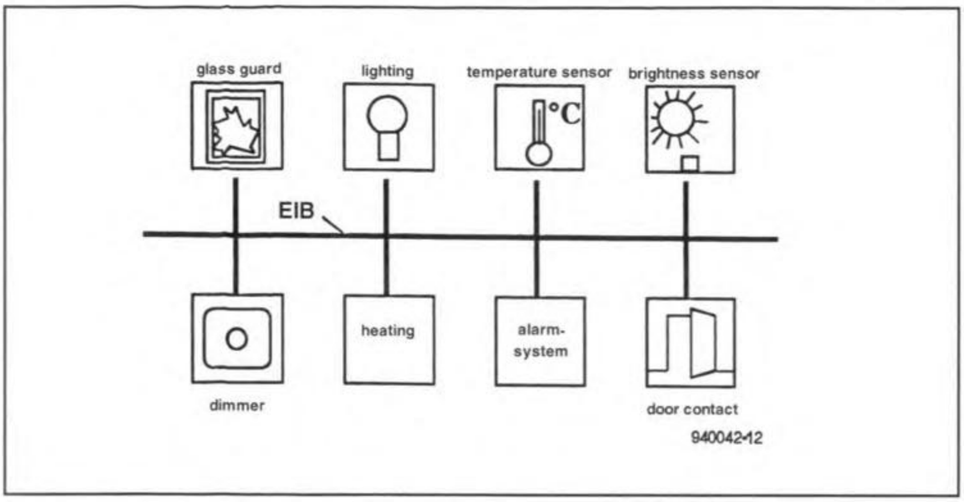


Fig. 2. The EIB enables information between system components to be distributed in a highly structured way.

The Cenelec standard

With the increasing use of mains signalling, the available frequencies have been reviewed and regulated in Cenelec standard ES 50065.

Frequency	User(s)
3 - 95 KHz	Electricity suppliers (out of doors).
3 - 9 KHz	Electricity suppliers only.
9 - 95 kHz	Electricity suppliers and licensees.
95 - 148.5 kHz	Users (indoors). (max. signal level 116 dBμV)
95 - 125 kHz	Free, no protocol.
125 - 140 kHz	Free, with protocol. carrier: 132.5 kHz max. transmit time 1 s min. wait time 125 ms each component recognises signals ≥8 dBμV
140 - 148 kHz	Free, no protocol.

directly to the mains, and have their own, local, on/off switch, instead of a master on/off switch at the control location.

- All users, sensors and actuators are interconnected via a special network, via which all data is exchanged.
- The operation of the network is not affected by the physical location of the devices in the network.
- The coupling of control components and users may be programmed via the network, allowing the electrical system to be given a new layout without having to change the wiring.
- All information carried via the network is available at each and every location.
- Inexpensive devices may be used to make conventional users and switching elements such as switches, pushbuttons and lamps compatible with the network. That allows an existing system to be modernized fairly easily.
- A central bus management ('control') is not present, in other words, there is no central computer in the system! The exchange of data is decentralized by making use of a microprocessor in each device connected to the bus. You, the user, will not notice a thing of the operation of these microprocessors.

Installing the bus system does not require special tools or highly qualified workers. The system can make do with two twisted wire pairs (one pair is reserved as a 'spare') of 0.8-mm^2 cross-sectional area. A SELV (safe extra low voltage) of 28 V d.c. is carried over the twisted pair to supply the bus system. These specifications make the EIB an excellent tool to realize electrical systems of today and tomorrow, and suitable for integrating automation systems in modern buildings. In practice, the high starting costs associated with the components that are endowed with some intelligence do not weigh up against the advantages of a really flexible installation. The cost of upturning an existing installation and slotting in new cable trunks if a building is electrically 're-organized' far exceeds the extra costs associated with the EIB system.

Hierarchical structure

There is much practical evidence that the tree structure as shown in **Fig. 3** is worth its salt when it comes to installing automation systems in buildings. That is mainly because such a structure need not comply with strict requirements. Installation engineers have a great deal of freedom in assigning functions and locations to the devices connected to the bus. Furthermore, the actual installation of the wires is uncritical because the EIB has no terminating resistors.

There is a good similarity between the tree structure and the division of a building. All components in a room are connected to a common line segment.

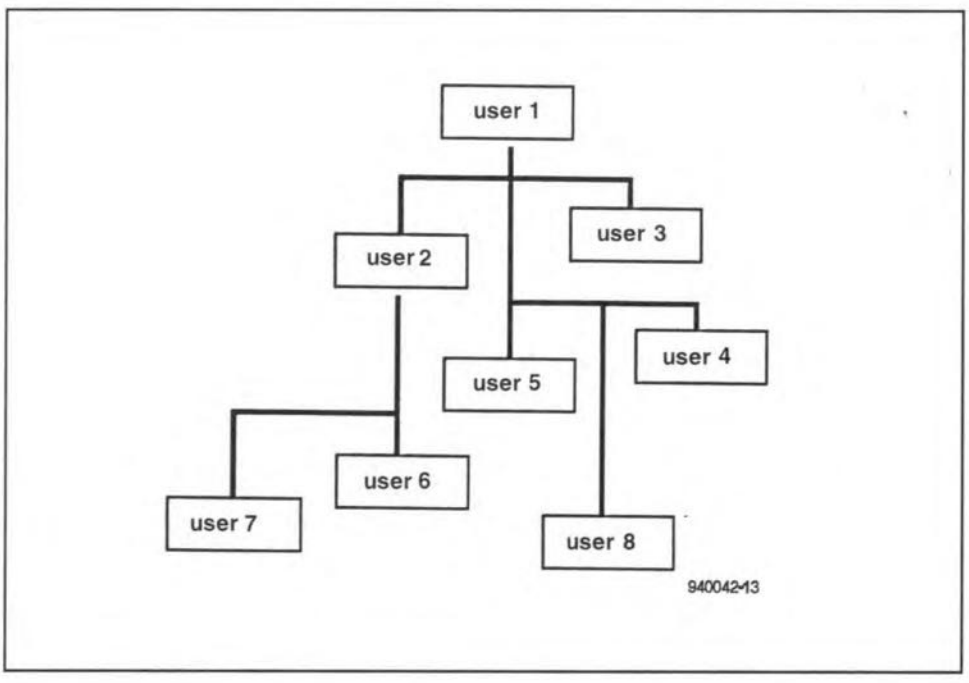


Fig. 3. In a tree structure, all users can get in touch with one another.

Similarly, the rooms in a building are interconnected. The structure can be extended to any size, covering complete floors and even buildings.

To make sure that the EIB system is a viable option under various circumstances (from a private home to an office complex), the hierarchical structure is marked by a great deal of flexibility. Although that does reduce the flexibility of the tree structure a little, it has the ever so important advantage of a clear lay-out, irrespective of the size of the system.

The tree built by the EIB has three principal lines: the bus line, the main line and the area line.

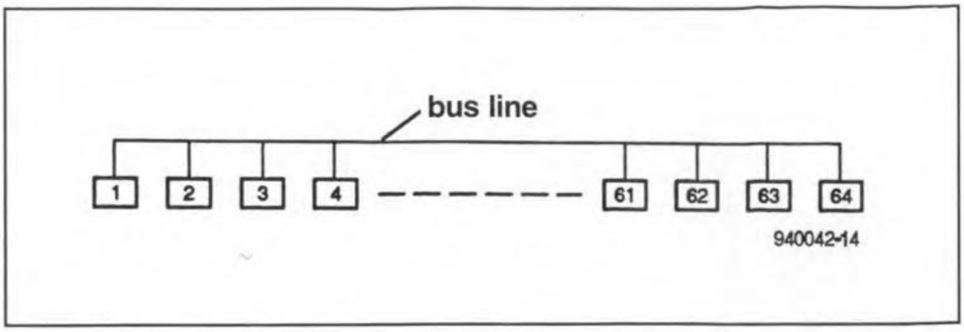


Fig. 4. The bus line is the lowest-level branch in the EIB system. Up to 64 users are directly interconnected.

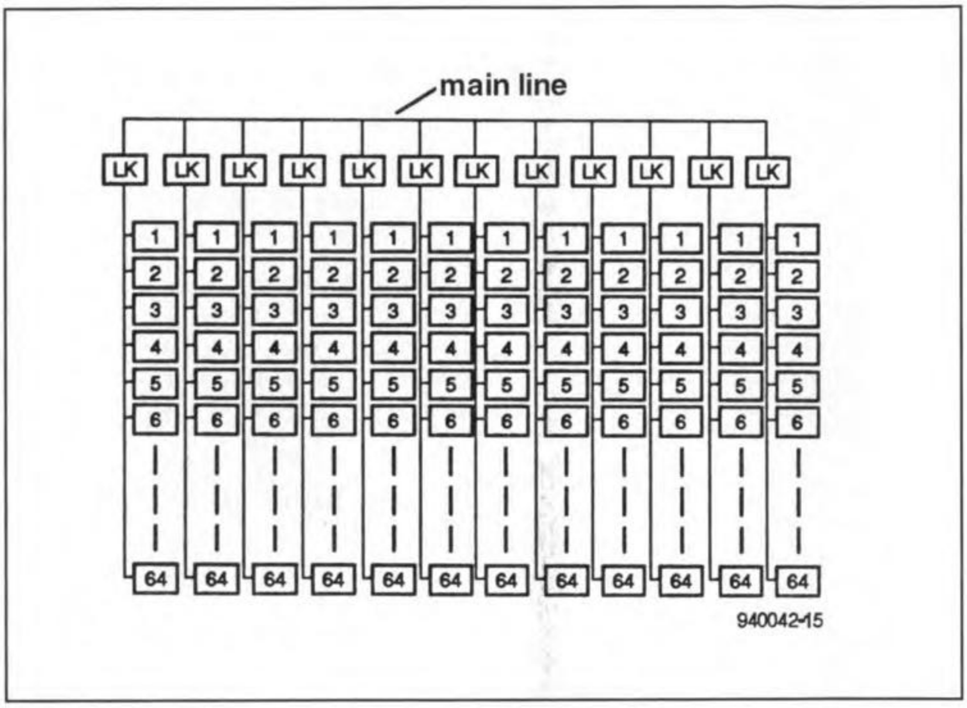


Fig. 5. The next higher step in the EIB hierarchy is the main line, which links up to 12 bus lines.

Bus line

Small automation projects may be realized with the aid of a single bus line. Since each bus line is limited as regards its range by the restricted transmit power of individual components, a maximum of 64 users may be connected to this bus. The structure of a simple network based on one bus line is shown in **Fig. 4**.

Main line

The next step in the hierarchical tree structure is combining a maximum of 12 individual lines to the main line via a line coupler (LK), see **Fig. 5**. Once that has been done, the individual lines can exchange information via the main line. Consequently, the information as regards the actual line to which the individual users are connect is irrelevant to the bus system.

Area line

The area line is at the top of the EIB's hierarchical structure. This line is connected to up to 15 main lines via an equal number of area couplers. This gives a maximum number of users of $64 \times 12 \times 15 = 11{,}520$. **Figure 6** shows the final structure of the system.

As indicated by the drawings, a full-blown EIB system is fairly complex. To ensure fast and efficient data exchange between different components, the line and system couplers contain a filter in addition to an amplifier. These filters ensure that

- information exchanged between two users of the bus line does not 'spill' on to the main line;
- information which is intended for users in one area only is exclusively available in that particular area.

This approach allows multiple messages to be conveyed at the lower levels without the available bandwidth turning into a restricting factor.

Access to the bus

A communication protocol needs to be defined to enable multiple users to communicate via a single line. This can be solved in a number of ways, depending on the area and the application.

The EIB makes use of CSMA/CA (carrier sense multiple access with collision avoidance), where the occurrence of a certain state indicates that users are busy exchanging data. The bus load remains low because no time is spent by the bus master polling (interrogating) bus users. Also, the response time to an individual message is short, allowing the message to be processed instantly in most cases. The response time of the whole system depends on (1) the response time of the relevant user to the command, (2) the wait time for a free bus, (3) the data transfer speed of the system and (4) the processing time of the receiver.

Conflicts may arise if multiple commands are given at the same time. To avoid conflicts, each transmitter checks if the bus is free (carrier sense). If any action is noted on the bus, the transmitter waits until it is ended, and only then attempts to claim the bus again. As soon as a user starts to transmit commands, it monitors the information stream on the bus. In this way, fault conditions arising from multiple users claiming the bus are immediately detected.

A special feature of the EIB is that only '0's are transmitted actively. The logic '1' is marked passively by a low level on the line. If several users claim the bus at the same time, the transmitter putting a '0' on the bus has priority over one that wants to transmit a '1'. Transmitters are capable of detecting that their own message is being corrupted, and automatically termi-

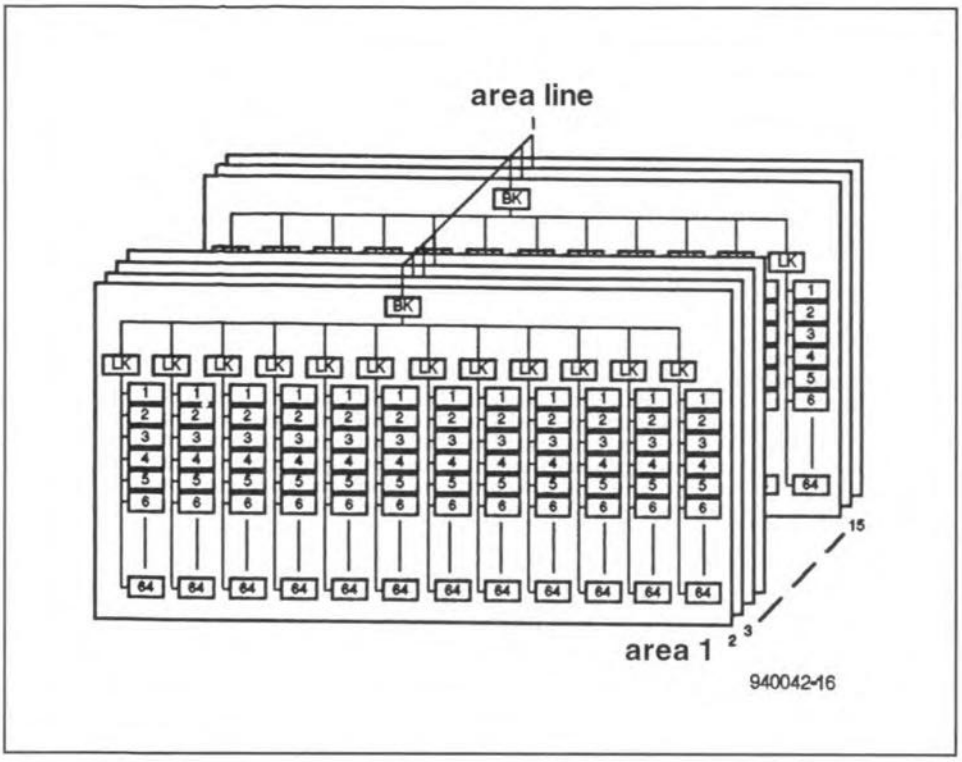


Fig. 6. The highest level is the area line. Each of up to 15 of these allows 12 bus lines to be connected, which, in turn, link up to 64 users.

nate the transmission procedure. The other transmitter can continue cheerfully. This approach keeps the response time low, even if there is a lot of activity non the bus. Also, the risk of a total bus blockage is eliminated.

Bus users

Components connected to the bus are powered via the data lines. To enable data signals to be separated from the supply voltage, a total of four inductors combined into a transformer is used as shown in **Fig. 7**. The transformer has two functions.

(1) Together with the associated capacitor, it forms a filter which prevents the high-frequency signals necessary for the data communication from reaching the supply circuit of the system components. The capacitor has across it the direct voltage which powers the system components. At the same time, the transformer (which forms a high impedance for alternating voltages) decouples the bus from the d.c. loads formed by the system components.

(2) The capacitor acts as a low impedance to the alternating voltage which contains the encoded information. Consequently, it is a virtual short-circuit between the two primary transformer windings. If a component is set up as a receiver, the transformer ensures that alternating voltages on the bus also appear across the primary winding. If a component acts as a transmitter, the transformer functions as a modulator, superimposing the signals to be transmitted on to the supply lines of the bus system. The balanced coupling causes the information to be transmitted differentially. That greatly boosts the noise immunity of the EIB system against electromagnetic and induced interference.

EIB — the origins

The European Installation Bus (EIB)is a field bus developed by Siemens for commercial applications such as automation networking in buildings. Introduced in 1992, the bus has gained wide support, including that of the European Installation Bus Association (EIBA). This organisation, located in Brussels, Belgium, represents about 70 companies responsible for the production of about 80% of electrical system components used in Europe.

Each system component has a Bus Control Unit based on a 6805. This microcontroller handles all protocols, and arranges all data traffic via eight pins. A transmission speed of about 9,600 baud is achieved. The only serious limitation of the EIB is the limited internal memory space of the processor used. That is why some system components require a second processor to be added.

Further information from
European Installation Bus Association EIBA), Avenue de la Tanche 5, B-1160 Brussels, Belgium.

or from one of the following UK members of the EIBA:
Crabtree Electrical Industries Ltd., Walsall.
Home Automation Ltd., Wiltshire.
Pillar Electrical Ltd., London.
Scholes Group plc, Wylex Ltd., Manchester.

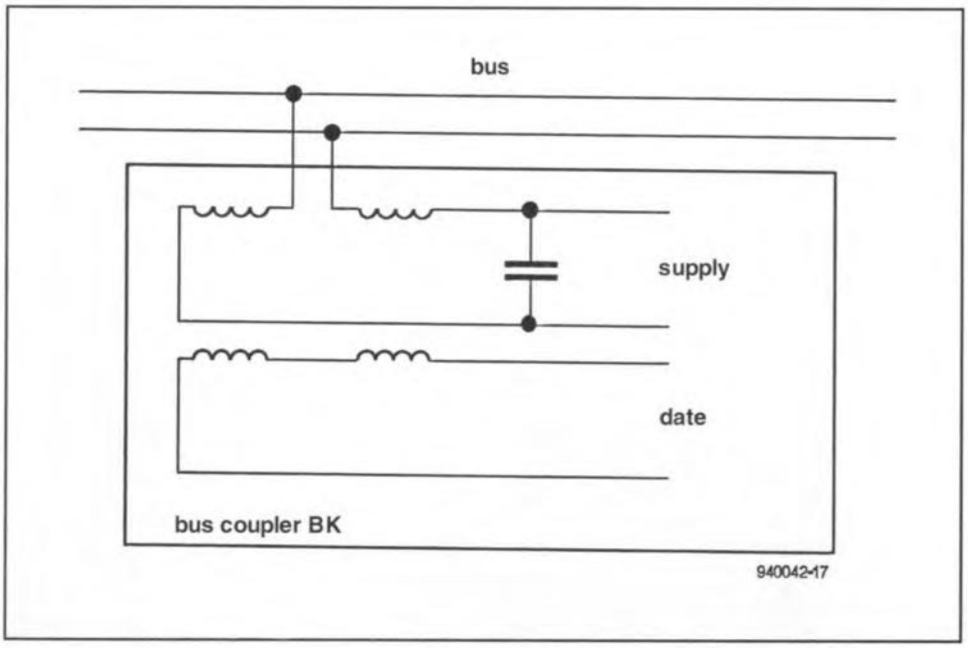


Fig. 7. Every system component ('user') is connected to the bus line via a transformer. Thanks to the inductive coupling, the supply voltage and the data signals can ride on the same two wires.

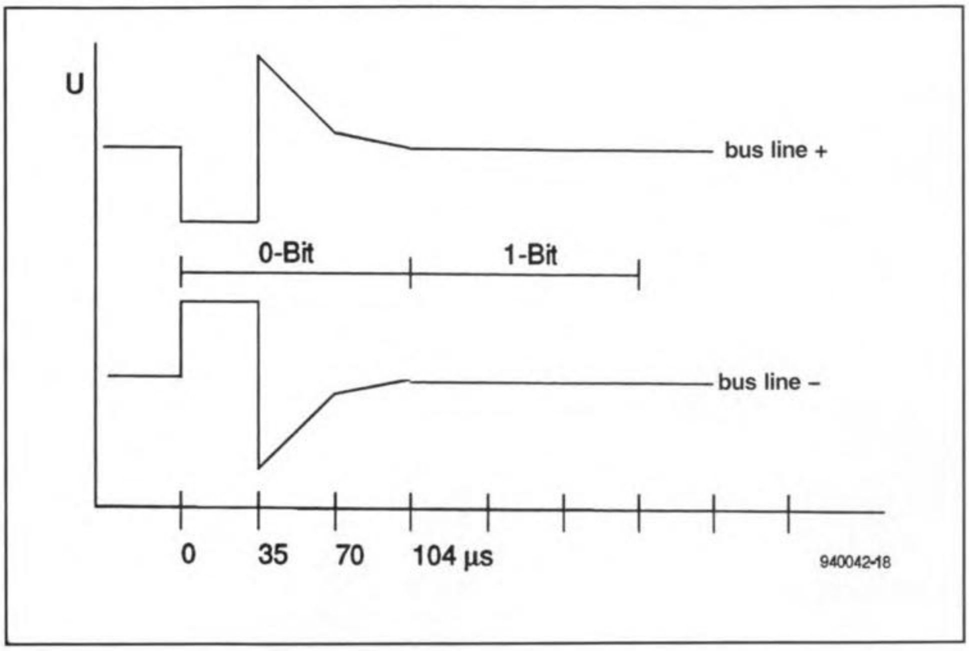


Fig. 8. Theoretical signal shapes of a logic '0' and a logic '1' transmitted via the EIB.

The bitstream

The actual encoding of a '1' and '0' on the EIB is shown in **Fig. 8**. The transmitter is only active during the first 35 µs when a '0' is being transmitted. The rest of the pulse is automatically created by the inductance of the bus system. A '1' is signalled simply by leaving the bus alone. In the event of a bus conflict, a '0' will always override any '1' simply because only a '0' is actively encoded.

In practice, the signal shape of the '0' will not be as clean as shown in **Fig. 8**. The irregularities are caused mainly by the capacitance introduced by the wiring. Hence, the maximum capacitance is defined as 200 nF, consisting of the wiring capacitance and the total input capacitance of the devices connected to the bus.

That concludes our introductory discussion of the main technical aspects of the EIB. For the more practical side you are referred to EIBA and the manufacturers of electrical system materials, who already have quite a few EIB modules in their product line.

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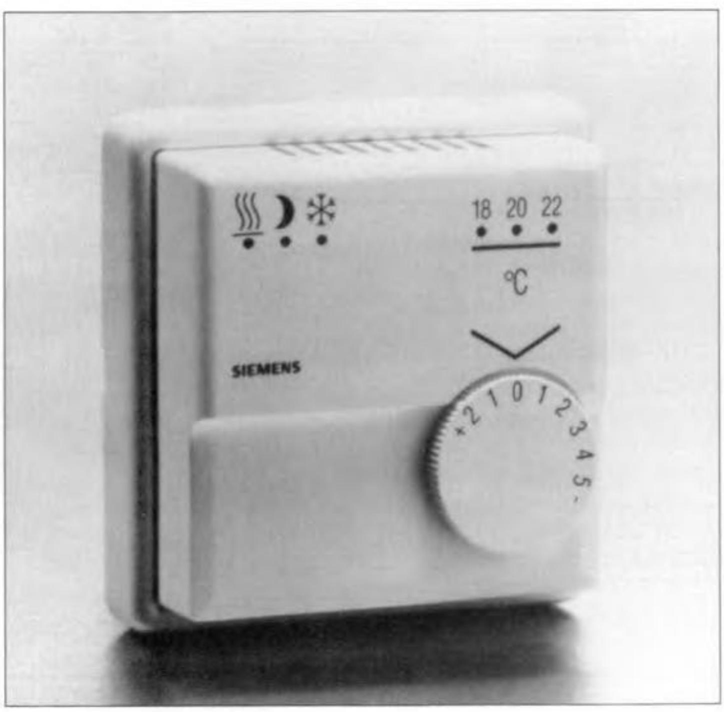


Fig. 9. An EIB compatible room thermostat (Siemens).

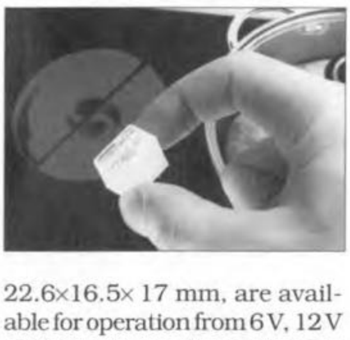
Fig. 10. This EIB supply provides the 28 V d.c. supply voltage for all devices connected to the bus (Siemens).

COMPONENTS SELECT

Programmable light sensor
Texas Instrument's TSL230 is the first programmable sensor that converts electromagnetic signals at wavelengths ranging from 300 nm to 1100 nmdirectly into digital data. It thus

makes possible the designof a microprocessor system that can measure the brightness of incident light directly. The dynamic range of the device is 160 dB.
Texas Instruments, Manton Lane, Bedford MK41 7PA, Telephone (0234) 63211

20 A miniature relay
Siemens has introduced the V23082 Series of of miniature relays with make or change-over contacts that are suitable for PCB mounting and can handle switch-on currents of up to 20 A. The contacts are rated at 5 A. The relays, measuring



22.6×16.5× 17 mm, are available for operation from 6 V, 12 V or 24 V. Their dissipation is 400 mW
Siemens Ltd, Siemens House, Windmill Road, Sunbury on Thames TW16 8HS.

Power comparator
A recent addition to the SGS Thomson catalogue is the Type L9907 power comparator. This device operates from supply voltages of up to 30 V, while its output can switch currents of up to 1 A. At that current, the

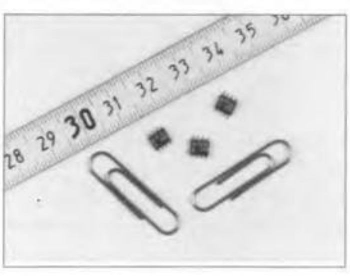
saturation voltages is ≤3.6 V.

The device was designed for electronically positioning the main beam of car headlights. To ensure satisfacytory operation in this demanding environment, the designers have paid particular attention to the temperature range and insensitivity of the IC to changes and noise in the supply voltage.

NTC for battery chargers
Philips have introduced a series of NTC resistors that are intended for protecting NiCd and NiMH batteries during charging. The resistors are compact and, by virtue of modern technology, have a R_{25} value with a tolerance of only 1%. Available in values of 4.7 kΩ or 10 kΩ, they can be supplied with a coating of lacquer, glass (for use in areas where chemicals or other substances might affect the laquer), or a coating suitable for use in humid conditions.
Philips Components (see page 19)

Linear optoisolator
Hewlett-Packard has recently introduced the Type CNR200/201 analogue optoisolator, which provides excellent linearity (non-linearity is only 0.01%). It is suitable for use with analogue signals up to 1 MHz. The amplification factor has a temperature coefficient of only 65 p.p.m., and a tolerance of ±15% (CNR200) or ±5% (CNR201).
Hewlett-Packard, 308–314 Kings Road, Reading RG1 4EJ.

SMD diode array
The SMDAxxC-8 Series of protection-diode arrays from General Instruments are surface-mount devices suitable for the protection of data lines, for



instance, an I/O port. They comply with IEC801-2 (ESD) and IEC801-4 (EFT) standards in respect of effective protection against transients. Manufactured in TransZorb-TVS technology, the arrays contain four independent protection components (both unidirectional

and bidirectional) in an SO-8 housing. The are available for operation from 5 V, 12 V, 15 V and 24 V supplies.
General Instrument, Regency House, 1–4 Warwick Street, London W1R 5WB.

Fastest DSP
Texas Instrument have achieved a breakthrough in DSP technology with its TMS320C80 digital processor. This device is intended particularly for video applications in the next generation of multi-media systems, but is also suitable for use in duplex systems for real-time video conferencing, wireless telephony and systems for document and image processing.

The processor contains, among others, four parallel operating 32-bit DSPs, an integrated RISC processor and coprocessor and a video controller. The device can handle more than 2×10^9 instructions per second.
Texas Instruments, Manton Lane, Bedford MK41 7PA, Telephone (0234) 63211

60 W microwave transistor
Philips have introduced a new silicon microwave transistor, the Type LFE15600X. This single 60 W device is intended for use in Class AB amplifiers in the frequency range 1.5–1.7 GHz. Later this year, a 1.7–2.0 GHz version will become available.

Because of the single base input, high power and diffused emitter resistance, parallel connected LFE15600Xs can develop transmit powers of hundreds of watts.
Philips Components (see page 19).

Multilayer varistors
Siemens' High Capacity (ceramic) Varistors (SHCV) are intended specifically for use in car engine management. They combine compactness with a high load capacity.

Because of improved ceramic materials, the varistors can handle currents of up to 1000 A (500 A in case of the SMD version), which makes them capable of withstanding pulse energies four times as high as those manufactured in traditional materials.

The SHCVs contain a varistor and a capacitor to protect electronic circuits against over-

voltage and electromagnetic interference pulses. The value of the capacitor may be specified as 0.47 µF, 1 µF or 1.5 µF.

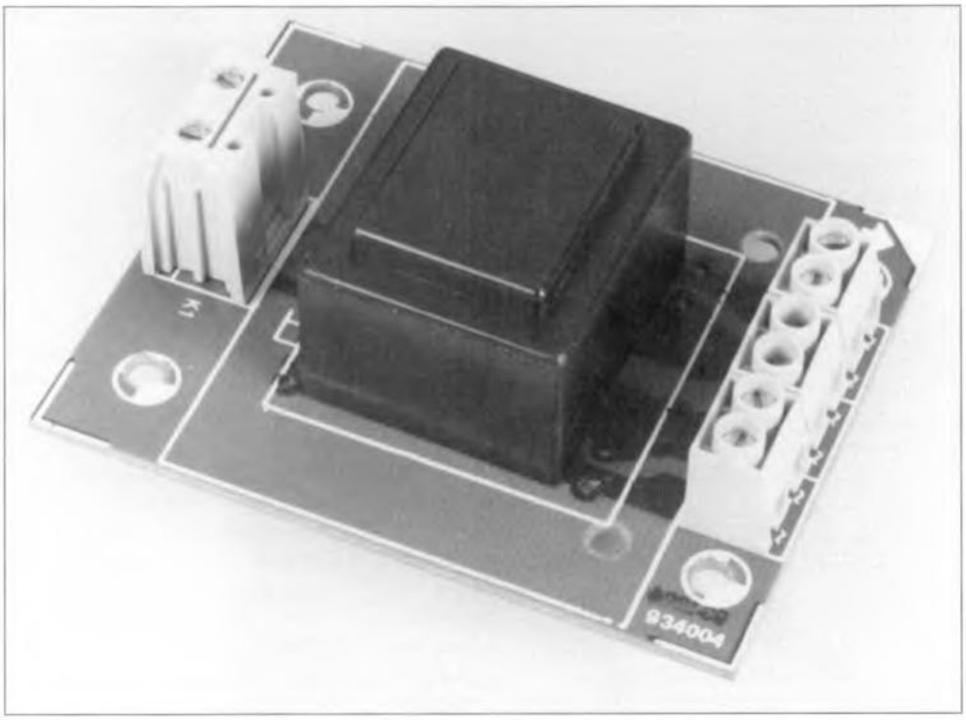
The SHVCs are available for operation with 14 V or 20 V systems.
Siemens Ltd, Siemens House, Windmill Road, Sunbury on Thames TW16 8HS.

Clock-RAM combination
SGS Thomson has expanded its range of compact integrated clock-RAM circuits with the Types M48T02 and M48T12. Both these devices contain a low-dissipation RAM with a capacity of 2K×8, a real-time clock, a quartz crystal and a lithium battery in a 28-pin housing.

The battery has a capacity of 120 mAh, which is sufficient to enable storing date and time data for at least 10 years.
When the supply voltage drops below a certain level, the RAM becomes read-only. This obviates damage to the stored data. Depending on the type, access time is 120 ns or 150 ns.

SWITCHABLE A.C. SUPPLY

Many experiments in the electronics workshop require alternating supply voltages of a certain value. If you need a relatively low voltage, an adjustable transformer is usually too bulky and also too expensive for the average hobbyist. The switchable a.c. supply presented here is a low-cost alternative.



Design by L. Pijpers

THE concept of a the present supply could not be simpler: connect a number of transformer secondary windings to a rotary switch, and you have your 'adjustable transformer' for low voltages. Such a supply is ideal for all sorts of experiments where you want to start as carefully as possible, which usually means seeing what happens at a safe, low, supply voltage. If everything seems in order, the supply voltage can be stepped up to the required level. Still on the subject of safety, the present supply is safer than a variable transformer because the latter is an 'autotransformer' which maintains a direct connection to the mains.

General transformer PCB

The basis of the present supply is the 'general transformer PCB' described in Ref. 1. Only the upper part of the circuit diagram in **Fig. 1** is used. Consequently, connector K3 is omitted from the printed circuit board (**Fig. 3**). Two 'general transformer PCBs' are used for the present application.

The actual circuit diagram of the switchable supply is shown in **Fig. 2**. Rotary switch $S_{1a/b}$ allows the secondary windings of the two transformers to be connected in series in any desirable configuration. The upshot is that you can select any of the stated a.c. voltages simply by setting the switch to the corresponding position.

Assuming that the indicated transformers are used (one with two 9-V windings, and one with two 6-V windings), the following voltages are available to choose from: 9 V and 18 V (from one winding, or two windings in series on the 9-V transformer); 6 V and 12 V (from one winding, or two windings in series on the 6 V transformer). Further voltages are: 3 V (one 6-V winding 'reverse connected' in series with one 9-V winding); 30 V (not used here; all windings in series); 21 V (not used; two 6-V and one 9-V winding in series); 24 V (two 9-V and one 6-V winding in series).

The principle should be familiar: if two or more windings are connected in series and in phase, the voltage on the outer wires is the sum of the voltages supplied by the individual windings. If one of the windings is connected 'the wrong way around' (i.e., in anti-phase with the others), its voltage should be subtracted from the rest instead of

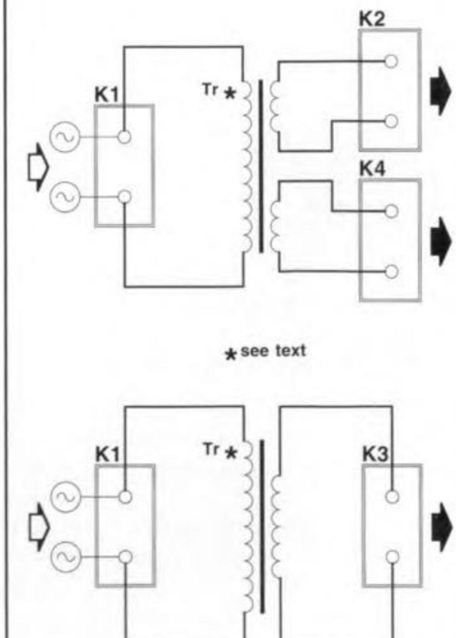


Fig. 1. Circuit diagram of the 'general transformer PCB'. Only the upper circuit is used in the present application.

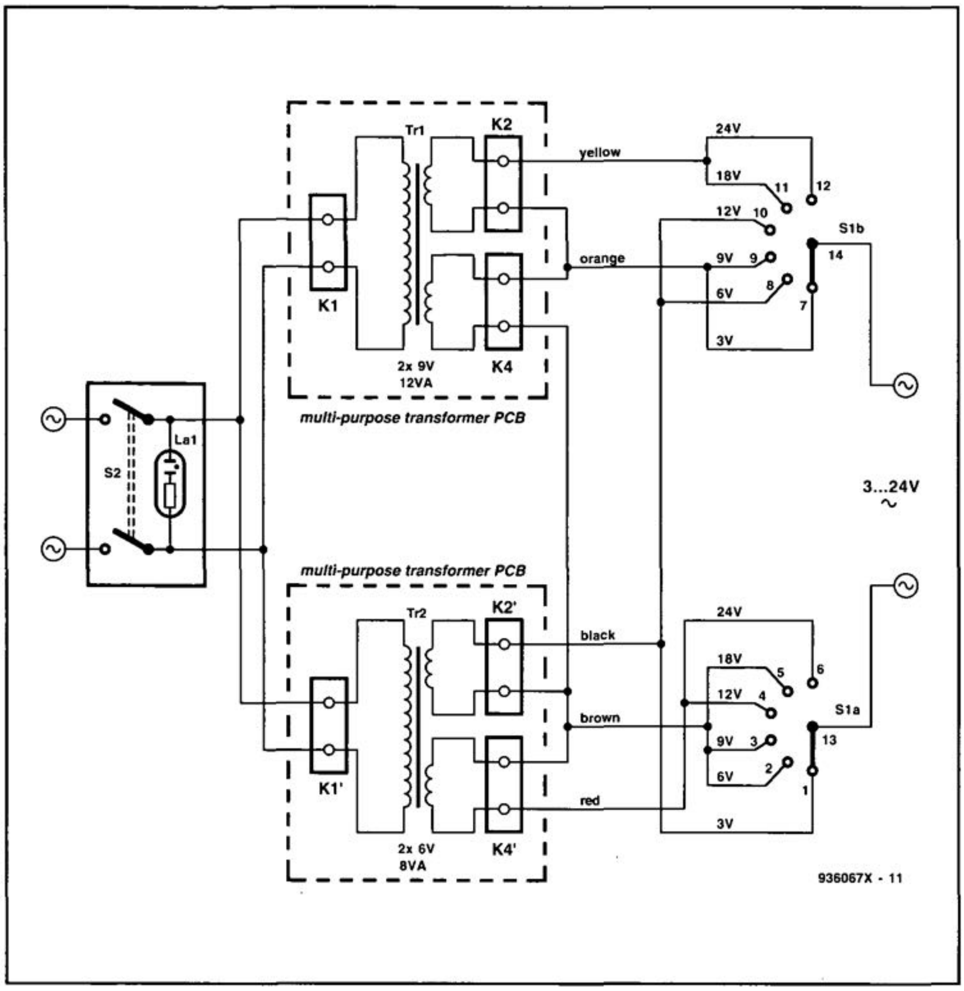


Fig. 2. Simple, low-cost and never to be missed again in the workshop: a multi-purpose a.c. supply capable of supplying a wide range of voltages.

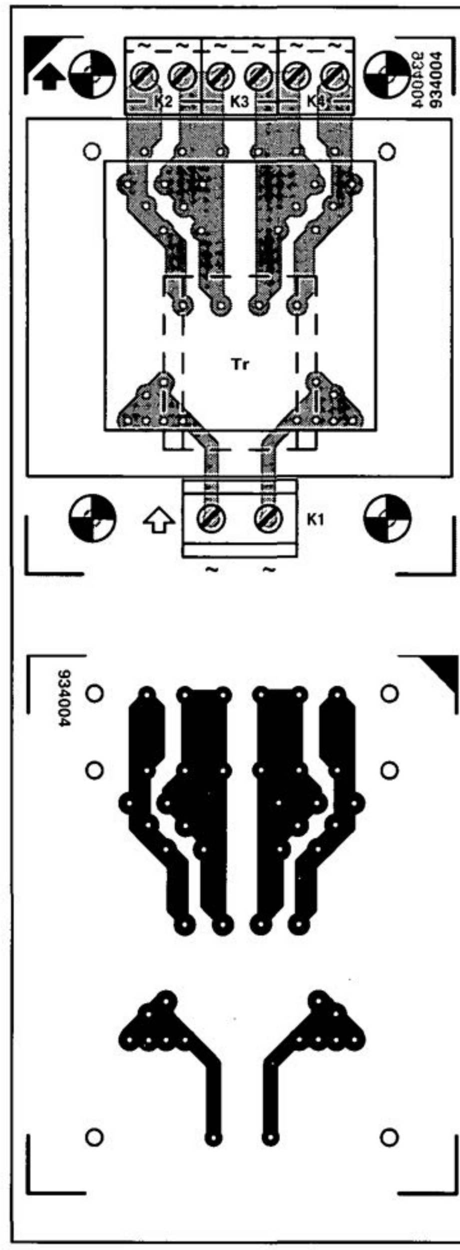


Fig. 3. Artwork of the 'general transformer PCB'. Connector K3 is not used this time. This board is available ready-made through the Readers Services under order code 934004.

added. The rotary switch shown in Fig. 2 skips two possible combinations, 21 V and 30 V, while all others, 3 V, 6 V, 9 V, 12 V, 18 V and 24-V, are available at the pole.

The prototype was built with a 2×9 V/12 VA and a 2×6 V/8 VA transformer because these supply roughly the same amount of current. Do remember, however, that the open-circuit (non-loaded) voltage supplied by each transformer may be much higher than the nominal (specified) value.

Construction

Although this a simple circuit by almost any standard, you should pay great attention to the wiring of the rotary switch. Also make sure that the primary windings of both transformers are connected the same way around to the mains; if not, the voltages at the secondary sides are in anti-phase and will not add to give the correct output values.

For obvious reasons the switch used may not be a 'make before break' type, since that causes a brief short-circuit when a different output voltage is selected. Also on the subject of short-circuits, although there are now available transformers which are specified as 'short-circuit resistant', it is still wise to use a fuse in the main line. That fuse is best located in the mains appliance socket on the rear of the case, together with the mains switch. That saves extra wiring with all the inherent risks of connections at mains potential. The 8 VA and 12 VA transformers used here require a single mains fuse rated at 125 mA, delayed action ('slow'). If you intend to use larger transformers, increase the fuse rating accordingly.

(936067)

Reference:

1. General transformer PCB. *Elektor Electronics* July/August 1993.

COMPONENTS LIST

S1 = 2-pole 6-way rotary switch.
S2 = double-pole mains on/off switch w. internal neon indicator.
Tr1 = 2x9V 12VA PCB mount transformer (see text).
Tr2 = 2x6V 8VA PCB mount transformer (see text).
K1;K1' = 2-pin PCB terminal block, raster 7.5mm.
K2;K2';K4;K4' = 2-pin PCB terminal block, raster 5mm.
K3 = mains socket with integral fuseholder.
PC1;PC2 = wander socket.
Two PCBs type 934004 (see page 78)

BATTERY CONDITION TESTER

Determining the amount of energy (charge residue) in a battery is fraught with pitfalls. If you want to keep things simple, an accurate voltage monitor is probably the best solution. Such a circuit is described here. It is compact enough to be permanently fitted even where space is tight, and affords a reliable 'live' test on the battery condition.

Design by K. Preiss

ALTHOUGH also suitable for car batteries and other applications, the simple battery tester described in this article was originally designed as a permanent add-on to the receiver battery in a model airplane. Contrary to what 'outsiders' may believe, the condition of the receiver battery in a model plane is far more important than that of the engine battery. The reason is simple: if the receiver battery goes flat during a flight, the model goes out of control, and almost certainly heading for a crash.

Before every flight, there is the dilemma of whether or not to exchange the receiver battery with a fresh (i.e., fully charged) one. If the model owner is not sure about the number of flight hours, he or she will usually decide to remain on the safe side. Unfortunately, if no spare battery is available, that means time lost on a charging cycle.

Obviously, a reliable, instant, test method for battery capacity should be of great interest to many modellers. There is little point in simply measuring the battery voltage using a voltmeter or a multimeter, since that does not give you the information you need. An non-loaded NiCd (nickel-cadmium) battery will keep up a nominal cell voltage of 1.2 V for a very long time, although it may well be three-quarters 'empty'. By contrast, an accurate voltmeter connected to the **loaded** battery does produce a usable readings. If this voltmeter is made small enough to mount it near the receiver or the battery, or, in any case, inside the model, you have a battery condition tester which operates under actual working conditions. A very simple test to see if the battery can go airborne is to actuate two servos while the model is still on the ground, and observe the voltmeter indication. If the voltage remains almost constant, you may safely assume that the battery will last for at least one flight.

Electronic magnifying glass

The open-circuit voltage variation of a NiCd battery is relatively small during a discharge cycle. The voltage variations at a number of discharge currents on a penlight (AA size) battery are shown in **Fig. 1**. The typical discharge current drawn from a model airplane receiver battery is between the 0.2*C* and 1*C* curves.

The graph shows that the entire discharge cycle covers only the voltage range between 1.1 V and 1.3 V per cell. A fully charged cell will produce an open-circuit voltage of about 1.4 V, which drops to about 1.3 V fairly quickly when a load is connected. A voltage of 1.2 V is typically maintained during the larger part of the discharge period, until the voltage starts to drop sharply. A voltage of 1.1 V means that the cell is nearly 'flat', while a voltage of 1 V indicates a completely exhausted cell.

The conclusion from the above examination is that the short voltage variation between about 1.1 V and 1.3 V tells the whole story on the battery condition. Thus, if we want to make a compact voltmeter for a set of, say, four series-connected penlight batteries, the voltage range should be 4.4 V to 5.2 V. Values over and under these limits are of little interest, because full is full, and flat is flat! The function of a voltmeter designed to select and monitor such a small voltage range may compared to that of a magnifying glass.

One IC and ten LEDs

The fact that only a very small voltage range needs to be watched allows an accurate indication to be obtained without resorting to an expensive and/or complex readout device.

The heart of the circuit is a Type LM3914 dot/bar display driver. The LM3914 is an 18-pin integrated circuit capable of converting an analogue input voltage into drive signals for a linear display consisting of 10 LEDs. As illustrated in **Fig. 2**, the internal circuit of the LM3914 is both ingenious and simple, consisting of ten comparators which are fed with a reference voltage via a precision resistor ladder network. The –inputs of the opamps are connected to the analogue

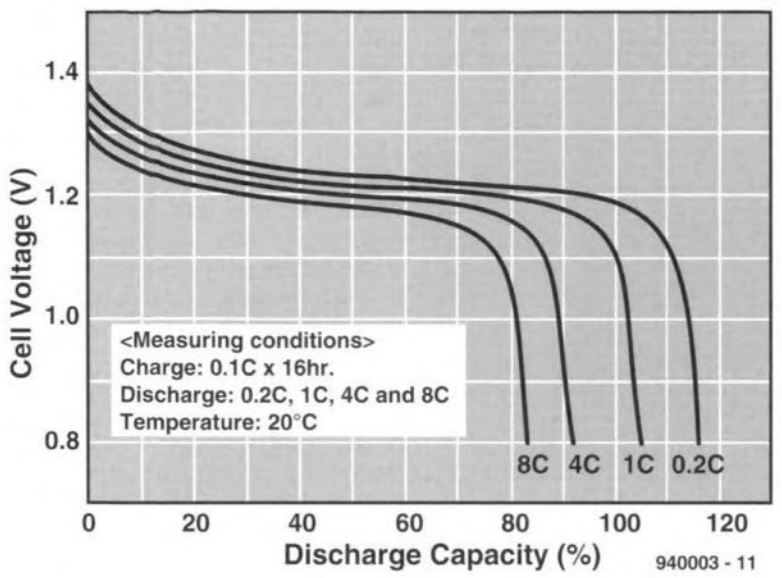


Fig. 1. Voltage supplied by a NiCd penlight battery at load currents of 8, 2, 1 and 0.2 times the nominal battery capacity (in mA/h). Although the exact condition of the battery is difficult to establish in the 'flat' area of the graph, it is clear that a cell voltage of 1.3 V means that the battery is 'full', while a value of 1.1 V indicates a flat battery.

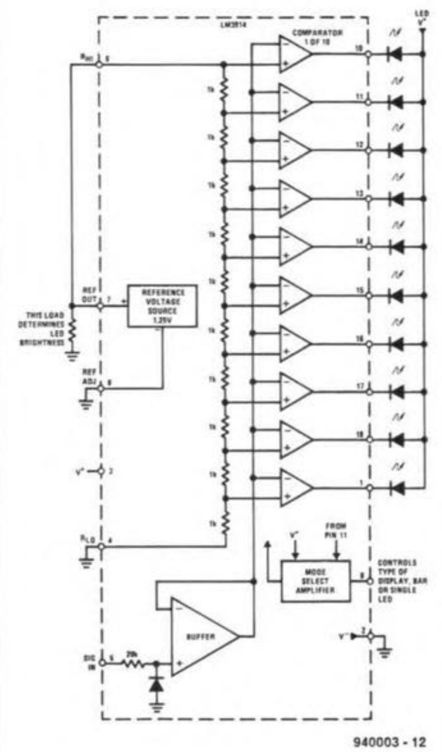


Fig. 2. Internal schematic of the LM3914 (courtesy National Semiconductor).

input via a buffer stage. The LEDs are driven directly by the comparator outputs. That is all. Pin 9 of the IC allows you to select between a bar-type and a dot-type readout. The internal circuit of the LM3914 arranges for a small overlap (approx. 1 mV) between the display segments. This is done to prevent all displays being out in certain drive conditions.

The complete circuit diagram of the battery tester is given in **Fig. 3**. As you can see, only very few external components have to be added to the LM3914.

The reference voltage is defined with the aid of voltage divider R_2-R_3. The total value of the two resistors also determines the brightness at which the LEDs light. Resistor R_4 takes the 'lower end' of the resistor ladder to ground, and thus determines the lower limit of the voltage window shown by the LEDs. The voltage divider formed by R_1 and R_5 is based on four series-connected penlight batteries. Diode D_1 and capacitor C_1 are added to eliminate sudden voltage variations.

If difficult to obtain, the LED bar may be replaced by ten discrete, rectangular, LEDs. The LED colour of the LED bar is indicated by the first letter of the type number: 'R' for red, 'G' for green, 'Y' for yellow, and 'O' for super-red. Here, the readout is used in 'dot' mode. If you prefer 'bar' mode, simply connect pin 9 of the LM3914 to pin 3 (V+). In 'bar' mode, the supply voltage should not be made to high, else the dissipation of the IC becomes too high.

Construction and use

Building the battery tester is simple if the printed circuit board shown in **Fig. 4** is used. Unfortunately this board is not available ready-made, so you have to produce it yourself, or have it produced. The only construction aspect that must be mentioned is that the LED bar is fitted at the solder side of the board. The bar may be sol-

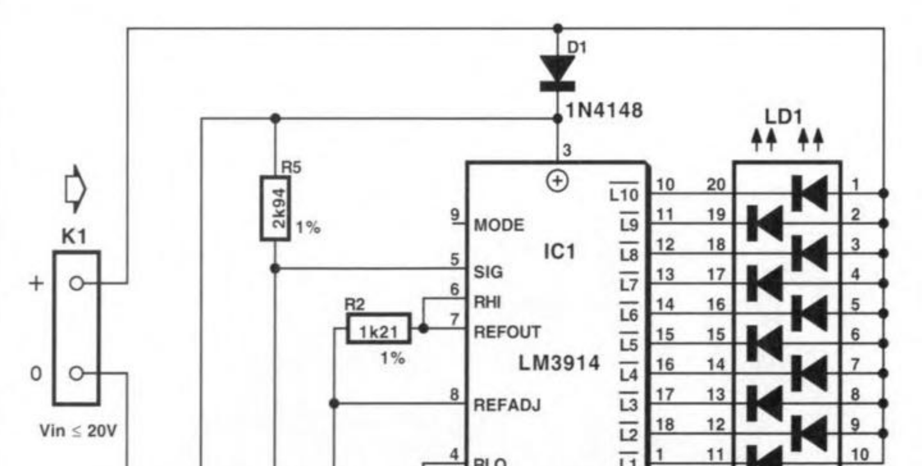


Fig. 3. Circuit diagram of the battery condition tester.

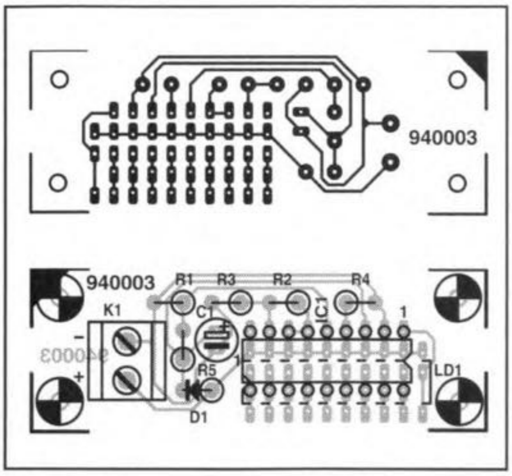


Fig. 4. Track layout and component mounting plan of the printed circuit board designed for the battery condition tester (not available ready-made through the Readers Services). Note that the LED bar must be fitted at the solder side of the board.

COMPONENTS LIST

Resistors:
R1 = 1kΩ00 1%
R2 = 1kΩ21 1%
R3 = 47kΩ5 1%
R4 = 53kΩ6 1%
R5 = 2kΩ94 1% (see table 1)

Capacitor:
C1 = 22µF 15V radial

Semiconductors:
D1 = 1N4148
IC1 = LM3914
LD1 = LED bar GBG1000 (green) or OBG1000 (red)

Miscellaneous:
2-way PCB terminal block, raster 5mm.

No. of cells	R_5	U_{min}	U_{step}	U_{max}
4	2kΩ94	4.4	0.08	5.2
5	3kΩ92	5.5	0.1	6.5
6	4kΩ87	6.6	0.12	7.8
7	5kΩ90	7.7	0.14	9.1
8	6kΩ81	8.8	0.16	10.4
9	7kΩ87	9.9	0.18	11.7
10	8kΩ87	11	0.2	13
Car battery (R_4=26kΩ1)	10kΩ5	11.1	0.41	15.2

Resistor values from E96 series.

Table 1. Design data to modify the circuit for other battery voltages.

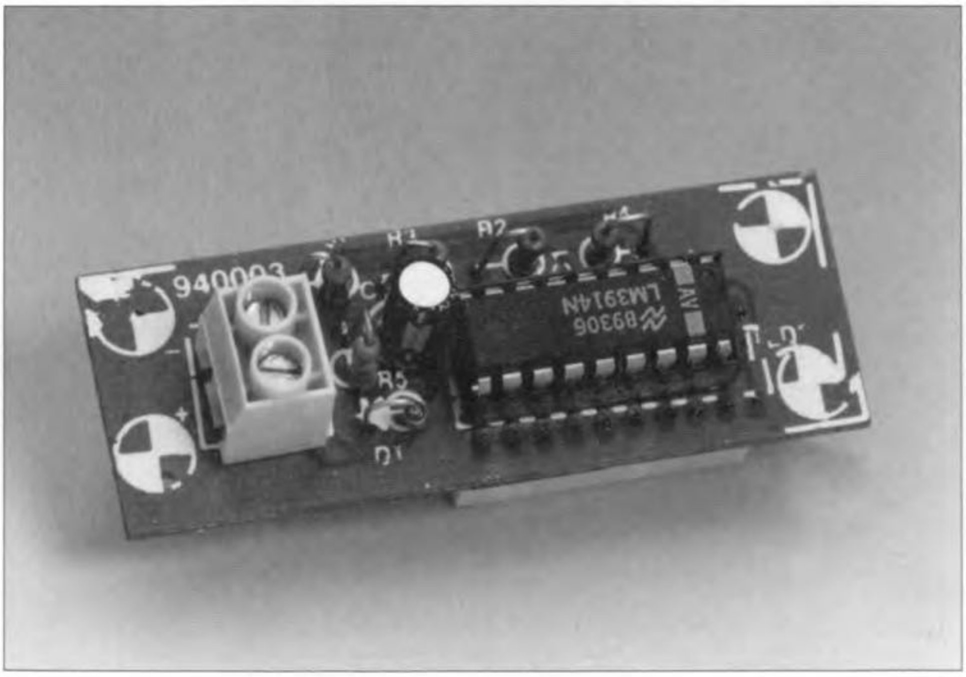


Fig. 5. Completed prototype photographed before it was fitted into a model aeroplane.

dered directly to the board, or inserted into a suitable socket. The completed prototype is shown in **Fig. 5**.

Since most modellers will know more about building the circuit into an model airplane than the engineers on the *Elektor Electronics* design staff, that subject is safely left to the specialists. As long as the LEDs are clearly visible from the outside, the exact location of the battery tester in the model is of little importance. Since the current consumption of the tester is only sa few milliamps in 'dot' mode, an on/off switch is not required, and the circuit may be permanently connected in parallel with the receiver.

The tester is simple to use. Switch on the receiver and check which LED lights — probably the upper one or the next lower one. Operate the transmitter so that at least two servos are actuated. The indication may not drop by more than two LEDs. If it drops below the centre of the bar, the battery voltage is dangerously close to 1.1 V per cell. The amount of energy left in the battery is then almost certainly insufficient for a flight.

Other voltages

As already mentioned, the component values indicated in Fig. 3 are based on the assumption that a four-cell NiCd battery pack is used with a nominal voltage of 4.8 V. The lower limit of the voltage window lies at about 4.4 V, and the upper limit at about 5.2 V.

Fortunately, the tester is easily modified for other battery voltages, since that only requires changing R_5. **Table 1** shows the value of R_5 for any series connection of four up to ten NiCd cells, and for use with a car battery. In the latter case, the value of R_4 needs to be changed, too. In the table, U_{min} and U_{max} indicate the bounds of the voltage window. For the sake of completeness, the 'step' size between two adjacent LEDs is also indicated.

If you modify the circuit for different battery voltages you may run into a small problem. When checking the size of the voltage window, you may find that the lower value is not as specified, although the upper limit corresponds to the value given in the table. This error is caused by production tolerances on the LM3914, and no cause for alarm. If you think the error is too large, the necessary correction is easily made by changing the value if R_4 a little.

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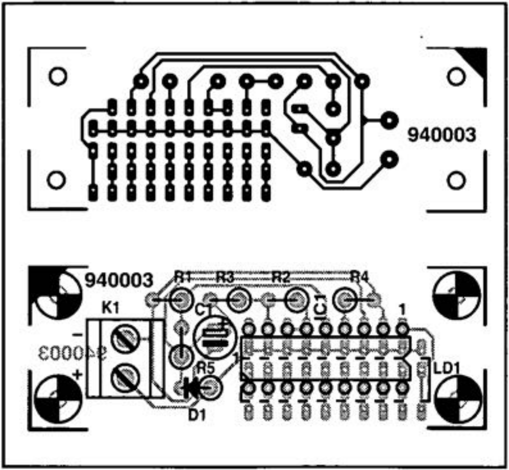
940003
940003
R1
R3
R2
R4
K1
C1
IC1
1
R5
LD1
D1