THE INTERNATIONAL MAGAZINE FOR ELECTRONICS ENTHUSIASTS

September 1990

UK £1·70

RS

Elektor Electronics

S-VHS RGB converter

High-current H_{fe} tester

LF-HF test probe

Isolation amplifier

Inter-IC communications

Make your own PCBs

RESULTS OF OUR SUMMER COMPETITION

ISSN 0268-4519

09

9 770268 451012

Elektor Electronics

CONTENTS

September 1990
Volume 16
Number 181

In our next issue:

- Guitar tuner
- Chopper opamps
- Negative resistance
- Measurement techniques
- Single/double symmetrical power supply
- μP controlled telephone exchange
- Medium power audio amplifier
- Dubbing mixer
- Selective preamplifier

Front cover

Electronic and electrical engineering involves people and is for people – it covers a wide spectrum of activities from the construction of the smallest piece of electronic equipment, hardly visible to the naked eye, to work on huge power stations generating electricity for a whole community.

Electricity and its associated technology is so much a part of our day-to-day life that we often take it for granted. It contributes to the manufacture of almost all products from paper to food, from oil cans to oil rigs, from push chairs to space vehicles. It helps to make our daily life more comfortable by running household appliances; it enables us to communicate, whether to our friends at the other side of town or to business colleagues on the other side of the world; it contributes to our health in the control and operation of hospital equipment.

There are career prospects for many able men and women in electronic and electrical engineering – see **Careers in Electronics** on page 60.

We regret the omission during printing of the following warning at the end of "Mini FM Transmitter" published in the July/August 1990 issue (3rd column, page 35).

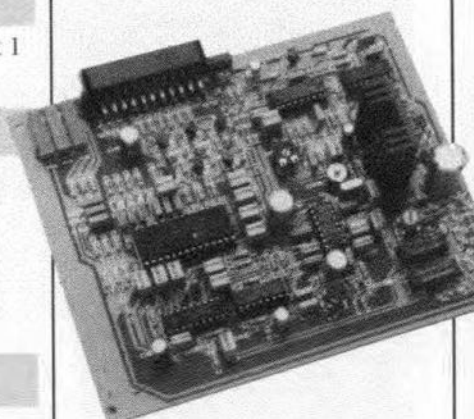
WARNING

Readers are advised that, although the use of the Mini FM Transmitter presents no problem in many countries, its use in the United Kingdom is not allowed. It may also not be in some other countries and advice should therefore be sought from the relevant authorities before construction is begun.

Anyone using the equipment in the United Kingdom faces possible prosecution under the relevant section of the Wireless Telegraphy Act.

As a consequence, printed-circuit board 896118 will be available only outside the UK.

Infra-red remote control – p. 23



S-VHS/CVBS/PAL-to-RGB converter p. 46

DIGITAL CAR ENGINE LOCK WITH ALARM

The circuit described here is a car theft deterrent that locks the starting motor until a pre-programmed code is recognized.

P.U. Mahesh

The code fed into the memory of the car engine lock is retained in a memory until it is intentionally cleared by the rightful owner of the car.

The operation of the circuit is relatively simple. Bistable IC1A-IC1B forms a debounce circuit for the clock pulses generated by S1 while the code is keyed in. The preset code is latched in memory IC2 and the code entered is decoded by IC3. If the preset code matches the code entered, and the ignition key is switched on, thyristor Th1 is provided with gate current, and fires so that the starting motor is powered. When no code or a wrong code is keyed in with the ignition switch on, Th2 fires and actuates the horn.

The operation of the circuit may be illustrated by assuming that code 0101 (example) is to be entered. The sequence in which the switches are pressed is as follows:

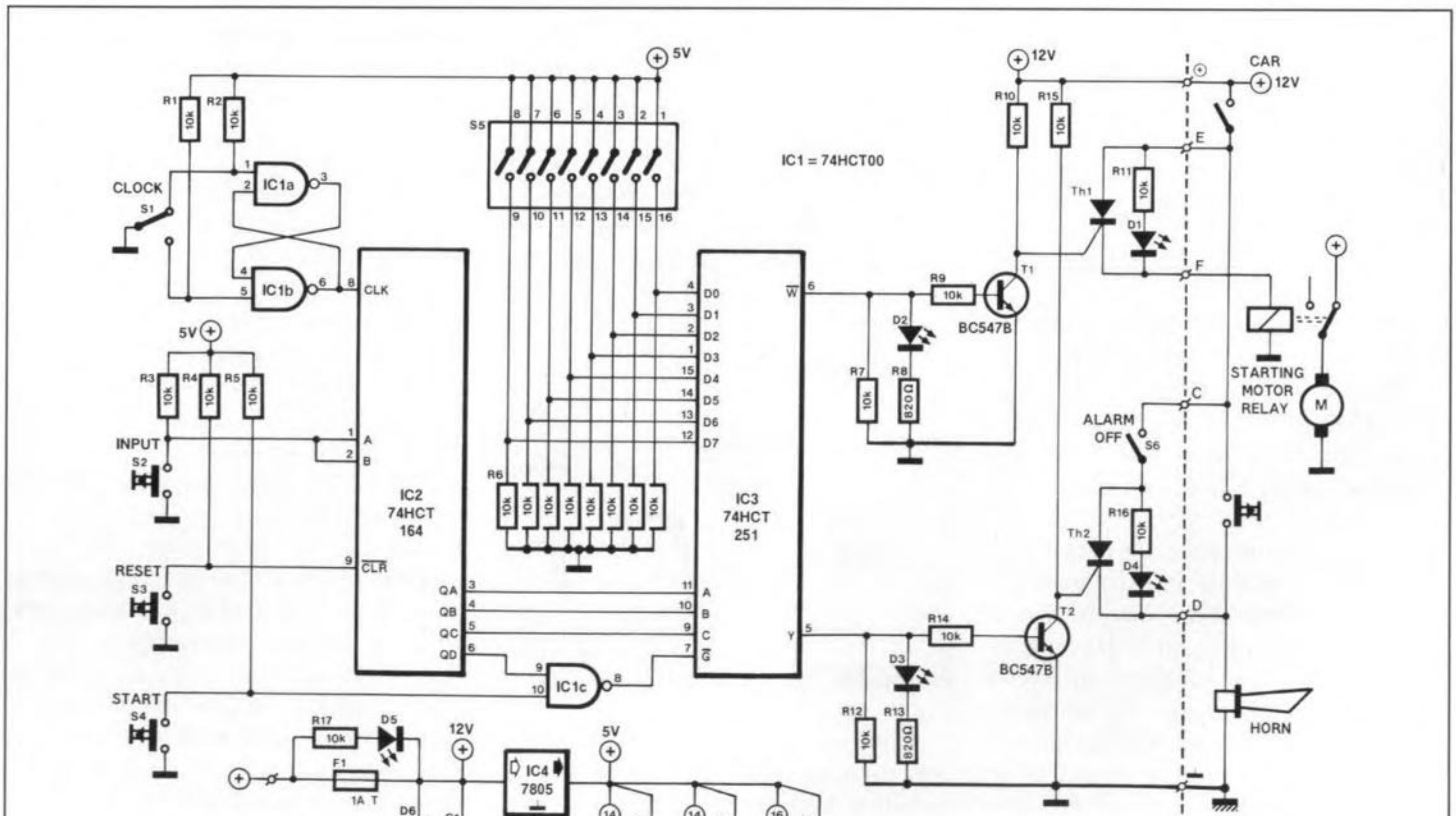
(S1) → (S1) → (S2, S1) → (S1) → (S4 , ignition)→ start

Here, (S2, S1) means that S2 is pressed, S1 is pressed, S1 is released, and S2 is released in that order. Note that pull-up resistor R3 ensures that a '1' is loaded when only S1 is actuated. The least-significant bit (LSB), which is keyed in first, is not used in IC4, so that the data is actually 010. Assuming that dataline D4 of IC3 is logic high because the associated switch in S5 is closed, the preset code matches the code entered. When the START switch, S4, is pressed while output QD of IC2 is high, multiplexer IC3 is enabled via its $\overline{G}$ input by a low level supplied by NAND gate IC1C. Since the code is right, the Y output (pin 5) goes high, and the $\overline{W}$ output goes low. A green LED, D3, lights to indicate that the correct code has been entered. Transistor T1 conducts and keeps the gate of Th2 low. At the same time, the low level at $\overline{W}$ of the multiplexer turns off T1 so that Th1 is fired via R10.

When the wrong code is keyed in, outputs $\overline{W}$ and Y of the multiplexer are high and low respectively. Upon turning the ignition key, Th2 is fired, and the horn sounds to alert passers-by and the owner of the car that someone is attempting to steal the vehicle.

Upon leaving the car, the owner must actuate the lock and the alarm by pressing S3. A standard 5 V regulator is incorporated into the circuit. LED D5 lights when the associated fuse, F1, blows as a result of a short circuit.

The complete circuit is easily built on a piece of veroboard. It is recommended to check the operation of the digital circuity before connecting the transistors and the thyristors. ■



INTER-IC COMMUNICATIONS: THE I^{2}C BUS

Reduction of the number of interconnections on a printed-circuit board results in lower production costs and increased reliability. Well aware of this fact, manufacturers of consumer electronics have sought ways to make inter-IC communications less complex. Philips Components have found a solution in the form of their patented I^{2}C bus, the main features of which are described in this article.

H. Stenhouse

An increasing number of complex integrated circuits, ranging from real-time clocks to frequency synthesizers, is provided with an I^{2}C interface. Not surprisingly, the I^{2}C bus is found in a wide variety of electronic equipment, including telephones, car radios, television sets and video recorders.

The aim of this article is to provide an introduction into the main features and communication protocols of the I^{2}C network. The acronym I^{2}C stands for Inter-IC Communication, and the network was developed by Philips to reduce the number of connections between integrated circuits. This proved feasible in practice mainly because many ICs have a large number of pins that carry information that is not time-critical and, therefore, suitable for conveying via a relatively slow serial bus with fewer connections than would be required for a high-speed parallel interface. The implementation of the I^{2}C bus on a real-time clock chip, for instance, may reduce the number of pins from 40 to as few as 8. This results in a much simpler PCB design with all the benefits of lower production cost and smaller risks of faults developing in equipment. However, a number of connections, including those for the supply voltage, for clock signals, etc., can not be replaced by a serial communication protocol. It will be clear that these connections remain necessary as before.

All ICs that use the I^{2}C bus are in principle connected to two lines as shown in the example application in Fig. 1. A central bus interconnects two microcontrollers, a memory, a gate array and an LCD driver.

In spite of their wide diversity as regards function and application, all I^{2}C-compatible integrated circuits have one common feature: all control commands and data are conveyed via a serial bus, according to a predefined communication protocol. The serial bus takes the form of three lines: ground, clock (SCL) and data (SDA).

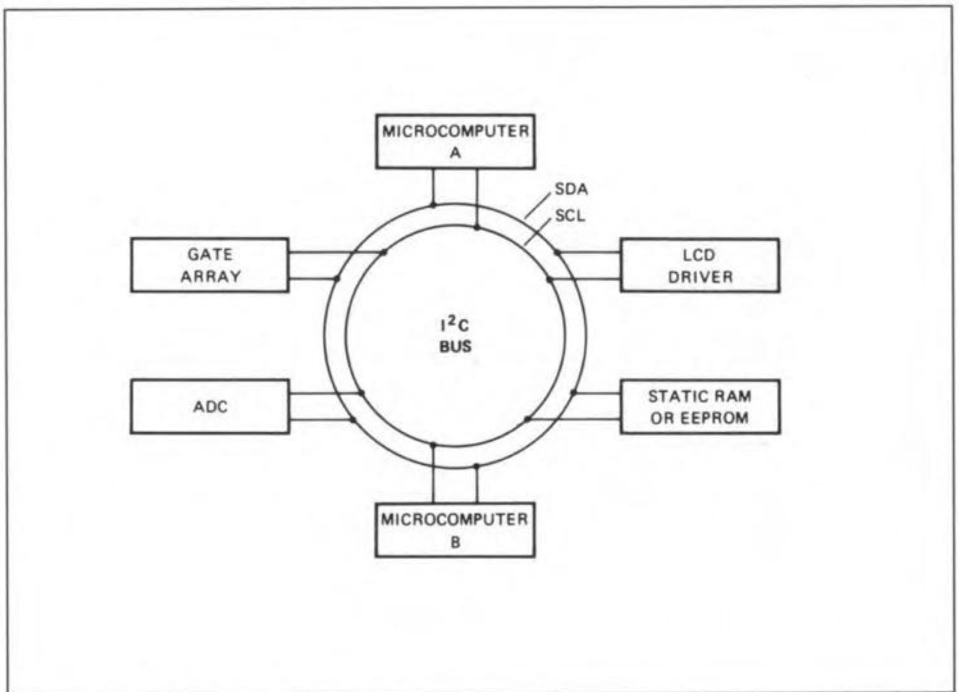


Fig. 1. Typical I^{2}C-bus configuration.

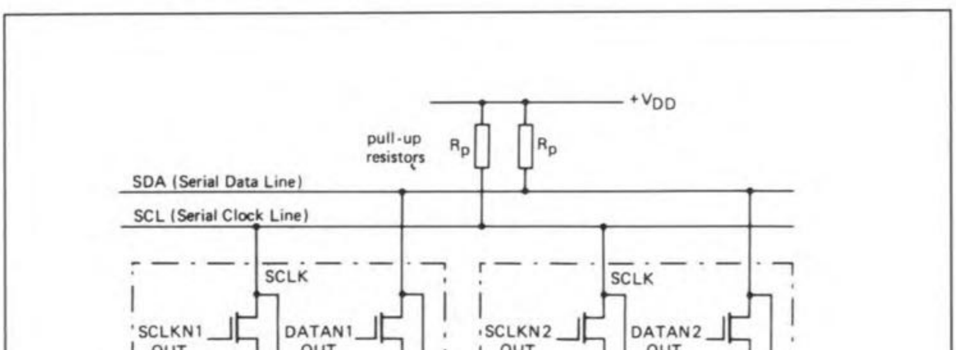


Fig. 2. Connection of I^{2}C interfaces to the I^{2}C bus.

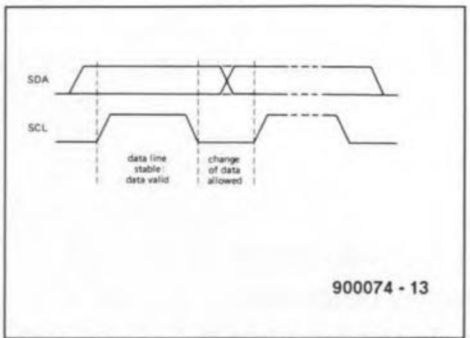


Fig. 3. Timing of bit transfer on the I^{2}C bus.

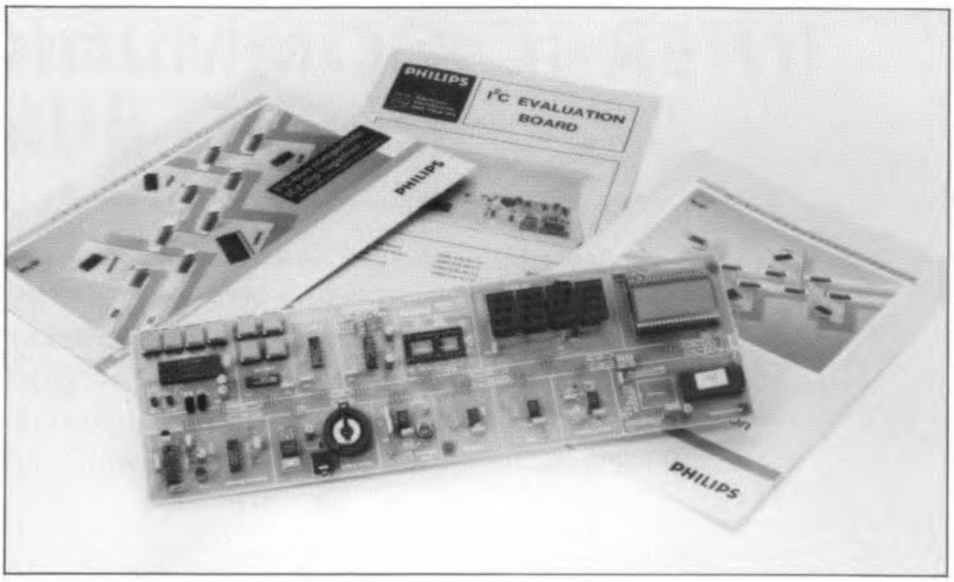


Fig. 5. Development kit for I^{2}C applications.

Normally, any I^{2}C configuration has at least one master (an IC capable of initiating the data exchange processes and generating a master clock signal) and one or more slaves (ICs that do the actual work). A master may be a microprocessor such as an 8048, an 8051 or a 68000, which are all available in special versions with a built-in I^{2}C interface. Two I/O port lines of the microprocessor are used as SDA and SCL lines. Together with the ground line, this implements an I^{2}C bus which allows serial communication between 'bused' devices at a data rate of up to 100 kbit per second.

Control programs for the I^{2}C bus

The two communication lines, SDA and SCL, are connected to open-drain or open-collector outputs, and have one, common, pull-up resistor. This arrangement is called a wired-AND structure. Adding or removing one or more I^{2}C components on the bus therefore does not affect the operation of already connected ICs, nor does it affect the software that runs on the system. In fact, the software is capable of automatic detection of the hardware configuration. This allows programs to be written for complex systems that do not provide certain features unless the relevant chips are connected to the bus. The absence of these chips is automatically detected by the master controller which interrogates certain addresses.

Existing software may be extended with subroutines written for add-on ICs without affecting the operation of the ICs already installed. This allows existing control programs to be used for a long time without the need of a completely new version every time the hardware is modified. This high level of compatibility is achieved by virtue of the fixed addresses of the ICs on the I^{2}C bus.

Two lines

Both SDA and SCL are bidirectional lines, connected to a positive supply voltage via a pull-up resistor (see Fig. 2). When all output transistors of connected devices are off, the bus is free, and both lines are high. When an IC is ready to transmit a data block, it pulls SDA low to mark a start condition. From that moment, all other ICs 'know' that the bus is in use. Arbitration procedures come into effect should two or more ICs claim access to the bus simultaneously. When the start condition is recognized, the SDA line is available for carrying databits. The clock line, SCL, determines the validity of the data levels on the SDA line as shown in Fig. 3.

The start of any data exchange via the bus is marked by SDA going low while SCL is high, i.e., by a start condition (see Fig. 4). The level on the SDA line is read by all ICs on the bus during the positive part of the clock pulse. However, only the IC selected by the transmitted address code responds to the information by actually loading the data and returning an acknowledge pulse. This pulse is generated by the addressed slave device pulling the data line low for one clock period, after the eight clock periods reserved for the databits (see Fig. 4).

When none of the ICs in the system responds to the transmitted data, the master does not receive an acknowledge pulse. This means that either the addressed slave is busy performing some real-time function, the address is wrong, or there is no device that responds at the particular address. The bus is free again after the transmission of the last data bit. Both SCL and SDA revert to high, and the bus may be used to convey the next data block.

The function of the SCL line is to generate one clock pulse for every transmitted databit. Each master must generate its own SCL signal. Although the frequency of this signal is not fixed, certain minimum timing specifications must be observed. In practice, the I^{2}C bus allows a maximum data speed of about 100 kBit/s.

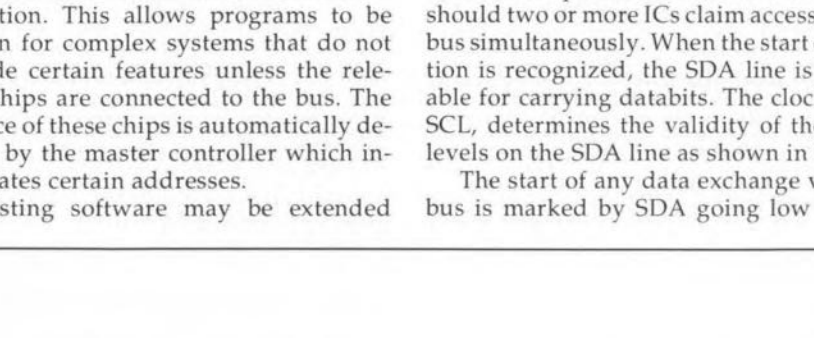


Fig. 4. Timing of data transfer on the I^{2}C bus.

Addressing

Each IC on the I^{2}C bus has its own, unique, 7-bit address, which is determined by the manufacturer and burned into the chip. The Type PCF8583 real-time clock chip, for example, is selected by sending the binary code 101000x. The last bit is not preset (x is 0 or 1) to allow two identical ICs to be used in parallel by tying their A0 inputs to ground or to the positive supply to set the address to 1010000 or 1010001 respectively. Similarly, certain ADCs, I/O chips and memories may be hard-wired to map them at one of up to eight addresses in a cluster.

The data blocks conveyed via the I^{2}C

THE RANGE OF I²C-BUS COMPATIBLE ICs

General-purpose ICs

LCD drivers

PCF8566	96-segment LCD driver; 1:1 – 1:4 MUX
PCF8576	160-segment LCD driver; 1:1 – 1:4 MUX
PCF8577(A)	64-segment LCD driver; 1:1 – 1:2 MUX
PCF8578/79	Row/column LCD dot-matrix driver; 1:8 – 1:32 MUX

I/O expanders

PCF8574	8-bit remote I/O port (I²C-bus to parallel converter)
PCF8584	8-bit parallel to I²C-bus converter
SAA1064	4-digit LED driver
SAA1300	5-bit high-current driver

Data converters

PCF8591	4-channel, 8-bit MUX ADC & one DAC
TDA8442	quad 6-bit DAC
TDA8444	octal 6-bit DAC

Memory

PCA8582B	256-byte EEPROM (automotive temperature range & error correction)
PCF8570	256-byte static RAM
PCF8570C	as PCF8570 with alternative slave address
PCF8571	128-byte static RAM
PCF8582A	256-byte EEPROM

Clocks/calendars

PCF8573	clock/calendar
PCF8583	256-byte RAM/clock/calendar

Application-oriented ICs

Video/radio/audio

PCF8200	voice synthesizer (male/female speech)
SAA1136	PCM-audio Ident word Interface (IDI) for compact disc
SAA1300	tuner switching unit
SAA3028	transcoder (RC-5) for IR remote control
SAA4700	data line processor for VPS
SAA5243*	enhanced computer-controlled teletext (ECCT) processor
SAA9050/51	digital PAL/NTSC colour decoder
SAA9055	digital SECAM decoder
SAA9062/63/64	digital deflection controller
SAA9068	picture-in-picture (PIP) controller
SAB3035/36/37	digital tuning circuits for computer-controlled TV
SAF1135	dataline-16 decoder for VCR
TDA8405/15	stereo/dual sound processor
TDA8421	audio processor with a loudspeaker channel and a headphone channel
TDA8425	audio processor with a loudspeaker channel only
TDA8433	deflection processor and sync controller
TDA8440	video/audio switch
TDA8461	PAL/NTSC colour decoder and RGB processor
TEA6100	FM/IF and digital tuning IC for computer-controlled radio
TEA6300	sound fader control and preamplifier/source selector for car radio
TEA6310T	sound fader control with tone and volume control for car radio
TSA5510	PLL frequency synthesizer for TV and VCR
TSA6057	PLL frequency synthesizer for radio

Telecomms

PCD3311/12	tone generator (DTMF/modem/music)
PCD3341	advanced 10/110-number repertory pulse/DTMF dialler with LCD control
PCD3343	microcontroller with 224-byte RAM/3K ROM
PCD3346	microcontroller with 128-byte RAM/4K ROM/256-byte EEPROM
PCD3348	microcontroller with 256-byte RAM/8K ROM
UMA1000T	data processor (DPROC) for mobile telephones
UMA1010T	frequency synthesizer (0.45 – 1.1 GHz) for mobile telephones
UMA1012T	frequency synthesizer (50 – 600 MHz) for mobile telephones

Table 1. Overview of I²C-compatible integrated circuits manufactured by Philips Components. Not shown here is a wide range of microcontrollers and memories.

bus invariably consist of 8 bits. The bit that follows the address indicates the start of a read or a write operation with the selected IC. Bit 8 is low for a write operation, and high for a read operation.

Applications

There is much more to the concept of the I²C bus than can be described in this article. The full specification of the system may be found in Ref. 1.

The I²C bus is relatively simple to implement on almost any microcomputer system that has at least one user port. If necessary, external buffers may have to be added to make such a port bidirectional. Interestingly, Philips Components recently introduced a special chip for this purpose: the PCF8584. Some microcomputers, including the Acorn Archimedes, even have an I²C interface as a standard feature. Developers of small stand-alone microprocessor systems may find the I²C version of the 8048, the PCF84C00T, a good starting point for the design of a dedicated control system.

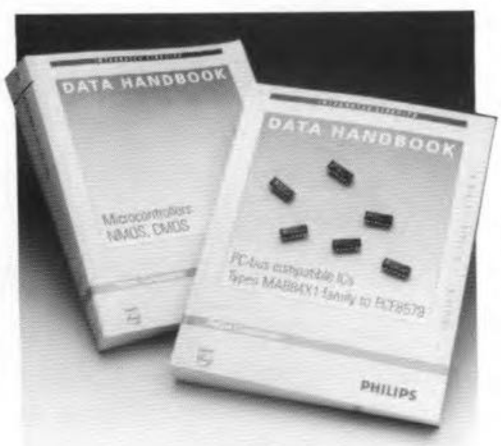


Fig. 6. Databooks on I²C devices, published by Philips Components.

Finally, an interesting example of the use of a 8051 microcontroller in combination with the SAA5243 I²C-bus Teletext decoder may be found in Ref. 2. ■

Reference:

1. *The I²C-bus Specification*. Philips Components publication.
2. *Computer-controlled teletext decoder*. Elektor Electronics October 1989

SOUND GENERATOR

The sound generator described here, designed and marketed as a kit by ELV, is capable of producing up 256 different siren-like sounds, including the popular Kojak-, FBI-, and Hawaii-Five-0 types. Compact, easy-to-build and suitable for use in conjunction with alarm systems in and on premises as well as on vehicles, the unit is complete with an on-board 20-watt amplifier.

The type of sound is selected with four slide switches on the front panel of the sound generator. Since each slide switch has four positions, a total number of 256 (4×4×4×4) different sounds are available. An output stage is included in the circuit to provide a solid 20 watts of audio power at a supply voltage of 12 V to 15 V. The slide switch at the extreme left on the front panel functions as a three-level volume control and as an on/off control.

Circuit description

Circuit IC1, a Type NE556, contains two multivibrators. One of these, IC1b, generates the basic siren sound. Switch S4 allows four different basic sounds to be generated by selecting one of four timing capacitors C7–C10. The output of IC1b, pin 9, drives the power output transistor, T1, direct via resistor R14. Depending on the position of volume switch S5, the loudspeaker is either disconnected ('off'), connected direct to the collector of T1 (volume level 3), or connected via series resistors R15 or R16 (volume levels 2 and 1).

Evidently, a single oscillator does not make a siren, let alone one capable of producing up to 256 different sounds. Circuit IC1b, is, therefore, frequency-modulated by applying a signal to its control voltage input, pin 11. This modulation signal is supplied by a second oscillator, formed by the parts to the left in the circuit diagram.

The second multivibrator in the circuit, IC1a, operates at a much lower frequency than IC1b. The oscillation frequency is determined by one of four capacitors C1–C4 connected to IC1a via the 'frequency' switch, S1. The other frequency-determining parts are R1 and R2, which set the charge and discharge periods respectively.

When S2 is set to the position shown in the circuit diagram, R3 is connected in parallel with R2, so that the input of buffer opamp IC2 receives a sawtooth signal. In the other extreme position, i.e., when S2 is set to the top position, R3 is not connected so that a triangular waveform is produced. The two centre positions of the switch produce a rectangular waveform and a combined rectangular/logarithmic waveform (as shown inset in Fig. 1). The latter is obtained with the aid of components C6, R7 and R3.

Opamp IC2 forms a buffer between the modulation waveform generator, IC1a, and the tone generator, IC1b. The level of the modulation signal fed to IC1b is determined by the position of switch S3, which connects one of four series resistors R8–R11 between the output of IC2a and pin 11 of IC1b. Switch S3 thus determines the modulation intensity.

Summarizing the above, the functions of the slide switches in the circuit are as follows (front panel marks in brackets):

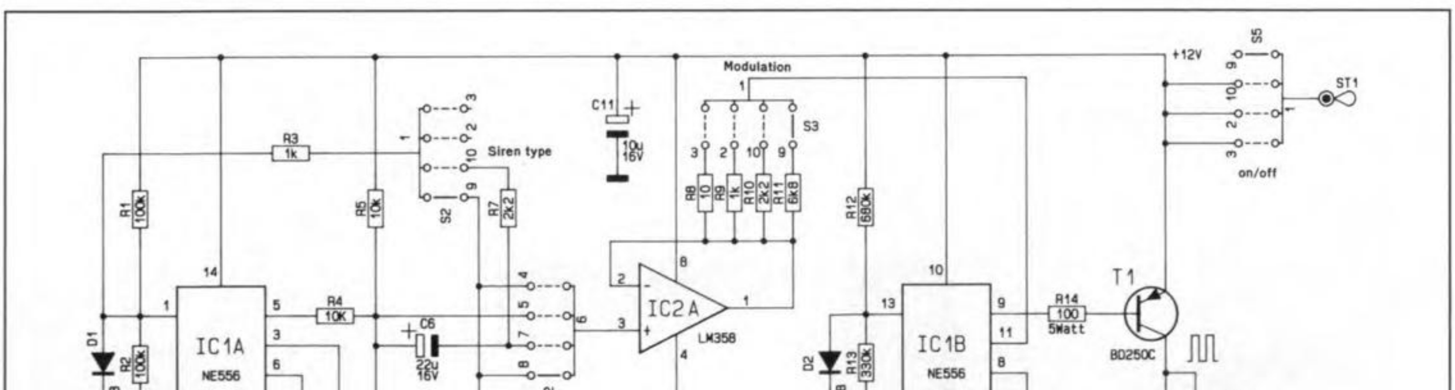


Fig. 1. Circuit diagram of the sound generator.

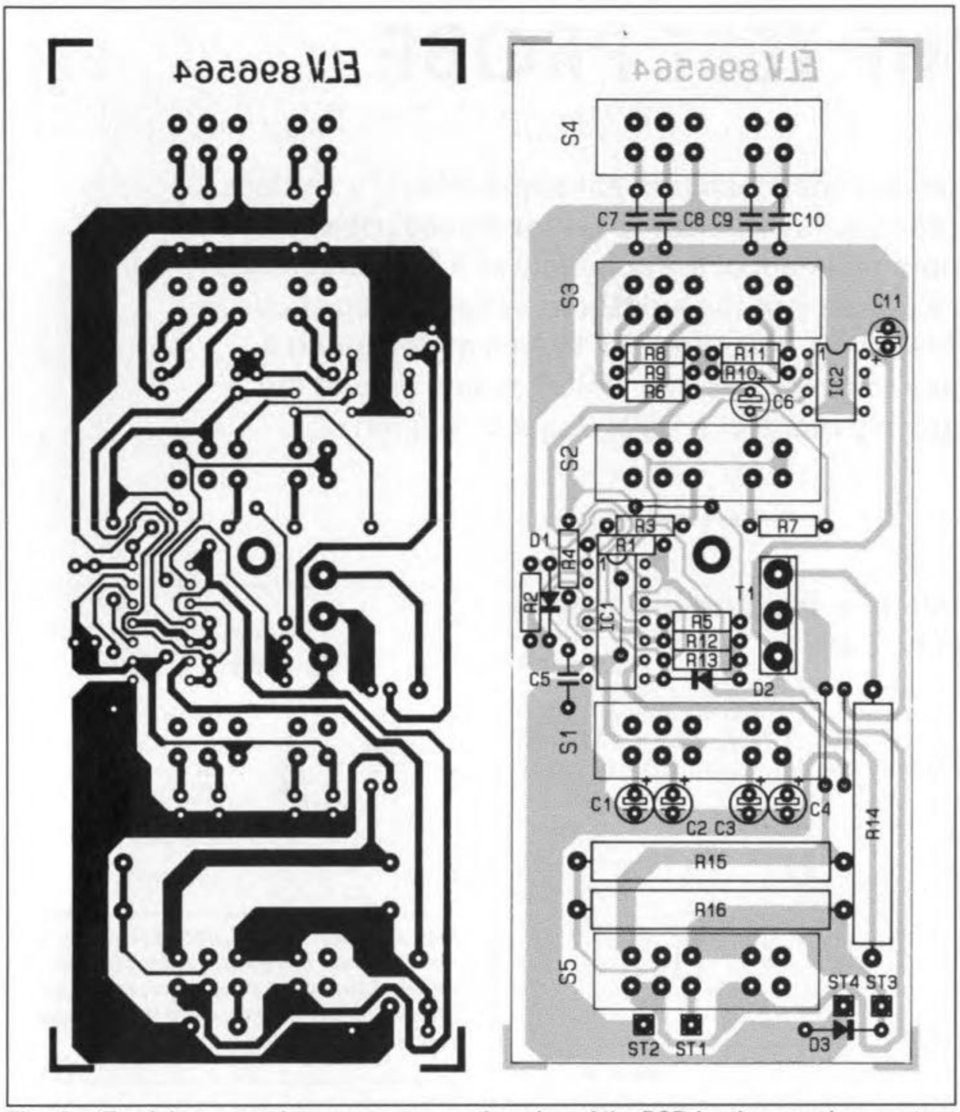


Fig. 2. Track lay-out and component mounting plan of the PCB for the sound generator.

COMPONENTS LIST

content of kit supplied by ELV France

Resistors:		
1	6.8Ω 5W	R15
1	10Ω	R8
1	22Ω 5W	R16
1	100Ω 5W	R14
2	1k	R3;R9
2	2k2	R7;R10
1	6k8	R11
1	9k1	R4
3	10k	R4;R5;R6
2	100k	R1;R2
1	330k	R13
1	680k	R12
Capacitors:		
1	1nF	C9
1	1n5	C10
1	2n2	C8
2	4n7	C5;C7
1	22nF	C11
1	1μF 16V	C4
1	2μ2 16V	C3
1	4μ7 16V	C2
1	10μF 16V	C1
1	22μF 16V	C6
Semiconductors:		
1	NE556	IC1
1	TLC271	IC2
1	BD250C	T1
1	1N4001	D3
2	1N4148	D1;D2
Miscellaneous:		
5	2-pole 4-way slide switch	S1–S5
4	solder pin	
1	printed-circuit board	
1	enclosure	

- S1 (frequency): modulation frequency
- S2 (siren type): modulation waveform
- S3 (modulation): modulation intensity
- S4 (basic sound): fundamental siren frequency
- S5 (volume): sound level and on/off control

The four switches S1–S4 allow 4^4=256 different sounds to be generated at three volume levels.

For the highest possible sound level (particularly in alarm systems), it is recommended to use a pressure-chamber type loudspeaker with a sufficiently high power rating (≥20 W). For other applications, standard loudspeakers may be used with good results. The minimum loudspeaker impedance is 4 Ω.

Construction

The sound generator is a relatively simple circuit which should not present difficulties in assembling. Moreover, the unit is supplied in kit form, which obviates problems with obtaining certain components.

Start the construction by fitting and soldering the low-profile parts, followed by the higher parts, on the single-sided printed circuit board shown in Fig. 2. The overlay printed on the component side of the board indicates the position of the parts mentioned in the parts list.

To assist in their cooling, the 5-W power resistors are mounted at a small distance above the printed-circuit board.

The use of a relatively flat enclosure makes it necessary to bend the power transistor, T1, towards the PCB surface as shown in the photograph of the completed board. By virtue of its low internal resistance, and the fact that it is driven at a fairly high level, T1 dissipates relatively little heat, even at full output power. Consequently, the transistor does not require a heat-sink.

After a careful visual check of the completed board, this may be fitted into the enclosure supplied with the kit. Connect the supply voltage to PCB terminals ST1 (+12 V to +15 V) and ST2 (ground). Connect the loudspeaker to terminals ST3 and ST4. Drill holes in the enclosure to pass the supply wires and the loudspeaker wires. Make knots in the wires at the inside of the enclosure to provide strain reliefs. Finally, fit the top half of the enclosure and secure it with the screws supplied.

Practical use

When a 4-Ω loudspeaker is used, the unit draws a peak current of up to 4 A. When used in a switched circuit, e.g., as a horn, the sound generator may be powered via a push-button or a relay with a suitable contact current rating. Use as a horn is possible because the siren starts to sound the moment is it powered. It should be noted, however, that in many countries the use of a siren as a sound actuator device in or on vehicles, and in some cases in or on premises as well, is restricted to emergency services. The use of a siren in general may also be subject to special licenses, rules or regulations as regards on-time, sound type and sound level. ■

A complete kit of parts for the sound generator is available from the designers' exclusive worldwide distributors:

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82837213
Fax: +33 82838180

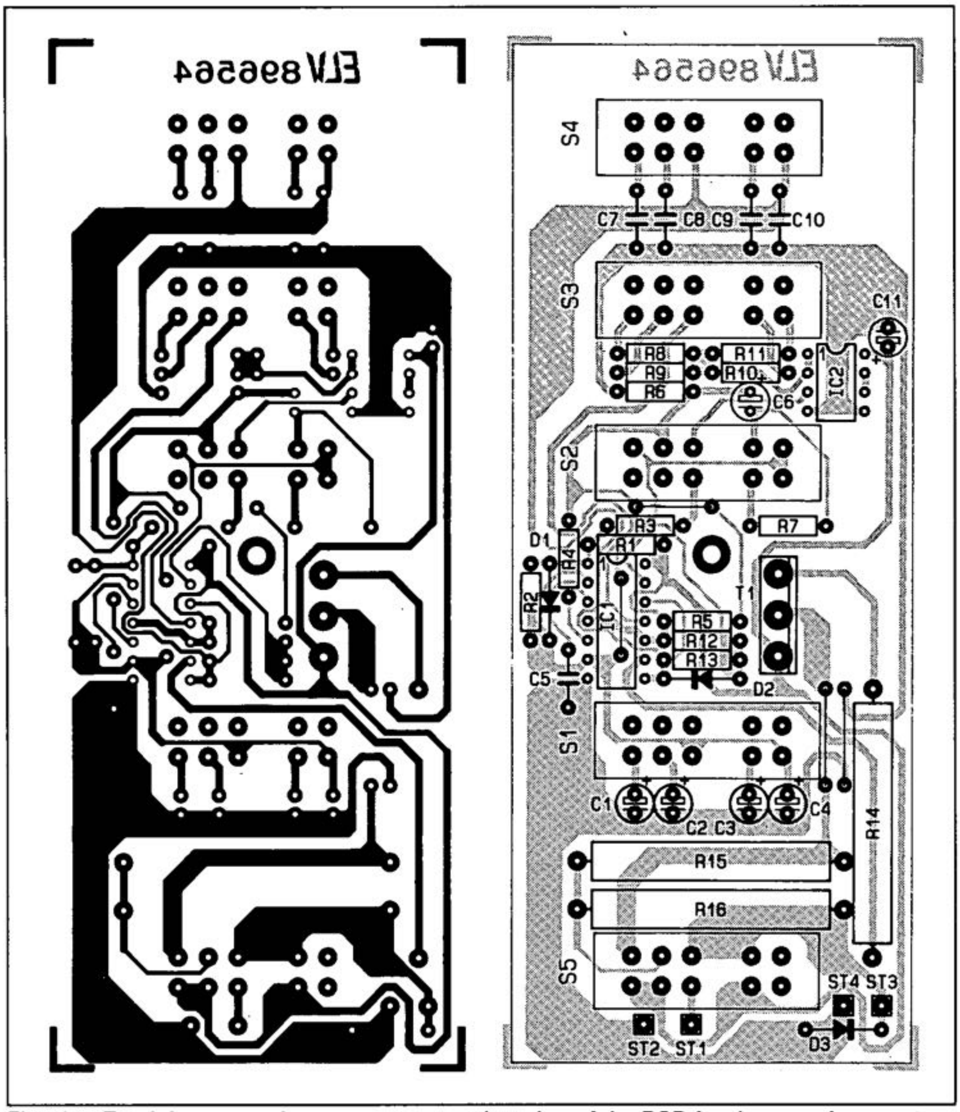


Fig. 2. Track lay-out and component mounting plan of the PCB for the sound generator.

- S1 (frequency): modulation frequency
- S2 (siren type): modulation waveform
- S3 (modulation): modulation intensity
- S4 (basic sound): fundamental siren frequency
- S5 (volume): sound level and on/off control

The four switches S1–S4 allow 4^4=256 different sounds to be generated at three volume levels.

For the highest possible sound level (particularly in alarm systems), it is recommended to use a pressure-chamber type loudspeaker with a sufficiently high power rating (≥20 W). For other applications, standard loudspeakers may be used with good results. The minimum loudspeaker impedance is 4 Ω.

Construction

The sound generator is a relatively simple circuit which should not present difficulties in assembling. Moreover, the unit is supplied in kit form, which obviates problems with obtaining certain components.

Start the construction by fitting and soldering the low-profile parts, followed by the higher parts, on the single-sided printed circuit board shown in Fig. 2. The overlay printed on the component side of the board indicates the position of the parts mentioned in the parts list.

To assist in their cooling, the 5-W power resistors are mounted at a small distance above the printed-circuit board.

The use of a relatively flat enclosure makes it necessary to bend the power transistor, T1, towards the PCB surface as shown in the photograph of the completed board. By virtue of its low internal resistance, and the fact that it is driven at a fairly high level, T1 dissipates relatively little heat, even at full output power. Consequently, the transistor does not require a heat-sink.

After a careful visual check of the completed board, this may be fitted into the enclosure supplied with the kit. Connect the supply voltage to PCB terminals ST1 (+12 V to +15 V) and ST2 (ground). Connect the loudspeaker to terminals ST3 and ST4. Drill holes in the enclosure to pass the supply wires and the loudspeaker wires. Make knots in the wires at the inside of the enclosure to provide strain reliefs. Finally, fit the top half of the enclosure and secure it with the screws supplied.

Practical use

When a 4-Ω loudspeaker is used, the unit draws a peak current of up to 4 A. When used in a switched circuit, e.g., as a horn, the sound generator may be powered via a push-button or a relay with a suitable contact current rating. Use as a horn is possible because the siren starts to sound the moment is it powered. It should be noted, however, that in many countries the use of a siren as a sound actuator device in or on vehicles, and in some cases in or on premises as well, is restricted to emergency services. The use of a siren in general may also be subject to special licenses, rules or regulations as regards on-time, sound type and sound level. ■

COMPONENTS LIST

content of kit supplied by ELV France

Resistors:

1	6.8Ω 5W	R15
1	10Ω	R8
1	22Ω 5W	R16
1	100Ω 5W	R14
2	1k	R3;R9
2	2k2	R7;R10
1	6k8	R11
1	9k1	R4
3	10k	R4;R5;R6
2	100k	R1;R2
1	330k	R13
1	680k	R12

Capacitors:

1	1nF	C9
1	1n5	C10
1	2n2	C8
2	4n7	C5;C7
1	22nF	C11
1	1µF 16V	C4
1	2µ2 16V	C3
1	4µ7 16V	C2
1	10µF 16V	C1
1	22µF 16V	C6

Semiconductors:

1	NE556	IC1
1	TLC271	IC2
1	BD250C	T1
1	1N4001	D3
2	1N4148	D1;D2

Miscellaneous:

5	2-pole 4-way slide switch	S1–S5
4	solder pin	
1	printed-circuit board	
1	enclosure	

A complete kit of parts for the sound generator is available from the designers' exclusive worldwide distributors:

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82837213
Fax: +33 82838180

LF/HF TEST PROBE

For most electronic test and measurement applications, a modern digital multimeter represents excellent value for money. However, in spite of the high input resistance, the accuracy of a DMM degrades rapidly when the frequency of the measured voltage rises above 400 Hz or so. This article describes the basics of designing a passive test probe to overcome this limitation and extend the usable frequency range of a DMM to about 100 MHz.

J. Bareford

The accuracy of digital multimeters is, in general, sufficient for all practical purposes. Although there are low-priced models that give usable results above the maximum input frequency of 400 Hz (stated by the manufacturer), DMMs with a guaranteed frequency range of 20 kHz, 50 kHz or even 100 kHz are rare and quite expensive.

An ideal passive signal rectifier flips each negative half-cycle of a sinusoidal voltage in between two positive half-cycles. The result is a direct voltage of which the peak value is roughly equal to that of the alternating voltage—irrespective of its frequency.

The inertia of a moving-coil meter in an analogue voltmeter provides an certain degree of integration of the measured alternating voltage. The result is a built-in averaging function that smooths the ripple on the direct voltage supplied by the rectifier. Since a DVM is an all-electronic instrument, a capacitor is required at the output of the rectifier to provide the required smoothing and ensure correct measurement results.

All diodes have a certain threshold voltage below which they do not conduct. For the application we are dealing with here, the threshold voltage and the reverse leakage current must be as small as possible. Small-signal germanium diodes of the point-contact type have the lowest threshold voltage, followed by Schottky diodes, normal germanium diodes and silicon diodes, in that order. The forward voltage drop is not a static characteristic but depends to some extent on the current passed by the diode, or, in other words, the load resistance at the output of the rectifier. Obviously, the high input resistance of the DMM (typ. 10 MΩ or more) is advantageous here, since apart from presenting a small load to the rectifier diode it also allows a small capacitor to be used for the previously mentioned averaging function. As a consequence, the smoothing capacitor may be a high-grade type, e.g., a polystyrene capacitor, avoiding large leakage currents typically introduced by, for instance, electrolytic capacitors.

Simple: the single-phase rectifier

A rectifier circuit in its simplest form is shown in Fig. 1. It consists of a diode, *D*, a buffer capacitor, *C*, and a resistor, *R*. The

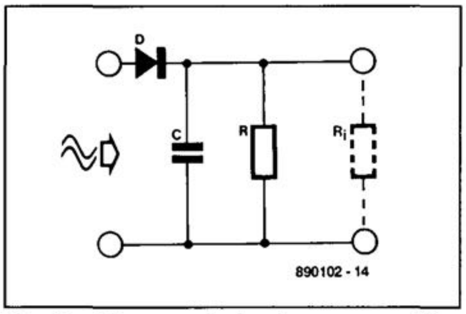


Fig. 1. The single-phase rectifier measures the peak value of the alternating voltage

input resistance, R_i, of the measuring instrument exists in parallel with *R*. When *R* is omitted, R_i alone determines the time constant of the *R-C* network. Since the diode passes the positive half cycles only, the capacitor is charged to the peak value of the alternating input voltage minus the forward drop across the diode. The *R-C* time constant determines the lower frequency limit and must, therefore, be large relative to the period of the input signal with the lowest expected input frequency. The time constant must, however, not be made too large to avoid an excessively slow meter response. The capacitor values shown in Figs. 2, 3 and 4 may be used for easy reference and as starting points for your experiments.

Provided a suitable diode is used, the single-phase rectifier will give good results. The use of a point-contact germanium diode results in a virtually linear frequency response up to about 10 MHz. Above this frequency, the response degrades slowly to about –3 dB at 100 MHz. The output voltage is 100 mV to 200 mV smaller than the peak value of the measured alternating voltage (the peak value equals 1.414 times the effective or root-mean-square value).

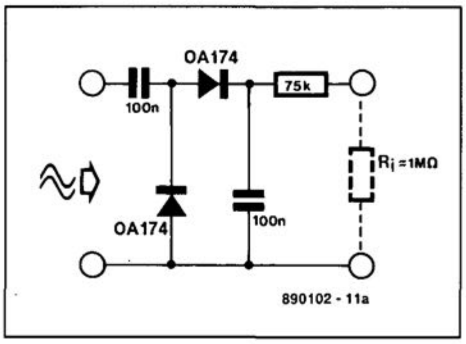


Fig. 2. The Villard circuit is a voltage doubler, i.e., it measures the peak-to-peak value of the alternating voltage. The circuit shown here is dimensioned for a frequency range of 10 kHz to about 1 MHz.

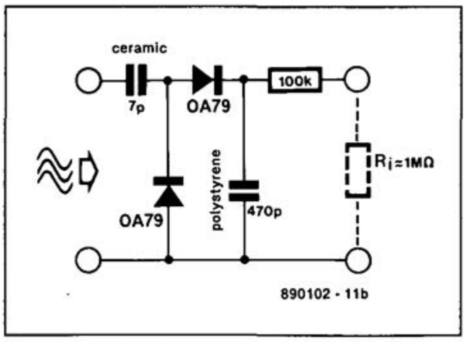


Fig. 3. RF version of Villard rectifier with point-contact germanium diodes.

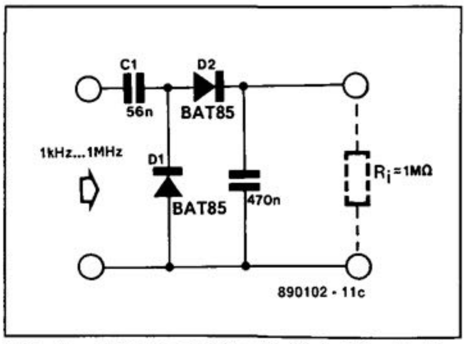


Fig. 4. Villard circuit with Schottky diodes for optimum linearity.

In the small-signal range, Schottky diodes such as the Type BAT85 are better than germanium types because they have a smaller reverse leakage current than point-contact germanium types. The one disadvantage of a Schottky diode, namely its slightly higher forward voltage drop (150 mV typ.), is made good by the better defined conduction voltage (the V-I curve shows a sharper rise than a germanium diode). In practice, the DVM reading for signals smaller than 1 V_{rms} is always a little lower than the actual effective value.

The Villard rectifier

The circuits in Figs. 2, 3 and 4 all contain two capacitors and two diodes. These parts provide voltage doubling. If ideal diodes were used (i.e., diodes with a threshold voltage of nought) the voltage across the output capacitor would be exactly two times that across R in Fig. 1. This voltage represents the peak-to-peak value of the input voltage, or

$$U_{pp} = U_{rms} \times 2 \times 1.4141$$

The operation of the Villard rectifier is best explained by assuming a point in time where a negative half cycle of the input voltage arrives at the input of the circuit in Fig. 4. Diode D_1 conducts, and capacitor C_1 is charged to the peak value of the half cycle. When the positive cycle starts, the voltage at C_1 is added to the peak value of it. This is because diode D_1 blocks, but D_2 conducts, so that C_2 is charged to the peak value plus the voltage across C_1. Thus, the voltage across C_2 represents virtually the peak-to-peak value of the input voltage:

$$U_{C2} = (U_p + U_{C1}) = U_{pp}$$

Capacitor C_2 and the resistance R_i of the DMM form the previously mentioned time constant.

The Villard/Delon rectifier not only supplies a higher output voltage than a single-phase rectifier, it is also more sensitive. Furthermore, its input resistance is roughly equal to that of the DMM at the output. Initially, the measured voltage is only briefly loaded as the capacitors are charged. After a few cycles of the input signal, the charge current virtually disappears, and the input signal is loaded only by R_i and the reverse leakage currents of the diodes.

Basically, the circuits in Figs. 2, 3 and 4 differ only in regard of the time constant, and, therefore, the frequency range. The circuits in Figs. 2 and 3 contain a series resistor at the output to protect the diodes against output short-circuits. In view of the high value of R_i (>1 MΩ) these resistors do not significantly affect the time constant.

The circuit in Fig. 2 is dimensioned for audio signal measurements and provides a linear peak-to-peak voltage reading for signals between 20 Hz and about 1 MHz, provided the DMM input resistance is not smaller than 10 MΩ. When a DMM with R_i = 1 MΩ is used, the capacitor values must be increased to 820 nF. The curves shown in Fig. 5 were recorded with a 1 MΩ DMM and 100 nF capacitors.

The circuit in Fig. 3 is basically the same but adapted to give a frequency range of 300 kHz to about 300 MHz, depending on the type of diode used. The 7 pF capacitor at the input allows the rectifier to be coupled lightly to the circuit under test, avoiding excessive loading of tuned circuits while these are adjusted.

Finally, the circuit in Fig. 4 is set up for a frequency range extending from audio to about 1 MHz. Its remarkably straight response curve is shown in Fig. 6. The sensitivity is also remarkable: the rectifier 'starts' at signal levels as low as 35 mV_{rms} or about 100 mV_{pp}, while at higher input levels (up to 2.5 V_{pp}) the direct output voltage is equal to the peak-to-peak value of the input voltage minus a constant difference of 0.1 V.

Practical notes

The combination of a DMM and a passive LF/HF probe may in many cases replace a much more expensive high-grade LF/HF millivoltmeter, and in addition prove useful for audio purposes, including filter adjustments, frequency response recordings and frequency response corrections (tape recorder calibration, adjustments on equalizers, etc.). The passive probe also allows loudspeaker enthusiasts to record, with the aid of a simple signal generator, the frequency response and steepness of cross-over filters. Provided a linear microphone is available (e.g., an electret reference microphone), it is possible to perform frequency response tests on loudspeakers.

In the RF range, a probe of the type described here enables small signals to be traced and measured. Critical adjustments on filters and oscillators no longer present problems caused by overloading and detuning effects.

The germanium diodes used in the circuits in Figs. 2 and 3 are obsolete types which may, however, be around somewhere in your junk-box. Recuperating an old TV set may also provide you with twenty-odd OA-type diodes of different power ratings. Use the ones that are physically the smallest since these, in general, have the lowest stray capacitance (a rectifier diode from somewhere around the TV's power supply is obviously not worth trying in a probe).

Alternative types with the prefix AA are still current components and may be used instead of the OA174. Arranged in order of decreasing sensitivity, these include: AA113, AA112, AA119, AA118, AA138 and AA137.

For AF applications, the difficult-to-obtain germanium diodes may be replaced by germanium transistors with an 'AC' prefix (AC151, AC152 and similar types). Simply cut off the emitter terminal. The collector becomes the anode, and the base the cathode, of your (admittedly relatively large) germanium diode. ■

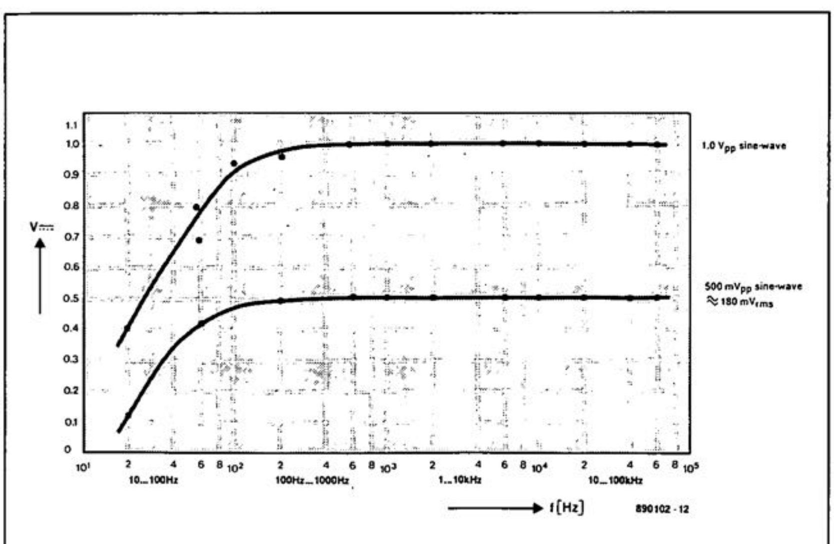


Fig. 5. Frequency response of the circuit in Fig. 2 at two typical AF voltage levels and a terminating resistance of 1 MΩ. The roll-off frequency will be a factor of 10 lower when a 10-MΩ DMM is used.

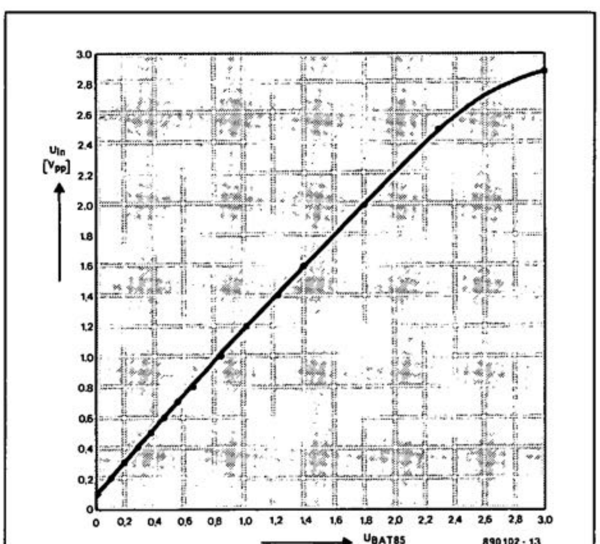


Fig. 6. Correlation between input voltage and output voltage for the circuit in Fig. 4. Excellent linearity is achieved by the use of a Schottky diode.

PROFILES:

by Bernard Hubbard

CRICKLEWOOD ELECTRONICS

After their management buy-out, Syd Wedeles and his partner Clem Clemence increased the turnover of Cricklewood Electronics by some 80 per cent.

Says Syd Wedeles, now a director of the company: "We simply concentrated on stocking the items that our customers wanted. That was the secret".

That was in 1981 and since then the former jazz drummer and his fellow director have successully increased the turnover of the London-based electronic components company year after year, during a decade when a large number of their rivals have gone out of business.

"At one time, there were about fifty electronics retailers between here and Paddington; now there are only a few left," says Syd.

Syd and Clem have achieved their success by a policy of stocking the possibly widest range of components in the UK within their 1000 square feet of retail premises on the Broadway, Cricklewood, NW London.

"Our range is unrivalled," says Syd, "we have over five million components in stock at any one time and if anyone contacts us and asks for new components, we endeavour to stock them".

All Cricklewood products are branded components and their prices are pitched at a highly competitive level.

Said Syd: "We offer a same-day dispatch service on most in-stock items—and that's the majority".

The range includes capacitors, connectors, resistors, semi-conductors, switches, relays, transducers, loudspeakers, tools, kits and test equipment.

Cricklewood Electronics publishes, at regular intervals, a catalogue—160 pages this year: its largest to date—which is mailed to thousands of customers. "These

A Cricklewood staff member with a bunch of newly arrived mail orders.

we have built up over the past nine years".

The majority of orders come over the phone, but on Saturdays there's normally a crush of customers in the shop. Between five and ten per cent of the orders come from overseas—particularly Yugoslavia, Nigeria and the Middle East. Staffed by a team of six, the business is always eager to give advice to customers, many of whom are companies rather than individual enthusiasts. The company is known for its generous discounts on volume sales.

Having steadily built up the business for the past fifteen years, Cricklewood Electronics is poised to enter a new era: it is looking at properties in other parts of the country in a bid to establish a chain of Cricklewood Electronics retail outlets.

For further information on Cricklewood Electronics, see the inside back cover of this issue.

LORD TRANSFORMERS

Difficulties in obtaining supplies of toroidal transformers resulted in Richard Lord launching another business. Already head of Lyon Force, an electronic component manufacturer of Rochford, Essex, Richard was busy hunting for such transformers and was experiencing such problems that he decided he ought to start manufacturing them for himself.

Richard told *Elektor Electronics* recently: "We were experiencing not only difficulties in the supply, but also an extremely bad attitude from existing suppliers; so much so that I decided there must be an opening in this market for us".

That was four years ago and since then Lord Transformers has grown to such an extent that it rivals the existing businesses in turnover.

Says Richard: "We supply standard laminated transformers and specialized ferrite and air core types for switchmode power supplies and PSWS and RF applications".

"We offer customers a fast turn-around of high-quality transformers at highly competitive prices".

Most of the materials used in the company's manufacture are purchased in the UK.

The majority of orders come from within the UK, but 35 per cent come from overseas, particularly France, Germany and Australia. "We manufacture certain sizes for stock, but the majority of our sales are made to order".

Richard maintains: "There has been a significant upsurge in the use of toroidal transformers in recent years, particularly of the flat pancake type".

In the future, Lord Transformers plans to specialize in the larger sizes of transformers, that is, the sizes most people have difficulty in obtaining.

For products and prices contact Lord Transformers, 3 Featherby Way, Purdeys Industrial Estate, ROCHFORD SS4 1LD, Telephone (0702) 544549, Fax (0702) 541075.

INFRA-RED REMOTE CONTROL

Two recently introduced integrated circuits from Plessey, the Types MV500 encoder/transmitter and the MV601 receiver/decoder, allow a versatile 16- or 10-channel infra-red remote control system to be built from a minimum number of components.

T. Giffard

As an example of its application, the infra-red (IR) system described here is used in conjunction with the recently published CMOS preamplifier (Ref. 1). The transmitter and the receiver are described separately below.

Infra-red transmitter

The circuit diagram in Fig. 1 shows that the Type MV500 integrated circuit from Plessey Semiconductors requires only a handful of additional parts to make an infra-red transmitter for remote control applications.

The MV500 contains an 8×4-line decoder for up to 32 keys, and a transmitter section based on PPM (pulse-position modulation). The on-chip oscillator works readily with an inexpensive 455-kHz ceramic resonator. The actual clock frequency is uncritical and may lie between 400 kHz and 1 MHz.

The MV500 operates from supply voltages between 3 V and 9 V. Since ABS enclosures with a compartment for a 9-V (PP3-size) battery are readily available, the circuit was designed to operate from a 9 V supply.

Three IREDs (infra-red emitting diodes) Type LD271 are connected in series and fitted with reflectors to ensure the highest possible infra-red efficiency given that the circuit is powered from 9 V. The current through the IREDs is pulsed by T1 under the control of the output signal supplied by the MV500. The IRED current is limited by a 10-Ω resistor, R2. Since the peak IRED current is about 400 mA, a buffer is required in the form of capacitor C2. Note, however, that the low duty factor of the IRED current pulses results in an average battery load of 1.4 mA to 1.8 mA only.

When no key is pressed, the MV500 switches itself to a 'sleep' state in which its current consumption is negligible. The chip is active only when a key is pressed, to which it responds by transmitting the associated code. The RATE inputs (pins 14 and 15) allow three different data transmission speeds to be programmed as shown in Table 1. The desired speed is set with the aid of wire links A, B, C and D (A and C for 1, B and D for 0).

The keyboard has several options. The printed-circuit board of the transmitter (see Fig. 3) accommodates ten push-buttons, but the relevant section may also be cut off and a separate 16-key membrane-type keyboard connected via K1. The connections are made in accordance with the matrix configuration in a Molex keyboard. For other keyboard makes, the wiring may have to be changed. The ten-key version of the transmitter was developed specifically for the CMOS preamplifier.

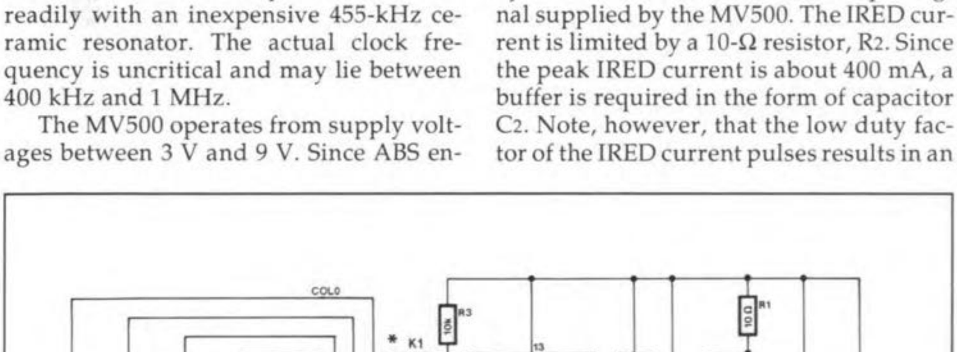


Fig. 1. Circuit diagram of the infra-red remote control transmitter based on Plessey's MV500. The shaded keys are not required in all cases.

Infra-red receiver

The infra-red receiver circuit (Fig. 2) consists of three parts:

- an input stage comprising IR sensor D1 and preamplifier IC1
- a remote control receiver/decoder Type MV601
- a level conversion circuit for the output signals

The IR signals received by D1 are fed to a

RATE INPUTS		RATE VALUE (CLOCK CYCLES)
B	A	
0	0	output inhibited
0	1	2048
1	0	1024
1	1	512

Table 1. Setting the transmission speed with the aid of wire jumpers.

Type SL486 high-gain amplifier which has an on-chip AGC (automatic gain control) circuit. The output signal supplied by IC1 is fed direct to the input of the MV601. The MV601 subsequently converts the PPM signal into a 5-bit dataword, accompanied by a 'data ready' and an 'output enable' signal. These signals, in combination with the momentary or latched modes of the MV601 (selected with wire link C) enable a simple link to be made to a microprocessor circuit.

The clock oscillator in the MV601 must operate at the same frequency as the transmitter with a maximum deviation of 4%. Resistor R2 prevents the on-chip oscillator operating at harmonic frequencies produced by some types of ceramic resonator. The RATE inputs of the MV601 must have the same logic configuration as those of the transmitter chip. A high level (logic 1) is established simply by leaving the relevant input open (jumpers A and B for data rate A and data rate B respectively).

The remarkably high noise immunity of the IR remote control system is ensured by the fact that the MV601 does not supply an output word until two identical PPM codes have been decoded. Reception of a data is signalled by LED D2.

Since most microprocessor interface circuits work with signal levels of 5 V, a level converter is provided on the receiver board. This function is assumed by two analogue multiplexers Type CD4051. The VEE (substrate) terminals of these ICs may be connected to a negative supply voltage.

Wire links E and F enable the outputs of IC3 and IC4 to supply either active-low or active-high signals. The 16 output lines of the two ICs correspond to the maximum number of keys in a (membrane) keyboard used at the transmitter side. Jumper D must be fitted when the level converters are used.

Voltage regulator IC5 in the power supply is required when the supply voltage is 6 V or higher; the maximum input voltage is 15 V.

The outputs of the level converters may be connected to the CMOS preamplifier, or to control inputs of equipment with a similar electronic control system.

Remember that output X0 of multiplexer IC3 (pin 13) is actuated when no signal is received. Hence, switch S1 is not used at the transmitter side.

The dimensions of the printed-circuit board for the receiver have been kept as

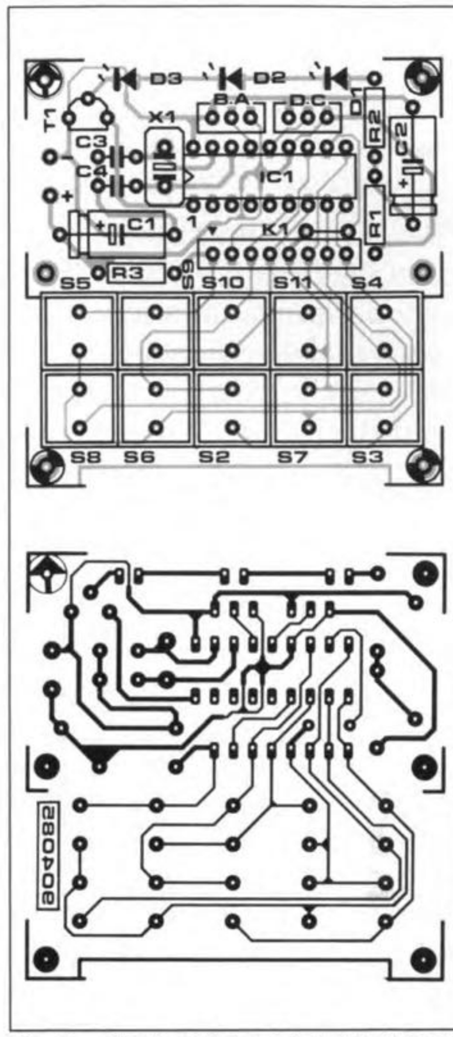


Fig. 3. Single-sided printed-circuit board for the IR transmitter.

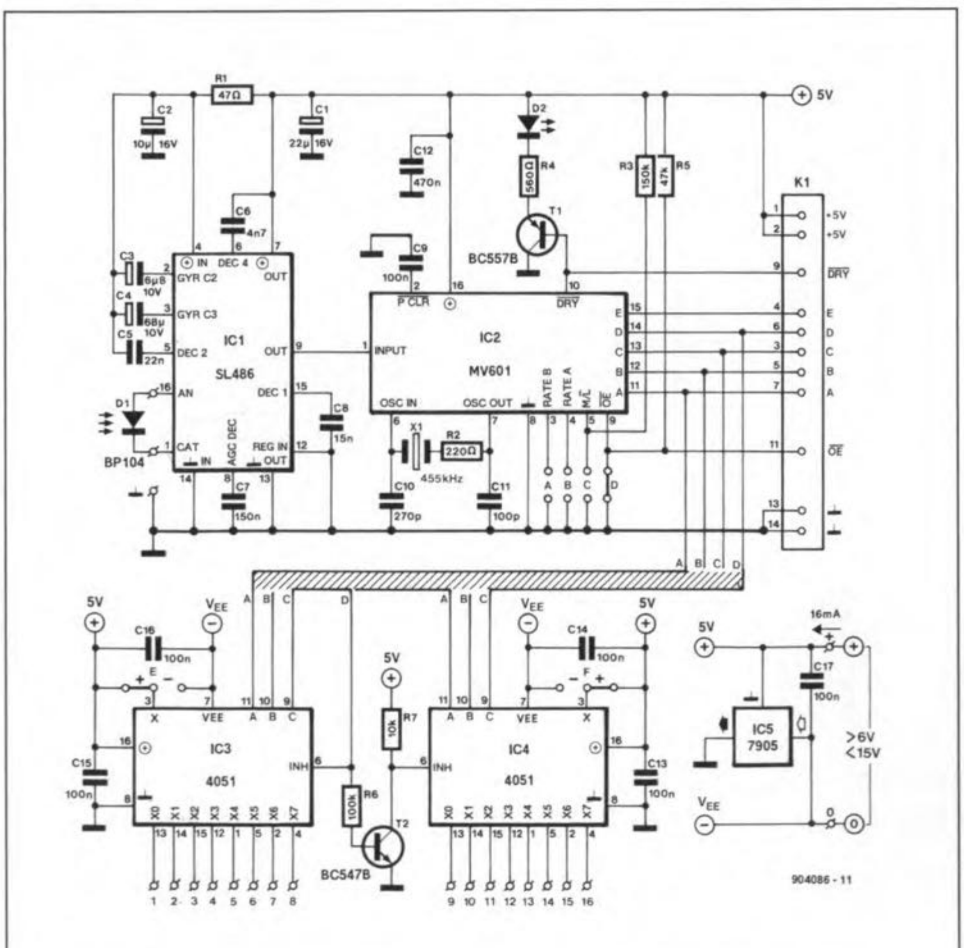


Fig. 2. Circuit diagram of the infra-red receiver. Two CMOS analogue multiplexers Type 4051 provide ready interfacing with a microprocesor system.

COMPONENTS LIST

INFRA-RED TRANSMITTER

Resistors:

2	10Ω	R1;R2
1	10k	R3

Capacitors:

2	100µF 10 V	C1;C2
2	100pF	C3;C4

Semiconductors:

3	LD271	D1;D2;D3
1	BC639	T1
1	MV500	IC1

Miscellaneous:

3	reflector for LD271	
1	ceramic resonator 455 kHz	X1
1	flatcable connector Molex 7583-08	K1
16	keyswitch, e.g., Monacor/ Monarch MS-660/SW or membrane keyboard (4×4 matrix Molex X-Y)	S1–S16
1	9-V PP3 battery with clip	Bt1
1	ABS enclosure with battery compartment, size approx. 100×60×25 mm	
2	3-way pin header	
2	jumper	

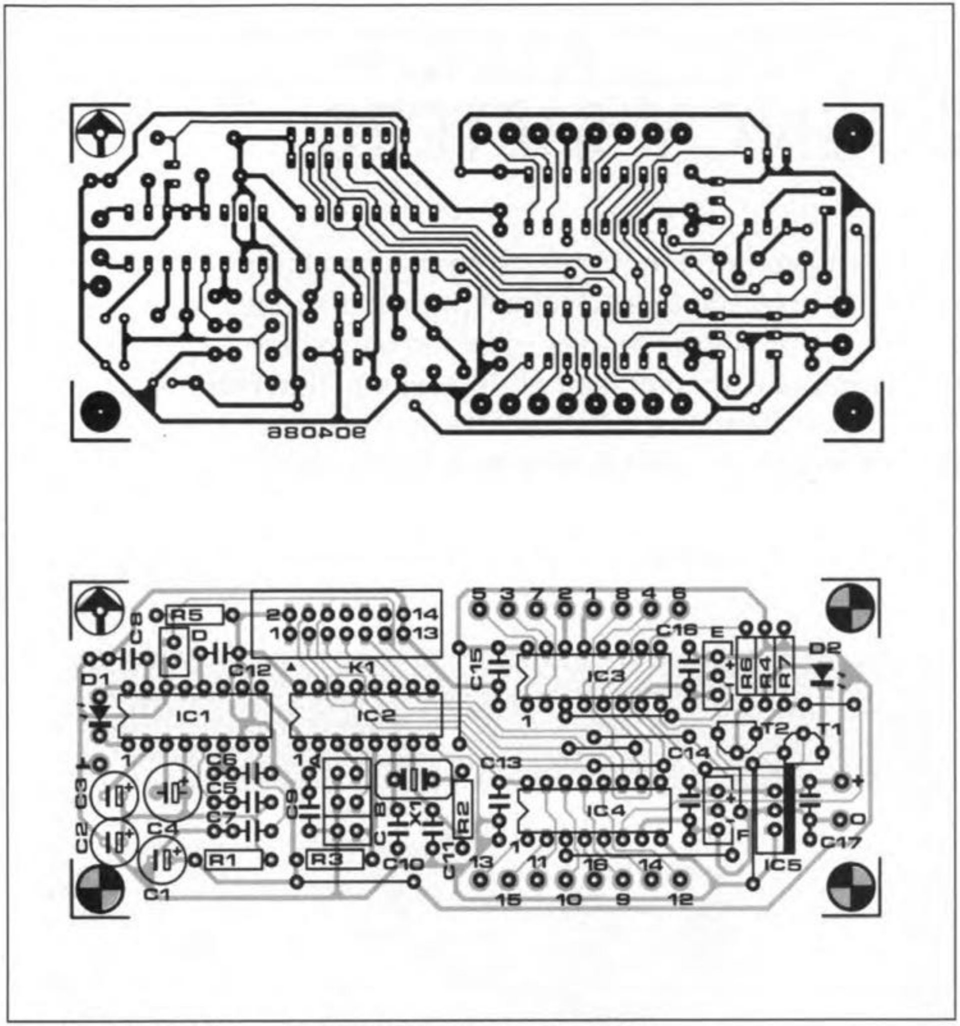


Fig. 4. Single-sided printed-circuit board for the IR receiver.

small as possible to enable ready installation into existing equipment. The quiescent current drawn by the receiver is about 16 mA. ■

Reference:

1. All-solid state preamplifier. *Elektor Electronics* December 1989, January 1990.

COMPONENTS LIST

INFRA-RED RECEIVER

Resistors:

1	47Ω	R1
1	220Ω	R2
1	150k	R3
1	560Ω	R4
1	47k	R5
1	100k	R6
1	10k	R7

Capacitors:

1	22μF 16V tantalum	C1
1	10μF 16V tantalum	C2
1	6μ8 10V radial	C3
1	68μF 10 V radial	C4
1	22nF	C5
1	4n7	C6
1	150nF	C7
1	15nF	C8
6	100nF	C9;C13–C17
1	270pF	C10
1	100pF	C11
1	470nF	C12

Semiconductors:

1	BP104	D1
1	LED	D2
1	BC557B	T1
1	BC547B	T2
1	SL486	IC1
1	MV601	IC2
2	4051	IC3;IC4
1	7905	IC5

Miscellaneous:

1	ceramic resonator 455 kHz	X1
1	14-way PCB-mount header	K1
1	printed-circuit board	904085/4086

RATE INPUTS		RATE VALUE (CLOCK CYCLES)
B	A	
0	0	output inhibited
0	1	2048
1	0	1024
1	1	512

Table 1. Setting the transmission speed with the aid of wire jumpers.

Type SL486 high-gain amplifier which has an on-chip AGC (automatic gain control) circuit. The output signal supplied by IC1 is fed direct to the input of the MV601. The MV601 subsequently converts the PPM signal into a 5-bit dataword, accompanied by a 'data ready' and an 'output enable' signal. These signals, in combination with the momentary or latched modes of the MV601 (selected with wire link C) enable a simple link to be made to a microprocessor circuit.

The clock oscillator in the MV601 must operate at the same frequency as the transmitter with a maximum deviation of 4%. Resistor R2 prevents the on-chip oscillator operating at harmonic frequencies produced by some types of ceramic resonator. The RATE inputs of the MV601 must have the same logic configuration as those of the transmitter chip. A high level (logic 1) is established simply by leaving the relevant input open (jumpers A and B for data rate A and data rate B respectively).

The remarkably high noise immunity of the IR remote control system is ensured by the fact that the MV601 does not supply an output word until two identical PPM codes have been decoded. Reception of a data is signalled by LED D2.

Since most microprocessor interface circuits work with signal levels of 5 V, a level converter is provided on the receiver board. This function is assumed by two analogue multiplexers Type CD4051. The VEE (substrate) terminals of these ICs may be connected to a negative supply voltage.

Wire links E and F enable the outputs of IC3 and IC4 to supply either active-low or active-high signals. The 16 output lines of the two ICs correspond to the maximum number of keys in a (membrane) keyboard used at the transmitter side. Jumper D must be fitted when the level converters are used.

Voltage regulator IC5 in the power supply is required when the supply voltage is 6 V or higher; the maximum input voltage is 15 V.

The outputs of the level converters may be connected to the CMOS preamplifier, or to control inputs of equipment with a similar electronic control system.

Remember that output X0 of multiplexer IC3 (pin 13) is actuated when no signal is received. Hence, switch S1 is not used at the transmitter side.

The dimensions of the printed-circuit board for the receiver have been kept as

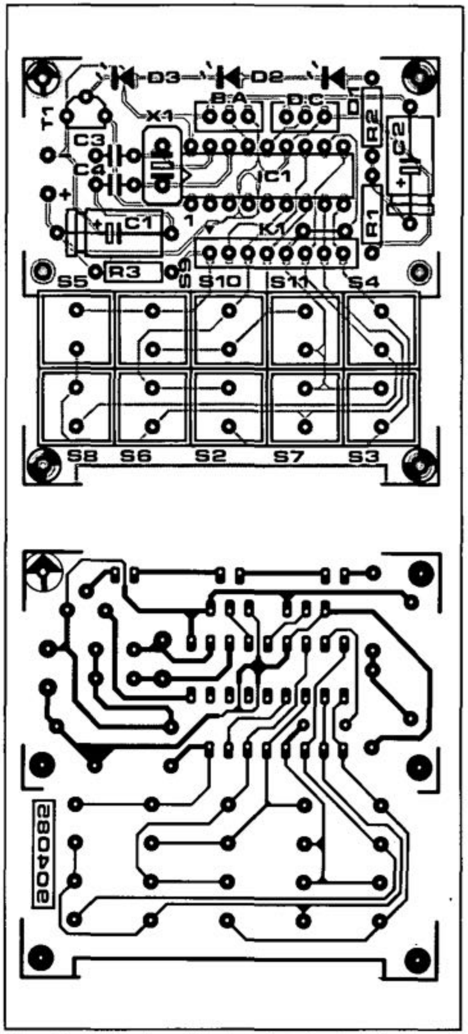
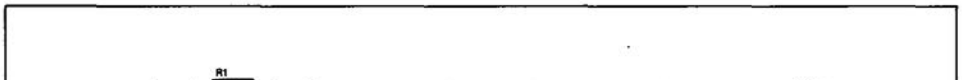


Fig. 3. Single-sided printed-circuit board for the IR transmitter.



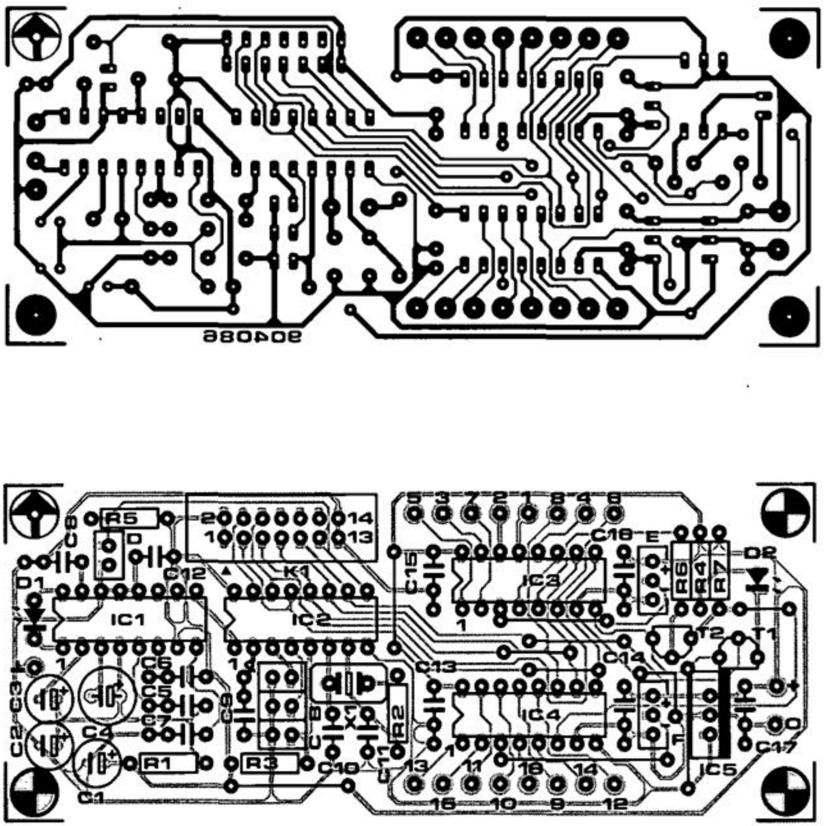


Fig. 4. Single-sided printed-circuit board for the IR receiver.

COMPONENTS LIST

INFRA-RED RECEIVER

Resistors:

1	47Ω	R1
1	220Ω	R2
1	150k	R3
1	560Ω	R4
1	47k	R5
1	100k	R6
1	10k	R7

Capacitors:

1	22µF 16V tantalum	C1
1	10µF 16V tantalum	C2
1	6µ8 10V radial	C3
1	68µF 10 V radial	C4
1	22nF	C5
1	4n7	C6
1	150nF	C7
1	15nF	C8
6	100nF	C9;C13–C17
1	270pF	C10
1	100pF	C11
1	470nF	C12

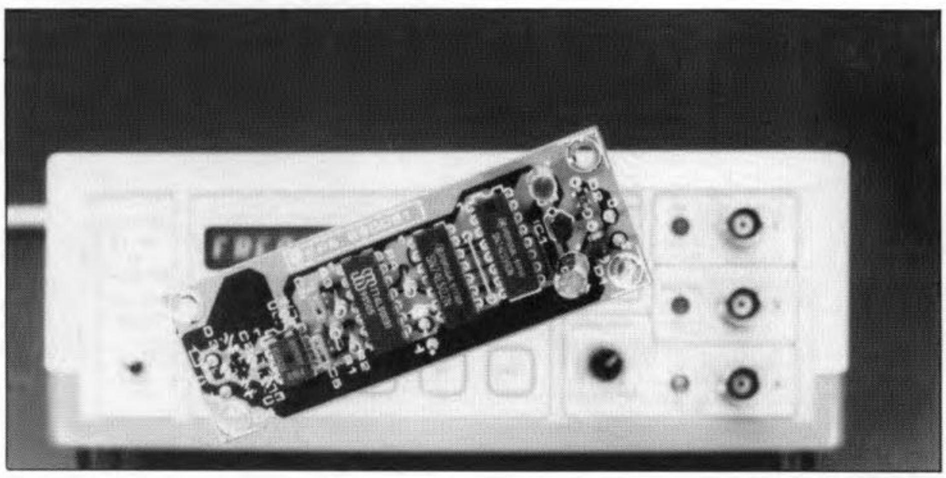
Semiconductors:

1	BP104	D1
1	LED	D2
1	BC557B	T1
1	BC547B	T2
1	SL486	IC1
1	MV601	IC2
2	4051	IC3;IC4
1	7905	IC5

1.5-GHz PRESCALER FOR FREQUENCY METERS

Most simple and inexpensive frequency meters have a maximum input frequency of 10 MHz or 100 MHz. Decimal prescalers that enable frequencies in the gigahertz ranges to be measured are expensive and usually require a special type of RF preamplifier. Not so with the SDA4212 from Siemens, which offers new ways of building a compact, low-cost 1.5-GHz divide-by-100 prescaler.

R. Bönsch



The SDA4212 features a high input sensitivity and excellent large-signal behaviour. Unlike many other prescalers, the SDA4212 does not require its input signal to be kept within certain limits for correct operation of the internal divider. Over the frequency range of 70 MHz to 1200 MHz, the SDA4212 accepts input signal levels between 5 mV and 400 mV (typical values). The prototype circuit constructed on the printed-circuit board shown here has a frequency range of 20 MHz to 1,600 MHz.

The internal structure of the SDA4212 is shown in Fig. 1. The input circuit is formed by a difference amplifier. Next, a high-speed divider divides the signal either by 64 (pin 5 'high') or 256 (pin 5 'low'). The ECL-level output signal (1 V_{pp}) is supplied at pin 6 or 7 by a symmetrical driver stage.

Circuit description

The circuit diagram of the prescaler is given in Fig. 2. The two Schottky diodes at the input protect the SDA4212 against excessive input voltages. The M (mode) input of the chip is tied to the positive supply line to achieve division by 64. The ECL-to-TTL level converter at the output of the SDA4212 is formed by LS-TTL gates N_1, N_2 and N_3. The first gate, N_1, is an AC-coupled linear amplifier. Its gain is set to about 10 times with the aid of feedback resistors R_1 and R_2. Gate N_2 provides further amplification and a well-defined TTL signal. The ECL-to-TTL level convertor used here is easily reproduced, requires no adjustment and is hardly affected by changes in the ambient temperature.

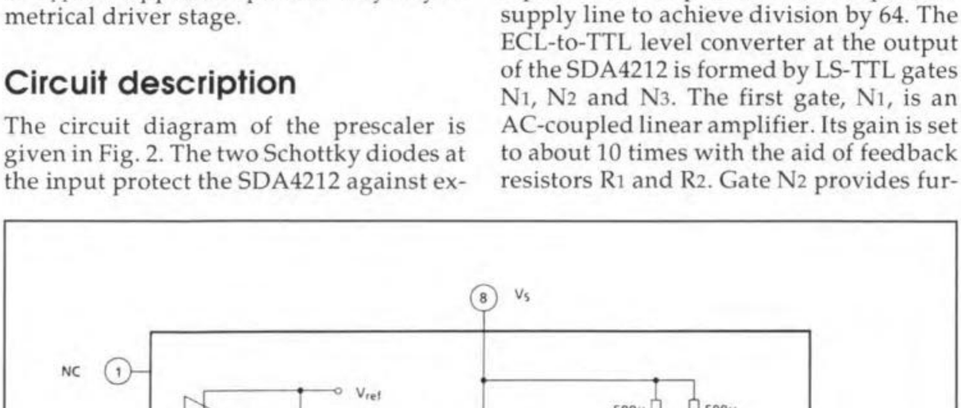


Fig. 1. Block diagram of the SDA4212 from Siemens.

The 74LS390 that follows the SDA4212 contains two divide-by-5 counters. The first Q_D output (pin 9) goes high on the fifth input pulse. At the same time, the input signal of the 74LS390 appears inverted at the output of N_5. This means that gate N_6 is blocked for one-fifth of the time and allows the signal to pass for four-fifths of the time. In other words: four of every five input pulses arrive at the input of the second divide-by-5 counter.

The above divide operation is duplicated in the second counter. Gates N_5 and N_7 are required to compensate signal delays in the dividers and to ensure that the 74LS390 can reach its maximum input frequency.

The output of the prescaler supplies a TTL signal whose frequency is

$$1/64 \times 4/5 \times 4/5 = 16/1600 = 1/100$$

times the frequency of the input signal. It should be noted that a dual-rate counter such as the one applied here supplies an output signal whose mark-space ratio is irregular. Fortunately, this is not a problem for most frequency meters. The out-

COMPONENTS LIST

Resistors:

1	220Ω	R1
1	2k2	R2
1	4k7	R3
1	10k	R4

Capacitors:

3	10nF SMA	C1;C2;C3
1	47nF SMA	C4
2	47nF ceramic	C5;C6
1	10µF 25V radial	C7
1	1µF 16V radial	C8

Semiconductors:

1	1N4001	D1
1	1N4148	D2
2	HP2800	D3;D4
1	78L05	IC1
1	SDA4212	IC2
1	74LS00	IC3
1	74LS02	IC4

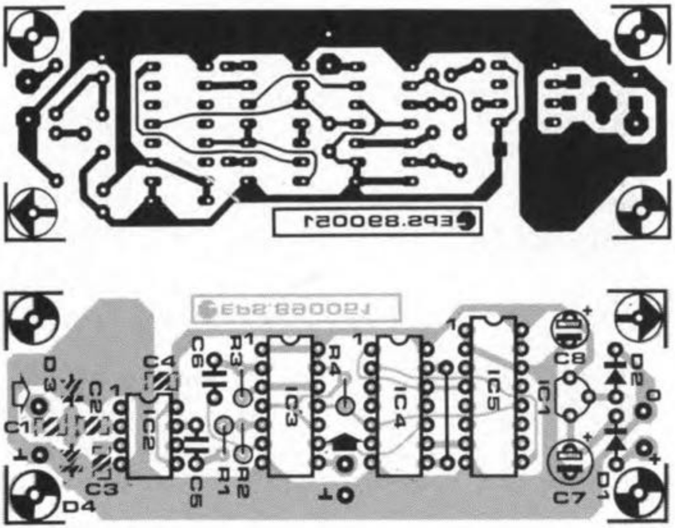


Fig. 3. Track layout and component overlay of the PCB for the prescaler.

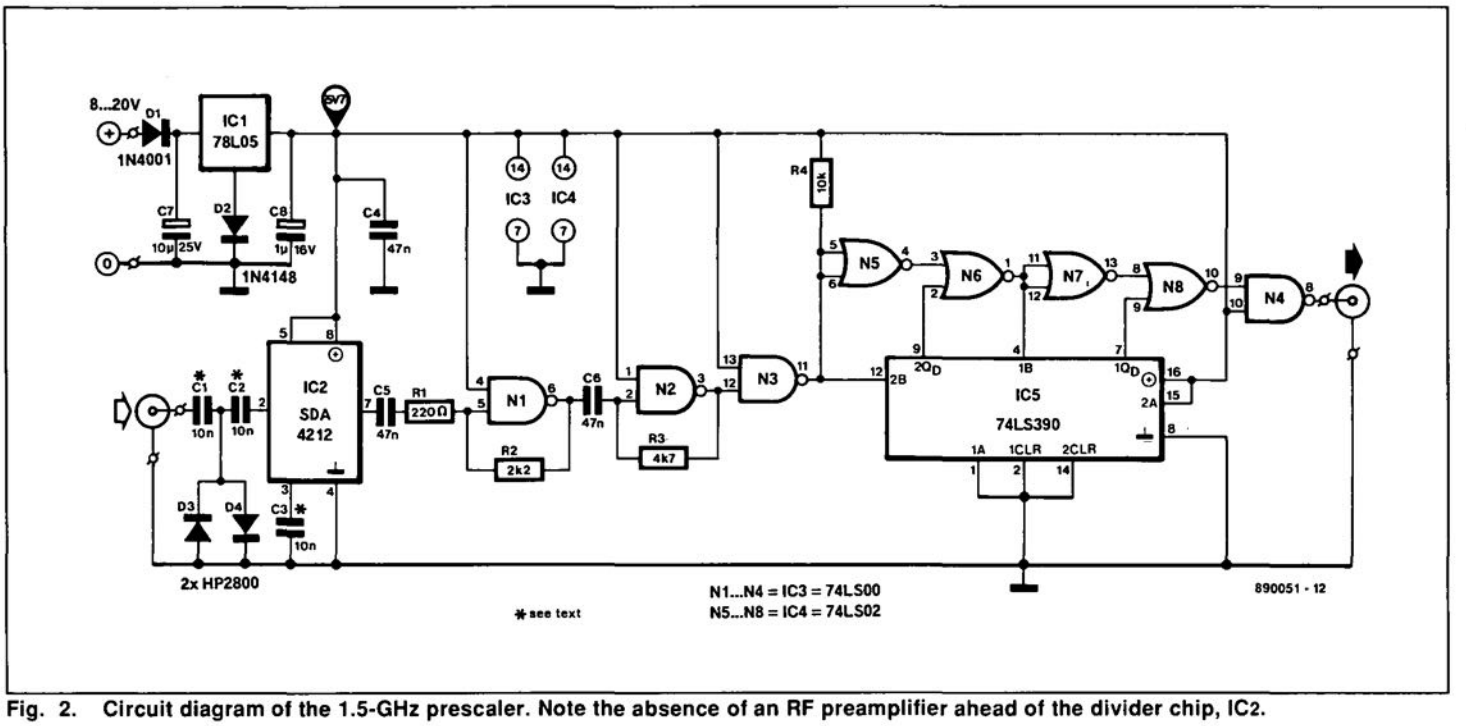


Fig. 2. Circuit diagram of the 1.5-GHz prescaler. Note the absence of an RF preamplifier ahead of the divider chip, IC2.

COMPONENTS LIST

Resistors:

1	220Ω	R1
1	2k2	R2
1	4k7	R3
1	10k	R4

Capacitors:

3	10nF SMA	C1;C2;C3
1	47nF SMA	C4
2	47nF ceramic	C5;C6
1	10µF 25V radial	C7
1	1µF 16V radial	C8

Semiconductors:

1	1N4001	D1
1	1N4148	D2
2	HP2800	D3;D4
1	78L05	IC1
1	SDA4212	IC2
1	74LS00	IC3
1	74LS02	IC4
1	74LS390	IC5

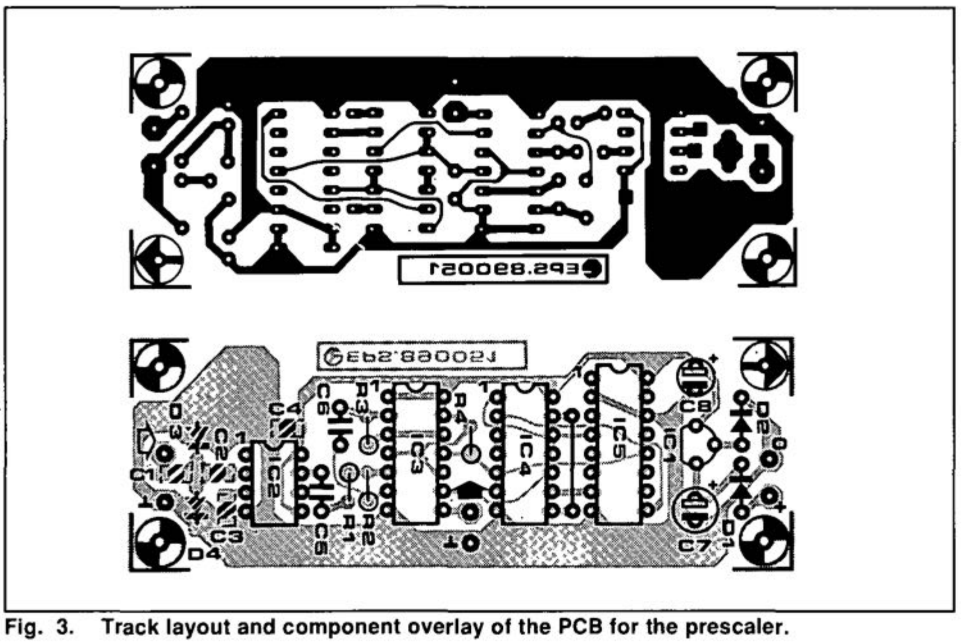


Fig. 3. Track layout and component overlay of the PCB for the prescaler.

put signal may not be suitable for PLL-based circuits, however.

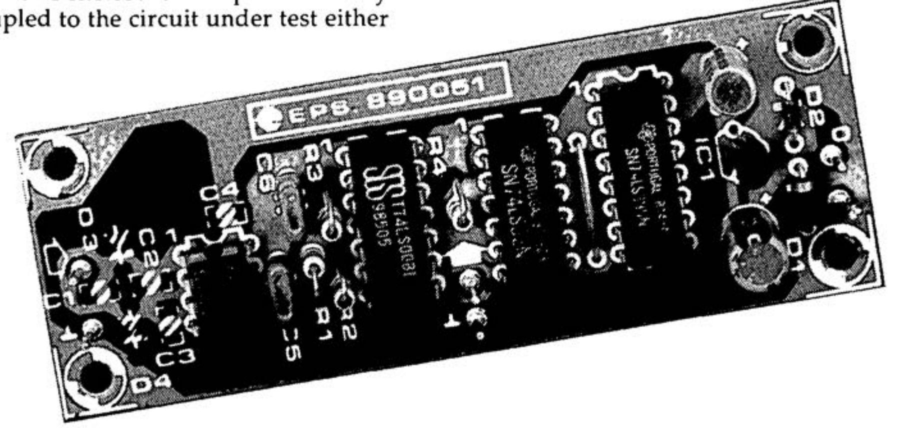
The prescaler is powered by a 5-V regulator Type 78L05 (IC1). Diode D2 at the common terminal raises the output voltage to about 5.7 V to push the highest input frequency of the SDA4212 to the maximum specification. Diode D1 protects the circuit against reverse supply voltages. The relatively low consumption of 30 to 50 mA enables the prescaler to be powered by a battery.

Construction

A number of surface-mount assembly (SMA) parts are accommodated on the printed-circuit board (Fig. 3) to ensure the shortest possible connection between the input of the circuit and the SDA4212. The SMA parts are fitted at the track side of the board.

The board is single-sided with a relatively large unetched copper area around the SDA4212 to ensure adequate screening and decoupling of the high-frequency input signals. The SDA4212 must be soldered directly on to the PCB.

Fit the completed circuit board in a small metal enclosure. The prescaler may be coupled to the circuit under test either capacitively by means of a short wire, or inductively (at a low impedance) by means of a wire loop. Both methods require the prescaler to be located fairly near the circuit under test, but avoid the risk of signal attenuation by the capacitance of long (coaxial) test cables. ■



ROM-COPY FOR 8052-BASIC COMPUTER

g00092

In an earlier article on our popular 8052-BASIC computer we showed a way to replace the 8052AH-BASIC microcontroller with the much cheaper 8032 or 80C32 and an external EPROM. This month we take a second look at unloading the BASIC interpreter from the 8052AH-BASIC and transferring it to an EPROM. We also avail ourselves of the opportunity to get to grips with serial communication protocols for the BASIC computer.

E. Vermeulen

Although many of our regular readers will be familiar with the BASIC computer published in Ref. 2, it does no harm to inform others that this computer is among the most popular construction projects of the past few years. The BASIC computer is an extremely versatile single-board microcontroller system based on Intel's 8052AH-BASIC processor. This chip from the MCS52 family of 8-bit processors features an on-board BASIC interpreter which can be accessed via a 3-wire serial link to an external RS232-compatible terminal. In addition to an on-board EPROM programmer, the system puts a number of easily controlled user ports and timers at the programmer's disposal, and is marked by versatility and simple construction. If you are not convinced of this, we suggest you look at the way the BASIC computer is used in the telephone exchange to be published in a forthcoming issue.

The current consumption of the BASIC computer may be reduced considerably by replacing the 8052AH-BASIC processor by the 80C32. Add an external EPROM that contains the machine code for the BASIC interpreter and you have a low-power BASIC computer suitable for battery operation.

```
1000 REM          **************************************************
1010 REM          **       8052 - AH   BASIC V 1.1 (INTEL)         **
1020 REM          **                                               **
1030 REM          **              R O M C O P Y                    **
1040 REM          **                           by E Vermeulen      **
1050 REM          **                                               **
1060 REM          **   THIS PROGRAM TRANSFERS THE BASIC            **
1070 REM          **   INTERPRETER IN THE ROM BURNED IN THE        **
1080 REM          **   8052AH-BASIC PROCESSOR TO A 2764 EPROM      **
1090 REM          **   BY USING THE ON-BOARD EPROM PROGRAMMER.     **
1100 REM          **   INTERPRETER  8052 IC:           0000H-1FFFH **
1110 REM          **   RAM ADDRESS (IC5) ON CARD:      2000H-3FFFH **
1120 REM          **   EPROM ADDRESS ON CARD (2764):  8000H-9FFFH **
1130 REM          **-----------------------------------------------**
1140 REM          **       8052-AH BASIC   Copyright by INTEL      **
1150 REM          ** For futher information refer to Intel's      **
1160 REM          **         MCS  BASIC - 52 User's Manual         **
1170 REM          **         Order Number : 270010-003             **
1180 REM          **         Intel Corporation (UK) Ltd.           **
1190 REM          **         Pipers Way, Swindon SN3 1RJ           **
1200 REM          **         phone (0793) 696000                   **
1210 REM          **                                               **
1220 REM          **************************************************
1230 REM
1240 PRINT"** PART 1: MOVE INTERPRETER TO RAM AT 2000H-3FFFH"
1250 PRINT:PRINT
1260 MTOP=1FFFH:                    REM : RESERVE RAM FROM 2000H
1270 DIM A(15):                     REM : DIMENSION ARRAY 1x16
1280 FOR X=0000H TO 1FFFH STEP 16:  REM : SET ROM ADDRESSING
1290    FOR Y=0 TO 15:              REM : IN BLOCKS OF 16 BYTES
1300       A(Y)=CBY(X+Y):           REM : READ 1 BYTE ROM, PLACE IN ARRAY
1310       Z=X+Y+2000H:             REM : COMPUTE CORRESPONDING RAM ADDRESS
1320       XBY(Z)=A(Y):             REM : WRITE BYTE IN RAM
1330       B=XBY(Z):                REM : READ IT BACK FROM RAM
1340                                REM : AND COMPARE WITH BYTE IN ROM
1350                                REM : ERROR? => REPORT AND STOP
1360       IF A(Y)<>B THEN ?"PROGRAM ERROR ,ROM= ",A,"RAM=",B:END
1370    NEXT Y:                     REM : 16-BYTE LOOP
1380                                REM : SHOW 16-BYTE LINE ON TERMINAL
1390    PH1. X,"     ",:PH0. A(0),A(1),A(2),A(3),A(4),A(5),A(6),A(7),
1400    PH0. A(8),A(9),A(10),A(11),A(12),A(13),A(14),A(15)
1410 NEXT X:                        REM : LOOP FOR ENTIRE ROM (0000H-1FFFH)
1430 PRINT"** PART2: SET INTERNAL MEMORY FOR PROGRAMMING PROCEDURE"
1450                                REM : INTERPRETER (size = 8 Kbyte)
1460                                REM : NOW IN RAM FROM 2000H TOT 3FFFH
1470                                REM : THE EPROM (also 8 Kbytes)
1480                                REM : OCCUPIES 8000H TO 9FFFH
1490 DBY(18H)=0FFH:                 REM : TARGET ADDRESS LOW BYTE-1
1500 DBY(19H)=00H:                  REM : SOURCE ADDRESS LOW BYTE
1510 DBY(1AH)=7FH:                  REM : TARGET ADDRESS HIGH BYTE
1520 DBY(1BH)=20H:                  REM : SOURCE ADDRESS HIGH BYTE
1530 DBY(1EH)=00H:                  REM : NUMBER OF BYTES LOW ADDRESS
1540 DBY(1FH)=20H:                  REM : NUMBER OF BYTES HIGH ADDRESS
1550 W=0.05:                        REM : W=WIDTH OF PROGRAMMING PULSE IN SEC
1560 R=65536-W*XTAL/12:             REM : R=RELOAD VALUE
1570 DBY(40H)=R/256:                REM : SET VALUE OF R (LOW BYTE)
1580 DBY(41H)=R .AND. 0FFH:         REM : SET VALUE OF R (HIGH BYTE)
1590 DBY(38)=DBY(38) .AND. 0F7H:    REM : RESET BIT 38.3 FOR 50mSEC PULSES
1600 PRINT:PRINT:PRINT SPC(25),"** SETTINGS OKAY **":PRINT:PRINT
1610
1620 PRINT"** PART 3: PROGRAMMING THE EPROM"
1630
1640 FOR R=1 TO 5:PRINT:NEXT R
1650 PRINT"** SWITCH ON THE PROGRAMMING VOLTAGE":PRINT
1660 PRINT"** BE SURE TO APPLY THE CORRECT VOLTAGE (12.5V OR 21 VOLT!":PRINT
1670 FOR R=1 TO 3:PRINT:NEXT R
1680 PRINT"PRESS ENTER TO START PROGRAMMING"
1690 PRINT"OR PRESS ESCAPE TO STOP":PRINT:PRINT
1700 PRINT"PLEASE TYPE <ENTER> OF <ESC>",
1710 K=GET:IF K=0 THEN 1710
1720 IF K=1BH  THEN PRINT:END:     REM ** 1BH is ASCII-code for Esc
1730 IF K<>0DH THEN 1710:          REM ** 0DH is ASCII-code for Enter
1740 PRINT:PRINT:PRINT"BUSY PROGRAMMING EPROM"
1750 PRINT:PRINT"THIS WILL TAKE ABOUT 7 MINUTES":PRINT
1760
1770 PGM                            :REM ** PROGRAMMING INSTRUCTION **
1780 PRINT
1790 PRINT:PRINT"** PART 4: CHECKING INTERNAL POINTERS FOR ERRORS"
1800 PRINT
1810 H=DBY(1AH):L=DBY(18H):HL=H*256+L
1820 IF (DBY(30).OR.DBY(31))<>0 THEN 1830 ELSE 1840
1830 PRINT"INCORRECT PROGRAMMING OF EPROM AT ADDRESS",:PH1.HL:END
1840 PRINT"PROGRAMMING FINISHED   ** NO ERRORS **":PRINT
1850 PRINT
1860 PRINT"** PART 5: DIRECT COMPARISON BETWEEN ROM AND EPROM":PRINT
1870 PRINT
1880 FOR X=0000H TO 1FFFH           :REM ** ADDRESS IN ROM
1890    Y=X+8000H                   :REM ** ADDRESS IN EPROM
1900    A=CBY(X):B=XBY(Y)           :REM ** READ OUT RAM AND EPROM
1910    PRINT"ROM",:PH1.X,:PRINT" =>",:PH0.A," =",
1920    PH0.B,:PRINT" <=",:PH1.Y,:PRINT" EPROM",CR,
1930    IF A<>B THEN PRINT"EPROM ERROR":END
1940 NEXT X:PRINT:PRINT
1950 PRINT"EPROM CORRECTLY PROGRAMMED":PRINT:PRINT
1960 PRINT"PROGRAM FINISHED, BYE !!!"
1970 END
```

This program is available on disk (5¼-inch 360K IBM format) as order number **1441** (see the Readers Services page).

Fig. 1. Type this listing of ROMCOPY.BAS into your favourite word processor.

Do you copy?

All that is needed to transfer the BASIC interpreter from the 8052AH-BASIC to a 27C64 EPROM is the program listed in Fig. 1. Simply get your original BASIC computer on line, and type in the listing on your terminal or PC. Check what you have on your screen against the printed listing. If this is all right, save the program. Plug an empty 27C64 EPROM in the ZIF socket on the BASIC computer board, apply the correct programming voltage, and RUN the program. The 8052AH-BASIC does the rest, i.e., it transfers its own BASIC interpreter to the EPROM. Some seven minutes later, the EPROM is ready for use in a 80C32-based version of the BASIC computer.

Procomm and the BASIC computer

Some problems may crop up when the BASIC computer is connected to an IBM PC or compatible running a communication program like Procomm. These problems arise mainly from the absence of any form of hardware handshaking on part of the BASIC computer. Also, while downloading a listing generated on the PC, the BASIC computer needs time to convert the ASCII data into token codes. This operation, carried out by the BASIC interpreter in the 8052AH-BASIC, starts after reception of a carriage return (CR) character from the terminal. During the token conversion, the 8052AH-BASIC does not monitor the serial port. The result is that characters may be lost in the downloading process when the terminal does not wait for the BASIC computer to complete its token conversion operation after every complete line of BASIC instructions. One way of solving the handshaking problems is to reduce the bit rate on the serial link to a value where the time needed for the token conversion is simply not noticed.

```
──────────────────────┤ TERMINAL SETUP ├──────────────────────
 1) Terminal emulation ... VT-52        10) Break Length (ms) .... 350
 2) Duplex ............... FULL         11) Enquiry (CTRL-E) ..... OFF
 3) Flow control ......... NONE
 4) CR translation (in) .. CR
 5) CR translation (out) . CR
 6) BS translation ....... DEST
 7) BS key definition .... BS
 8) Line wrap ............ OFF
 9) Scroll ............... ON
──────────────────────────────────────────────────────────────
OPTION ─►                                              ESC► Exit
```

Fig. 2. Screendump of Procomm's terminal setup menu with parameters set for the BASIC computer.

```
──────────────────────┤ ASCII TRANSFER SETUP ├──────────────────────
            ASCII UPLOAD
 1) Echo locally ......... NO
 2) Expand blank lines ... NO

 3) Pace character ....... 62    (ASCII)
 4) Character pacing ..... 0     (1/1000 sec)
 5) Line pacing .......... 5     (1/10 sec)

 6) CR translation ....... NONE
 7) LF translation ....... NONE
            ASCII DOWNLOAD

 8) CR translation ....... NONE
 9) LF translation ....... NONE
────────────────────────────────────────────────────────────────────
OPTION ─►                                                   ESC► Exit
```

Fig. 3. Screendump of Procomm's ASCII transfer setup menu. Pace character 62 forces the PC to wait for the prompt transmitted by the BASIC computer.

Waiting for 62

A fairly reliable method of ensuring software handshaking is to have the PC wait for the prompt (>) transmitted by the BASIC computer. This can be achieved by programming ASCII value 62 for the pace character in Procomm's ASCII transfer setup menu. First, however, go to the line settings menu (ALT-S) and select option 12 to set the serial data speed and format to 19,200 bits/s, no parity, 8 databits and 1 stopbit. If necessary change the setting as required for the COM port you intend to use. Save the settings.

Next, call up the setup screen by typing ALT-S. Select item 2, the terminal setup. The parameters necessary for the BASIC computer are shown in the screendump in Fig. 2. Return to the setup screen, and select item 6, the ASCII transfer configuration. The settings used by the author are shown in Fig. 3. The pace character (option 3) is set to 62 to force the PC to wait for the BASIC computer after sending a line of BASIC. The character pacing may be left at 0 in most cases. The line pacing is best set to 5 as shown, although this is not critical.

When the pace character is set to 0, Procomm ignores any character returned by the BASIC computer. This setting is still useful, however, when the software handshaking is not used, and the serial link runs at a relatively low baud rate. When software handshaking is not used, some experimenting with the character pacing and line pacing parameters may be required for best results. A lot depends on the length of the cable between the PC and the BASIC computer, and also whether or not this cable is screened. Fortunately, the wait times that may be programmed in Procomm will allow you to find a compromise between fast data transfer and as few as possible errors.

Uploading the ROMcopy program

Assuming that you have typed the listing in Fig. 1 into your favourite word processor (set to ASCII output), the procedure to send the program to the BASIC computer is as follows. First, save the file as ROMCOPY.BAS on floppy disk or hard disk. Next, run Procomm, and get the BASIC computer on line by pressing the space bar. Type PgUp to enter the upload menu. Select option 7, ASCII PROTOCOL. Type the filename (if necessary, preceded by the drive station and/or the path, e.g. A:\ROMCOPY.BAS) followed by a return. The upload process can be followed on the screen. On completion, type LIST to check the loaded file against the listing. RUN the program and you have your BASIC interpreter available in EPROM after about seven minutes. ■

References:

1. CMOS replacement for 8052AH-BASIC. *Elektor Electronics* January 1990

2. BASIC computer. *Elektor Electronics* November 1987.

890114

MAKE YOUR OWN PCBs

Experimenting with electronics is great fun and many of you will have no objection to a drawer full of projects that never made it beyond the stripboard stage. However, once a circuit is known to function reliably, the last step is inevitably to put it on a printed-circuit board. In this magazine we monthly offer the PCB artwork for numerous projects, so all the designing and checking has been done for you. The next logical step is to make a flying start with the project by buying one of our ready-made PCBs produced to professional standards. However, in case a PCB is not available ready-made, there is no other way than to produce it yourself.

This article discusses the basic steps involved in making one's own PCB from available artwork, i.e., the design (drawing) and design checking stages are not covered. It is assumed that artwork of the type normally published in this magazine is used to produce a printed-circuit board.

The first thing to note is that the track-layouts are printed as a reverse (mirror image) of the original, i.e., as if they were viewed from the component side. The reason for doing this will be reverted below. Second, the lay-out is always at full-size, i.e., at 100%. Third, every track lay-out has an associated component mounting plan (or overlay), which shows the positions of the components at the component side of the board. The values of the components and the type indications are in accordance with the parts list printed next to, or below, the track lay-out and the overlay. Unfortunately, you can not produce your own component overlay unless you have fairly sophisticated photographic equipment. The overlays shown in the articles are used for printing on the PCBs supplied ready-made through the Readers Services.

Fig. 1. Home-made UV exposure unit suitable for use with a vacuum pump for optimum contact between the photosentive layer on the board and the emulsion on the film.

The artwork

The term artwork for a PCB commonly refers to the drawings that 1) show the pattern of the copper tracks at the track side of the board and 2) show the positions of the components at the component side of the board.

Several things must be kept in mind before deciding to make one's own PCB for a particular project. First, the track lay-out must be relatively simple. We do not recommend making PCBs with a high track density, as in that case the artwork is usually produced on a computer. To keep the board as small as possible, the tracks are then relatively thin and difficult to reproduce reliably with simple equipment. This means that a PCB with many ICs, or a double-sided PCB, is better purchased ready-made than made at home with all the risks of almost invisible track interruptions and short-circuits in between IC pins which can take hours to trace. The results in not a few cases are disappointment, burnt dinners and an ugly PCB with lots of wires running at the track side. Also bear in mind that a ready-made PCB has a component overlay which consists of the component symbols and reference numbers printed in white, and the track lay-out printed in black. The latter makes it easy to follow the track pattern while viewing the component side. Most artwork shown in the magazine is produced to certain conventions you have come to recognize and appreciate over the years.

The materials

The minimum bill of materials for making a printed-circuit board from a lay-out in the magazine looks like this:

- transparent (mylar) film
- a piece of (positive) photo-resist copper clad board (single- or double-sided)
- an UV (ultra-violet) light exposure unit
- a quantity of sodium hydroxide (caustic soda) pellets (developer)
- a quantity of ferric chloride pellets (etchant)
- two polystyrene or glass trays as used in photographic darkrooms
- a small heating plate
- a thermometer
- rubber gloves
- lots of clean water
- plastic flasks or other containers for the developer and the etchant
- a pair of polystyrene tongs
- a saw or another suitable cutting device
- a high-speed drill on a stand
- a can of solder spray
- tin-plating materials (optional)

All of the above tools and the chemicals are available in various sizes and quantities from electronic mail order firms. In many cases, complete PCB production kits are offered, although a drill on a stand is not usually included.

Exposing and developing

Although some of you appear to get away successfully with photocopies on paper, it will be necessary in most cases to transfer the lay-out found in the magazine to a transparent film. If you do not have the necessary dark-room equipment to do this, your local photographer or printer may be able to help you. The film used should not be too thick. Interestingly, some photocopying machines are capable of copying on to film transparencies or peel-off film material such as Frisket. Whether or not this method of producing a film from a lay-out printed in the magazine is usable depends largely on the thickness of the film and the quality of the toner. Do not be surprised if the machine 'gobbles up' your film and keeps it hidden somewhere inside until a well-trained service technician arrives! Normal transparent foil must never be fed into a photocopying machine because the heat applied to burn the toner into the material will cause it to melt and remain stuck on the drum.

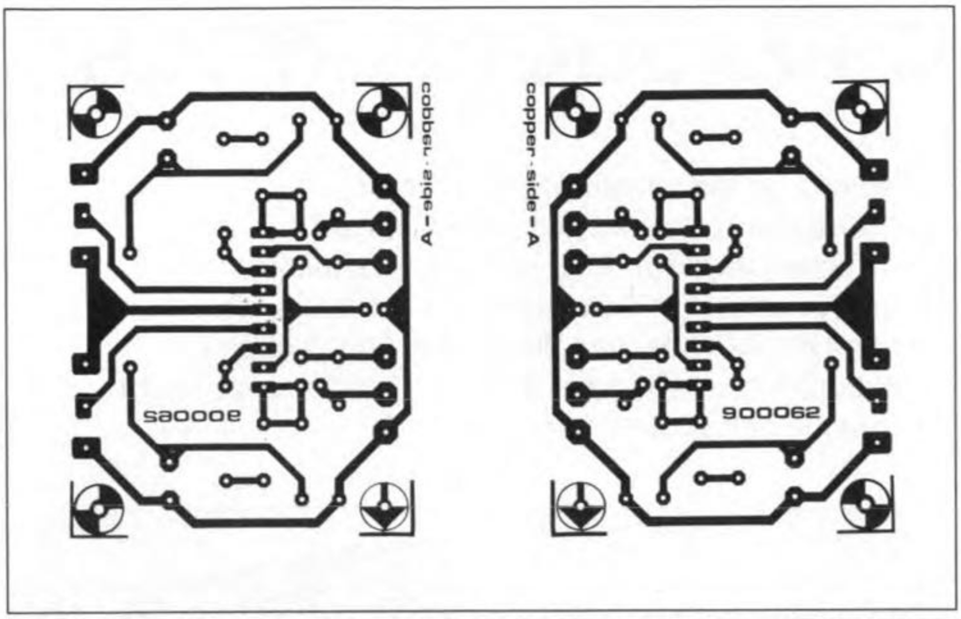


Fig. 2. Look at the letters and numbers to see the difference between the mirror image of a PCB track lay-out (left) and the 'real' track lay-out (same PCB shown to the right). The lay-outs in this magazine are usually printed as reverse images to prevent parallax effects during the exposure period.

The rear side of the film is secured on the glass plate of the UV exposure unit, i.e., the emulsion side of the film is in direct contact with the copper side of the circuit board placed on top. The track-layouts found in the magazine are printed as mirror images to enable this direct contact so as to prevent parallax effects (small image-shifting) with relatively thick films or paper. Before putting the PCB on top of the film, peel off the light-resistant plastic cover. It is very important for the board to be pressed as firmly as possible on to the film, particularly if the board is fairly large.

UV exposure units may be purchased ready-made, usually complete with two or four UV tubes. The photograph in Fig. 1 shows a home-made type with four UV tubes, a built-in mechanical timer and a connection for a vacuum pump (not shown). The pump is capable of creating a fairly strong vacuum, ensuring a high and equally distributed contact pressure between the photosensitive layer on the board and the emulsion side of the film. The UV unit shown has been used successfully many times for the production of relatively large circuit boards. Remember that ultra-violet light is harmful to the eyes, so make sure that the UV unit can be closed properly.

In general, the exposure time is fairly uncritical. As a rule of thumb, a unit with four 20-watt tubes and the film at a distance of about 5 cm will be on for about 5 minutes. The optimum exposure time depends on the type of film and the photosensitive board, and will have to be found empirically.

The developer may be made during the exposure time. Depending on the size of the board, a quantity of sodium hydroxide crystals is dissolved in water in accordance with the instructions on the package. Use water of about 25°C and stir the solution to make sure all crystals are dissolved. After exposure, the board is immersed in the developer, and the tray is gently agitated. After one to two minutes, the track pattern should become visible as the solution turns light blue or purple. Clean the board with plenty of running water. Although the developer is reusable, it is best disposed of after about ten times' use.

Etching and drilling

The etchant, a ferric chloride ($FeCl_3$) solution, is prepared at the concentration and temperature recommended by the manufacturer. Always wear rubber gloves while working with the etchant: although most of you will be aware of it, it is an aggressive, highly corrosive and toxic solution. Never use metal trays, spoons, tongs or flasks in combination with the etchant.

Depending on the temperature, the concentration of the etchant, and the amount of copper to be etched away, the etching process will take between 10 and 20 minutes. The glass or polystyrene tray must be agitated gently during the entire etching process, while the temperature must be kept between 40ÉC and 50°C. Higher temperatures do not make the etching process any faster. If necessary add a little water to compensate the loss due to evaporation. A small aquarium pump that feeds bubbles into the etchant is a good alternative to having to agitate the etching tray. Likewise, a glass, immersion-type heating element as used in aquariums may be used to ensure the right etching temperature. As you can see, it is worthwhile to visit your local pet shop!

When the board is ready, use a pair of polystyrene tongs to remove it from the tray. Clean the board thoroughly under running water to remove all etchant. The etchant solution is reusable and may be poured back in its container. Although ferric chloride etchant is known to be usable even when it is thick with dissolved copper, its concentration may be kept at a usable level by adding a little hydrochloric acid from time to time.

Allow plenty of time for the PCB to dry before spraying the track side with solder lacquer.

Even a perfectly produced board may be spoilt when it is not drilled properly. In most cases, you will find that a high-speed drill on a stand is a must after all the effort you have put into making the PCB. The optimum drill speed depends on the drill diameter and the board material. In general, it is best to purchase a set of special high-speed PCB drills with diameters between 0.7 mm and 1.5 mm. Normal high-speed steel (HSS) drills may be used for the larger holes (e.g., the ones in the corners of the PCB). As you will soon discover, drilling a PCB is a fairly boring and time-consuming job. Brush off the dust from time to time and check for burrs at the component side.

The PCB is now ready for assembly. Use the component overlay found in the magazine as an orientation aid to fit the parts at their respective positions. ■

WARNING

During the production of printed-circuit boards chemicals are used which are possibly aggressive and corrosive. Strictly follow the manufacturers' or suppliers' instructions for handling and storage of these chemicals. Wear goggles and protective gloves, and ensure that the space you are working in is amply ventilated. **Contact your local council on environmentally safe ways to dispose of used developer, etchant, empty cans, and other packaging materials.**

SUPPLIERS

A wide range of materials, tools and equipment for making printed-circuit boards is available from UK-based electronics mail-order companies including

- **Maplin Electronics PLC, (0702) 554155**
- **ElectroMail, (0536) 204555**
- **Cirkit Distribution Ltd., (0992) 444111**
- **VeroSpeed, (0800) 272555**
- **ElectroValue Ltd., (0784) 33603 or (061 432) 4945**
- **Cricklewood Electronics Ltd., (081 452) 0161**

AUTOMATIC HEADLIGHT SWITCH

This circuit automatically switches on the headlights of a car after the engine is started. In countries like Sweden and Canada, car drivers must have their day-running lights on, even in broad daylight. Accordingly, cars manufactured or imported into these countries often have an automatic lights-on control which, although it can not be disabled, leaves the normal light switch in function. A similar circuit with automatic switch-on, switch-off and reduced light intensity is described here.

J. Ruffell

Although a motorcyclist driving with his headlight on is a familiar sight even in broad daylight, many of you will find it strange at first to see car lights during daylight hours. In Sweden, however, it is not allowed to drive a car or any other motor vehicle without switching the headlights on. This traffic rule is imposed for reasons of safety, and has prompted car manufacturers exporting to Sweden to devise circuits that switch the lights on and off at the same time as the engine. Similar regulations for the use of car lights are currently proposed by other countries like Holland.

Evidently, the switching on action of an automatic lights controller serves safety, and the switching off action prevents the annoyance of a flat battery after, say, a few hours of shopping with the car left on a misty parking lot.

To extend the life of the bulbs, the automatic light switch reduces the operating voltage to about 80% of the battery voltage. Tests have shown that the reduction in light intensity is not noticed under daylight conditions. The circuit allows the driver to switch on the headlights when he thinks this is necessary. When the 'normal' light switch takes over, the lamps are powered with the full battery voltage.

Circuit description

Figure 1 shows the circuit diagram of the headlight switch. The number of parts has been kept to a minimum to enable the circuit to be fitted in a compact, aluminium, enclosure, for which there should be room in almost any car. The circuit is connected to the car's electrical system without the need to break or modify existing connections.

When the engine is started, the car's headlights and rear lights are switched on automatically. Pulse-width modulation is used to power the lamps at about 80% of their nominal working voltage, which is automatically restored to 100% when the driver switches on the lights.

The heart of the circuit is formed by opamp IC1, a CMOS type TLC271. The oscillator it forms part of supplies a 50 Hz signal of which the duty factor can be set to a value of between 0.7 and 0.9 with the aid of preset P1. For other ratios, resistors RR6 and R7 have to be given different values. The output signal of the opamp controls output transistor T1 via a buffer. The output transistor is a power MOSFET Type BUZ11 capable of switching loads of several tens of ampères (30 A nominally). When the FET conducts, the 'on' resistance of the drain-source junction is as low as 0.04 Ω, so that the dissipation remains below 2 W even in worst-case conditions. The MOSFET is protected against voltage surges by a VDR (voltage-dependent resistor), R14.

Transistor T2 switches the opamp-based oscillator on and off. When the en-

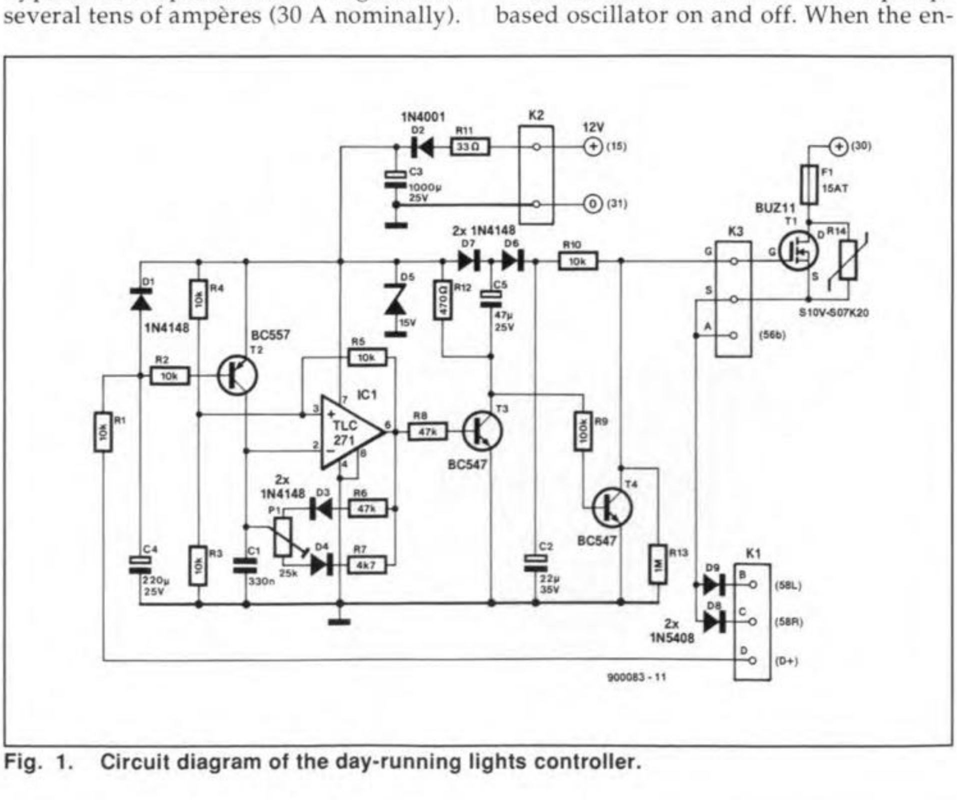


Fig. 1. Circuit diagram of the day-running lights controller.

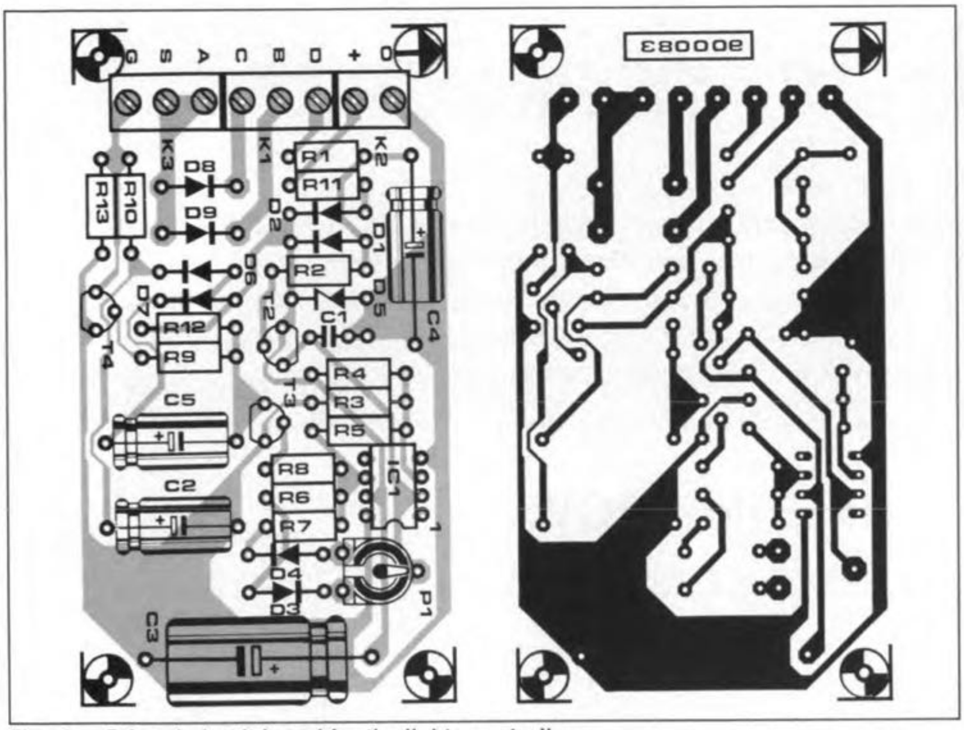


Fig. 2. Printed-circuit board for the lights controller.

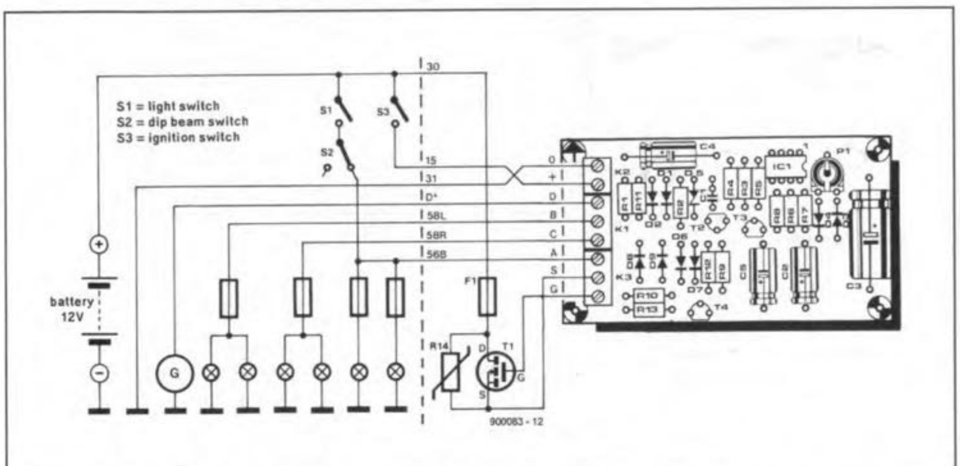


Fig. 3. Connection diagram. The terminal numbers follow the Bosch convention.

COMPONENTS LIST

Resistors:

6	10kΩ	R1 – R5;R10
2	47kΩ	R6;R8
1	4k7	R7
1	100kΩ	R9
1	33Ω	R11
1	470Ω	R12
1	1MΩ	R13
1	SIOV-S07K20	R14
1	25kΩ preset H	P1

Capacitors:

1	330n	C1
1	22µF 35V	C2
1	1000µF 25V	C3
1	220µF 25V	C4
1	47µF 25V	C5

Semiconductors:

1	TLC271	IC1
1	BUZ11	T1
1	BC557	T2
2	BC547	T3;T4
5	1N4148	D1;D3;D4;D6;D7
1	1N4001	D2
1	15V 0.4W zener	D5
2	1N5408	D8;D9

Miscellaneous:

2	3-way PCB terminal block	K1;K3
1	2-way PCB terminal block	K2
1	15A (slow) fuse with cable holder	
1	enclosure Hammond 1590B	
1	insulation set for T1	

gine does not run, the alternator does not run either, so that terminal D+ is at a low potential. Consequently, T2 conducts, C1 is kept fully charged, and the oscillator is disabled. When the alternator voltage is present at terminal D+, T2 will block after a few seconds, enabling the oscillator to start working.

The rectangular oscillator signal is buffered and inverted by T3, which, together with T4, C2, C5, D6 and D7, forms a voltage doubler to ensure that the MOSFET is driven with a sufficiently high gate voltage. Here, this voltage is about 21 V, or 6 V higher than the voltage across the lamps when the engine runs.

Diodes D8 and D9 serve to keep the parking lights (if available on the car) in function. The battery voltage is filtered and limited to 15 V by components R11-C3 and D5 respectively. The circuit is connected to the electrical system of the car via terminals 56b, 58L and 58R.

Bosch code	Function
15	battery voltage (switched)
30	battery voltage (unswitched)
31	chassis
56b	dimlights
58L	parking light and left tail light
58R	parking light and right tail light
D+	alternator voltage

Construction

The circuit is best constructed on a printed circuit board of which the artwork is shown in Fig. 2. The only parts not accommodated on the board are the MOSFET, T1, and the associated VDR, R14. Both are mounted on one side of the metal enclosure. For optimum suppression of voltage surges, the VDR must be fitted as close as possible to the transistor. The photograph shows the prototype fitted in a Hammond enclosure, for which there should be ample space either in the engine compartment or below the dashboard.

Spade receptacles and short, heavy-duty wires are used to connect the unit to the appropriate points in the car's electrical system. In most cars, an extra connection to any of these points can be made without problems by fitting two-way spade terminals.

The completed unit is fitted at a suitable location behind or underneath the dashboard. As already noted, the connection to the car's electrical system is in parallel, i.e., you need not cut any of the existing wiring.

The codes with the terminals indicated in the circuit diagram follow the Bosch convention which is used in many cars. Table 1 lists the functions of the relevant terminals to assist owners of cars in which the Bosch code is not used.

The electrical connection of the automatic headlight switch is illustrated in Fig. 3. The unit is powered when the ignition switch is closed. Upon detection of a sufficiently high alternator voltage at terminal D+, the unit powers the headlights with a pulse-width modulated voltage. When the dipped lights are switched on by the driver, terminal 56b carries the battery voltage. Consequently, transistor T1 is short-circuited so that the headlight switch has no effect on the lamps, which then light at their full intensity. ■

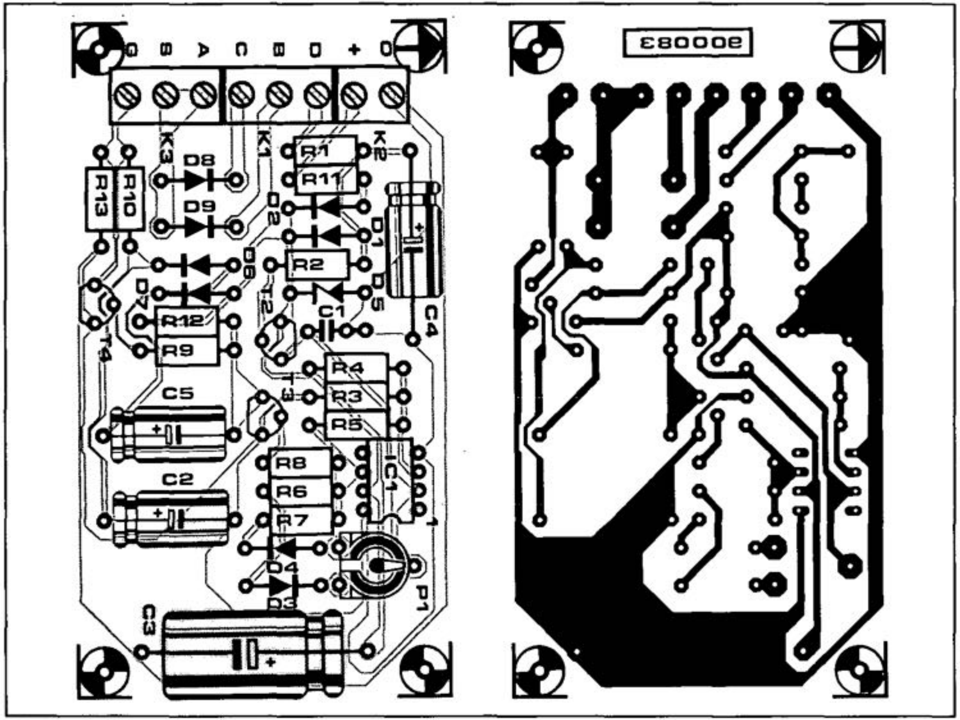


Fig. 2. Printed-circuit board for the lights controller.

COMPONENTS LIST

Resistors:

6	10kΩ	R1 – R5;R10
2	47kΩ	R6;R8
1	4k7	R7
1	100kΩ	R9
1	33Ω	R11
1	470Ω	R12
1	1MΩ	R13
1	SIOV-S07K20	R14
1	25kΩ preset H	P1

Capacitors:

1	330n	C1
1	22µF 35V	C2
1	1000µF 25V	C3
1	220µF 25V	C4
1	47µF 25V	C5

Semiconductors:

1	TLC271	IC1
1	BUZ11	T1
1	BC557	T2
2	BC547	T3;T4
5	1N4148	D1;D3;D4;D6;D7
1	1N4001	D2

INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.

THREE IN A ROW

This month's project is an all-electronic version of the well-known one-armed bandit found in amusement arcades. Apart from the fun in building and using this portable mind-teaser, you have your free introduction into pseudo-random number generation.

T. Giffard

A one-armed bandit is a gambling machine that is played by inserting a coin, pulling a heavy lever (or pressing a button) and waiting anxiously for three identical fruit symbols to appear behind the glass cover as the rolling drums come to a standstill. Unfortunately, this rarely happens, and the owner of the machine has the financial advantage. However, most of us will admit that it is great fun to play a few games, if only for the thrill of it, the noise of the lever, the clicking sounds as the wheels stop, and, of course, the chance of winning.

In terms of electronics, not much is needed to build a circuit that does basically the same as the one-armed bandit, with the exception, of course, of insisting on your money! The fruit symbols are simply replaced by LEDs (light-emitting diodes), and the drums on which they revolve are replaced by counter circuits. Oscillators form the electronic equivalent of a mechanical assembly that powers the wheels.

The result of these 'transformations' is a circuit in which three counters, upon being triggered, start to operate. Each counter has a number of associated LEDs at its outputs that indicate the counter states. These LEDs are arranged vertically, so that one horizontal row consists of three LEDs, each of which belongs with

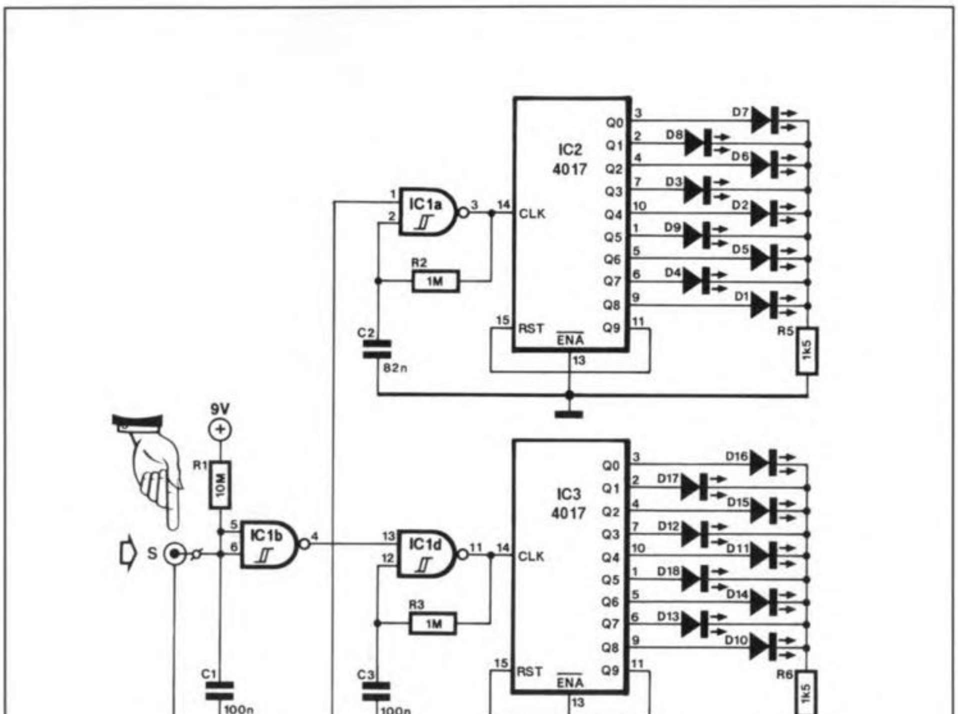


Fig. 1. Circuit diagram of the electronic game.

a different counter. The upshot is that we have three independently operating running lights, of which the indicators, in this case LEDs, start to flash when a button is pressed. When the button is released, each of the three running lights will stop at a particular LED. Statistically, one in 81 runs will result in three LEDs in a row remaining on when the 'play' button is released.

The LEDs may be provided with labels or fruit symbols similar to those used in a real one-armed bandit. Alternatively, numbers may used, and the rules of the game modified accordingly, for instance, a jackpot may be won when three nines appear.

Circuit description

The circuit diagram of the 'three-in-a row' game is given in Fig. 1. The indicators are formed by LEDs D1–D27 which are arranged in three vertical rows of nine LEDs each. The counters are formed by three integrated circuits Type 4017 (IC2, IC3 and IC4). Three of the four NAND Schmitt trigger gates contained in the Type 4093 package are set up as oscillators. The remaining gate, IC1b, functions as an input buffer for the touch-sensitive 'play/stop' button.

When contact S is not touched, the inputs of IC1b are held at a high potential by resistor R1. Since the gate is an inverter, it supplies a low output level which keeps the rest of the circuit disabled.

When contact S is touched, the skin resistance of the finger takes the two inputs of IC1b low. This happens because the skin resistance, although it can take values from a few kilo-ohms to a few hundred kilo-ohms depending on factors such as the degree of moisture, is much lower than 10 MΩ, being the value of R1.

The result of touching contact S is a high level at the output (pin 4) of IC1b. Consequently, the oscillators built around gates IC1a, IC1b and IC1c start to operate. Their output signals have a frequency of about 10 Hz and are fed to the clock inputs (CLK, pin 14)) of the respective counters.

The counters are divide-by-ten (decade) types of which the operation is illustrated by Fig. 2. As shown by the inset timing diagram, outputs Q0–Q9 go high one after another on arrival of the clock pulses. The desired 'running light' effect is therefore simple to create by connecting LEDs to the Q outputs of the counter.

Resistors R5, R6 and R7 limit the LED currents to a safe value. Note that only one LED in a group lights at a time, and that the 4017s supply virtually the battery voltage when an output goes high. Since a red LED drops about 1.6 V when it is on, each series resistor will have to be dimensioned for a drop of 7.4 V (assuming that the battery voltage is 9 V). The values shown in the circuit diagram result in a LED current of 7.4 V/1.5 kΩ ≅ 5 mA. This is, in general, ample current to make a red LED light visibly. In case you want to increase the brightness, use somewhat lower values for resistors R5, R6 and R7. Bear in mind, however, that you can not go much higher than 5 mA because this is about the maximum current that can be supplied by the outputs of the 4017.

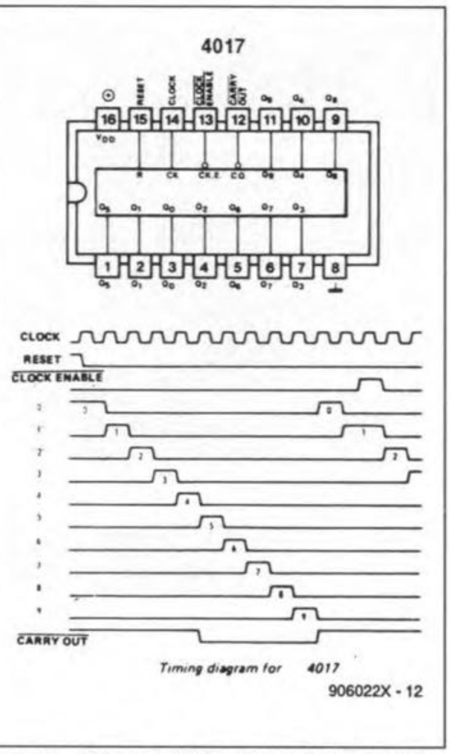


Fig. 2. Pinning, internal structure and timing diagram of the 4017 decade counter.

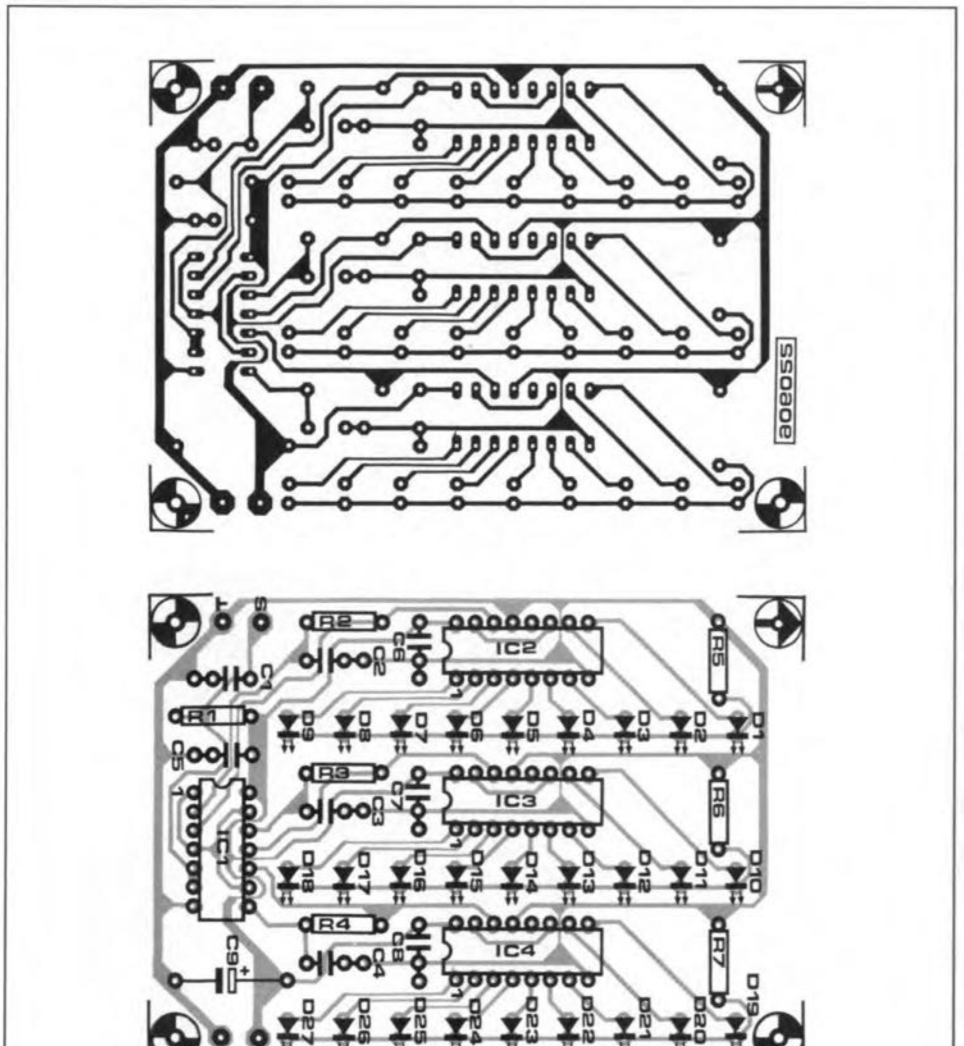


Fig. 3. PCB track layout (mirror image) and component mounting plan.

COMPONENTS LIST

Resistors:		
1	10MΩ	R1
3	1MΩ	R2;R3;R4
3	1k5	R5;R6;R7
Capacitors:		
6	100nF	C1;C3;C5–C8
1	82nF	C2
1	120nF	C4
1	10µF 16V	C9
Semiconductors:		
27	red LED	D1–D27
1	4093	IC1
3	4017	IC2;IC3;IC4
Miscellaneous:		
1	6V or 9V battery	Bt1
1	ABS enclosure with battery compartment. Approx. size: 86×140×32 mm, e.g. Pactec.	

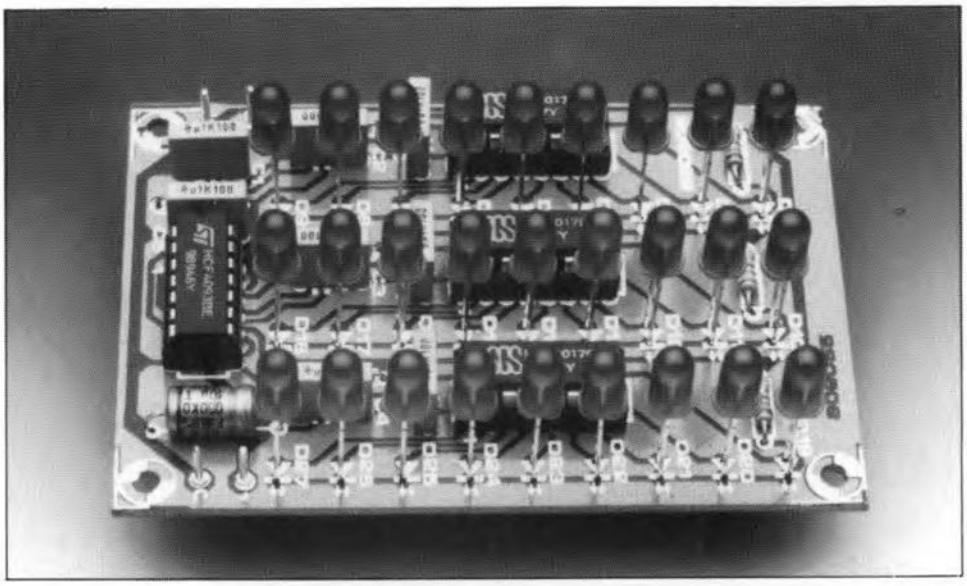


Fig. 4. Completed circuit board before it is fitted into the enclosure.

The Q9 output of each counter is connected to the reset (RST) input, so that the counting process continues as long as contact S is touched.

The frequency-determining capacitors in the oscillators have different values so that the LEDs in the associated group (column of nine each) do not 'run' at the same speed. This is done to prevent players with extremely fast finger movements from winning the game over and over again.

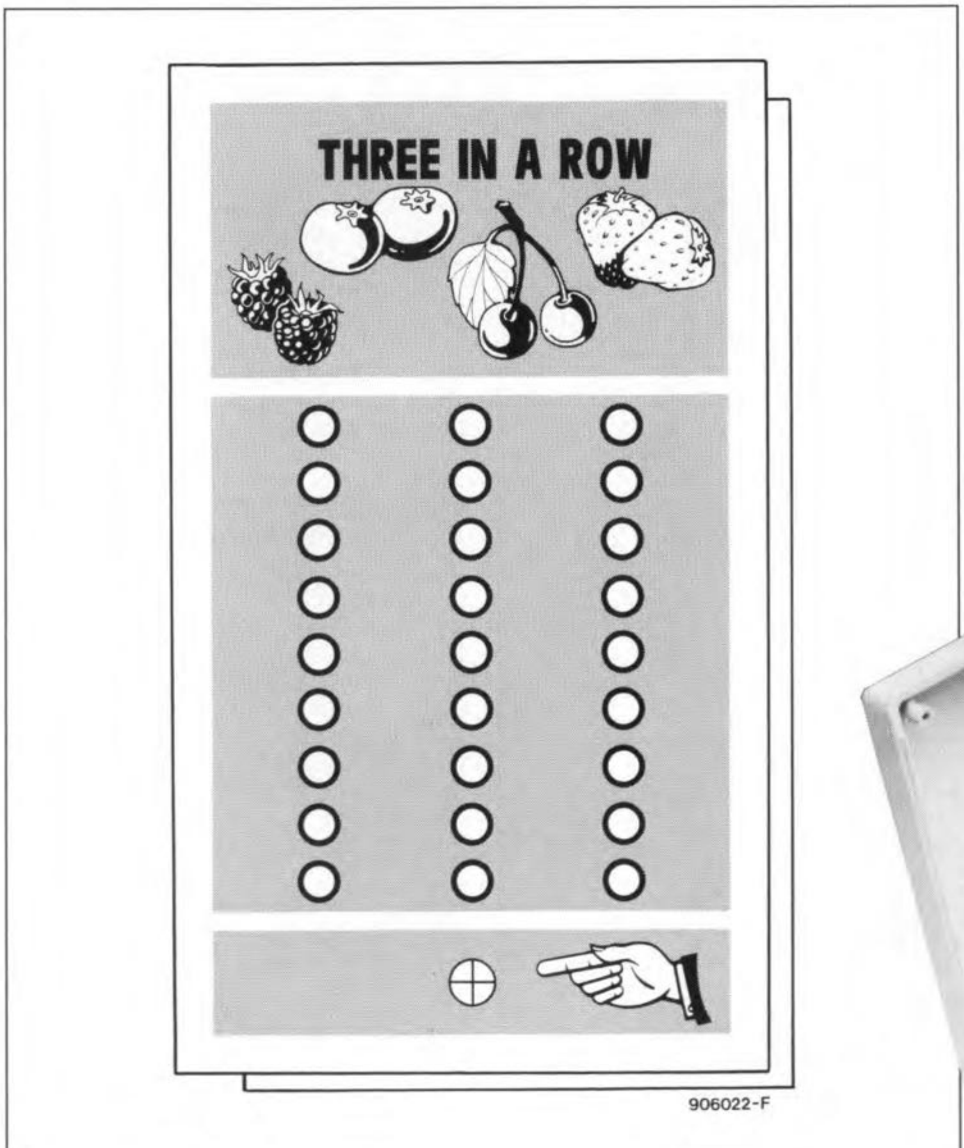


Fig. 6. Suggested front-panel design.

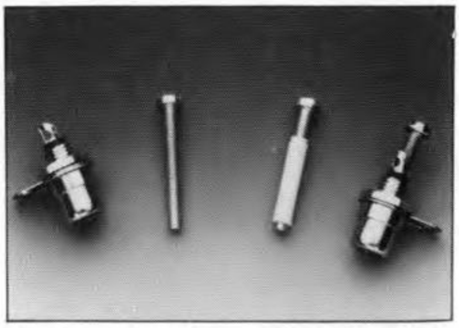


Fig. 5. Parts needed to make the touch-sensitive play/stop control.

Finally, the value of capacitor C_1 is such that the LEDs will keep flashing about one second after releasing the touch-sensitive contact. This is done to make the game more realistic.

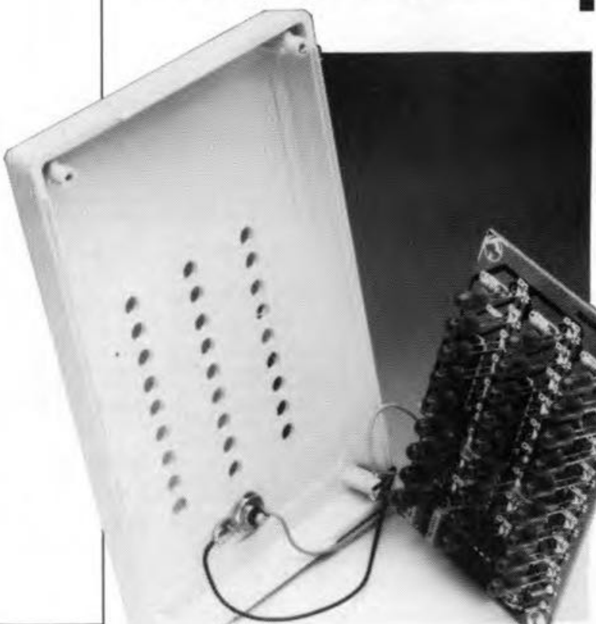
Construction

The printed-circuit board for this project is etched and drilled with the aid of the mirror-image of the track layout shown in Fig. 3. Fitting the components on the board as per the overlay (also shown in Fig. 3) and the parts list is fairly straightforward. Make sure the ICs are fitted the right way around, and also observe the polarity of the LEDs—in most cases, the short terminal is the cathode.

A game like 'three in a row' will not be a success unless the electronics are housed in an attractive enclosure. Make a copy of the component overlay (Fig. 3) and use it as a template for drilling the holes for the 27 LEDs in the top half of the enclosure.

The touch-sensitive switch to start and stop the game is made from a phono socket and an M3 bolt as shown in Fig. 5. A section of the threaded part of the bolt is covered with thin insulating tape to make it fit tightly inside the phono socket.

Finally, the game may be powered either by a pack of four 1.5 V batteries or a single 9 V PP3 battery. The average current consumption is about 15 mA, so that a 9 V battery will last for about 15 hours. ■



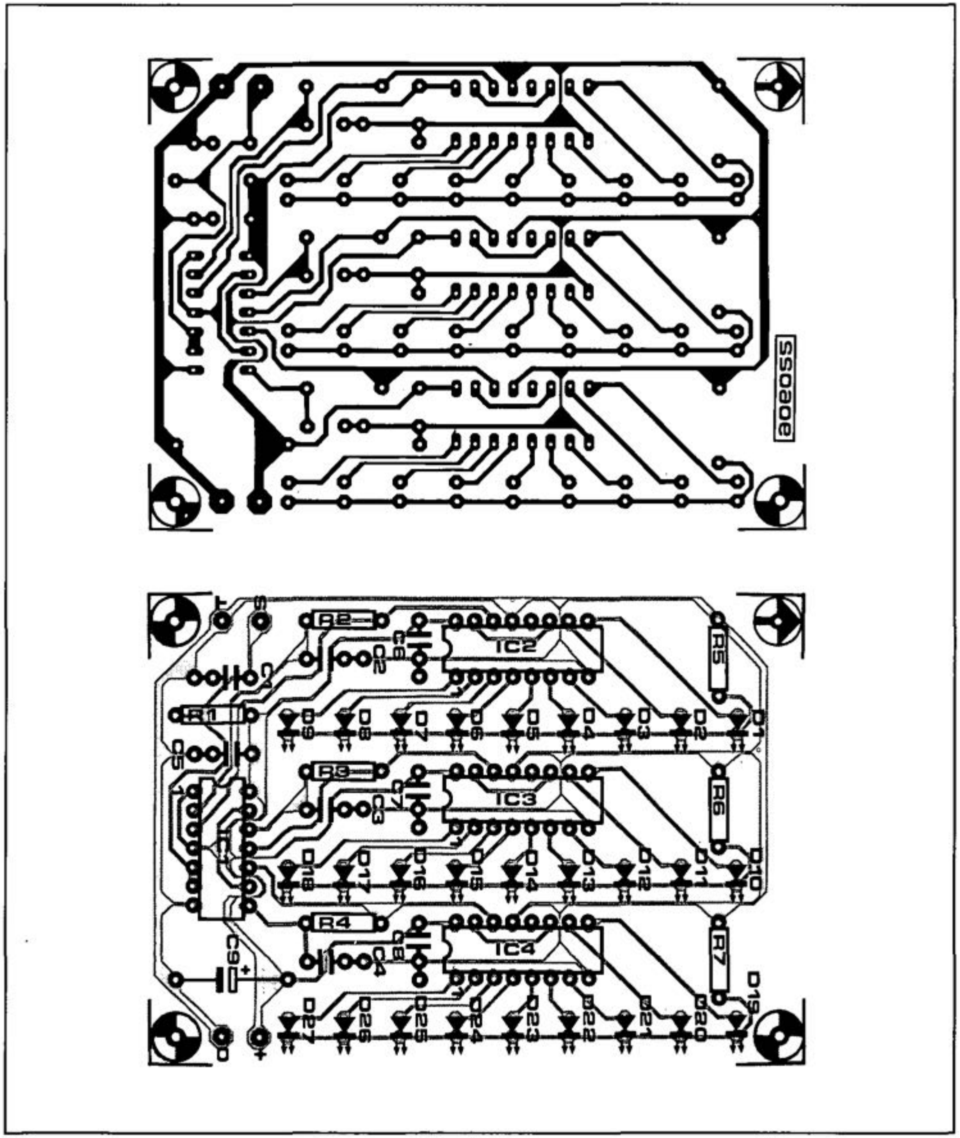


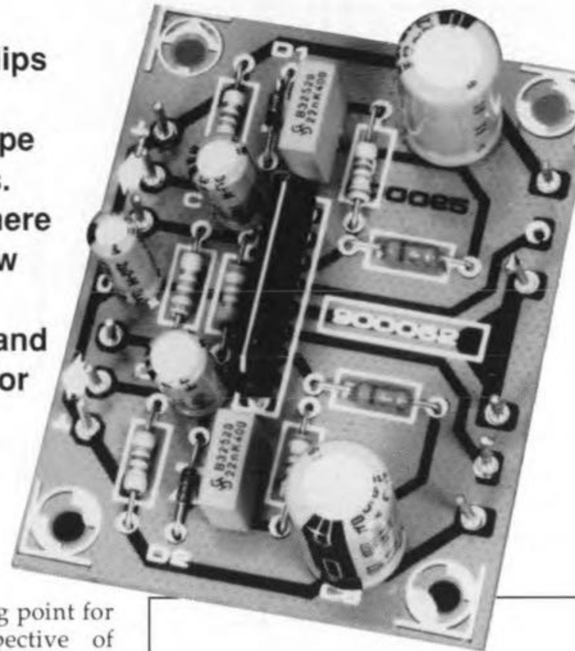
Fig. 3. PCB track layout (mirror image) and component mounting plan.

the 4017s supply virtually the battery voltage when an output goes high. Since a red LED drops about 1.6 V when it is on, each series resistor will have to be dimensioned for a drop of 7.4 V (assuming that the battery voltage is 9 V). The values shown in the circuit diagram result in a LED current of 7.4 V/1.5 kΩ ≅ 5 mA. This is, in general, ample current to make a red LED light visibly. In case you want to increase the brightness, use somewhat lower values for resistors R5, R6 and R7. Bear in mind, however, that you can not go much higher than 5 mA because this is about the maximum current that can be supplied by the outputs of the 4017.

COMPONENTS LIST

Resistors:		
1	10MΩ	R1
3	1MΩ	R2;R3;R4
3	1k5	R5;R6;R7
Capacitors:		
6	100nF	C1;C3;C5–C8
1	82nF	C2
1	120nF	C4
1	10µF 16V	C9
Semiconductors:		
27	red LED	D1–D27
1	4093	IC1
3	4017	IC2;IC3;IC4
Miscellaneous:		
1	6V or 9V battery	Bt1
1	ABS enclosure with battery compartment. Approx. size: 86×140×32 mm, e.g. Pactec.	

STEREO PLAYBACK PREAMPLIFIER



The TDA1522 from Philips Components is a preamplifier chip for tape and cassette recorders. The circuit presented here is remarkable for its low noise level, built-in click-free mute circuit and direct-coupled inputs for the playback heads.

T. Giffard

MAIN FEATURES	
• Supply voltage	7.5 V – 23 V (8.5 V typ.)
• Current consumption:	5 mA (typ.)
• open-loop gain:	90 dB
• Output voltage:	1.5 Vrms (typ.) 720 mVrms (at 0.3 mV in and 315 Hz)
• Distortion:	0.05% (720 mVrms, 1 kHz, R_L = 4k7)
• Input resistance:	> 200 kΩ
• Output resistance:	< 1 kΩ

The TDA1522 is a good starting point for a playback amplifier irrespective of whether this is installed on a 'bare' chassis (reels, motor, drive, etc.), or fitted as an upgrade in an existing recorder. As regards component count and complexity of the circuit, the TDA1522 compares favourably with many opamp-based alternatives.

The TDA1522 contains two high-gain, low-noise amplifiers and an internal mute circuit that allows totally click-free control. The chip does not require bias setting networks, provided its internal regulator receives a supply voltage of 7.5 V or greater. There are no special requirements as regards supply voltage filtering, since the suppression of 100 Hz hum (from a bridge rectifier) is about 95 dB.

The playback heads are connected direct to the inputs of the TDA1522, i.e., without the usual coupling capacitors. The advantage of this direct connection is the absence of clicks and other noises on the tape. Such clicks are usually caused by a coupling capacitor, whose charging current flows through the head.

High amplification

Figure 1 shows the circuit diagram of the playback amplifier. The two playback heads (left and right channel) are connected direct to the respective inputs, pins 4 and 6, of the TDA1522. Playback heads typically have a resistance of 300 Ω and an inductance of 80 mH. Since the input resistance of the amplifiers in the TDA1522 is about 200 kΩ, the signal induced in the heads is hardly loaded.

Since the heads supply a very small signal, the amplifiers must produce little noise themselves, and in addition provide a high amplification factor. The noise level of the TDA1522 is specified at 5 nV/√Hz (at R_G = 0 Ω), while the amplification is determined by a feedback network. Since the amplifiers in the TDA1522 are non-in-

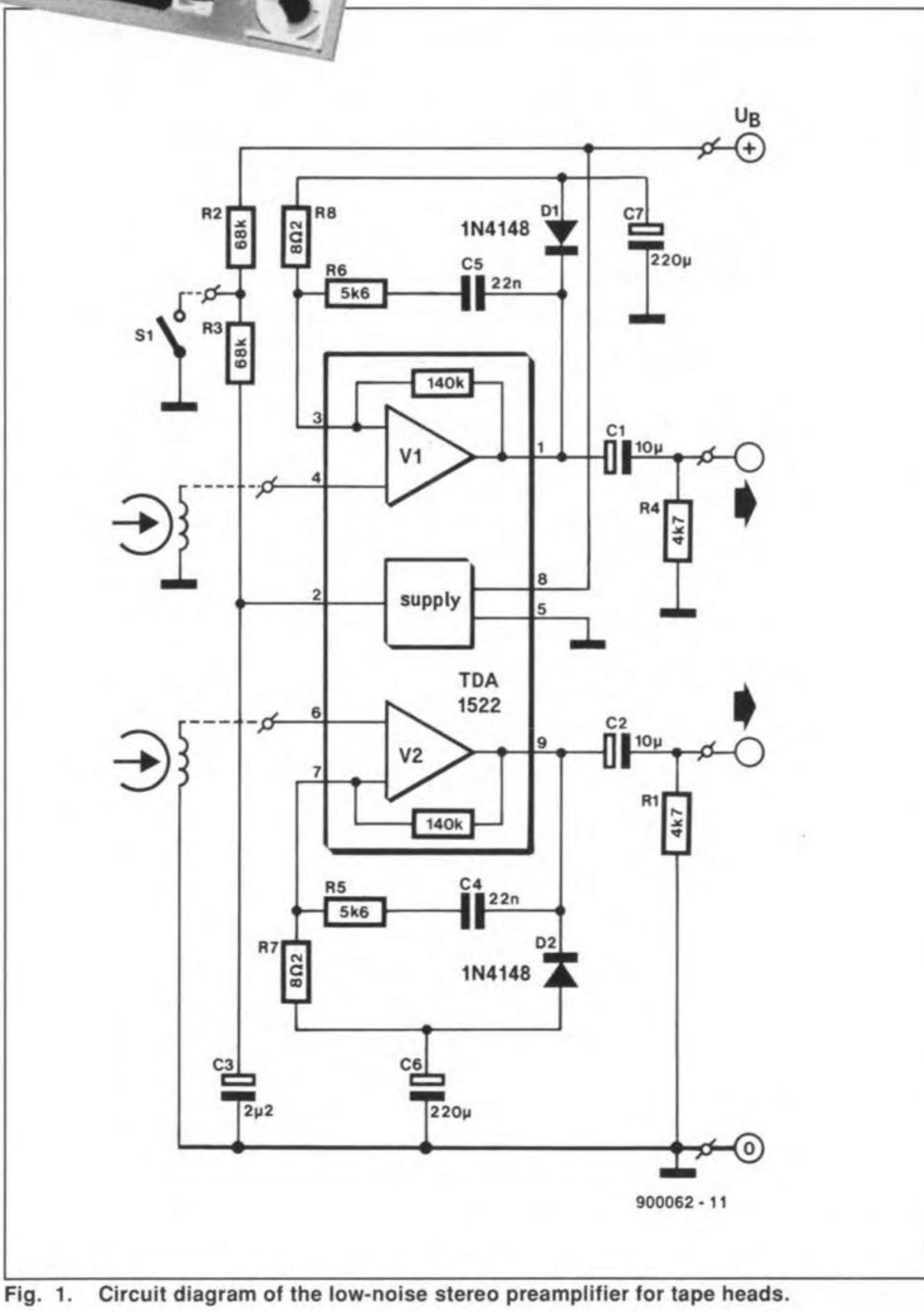


Fig. 1. Circuit diagram of the low-noise stereo preamplifier for tape heads.

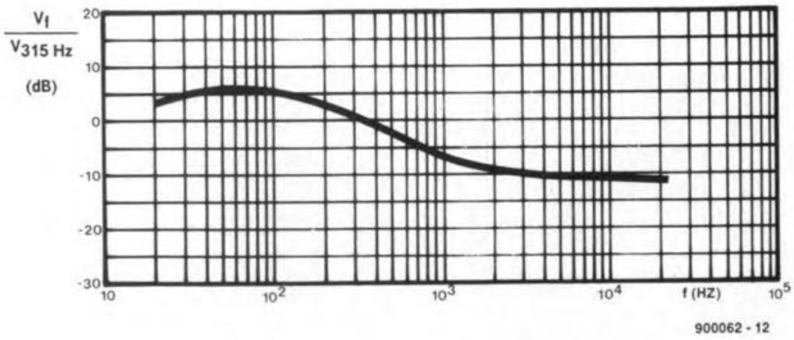


Fig. 2. Frequency response of the preamplifier (120 μs de-emphasis).

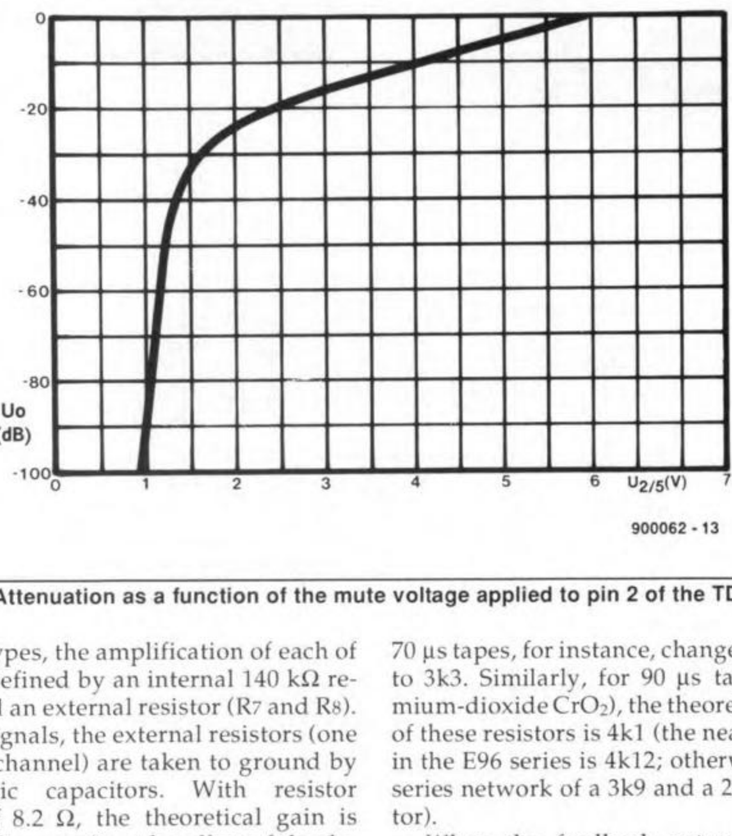


Fig. 3. Attenuation as a function of the mute voltage applied to pin 2 of the TDA1522.

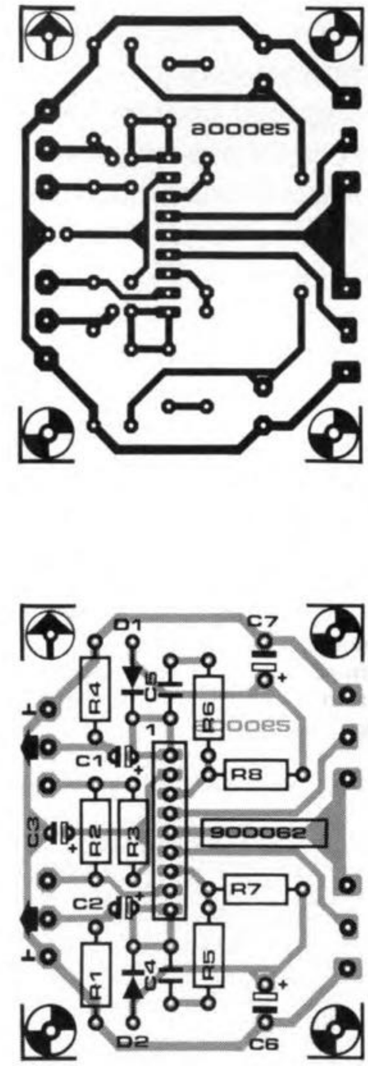


Fig. 4. Track layout and component mounting plan of the PCB for the preamplifier.

verting types, the amplification of each of these is defined by an internal 140 kΩ resistor and an external resistor (R7 and R8). For AC signals, the external resistors (one for each channel) are taken to ground by electrolytic capacitors. With resistor values of 8.2 Ω, the theoretical gain is 84.6 dB, disregarding the effect of the de-emphasis networks R5-C4 and R6-C5.

Frequency response

A network consisting of a 5.6 kΩ resistor (R5; R6) and a 22 nF capacitor is connected in parallel with the 140 kΩ resistor in the TDA1522. This network makes the gain of the preamplifier frequency-dependent (about 68 dB at 315 Hz). Since the reactance, X_c, of the capacitor drops as the frequency rises, the degree of feedback rises so that the amplification drops. Figure 3 shows the frequency response for a pre-emphasis of 120 μs, the standard value for tape type 1 ('ferro' or iron-oxide). The 0 dB level is defined at a frequency of 325 Hz. Other tape types require a corresponding change in the time constant as advised by the manufacturer. The circuit is simple to modify: for 70 μs tapes, for instance, change R5 and R6 to 3k3. Similarly, for 90 μs tapes (chromium-dioxide CrO_2), the theoretical value of these resistors is 4k1 (the nearest value in the E96 series is 4k12; otherwise, use a series network of a 3k9 and a 220 Ω resistor).

When the feedback networks R5-C4 and R6-C5 are omitted, the frequency response becomes linear and the amplification, α, depends on one resistor only:

$$\alpha = 1 + 140k/R_7$$

where R7 is substituted by R8 for the amplification in the other channel.

Without the feedback parts, the circuit may be used as a low-noise stereo preamplifier with high-impedance inputs and low-impedance outputs. One possible application is, therefore, with dynamic microphones.

Mute circuit

The click-free mute circuit in the TDA1522 may be used together with, for instance, a mute switch, a VOX (voice-operated control) or a squelch circuit. The mute func-

COMPONENTS LIST

Resistors:

2	4k7	R1;R4
2	68k	R2;R3
2	5k6	R5;R6
2	8Ω2	R7;R8

Capacitors:

2	10μF 10V radial	C1;C2
1	2μ2 25V radial	C3
2	22nF	C4;C5
2	220μF 10V radial	C6;C7

Semiconductors:

2	1N4148	D1;D2
1	TDA1522	IC1

Miscellaneous:

1	on/off switch	S1

tion is controlled with a direct voltage applied to pin 2 of the TDA1522. The relation between the control voltage and the signal attenuation is shown in Fig. 3. The IC operates at full amplification when pin 2 is not connected, or held at a voltage greater than 6 V. Below this level, the output signal is lightly attenuated at first. Below 2 V, however, the attenuation increases rapidly as shown by the curve. The output signal is virtually blocked when pin 2 is connected to ground or made low by a digital control signal. The most important characteristic of the mute circuit, however, is that it operates in a click-free manner. In the application circuit (Fig. 1), pin 2 is connected to an electrolytic capacitor, C3, and to the positive supply line via R2 and R3. This network forms a power-on delay that prevents switch-on noises, including those produced by circuits ahead of the preamplifier. The mute function is controlled by an external single-pole on/off switch or a push-button connected to junction R2-R3.

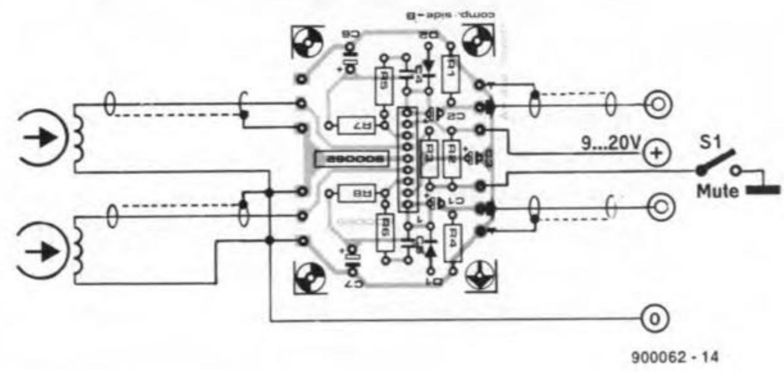


Fig. 5. Wiring digram showing how to connect the preamplifier board in an existing tape recorder. Screened cables must be used at the input to prevent noise and hum.

Printed-circuit board

The circuit shown in Fig. 1 is built on the printed-circuit board shown in Fig. 4. The external connections are simple—see Fig. 5. Note that the two centre terminals at the input side of the circuit board (near R7 and R8) must be connected to ground (0 V of the power supply), since these terminals form the ground path of pin 5 of the TDA1522.

The connections between the heads and the PCB must be kept as short as possible and made in single screened cable. Also make sure that the heads are connected in phase.

Although the wiring at the low-impedance output of the preamplifier is less critical, it is recommended to use screened cable for relatively long connections.

Test voltages in the circuit are the supply voltage at pin 8 of the TDA1522, and the voltage between pin 1 and pin 9, which should be about 3.7 V when the chip is functioning. ■

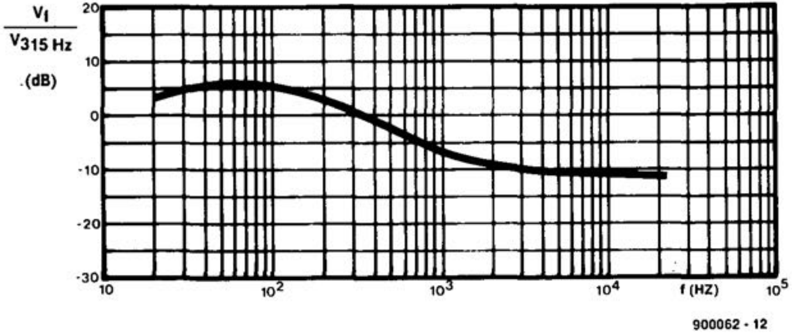
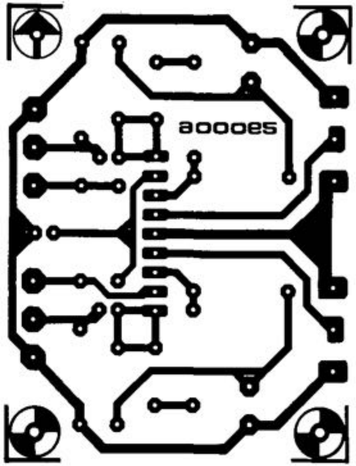


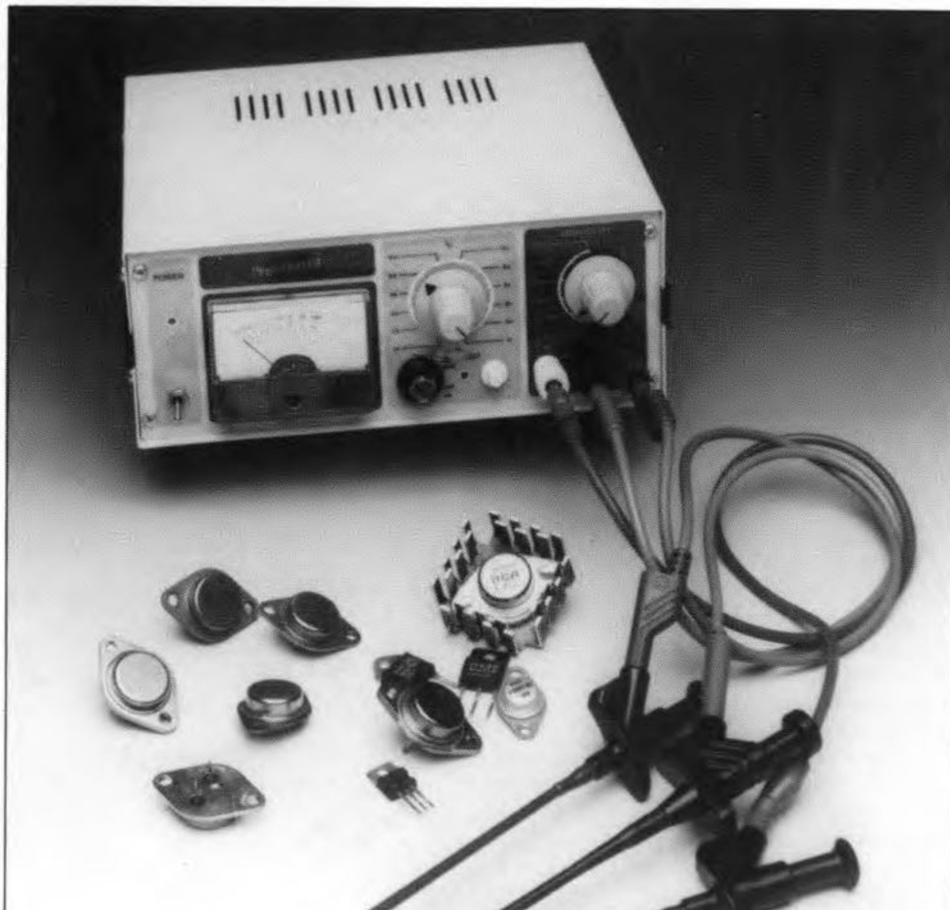
Fig. 2. Frequency response of the preamplifier (120 μs de-emphasis).



HIGH-CURRENT h_{FE} TESTER

A conventional current gain tester is too unreliable for measuring and comparing the d.c. gain of power transistors because it fails to set the collector current at a value that reflects the actual working conditions in, say, the output stage of an audio power amplifier. The tester described here measures the large-signal gain, h_{FE}, of n-p-n as well as p-n-p transistors at a collector current of up to 10 A, yet does not require a bulky power supply.

from an idea by C. Sanjay



Complementary power transistors with matching d.c. gain characteristics are often used in power amplifiers to prevent problems with thermal drift, unbalanced operation and, worst of all, unacceptable distortion. Similarly, in series-regulated power supplies, it is often required that two or more parallel-connected power transistors have virtually the same gain to ensure equal current distribution.

Selecting power transistors with matching gains from a batch is definitely worthwhile for the above applications. Two problems may arise, however. First, the conditions under which the power transistors are to operate differ considerably from those presented by a conventional h_{fe} tester. The tester typically measures the small-signal gain of the transistor at a collector current not higher than 100 mA or so, while in 'real life' the power transistor may carry up to 100 times more current, causing its junction temperature to rise and the d.c. gain to change. This makes low-signal h_{fe} testing of little use for power transistors.

The second difficulty follows from the first: assuming there is a need to test at high collector currents (several ampères), the cost and size of a suitable power supply and a heat-sink for the transistor under test may be prohibitive factors. Also, an ammeter to read the collector current during the test may not be available.

Principle of operation

The present circuit overcomes the above problems in an elegant way by pulsing the collector current at a duty factor of about 0.01. This enables the circuit to be powered by a supply of modest current rating—say, 1 A or so. An impractical collector current meter is also avoided by reversing the principle of current gain measurement: set a certain collector current and measure the (much smaller) base current that produces it (see the inset column on the design background). This measurement is performed during the 'on' time of the collector current, and the values obtained are processed by a sample-and-hold circuit whose timing is controlled by the same generator that supplies the pulses to the transistor under test.

Basic test circuit

The basic test circuit for n-p-n transistors is shown in Fig. 1. Pulses with a duty factor of about 1/100 cause T_1, T_4, T_5 and the TUT (transistor under test) to conduct for about 1.5 ms and block for about 150 ms. Because of the presence of a 3.9-V zener diode, D_z, transistor T_4 limits the voltage across the collector resistor, R_c, of the TUT to about 3.3 V. If the voltage rises above this level, T_5 is blocked and consequently the TUT also. The maximum voltage allowed across R_c thus establishes a constant collector current, at which the base current may be measured to obtain the h_{FE} value.

Capacitor C_p between the base and the collector of T_4 prevents oscillation in the current source by allowing some time for T_5 to start conducting. The disadvantage

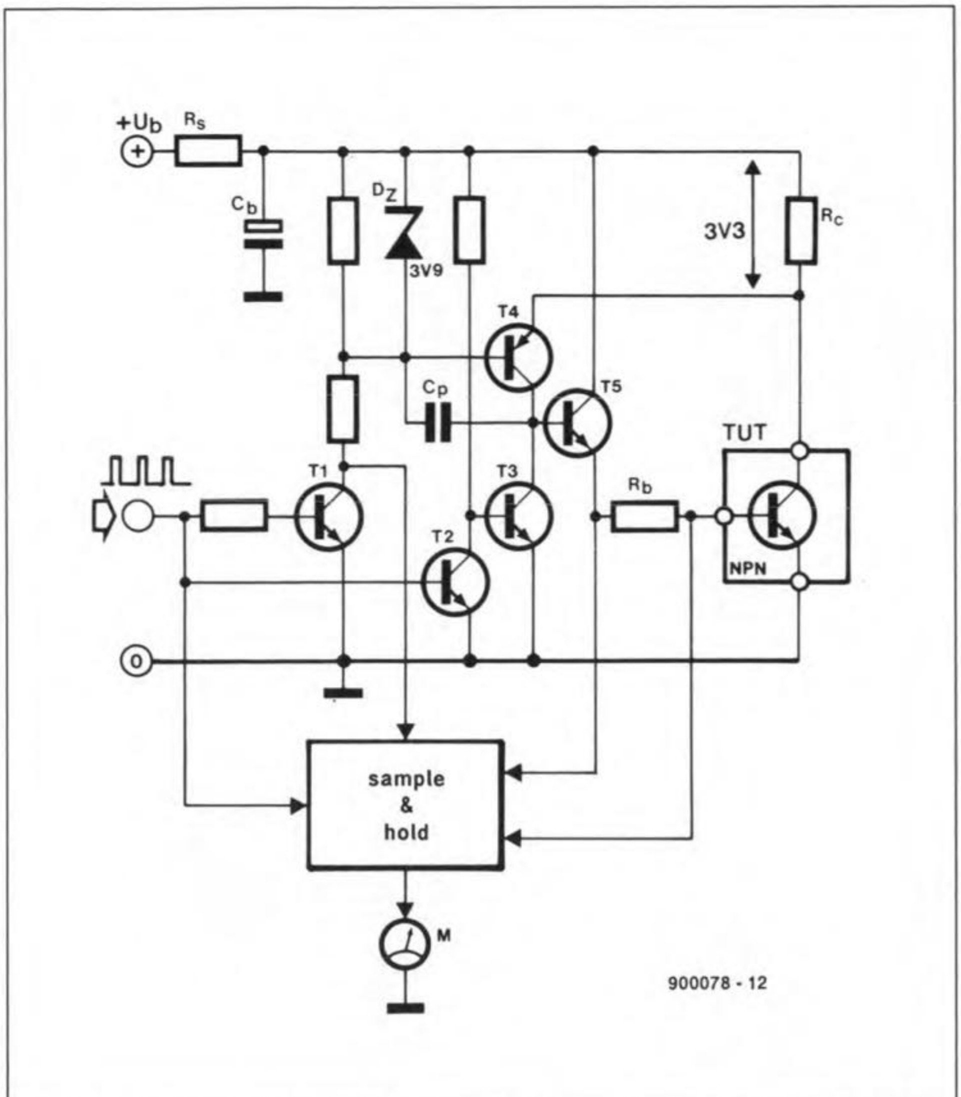


Fig. 1. Basic circuit of the current-gain tester configured for n-p-n transistors. The transistor under test is allowed to draw a relatively high current with a small duty factor from buffer capacitor C_b via collector resistor R_c. A sample-and-hold measurement circuit derives the current gain parameter, h_{FE}, from the level of the pulsed base voltage. The test circuit is in principle the same for p-n-p transistors.

of this form of compensation is that C_p is briefly connected to a higher voltage when T_4 switches off. However, transistors T_2 and T_3 prevent current surges through T_5 and the TUT by pulling the base of T_5 to ground in between pulses.

The (high) current that flows through the TUT during the 1.5-ms long 'on' time is supplied by a reservoir capacitor, C_b.

For p-n-p transistors, the basic test circuit is duplicated with complementary transistor types. The same zener diode D_z and the same power resistor R_c are used, however, to enable comparative gain tests to be carried out on pairs of complementary TUTs.

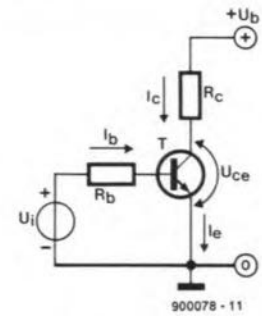
A sample-and-hold circuit measures the voltage drop across the base resistor, R_b, of the TUT. This measurement is performed under the control of the generator pulses. The output voltage produced by the sample-and-hold circuit is fed to a moving-coil meter, M, provided with a 1-to-infinite scale. The indicated value is multiplied by two factors, the collector current multiplier and the sensitivity, to arrive at the h_{FE} value.

Circuit description

Figure 2 shows the circuit diagram of the h_{FE} tester for n-p-n and p-n-p power transistors. The pulse generator is formed by opamp IC_{1a}, which is wired as an astable multivibrator. Two series-connected LEDs, D_3 and D_4, supply the 3.9-V reference voltage for the current source. The red LED, D_3, drops about 1.8 V, and the green LED, D_4, about 2.1 V.

The base current drive circuits consist of T_1–T_5 for n-p-n TUTs, and T_1-T_6–T_{10} for p-n-p TUTs. Depending on the range selected, the base of the TUT develops a voltage across one of the six resistors connected to the contacts of switch S_3. During the 1.5-ms long measurement period, electronic switches IC_{3c} and IC_{3d} are closed, while IC_{3a} and IC_{3b} are open. This results in the voltage developed across the base resistor being stored in capacitor C_3. The electronic switches change state during the 'off' period of the generator output signal, so that the measured voltage exists with reference to ground and can be fed to the meter driver, opamp IC_{1B}. Capacitor C_4 smooths the output voltage supplied

DESIGN BACKGROUND



The static forward current transfer ratio, h_{FE}, of a bipolar transistor in a common-emitter circuit is defined as the ratio of d.c. output current to the d.c. input current, or

$$h_{FE}=\frac{I_c}{I_b}$$

provided that I_{ceo}, the collector current with the base open-circuited, is much smaller than I_c.

In the test circuit,

$$I_c=\frac{3.3V}{R_c} \quad \text{and} \quad I_b=\frac{U_{Rb}}{R_b}$$

Substituting:

$$\boxed{h_{FE}=\frac{I_c \times R_b}{U_{Rb}}}$$

The values of R_c result in I_c values of 1 A, 2 A, 3 A, 4 A, 6 A and 10 A.

Examples:

1. When I_c = 1 A; R_b = 10 Ω and U_{Rb} = 2.5 V (full-scale deflection):

$$h_{FE}=\frac{1 \times 10}{2.5}=4$$

2. When I_c = 1 A; R_b = 24.9 Ω and U_{Rb} = 2.5 V (full-scale deflection):

$$h_{FE}=\frac{1 \times 24.9}{2.5} \approx 10$$

3. When I_c = 3 A; R_b = 249 Ω and U_{Rb} = 1.75 V:

$$h_{FE}=\frac{3 \times 249}{1.75} \approx 427$$

Note:

The d.c. gain parameter, h_{FE}, must not be confused with the common-emitter small-signal short-circuit forward current transfer ratio, h_{fe}, which is defined in terms of a.c. as

$$h_{fe}=\frac{\delta I_c}{\delta I_b}$$

with the output short-circuited to a.c.

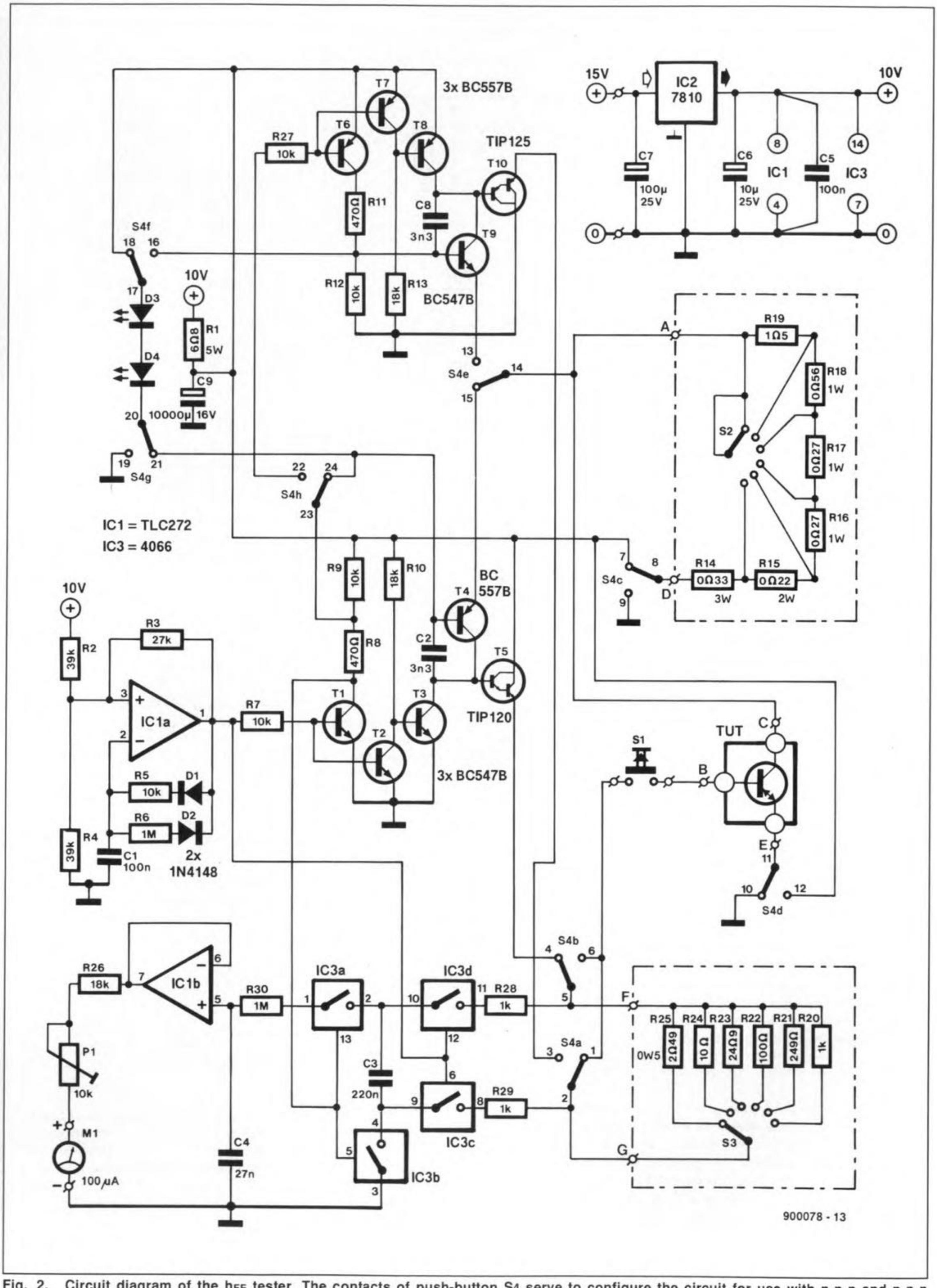


Fig. 2. Circuit diagram of the h_{FE} tester. The contacts of push-button S4 serve to configure the circuit for use with n-p-n and p-n-p transistors.

by the sample-and-hold circuit.

Preset P1 at the output of IC1B serves to set full-scale deflection on the moving-coil meter when the base resistor drops 2.5 V. Note that the meter deflection is inversely related to the h_{FE} value, creating the need for a meter scale from infinite (no deflection) to 1 (full deflection), similar to that used on an analogue ohm-meter.

The collector currents for the six available ranges are determined by power resistors connected to the contacts of switch S2. The circuit is switched from n-p-n testing to p-n-p testing by eight toggle contacts on switch S4.

The power supply of the h_{FE} tester is conventional and based on a three-terminal 10-V regulator Type 7810 (IC2). The unregulated input voltage to the circuit should be between 15 V and 20 V.

Construction

Start the construction by fitting the six base resistors (1% types from the E96 series) on the 'sensitivity' switch, S3. Each of the resistors is soldered between its terminal on the switch and a common junction. Resistor R25 goes to the switch terminal marked '1', R24 to switch terminal '2' and so on. Next, connect approximately 10-cm long light-duty wires to the pole and the resistor junction.

The collector current switch, S2, also has the associated resistors, in this case R14–R19, fitted direct at its terminals. Resistor R19 is connected between switch terminals '1' and '2', resistor R18 between switch terminals '2' and '3', and so on. The pole of the switch and the last resistor, R14, are connected to the circuit board by two 5-cm long insulated wires. The respective PCB terminals are marked 'A' and 'D'.

Fit the ten wire links on the printed-circuit board as indicated by the component overlay (see Fig. 4). Next, fit all parts on the board as indicated by the parts list and the component overlay.

LED D4 is not fitted on the board, but

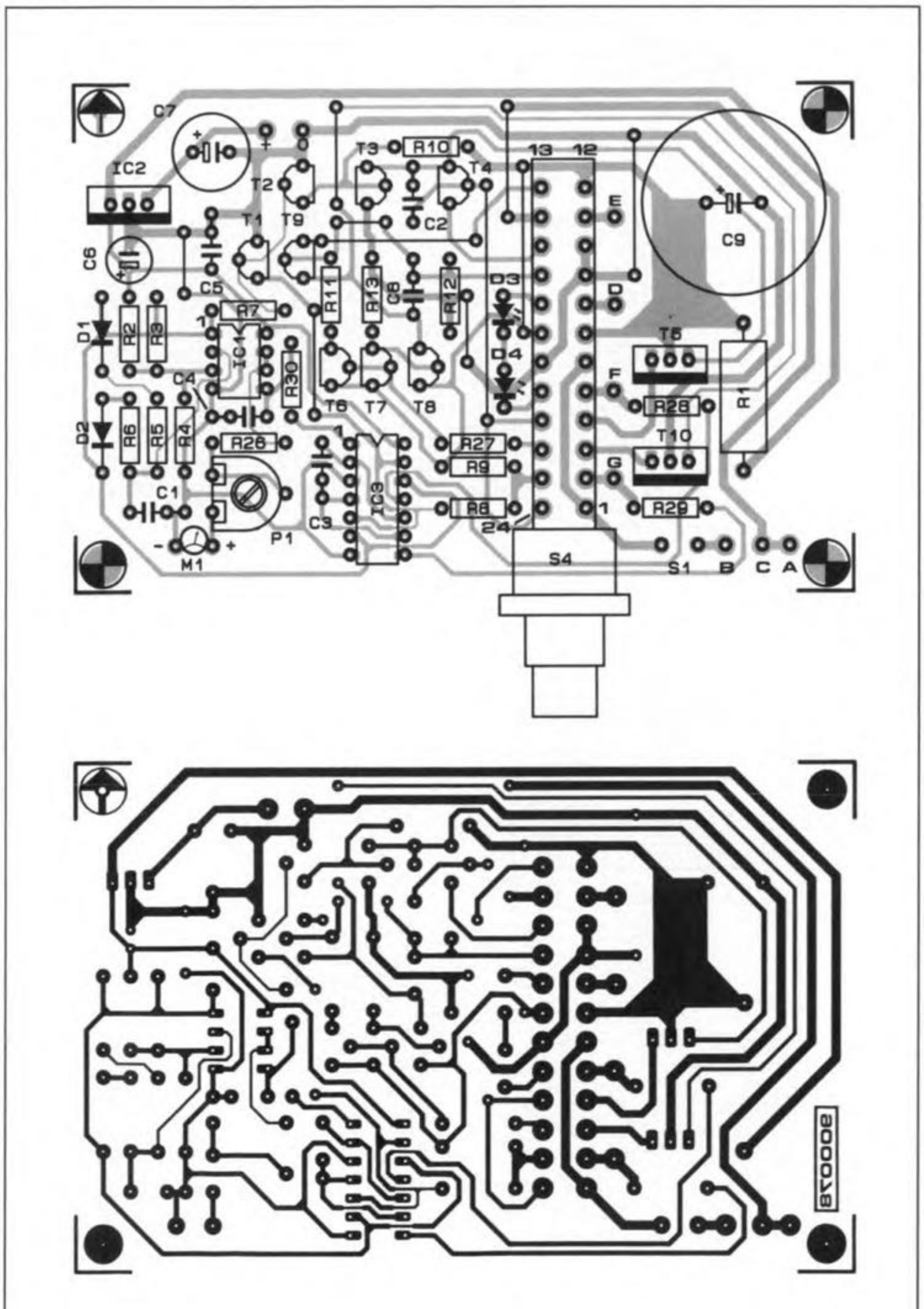


Fig. 3. Track lay-out (mirror image) and component mounting plan.

COMPONENTS LIST

Resistors:		
1	6Ω8 5W	R1
2	39kΩ	R2;R4
1	27kΩ	R3
5	10kΩ	R5;R7;R9;R12;R27
2	1MΩ	R6;R30
2	470Ω	R8;R11
3	18kΩ	R10;R13;R26
1	0Ω33 3W	R14
1	0Ω22 2W	R15
2	0Ω27 1W	R16;R17
1	0Ω56 1W	R18
1	1Ω5	R19
1	1kΩ 1%	R20
1	249Ω 1%	R21
1	100Ω 1%	R22
1	24Ω9 1%	R23
1	10Ω 1%	R24
1	2Ω49 1%	R25
2	1kΩ	R28;R29
1	10kΩ preset H	P1
Capacitors:		
2	100nF	C1;C5
3	3nF3	C2;C8
1	220nF	C3
1	27nF	C4
1	10μF 25V radial	C6
1	100μF 25V radial	C7
1	10,000μF 16V radial	C9
Semiconductors:		
2	1N4148	D1;D2
1	green LED (3 mm)	D3
1	high-efficiency red LED (3 mm)	D4
4	BC547B	T1;T2;T3;T9
4	BC557B	T4;T6;T7;T8
1	TIP120	T5
1	TIP125	T10
1	TLC272	IC1
1	7810	IC2
1	4066	IC3
Miscellaneous:		
1	push-button with 1 make contact	S1
2	1-pole 6-way rotary switch	S2;S3
1	locking 8× changeover switch Schadow type F-8U/EE. Accessory parts: indicator button FA200, upper and lower cap half M14, spring washer, plastic nut and front panel cap (parts set type ZFA)	S4
1	100-μA moving-coil meter e.g. Monacor (Monarch) type PM-2	M1
1	printed-circuit board	900078
1	front-panel foil	900078-F
1	enclosure Telet LC850	

connected to it via 10-cm long wires. The LED is an used on the front panel as a 'test' indicator that flashes when the associated button is pressed.

The vertically mounted power transistors, T5 and T10, and the voltage regulator, IC2, do not require heat-sinks. Use IC sockets for the TLC272 and the 4066. The completed printed-circuit board and the wired range switches are shown in Fig. 5. All wires are soldered direct to the board—solder terminals must not be used.

Testing

It is recommended to test the completed board before fitting it into an enclosure. Connect the moving-coil meter, the three switches and the power supply (15–20 VDC) to the board.

Temporarily disconnect R30 from pin 1 of IC3, and connect it to the junction of a voltage divider that supplies 2.5 V (e.g., a 1-kΩ resistor and a 500Ω preset connected between the +10-V line and ground). Switch on and check the output voltage of the potential divider. Adjust the preset until +2.5 V is produced. Next, adjust P1 for full-scale meter deflection. Remove the potential divider.

Connect an oscilloscope and check the presence of a 10-V_{pp} rectangular signal at pin 13 (negative-going) and pin 12 (positive-going) of IC3. Because of the small duty factor of the signal, it may be necessary to switch the input of the scope to DC-coupled.

Connect a test transistor to the circuit. Check that the two LEDs drop 3.9 V or a voltage within 10% of that value. If they do not, use other LEDs (e.g., a high-efficiency type for the red one).

Finally, connect R30 to pin 1 of IC3 again.

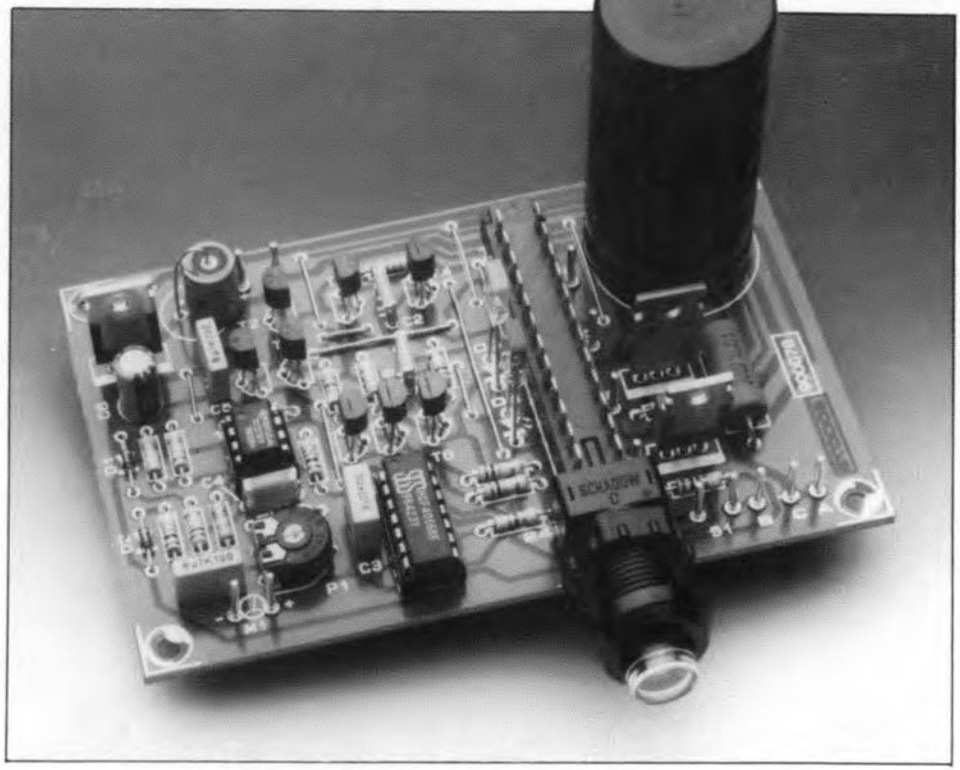


Fig. 4. Completed PCB ready for fitting into the enclosure.

Assembly

The instrument is relatively simple to build in an aluminium enclosure Type LC850 from Telet. A ready-made front panel (Fig. 7) finished in light and dark blue is available to give the tester an attractive appearance, in line with previous instruments in this series.

First, remove the existing scale from your moving-coil meter and replace it with the scale cut out from the front panel foil. Carefully re-assemble the meter and connect short wires to it.

Next, cut and drill the metal front panel of the LC850 case, using the drilling template supplied with the front panel foil

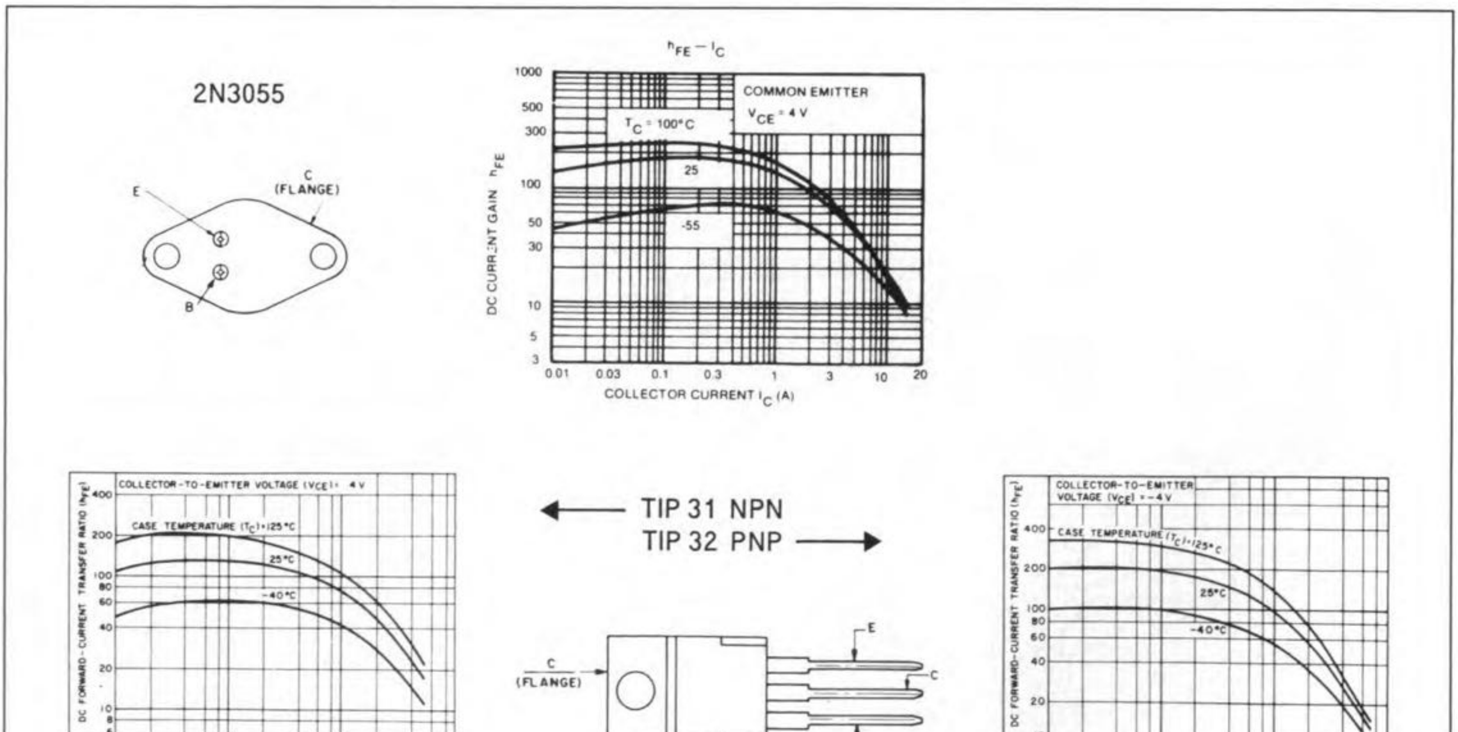


Fig. 5. For your reference: h_{FE} curves of three well-known power transistors (courtesy Harris Semiconductor, formerly RCA).

to mark the holes.

Before applying the self-adhesive foil to the metal front panel, make sure all drilled holes have the right diameter. This is easily checked by temporarily fitting the front panel controls, the two LEDs (the 'power' indicator and the 'test' indicator), the three wander sockets and the meter.

The pnp/npn switch, S4, secures the PCB to the front panel. Determine how far the PCB can be moved towards the front panel without the rear side of the meter touching any component on the board. Check that you can secure the switch to the front panel with the aid of the locking nut at the inside and the threaded cap at the outside. Mark the final position of the board on the bottom plate and drill the four holes for the M3 screws in the corners. Use short PCB spacers or a couple of nuts and washers to fit the board at the correct height above the bottom plate.

Next, remove all parts from the front panel, and carefully apply the self-adhesive foil. The two 3-mm LEDs are glued in position at the rear side of the panel. They do not protrude from the front panel foil.

Fit the range switches (complete with the resistors soldered at the terminals), the wander sockets, the 'test' push-button, the on/off switch and the meter on the front panel. Cut the spindles of the range switches to the required length, and fit the collet knobs. The second pointer on the collector current switch, a small solid triangle, is applied as a rub-off symbol on the collet.

Secure the PCB on the bottom plate of the enclosure. Next, fit the front panel with all parts on it. Secure it to the case with the four self-tapping screws supplied. Finally, tighten the threaded cap of the npn/pnp switch.

The remainder of the wiring work is straightforward. The 'power' indicator, a 3-mm LED, and the associated on/off switch are connected to a DC adapter socket on the rear panel of the enclosure. A 1-kΩ series resistor is used for the LED.

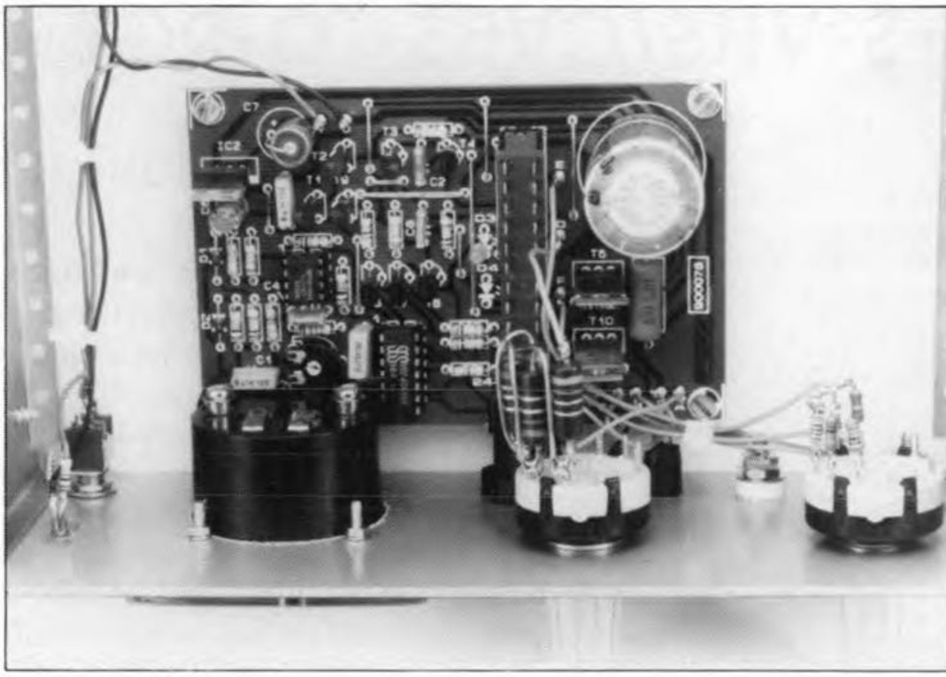


Fig. 5. A look inside the completed prototype. Note that the resistors are soldered direct to the terminals of the rotary switches.

The input voltage connections are made in light-duty red and black wires.

Practical use

A few points must be made as regards the use of the h_{FE} tester. First, always make sure the n-p-n/p-n-p switch is in the right position. You may destroy a transistor if you do not stick to this rule. Second, make sure you get the emitter, base and collector terminals right. Third, always start the test in the lowest collector current range, 1 A, and the lowest sensitivity, 1×. Increase the sensitivity until a usable meter indication is produced. The h_{FE} value is calculated from

meter value × I_c multiplier × sensitivity

Familiarize yourself with the operation of the instrument by measuring the current gain of a few power transistors from your junkbox. The curves in Fig. 6 show the relation between the collector current and the current gain of the well-known 2N3055 and the complementary pair TIP31/TIP32. Bear in mind the maximum collector current of the transistor under test—a medium-power transistor such as the BD139, for instance, will almost certainly be destroyed when the tester is set to the 10 A range.

Having built the instrument you are in a position to select matching pairs of complementary power transistors for your next power supply or AF amplifier. Because of the low duty factor of the collector current passed through the transistor under test, it is not normally required to fit this on a heat-sink, not even while testing in the 10 A range. ■

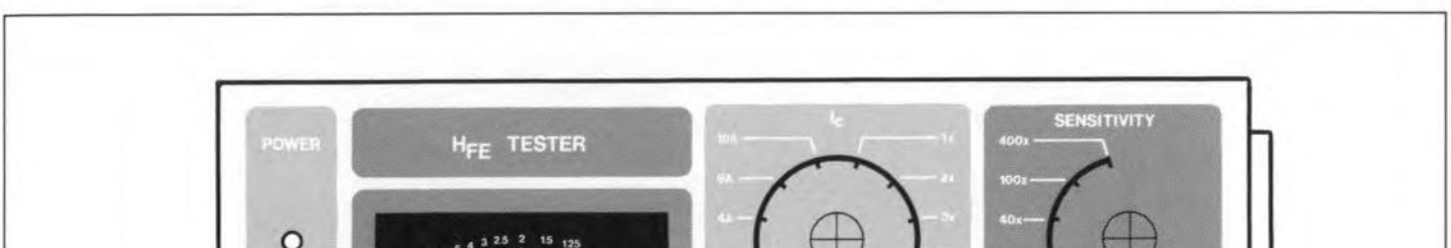


Fig. 7. This front panel foil (shown here at 66% of true size) is available ready-made to give your current gain tester a professional appearance. The meter scale is cut out from the foil and applied to the moving-coil meter.

by the sample-and-hold circuit.

Preset P1 at the output of IC1B serves to set full-scale deflection on the moving-coil meter when the base resistor drops 2.5 V. Note that the meter deflection is inversely related to the h_{FE} value, creating the need for a meter scale from infinite (no deflection) to 1 (full deflection), similar to that used on an analogue ohm-meter.

The collector currents for the six available ranges are determined by power resistors connected to the contacts of switch S2. The circuit is switched from n-p-n testing to p-n-p testing by eight toggle contacts on switch S4.

The power supply of the h_{FE} tester is conventional and based on a three-terminal 10-V regulator Type 7810 (IC2). The unregulated input voltage to the circuit should be between 15 V and 20 V.

Construction

Start the construction by fitting the six base resistors (1% types from the E96 series) on the 'sensitivity' switch, S3. Each of the resistors is soldered between its terminal on the switch and a common junction. Resistor R25 goes to the switch terminal marked '1', R24 to switch terminal '2' and so on. Next, connect approximately 10-cm long light-duty wires to the pole and the resistor junction.

The collector current switch, S2, also has the associated resistors, in this case R14–R19, fitted direct at its terminals. Resistor R19 is connected between switch terminals '1' and '2', resistor R18 between switch terminals '2' and '3', and so on. The pole of the switch and the last resistor, R14, are connected to the circuit board by two 5-cm long insulated wires. The respective PCB terminals are marked 'A' and 'D'.

Fit the ten wire links on the printed-circuit board as indicated by the component overlay (see Fig. 4). Next, fit all parts on the board as indicated by the parts list and the component overlay.

LED D4 is not fitted on the board, but

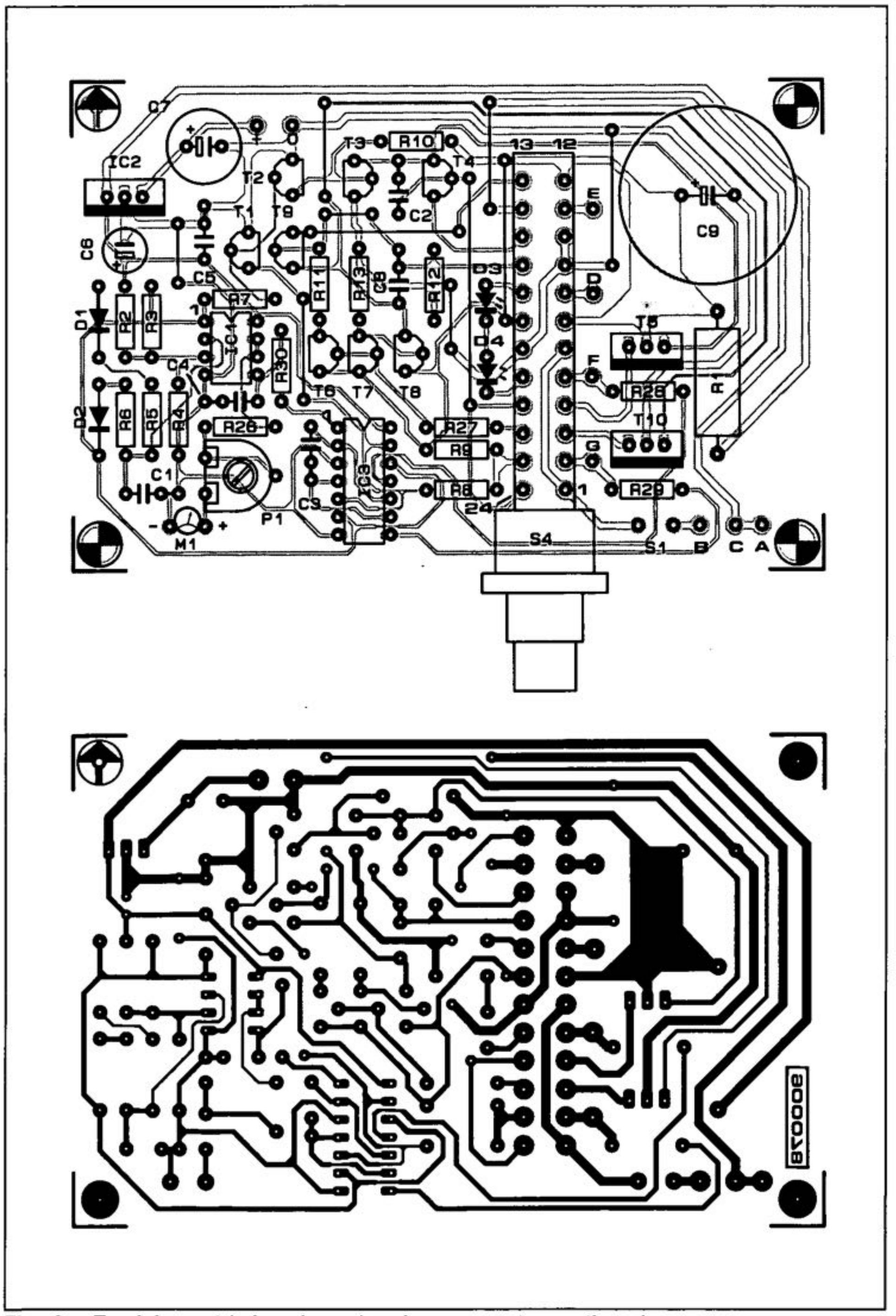


Fig. 3. Track lay-out (mirror image) and component mounting plan.

COMPONENTS LIST

Resistors:		
1	6Ω8 5W	R1
2	39kΩ	R2;R4
1	27kΩ	R3
5	10kΩ	R5;R7;R9;R12;R27
2	1MΩ	R6;R30
2	470Ω	R8;R11
3	18kΩ	R10;R13;R26
1	0Ω33 3W	R14
1	0Ω22 2W	R15
2	0Ω27 1W	R16;R17
1	0Ω56 1W	R18
1	1Ω5	R19
1	1kΩ 1%	R20
1	249Ω 1%	R21
1	100Ω 1%	R22
1	24Ω9 1%	R23
1	10Ω 1%	R24
1	2Ω49 1%	R25
2	1kΩ	R28;R29
1	10kΩ preset H	P1
Capacitors:		
2	100nF	C1;C5
3	3nF3	C2;C8
1	220nF	C3
1	27nF	C4
1	10µF 25V radial	C6
1	100µF 25V radial	C7
1	10,000µF 16V radial	C9
Semiconductors:		
2	1N4148	D1;D2
1	green LED (3 mm)	D3
1	high-efficiency red LED (3 mm)	D4
4	BC547B	T1;T2;T3;T9
4	BC557B	T4;T6;T7;T8
1	TIP120	T5
1	TIP125	T10
1	TLC272	IC1
1	7810	IC2
1	4066	IC3
Miscellaneous:		
1	push-button with 1 make contact	S1
2	1-pole 6-way rotary switch	S2;S3
1	locking 8× changeover switch Schadow type F-8U/EE. Accessory parts: indicator button FA200, upper and lower cap half M14, spring washer, plastic nut and front panel cap (parts set type ZFA)	S4
1	100-µA moving-coil meter e.g. Monacor (Monarch) type PM-2	M1
1	printed-circuit board	900078
1	front-panel foil	900078-F
1	enclosure Telet LC850	

S-VHS/CVBS-TO-RGB CONVERTER

PART 1: INTRODUCTION

Although the technical advantages of the Super-VHS video system are well proven, many owners of an S-VHS video recorder balk at the expense of a compatible monitor or TV set with separate luminance and chrominance inputs. This article describes an obvious missing link in the apparently ever-incompatible field of video equipment connections. An advanced circuit is discussed that converts S-VHS or CVBS (composite) video signals into RGB components. The upshot is that you can use your existing monitor with an RGB input (i.e., with a SCART or Euro-AV connection) to benefit from the improved picture resolution offered by an S-VHS video recorder. This month we discuss the basics of the video standards involved.

H. Reelsen

The compatibility issue has played a significant role in the development of both the NTSC and the PAL TV transmission systems. In both cases, there were two conflicting aspects: on the one hand, existing monochrome TV sets were not to be affected by colour transmissions; on the other hand, existing bandwidths of about 5 MHz for the luminance (brightness) signal were to be maintained.

The compatibility requirement automatically dictates that the black-and-white information (luminance or 'Y' signal) must also be conveyed in colour transmissions. The Y signal forms the sum of all basic colours, red (R), green (G) and blue (B), but only as far as their relative brightness is concerned. From perception experiments, the brightness appears to determine the overall sharpness of the picture. Hence, the luminance bandwidth must be as large as possible (up to 5 MHz) for monochrome as well as colour TV sets. However, this raises the problem of where to put the colour information.

Colour components and transmission

Any colour can be reproduced on a picture tube by actuating in the correct proportion the basic colours it is composed of. The final colour is obtained by controlling the intensity at which the RGB pixels at the inside of the picture tube light up. To the human eye, the three individual basic colours in a pixel group appear as one, composite, colour or hue at a particular saturation.

The need to convey R, G and B, is, therefore, obvious. Since the sum of the equivalent luminance values of all three is already contained in the Y signal, only two further signals, R–Y and B–Y, are generated by means of a differential operation with the Y signal. R–Y and B–Y are therefore referred to as the colour difference signals. Before these signals are transmitted, they are given relative brightness factors. The resulting chrominance signals may be written as

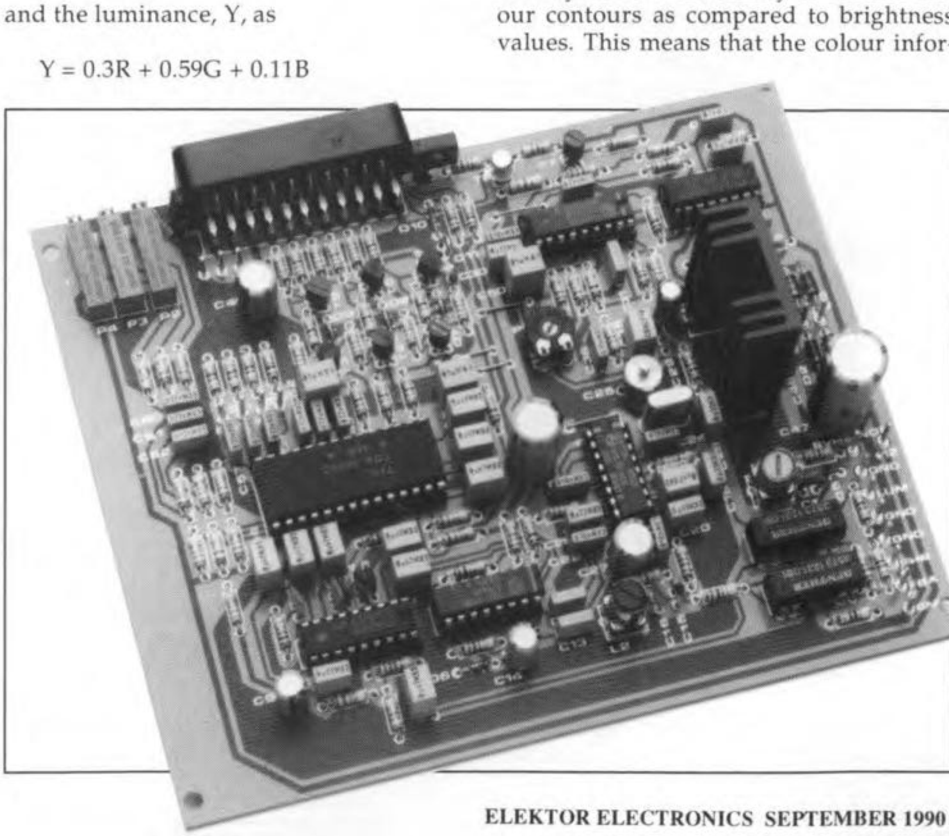
$$U = 0.49(B-Y)$$
$$V = 0.88(R-Y)$$

and the luminance, Y, as

$$Y = 0.3R + 0.59G + 0.11B$$

The RGB intensity information required to control the respective electron guns in the picture tube is obtained from the R–Y, B–Y and Y information with the aid of an addition operation in a matrix circuit.

A problem that remains to be solved is how to include the colour difference signal in the bandwidth already occupied by the Y signal, without causing interference on monochrome TV sets, or reducing the picture sharpness on colour sets. At this point, design engineers are in a position to profit from a characteristic of human eye, namely its reduced ability to resolve colour contours as compared to brightness values. This means that the colour infor-



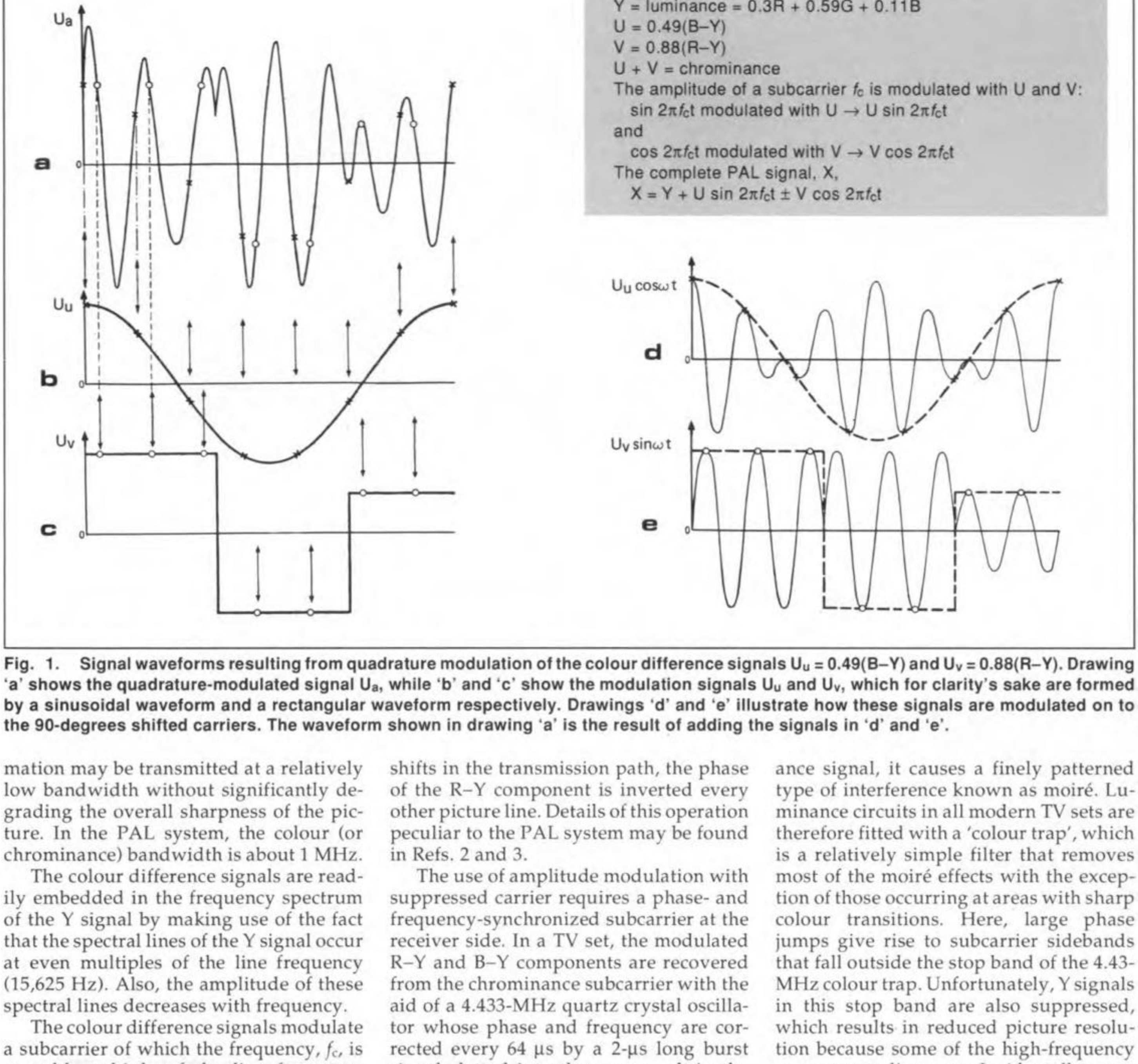


Fig. 1. Signal waveforms resulting from quadrature modulation of the colour difference signals U_u = 0.49(B–Y) and U_v = 0.88(R–Y). Drawing 'a' shows the quadrature-modulated signal U_a, while 'b' and 'c' show the modulation signals U_u and U_v, which for clarity's sake are formed by a sinusoidal waveform and a rectangular waveform respectively. Drawings 'd' and 'e' illustrate how these signals are modulated on to the 90-degrees shifted carriers. The waveform shown in drawing 'a' is the result of adding the signals in 'd' and 'e'.

mation may be transmitted at a relatively low bandwidth without significantly degrading the overall sharpness of the picture. In the PAL system, the colour (or chrominance) bandwidth is about 1 MHz.

The colour difference signals are readily embedded in the frequency spectrum of the Y signal by making use of the fact that the spectral lines of the Y signal occur at even multiples of the line frequency (15,625 Hz). Also, the amplitude of these spectral lines decreases with frequency.

The colour difference signals modulate a subcarrier of which the frequency, f_c, is an odd multiple of the line frequency divided by four, plus the picture refresh frequency (see Ref. 1):

$$f_c = 1135 \times (15{,}625/4) + 25 \quad \text{(Hz)}$$

This causes the spectral lines of the colour difference signal to be slotted in between those of the Y signal. The colour subcarrier frequency is set at 4.43361875 MHz, and the colour difference signals are quadrature-amplitude modulated (QAM). The B–Y and R–Y components modulate the amplitude of the colour subcarriers of 0 degrees and 90 degrees respectively (see Figs. 1d and 1e). The carrier itself is suppressed, so that it has an amplitude of nought in the absence of a colour difference signal. This is done to keep the picture free from interference caused by the otherwise continuously present subcarrier.

In order to eliminate the risk of phase shifts in the transmission path, the phase of the R–Y component is inverted every other picture line. Details of this operation peculiar to the PAL system may be found in Refs. 2 and 3.

The use of amplitude modulation with suppressed carrier requires a phase- and frequency-synchronized subcarrier at the receiver side. In a TV set, the modulated R–Y and B–Y components are recovered from the chrominance subcarrier with the aid of a 4.433-MHz quartz crystal oscillator whose phase and frequency are corrected every 64 µs by a 2-µs long burst signal slotted into the rear porch in the blanking period at the end of every picture. The burst consists of 8 to 11 cycles of the colour subcarrier frequency and follows the line sync pulse as shown in Fig. 2. A phase comparator is used to keep the crystal oscillator synchronized to the received burst, which also contains the PAL switch signal for the line-by-line R–Y phase reversal. This arrangement ensures that the R–Y signal in the receiver is inverted in synchronism with that at the transmitter side to ensure that the demodulation operation can work correctly.

Pitfalls...

In practice, the 'packaging' of the luminance and the chrominance information into a single CVBS (chrominance-video-blanking-synchronisation) signal is not without problems. Since the colour subcarrier falls in the spectrum of the luminance signal, it causes a finely patterned type of interference known as moiré. Luminance circuits in all modern TV sets are therefore fitted with a 'colour trap', which is a relatively simple filter that removes most of the moiré effects with the exception of those occurring at areas with sharp colour transitions. Here, large phase jumps give rise to subcarrier sidebands that fall outside the stop band of the 4.43-MHz colour trap. Unfortunately, Y signals in this stop band are also suppressed, which results in reduced picture resolution because some of the high-frequency components disappear. Incidentally, most monochrome sets also contain a colour trap to eliminate moiré.

The (possible) interference between chrominance and luminance also works the other way around: since the lumin-

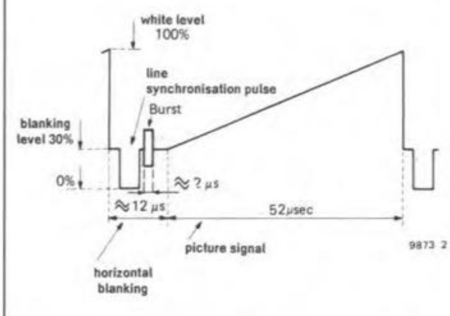


Fig. 2. Structure and timing of a composite video signal (PAL standard).

ance band includes the frequency range for the colour subcarrier, high-frequency Y signals can cause interference in the frequency range around 4.43 MHz. The result is a quasi-random type of patterning and colouring in and around picture areas of fine detail. Notorious examples of this happening can be seen virtually every evening in jackets, shirts or ties of people on television.

Standard VHS video recorders

Some 15 years ago, during the development of the VHS video system, a luminance bandwidth of 3 MHz was deemed satisfactory for VCRs considering the technical limitations imposed by the drum head speed and the tape consumption. In the original VHS system, the colour subcarrier is mixed down to 627 kHz to keep it well way from the lower end of the spectrum of the Y information, which is recorded as a frequency-modulated (FM) signal (see Fig. 3)

The FM recording improves the signal-to-noise ratio of the Y signal and makes it largely independent of amplitude variations of the tape signal. The frequency sweep ranges from 3.8 MHz to 4.8 MHz.

Returning to the colour information, this is recorded as an analogue signal in 'helical scan' mode (Ref. 3). The different frequencies used allow ready separation of the two signals. However, the bandwidth of the colour information is inevitably reduced to about 500 kHz. The result is noticed as 'smeared' colour transitions, to which the reduced (3-MHz) luminance bandwidth adds an impression that the picture is blurred.

These imperfections of the original VHS system were soon recognized by VCR manufacturers. Their answer, the HQ video recorder, was based on small improvements to the recording method and a better edge definition of the Y signal. The resultant picture quality improvement was marginal and not really a step forward. It was, however, the best that could be achieved given the need for continued compatibility. Clearly, real improvements to the picture quality offered by VCRs could be achieved only by changing some of the standards.

The Super-VHS system

The bandwidth of the recorded video signal was increased significantly (at the existing relative speed of 4.85 m/s between the tape and the head) by virtue of two technological developments. First, new metallurgic techniques allowed the size of the air gap of the video head to be reduced. Second, tapes with a very high magnetic particle density became available.

To maintain compatibility with older VHS recorders, the S-VHS system is based on the same method of colour recording (see Fig. 3). However, the frequency sweep of the Y signal is shifted up to a band from 5.4 MHz to 7.0 MHz to give a much higher noise margin. At the same time, the frequency of the FM subcarrier allows the luminance signal to be recorded at its full bandwidth of about 5 MHz.

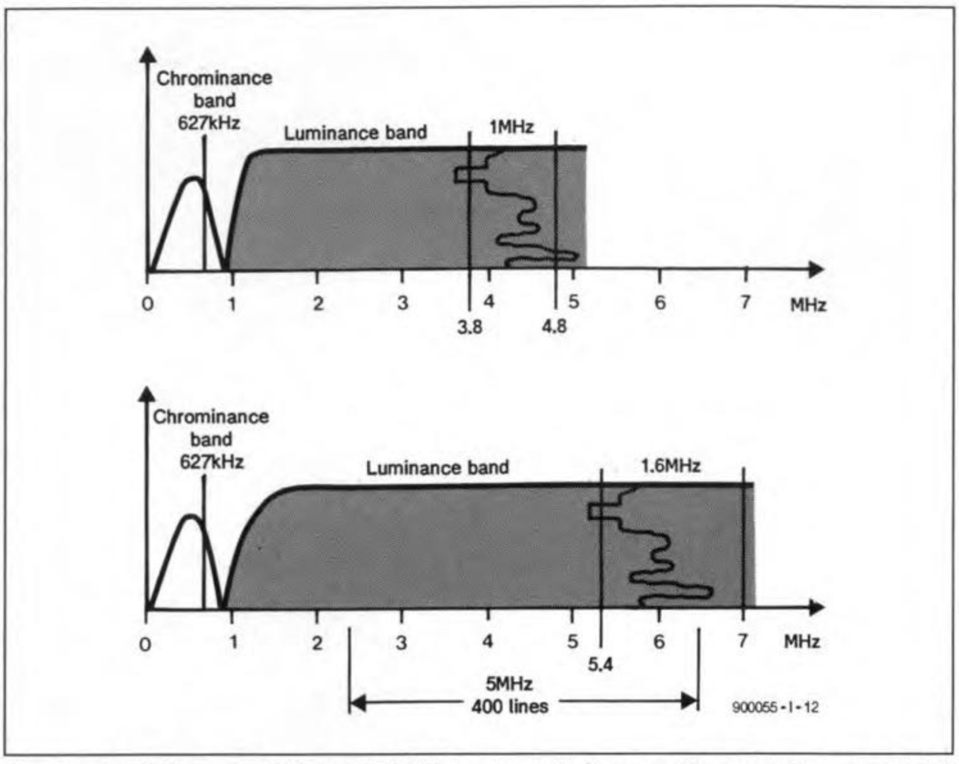


Fig. 3. Typical standard-VHS and S-VHS spectra. In both cases, the quadrature-modulated colour signal is recorded with the aid of a carrier which is mixed down to 627 kHz, while the luminance signal (Y) is recorded in FM. S-VHS recorders use a luminance carrier frequency of 5.4 MHz and a frequency sweep of 1.6 MHz. This offers a bandwidth of 5 MHz for the Y signal, as opposed to about 3 MHz for the standard-VHS video recorder.

In most standard VHS video recorders, the chrominance and luminance signals are processed separately until they are combined to give a CVBS signal with all the previously mentioned risks of running into trouble with interference.

By contrast, the S-VHS system is based on separate chrominance and luminance signals right up to the two associated outputs on the VCR. Evidently, this separation is not perfect when, for instance, a TV programme is recorded, since then the chrominance and luminance components must be extracted from the composite signal before they can be recorded, played back and fed separately to a monitor. The process of extracting the components has pitfalls as described before. Not so, however, with video sources that do supply the components separately. Examples include S-VHS cameras, some prerecorded S-VHS video tapes and MAC decoders.

Connection problems

So far, so good. A look at the rear panel of the TV set, however, reveals that there is at best a SCART connector, which does not allow luminance and chrominance signals to be taken in separately. The TV set is, therefore, not S-VHS compatible. This unfortunate discovery forces owners of S-VHS recorders to connect the monitor and the recorder via a CVBS link, forgoing most of the advantages of better picture reproduction offered by the new video system.

Considering the cost of an S-VHS compatible monitor, the only way to benefit from the separate chrominance and luminance signals supplied by S-VHS recorders and other video sources is to convert these to RGB signals that can be applied to the existing monitor or TV set via its SCART input. Next month's second instalment of this article discusses a circuit to accomplish this. In addition, the circuit provides a colour transition improvement (CTI) function, and is capable of converting CVBS to RGB.

From composite to RGB

Although most standard video recorders have a SCART socket, this rarely supplies RGB signals. Likewise, most set-top TV tuners and indoor units for satellite TV reception supply a CVBS (composite video) signal only. A problem arises when this equipment is to be connected to a high-resolution colour monitor with analogue RGB inputs, or a TV set with a SCART (Euro-AV) input. In both cases, the converter to be described next month can link this equipment and ensure optimum picture quality. ❑

References:

1. Chrominance-locked clock generator. *Elektor Electronics* July/August 1988.
2. Video line selector. *Elektor Electronics* April 1990.
3. *Video Handbook* (second edition). by R. van Wezel. Published by Heineman Newnes, ISBN 0 434 92189 0.

CORRECTIONS

400-watt laboratory power supply

October 1989 and November 1990

A number of constructors of this popular project have brought the following problems to our attention.

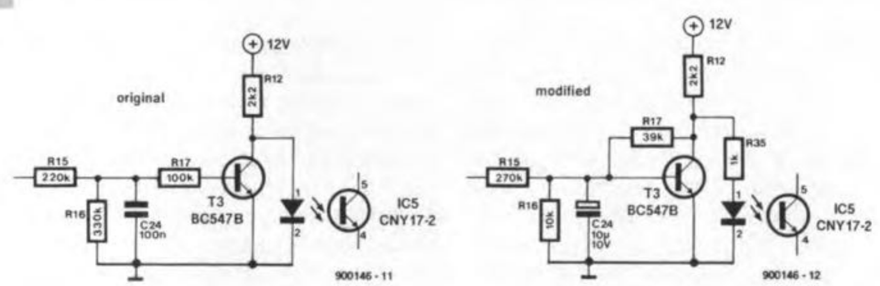
1. The onset point of the current limit circuit lies at about 3 A, which is too low. Solve this problem by replacing T1 with a Type BC517 darlington transistor, and R20 with a 82kΩ resistor.

2. Depending on the current transfer ratio of the optocoupler used, the transformer produces ticking noises. This effect, which is caused by overshoot in the pre-regulation circuit, may be traced with the aid of an oscilloscope monitoring the voltage across C26 at a moderate load current. The capacitor must be charged at each cycle of the mains frequency, and not once every five cycles. The problem is best solved by reducing the amplification of the regulation circuit. Replace R17 with a 39 kΩ resistor, and create feedback by fitting it between the base and the collector of T3. Also add a resistor in series with the optocoupler. These two changes are illustrated in Figs. 1 and 2. Lower R16 to 10 kΩ, increase C24 to 10 µF, and increase R15 to 270 kΩ.

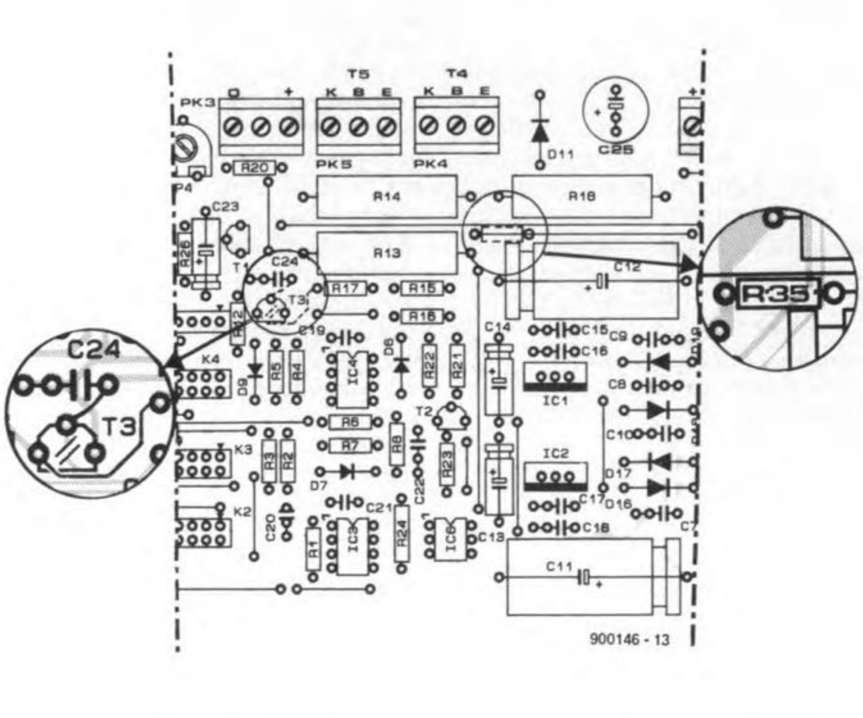
3. Excessive heating of the transformer is caused by a d.c. component in the primary winding. This is simple to remedy by fitting a capacitor of any value between 47 nF and 470 nF, and a voltage rating of 630 V, across the primary connections. This capacitor is conveniently mounted on to the PCB terminal block that connects the transformer to the mains.

4. One final point: when using LED DVMs for the voltage/current indication, their ground line must be connected to the **positive** terminal of C12.

1



2



Hard disk monitor

December 1989

In some cases, the circuit will not reset properly because the CLEAR input of IC3A is erroneously connected to ground. Cut the ground track to pin 3 of IC3, and use a short wire to connect pin 3 to pin 16 (+5 V).

Microprocessor-controlled telephone exchange

October 1990

In some cases, the timing of the signals applied to IC17 causes a latch-up in the circuit, so that the exchange does not detect the state of the connected telephones properly. Solve this problem by cutting the track to pin 1 of IC17, and connecting pin 1 to ground (a suitable point is the lower terminal of C6).

The text on the fitting of wires on the BASIC computer board (page 19, towards the bottom of the right-hand column) should be modified to read: 'Finally, connect pin 6 of K2 to pin 7 of IC3 ($\overline{Y7}$ signal).'

S-VHS/CVBS-to-RGB converter (2)

October 1990

The capacitor marked 'C37', next to R21 on the component overlay (Fig. 7b and ready-made printed circuit board), should be marked 'C39'.

In case they are difficult to obtain locally, inductors type 119-LN-A3753 (L1) and 119-LN-A5783 (L2) may be replaced with the respective types 119-ANA-5874HM and 119-ANA-5871HM, also from Toko, Inc. Suggested suppliers are Cirkit Distribution Ltd., and C-I Electronics.

EPROM simulator

December 1989

Counters IC3 and IC4 may not function properly owing to a too low supply voltage. This problem may be solved by replacing IC12 with a 7806. Alternatively, use BAT85 diodes in positions D1 and D2.

Programmer for the 8751

November 1990

The ready-programmed 8751 for this project is available at £35.25 (plus VAT) under order number **ESS 7061**, not under order number ESS 5951 as stated on the Readers Services pages in the November and December 1990 issues.

CORRECTIONS

6-metre band converter

April 1991, p. 38-43

The components list and the inductor overview in the top left hand corner of the circuit diagram should be corrected to read:

L_1, L_2 = 301KN0800.

Capacitor C_{16} (4.7 pF) must not be fitted on the board.

Finally, a few constructional tips:

- Fit a 10 nF ceramic decoupling capacitor at junction L_7-R_{36}.
- Fit a 18 kΩ resistor between the base of T_3 and ground. This reduces the Q factor of L_2, and prevents too high signal levels at the base of T_3.
- For improved tuning, inductor L_9 may be replaced by a Toko Type 113KN2K1026HM.

Multifunction measurement card for PCs

January and February 1991

We understand that the 79L08 (IC_{17}) is no longer manufactured and, therefore, difficult to obtain. Here, the IC may be replaced by a 7908, which, although physically larger than the 79L08, is pin-compatible, and should fit on the PCB.

Dimmer for halogen lights

April 1991, p. 54-58

In the circuit diagram of the transmitter, Fig. 2, pin 14 of the MV500 should be shown connected to pin 13, not to junction R1-R_2-C_2. The relevant printed-circuit board (Fig. 6) is all right.

RDS decoder

February 1991, p. 59

Line A0 between the 80C32 control board and the LC display is not used to reset the display, but to select between registers and data.

We understand that the SAF7579T and the associated 4.332 MHz quartz crystal are difficult to obtain through Philips Components distributors. These parts are available from C-I Electronics, P.O. Box 22089, 6360 AB Nuth, Holland. For prices and ordering information see C-I's advertisement on page 6 of the May 1991 issue.

S-VHS-to-RGB converter

October 1990, p. 35-40

Relays Re_1 and Re_2 must be types with a coil voltage of 5 V, not 12 V as indicated in the components list. Constructors who have already used 12-V relays may connect the coils in parallel rather than in series.

Suitable 5-V relays for this project are the 3573-1231.051 from Günther, and the V23100-V4305-C000 from Siemens.

The components list should me modified to read:

6	33nF	C57-C62

ISOLATING AMPLIFIERS

Until recently, isolating amplifiers were generally accessible to professional engineers only. These integrated circuits, widely used in laboratory and industrial measuring instruments, were simply too expensive for amateurs. Now, Burr Brown have available a series of isolating devices that, without any relaxation of specification, are available at prices that are affordable for most.

An isolating amplifier, as its name suggests, is a circuit between whose input and output no electrical connection exists (at least in theory). Normally, such a device consists of an input amplifier, a modulator, an isolating barrier, and a demodulator with a voltage follower at the output. The signal paths in the input and output sections are electrically fully isolated from each other. An important feature of an isolating amplifier is that it has a completely floating input, which helps eliminate cumbersome connections to source ground.

Fundamentals

The block diagram of a typical isolating amplifier is shown in Fig. 2. The input section may take one of many forms, from a complete instrumentation amplifier with programmable gain to a simple impedance converter that uses only one input pin.

The signal at the output of the input section is superimposed on to an HF carrier to enable it being transferred across the inductive or capacitive isolating barrier.

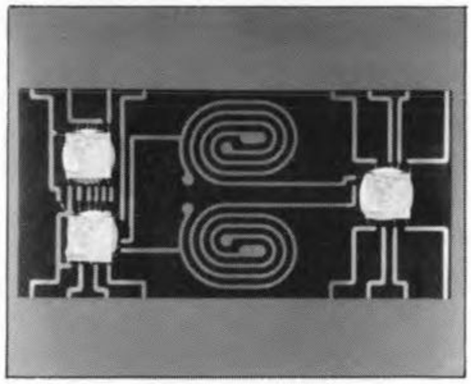


Fig. 1. Capacitive isolating barrier.

There are isolating amplifiers with an optical barrier. These devices do not need a modulator or (external) low-pass filters for reducing the modulation residue at the output.

Of the three types of isolating barrier already mentioned, inductive, capacitive and optical, the inductive one is the oldest and most widely used. Although it has not the bandwidth of the less expensive optical type, it has some important advantages: a very high isolating voltage (up to 8 kV) and great accuracy (see later).

The barrier of a typical capacitive type is shown in Fig. 1. The two spirals in the centre form the 1 pF coupling capacitors to which the modulated signal is applied in push-pull. Since the capacitors and the two signals are of equal value, the resultant sum signal is zero. This is vital as otherwise energy might be transferred from the input to the output, which would manifest itself as interference. This technique affords good bandwidth without detriment to the precision of operation or the maximum isolating voltage.

Theory of operation

The description that follows is based on Burr-Brown's Type ISO 122P isolating amplifier. This device uses an input and an output section that are galvanically isolated by matched 1 pF isolating capacitors built into the plastic package. The input is duty-factor modulated and transmitted digitally across the barrier.

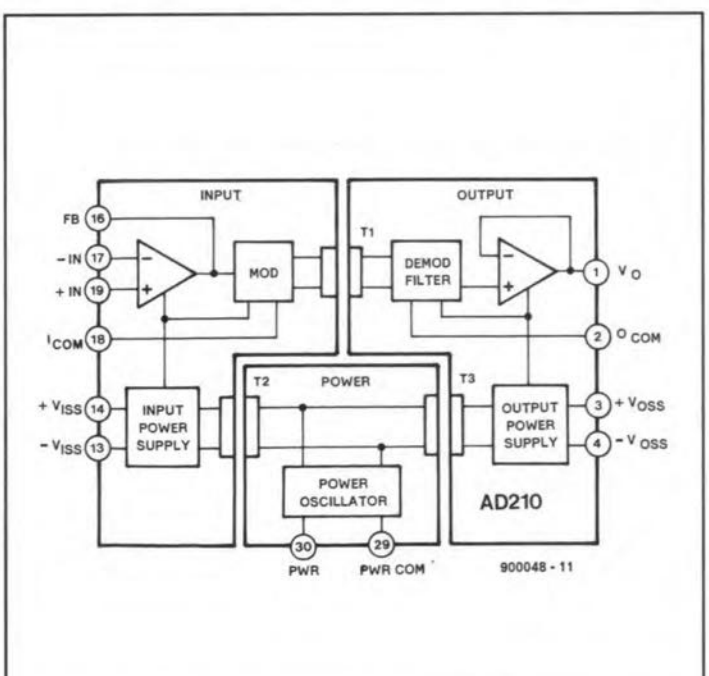


Fig. 2. Block diagram of a typical isolating amplifier.

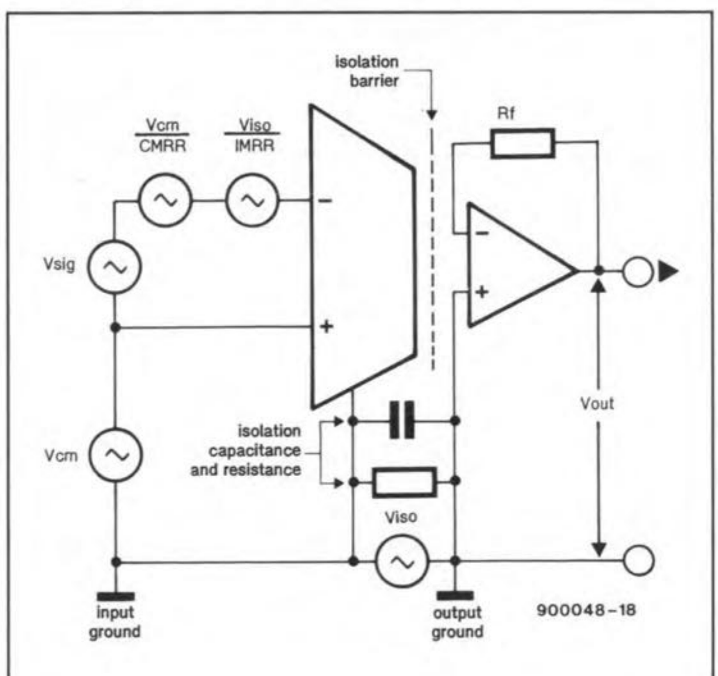


Fig. 3. The isolating voltage exists between the two earths.

The output section receives the modulated signal, converts it back to an analogue voltage and removes the ripple component inherent in the demodulated signal.

The input and output sections are laser trimmed for accurate circuit matching, after which they are mounted at opposite ends of the package with the isolating capacitors mounted between the sections.

Modulator. The input amplifier, A1 in Fig. 4, integrates the difference between the input current (V_{in}/200 kΩ) and a switched ±100 μA current source. This current source is implemented by a switchable 200 μA source and a fixed 100 μA current sink.

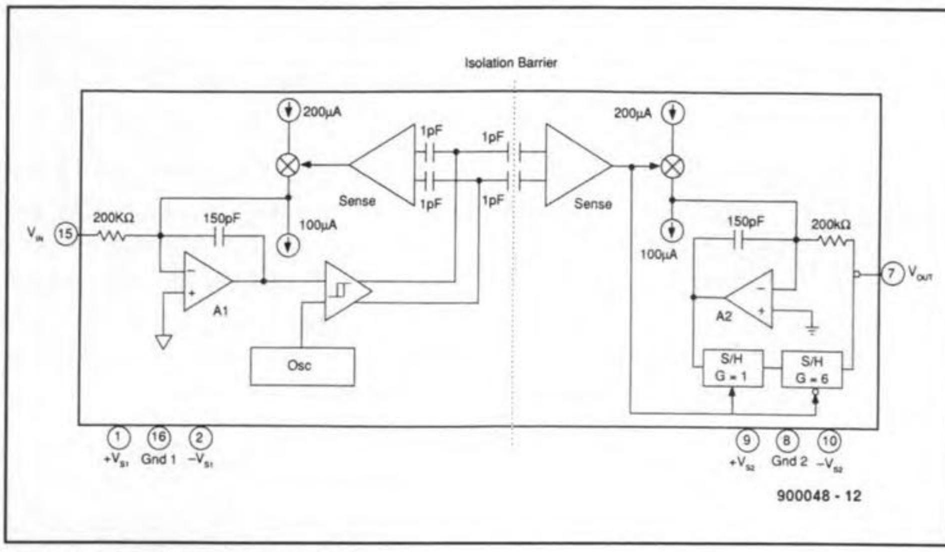


Fig. 4. Block diagram of the ISO 122P.

To understand the basic operation of the modulator, assume that V_{in} = 0.0 V. The integrator will ramp in one direction until the comparator threshold is exceeded. The comparator and sense amplifier will force the current source to switch; the resultant signal is a triangular waveform with a 50% duty factor. The internal oscillator forces the current source to switch at a frequency of 500 kHz. If V_{in} changes, the duty factor of the integrator will change to keep the average d.c. value at the output of A1 near zero volts.

Demodulator. The sense amplifier drives a switched current source into integrator A2. The output stage balances the duty-factor modulated current against the feedback current through the 200 kΩ feedback resistor, resulting in an average value at the V_{out} pin equal to V_{in}. The sample and hold amplifiers in the output feedback loop serve to remove undesired ripple voltages inherent in the demodulation process.

Signal and power connections. Each power supply pin should be bypassed with 1 μF tantalum capacitors located as close to the amplifier as possible. The frequency of the modulator/demodulator is set at 500 kHz by an internal oscillator. Therefore, if it is desired to minimize any feedthrough noise (beat frequencies) from a d.c./d.c. converter, use a pie filter on the supplies as shown in Fig. 5.

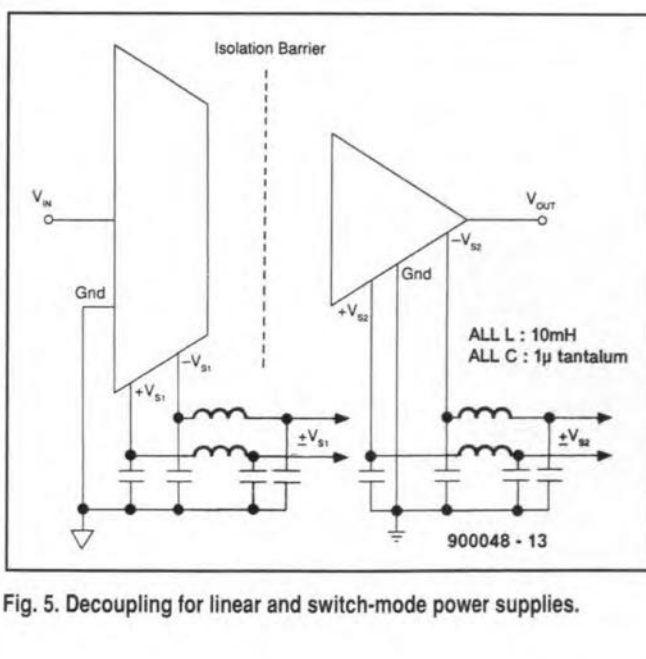


Fig. 5. Decoupling for linear and switch-mode power supplies.

Parameters

Although it is as easy to work with most isolating amplifiers as it is with opamps, there are a few parameters that need closer examination or that do not exist in opamps. Typical voltages in isolating amplifiers are shown in Fig. 3.

V_{sig} is, as in opamps, the differential input voltage; its level is ±10–15 V.

V_{CM} is the common-mode voltage, that is, the voltage between the signal inputs and the input earth. The maximum level at either input pin, with respect to earth, should not be higher than V_{CM}—in practice about ±10 V. If levels higher than that are needed, there can be no earth at the input, only at the output. It may also be impossible to use an earth. In that case, V_{ISO} becomes the reference.

V_{ISO} is the maximum isolating voltage between the reference earths of the input and output signal. Its level may be several kilovolts.

CMRR, the common-mode rejection ratio, shows the change in output voltage with respect to output earth for simultaneous changes in input voltages referred to the input earth, that is, V_{CM}.

IMRR, the isolation-mode rejection ratio, is $\Delta V_{ISO}/\Delta V_{out}$.

The foregoing parameters can now be used to express the amplication function:

$$V_{out}=\alpha(V_{sig}\pm V_{CM}/CMRR\pm V_{ISO}/IMRR)$$

where α is the amplification factor.

Another important parameter is the **accuracy**, which takes acoount of temperature stability, long-term stability, amplification error and non-linearity, which are well known from opamps. Peculiar to isolating amplifiers is the **leakage current**, which expresses the input error current as a function of the isolating voltage and frequency. In data sheets these are normally given as 240 V and 60 Hz respectively; the leakage current is expressed in μA.

WARNING. In medicine, isolating amplifiers are used primarily for ground loop elimination. Readers are warned not to use isolating amplifiers (for instance, for mains isolation) in equipment that is in frequent contact with their bodies.

Basic circuits

Isolating amplifiers, like all circuits that combine digital and analogue techniques, are particularly sensitive to external interference. Reference has already been made to the need of decoupling capacitors at the power supply pins.

Since the ISO 122P superimposes the signal on to a 500 kHz carrier, the transfer function for signals at frequencies up to 25 kHz may be considered linear. At higher frequencies, the output contains more residual modulation as may be seen from Fig. 6. A sinusoidal input at a level of 10 V and a frequency, *f*, of 2 kHz results in an undistorted output signal. The

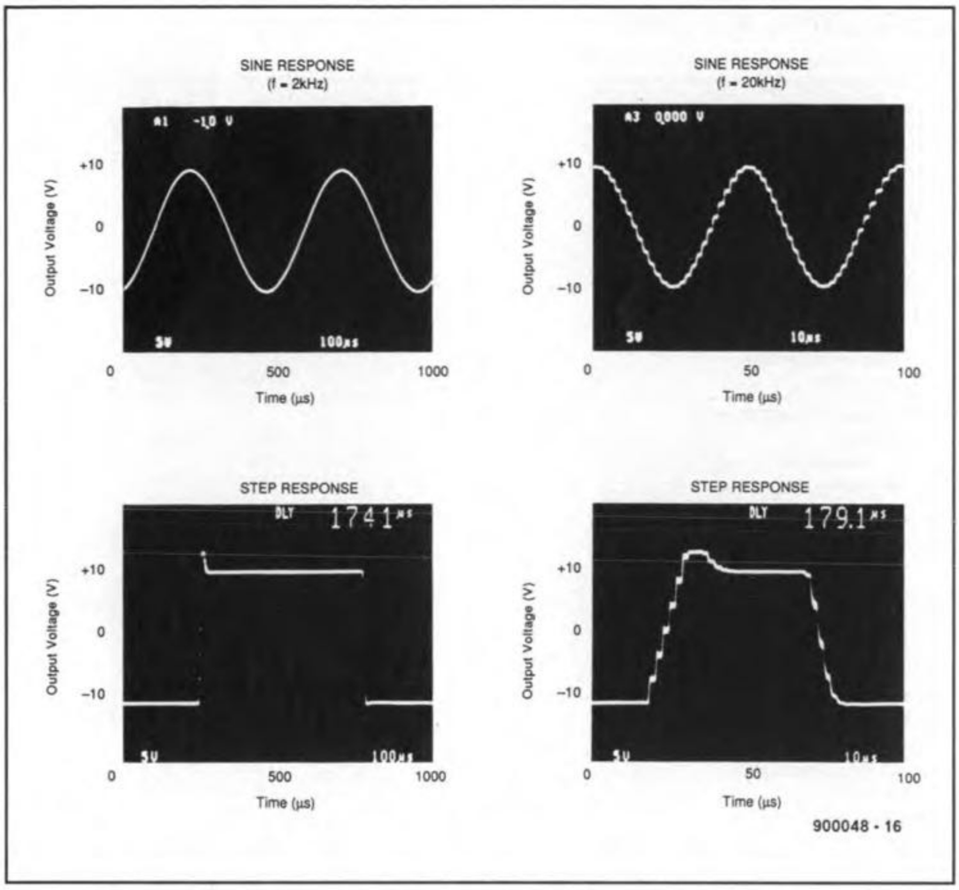


Fig. 6. The higher the input voltage, the greater the modulation residue and distortion.

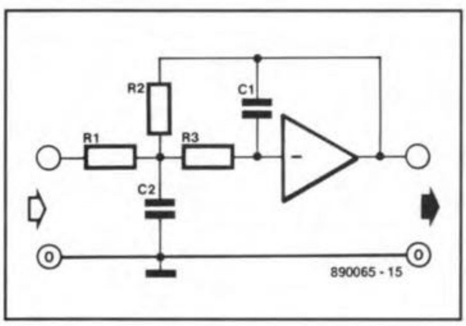


Fig. 7. Active low-pass filter for the suppression of modulation residue.

same is true for a rectangular signal even though the leading edge rising to +10 V is seen as a small overshoot.

When the frequency is increased by a factor of 10, the modulation residue on the waveforms is clearly visible. Furthermore, the edges of the rectangular signal have become less steep and the input rise time has increased appreciably.

Most of the modulation residue may be removed with the aid of an active low-pass filter of the first or second order as shown in Fig. 7. The cut-off frequency of that filter is 100 kHz, when R1=R2=13 kΩ; R3= =385 Ω; C1=100 pF; C2=4700 pF. For most applications, this is a good compromise between effective bandwidth and modulation suppression.

Error sources

Normally, the maximum signal frequency of a scanning circuit should be limited to half the scanning rate—at least, according to the relevant Nyquist or Shannon theorems. However, here there is a little leeway: up to 50 kHz, the specified bandwidth, the output signal is identical to the input signal as far as frequency and level are concerned. Over the range 50–250 kHz the amplification factor drops from 1 to 0.063 but there is no discernible increase in distortion.

At even higher frequencies, the isolating amplifier produces a relatively small noise signal at a frequency below 250 kHz in addition to the normal output signal. This behaviour may be explained with the aid of Fig. 8. The composition of the whole output signal may be considered in steps for which on the one hand the frequency behaviour and on the other the amplitude must be taken into account.

The triangular characteristics show the relation between input frequency and the interference frequency at the output (straight *y*-axis). When the input frequency lies between 250 kHz and 500 kHz, the interference frequency drops, rises again when the input frequency increases to 750 kHz, drops until the first harmonic (1 MHz) is reached, then rises again, and so on.

The amplification factor may also be evaluated from Fig. 8, and it is seen that at frequencies above that of the carrier, it

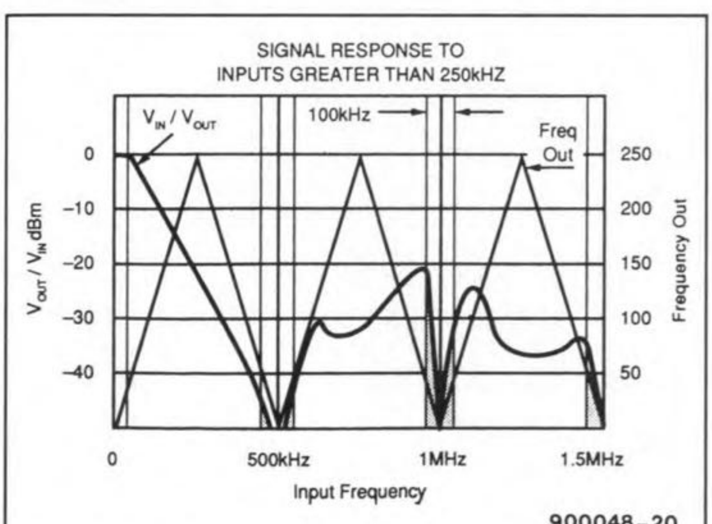


Fig. 8. For input signals above 250 kHz, operation of the isolating amplifier is no longer linear.

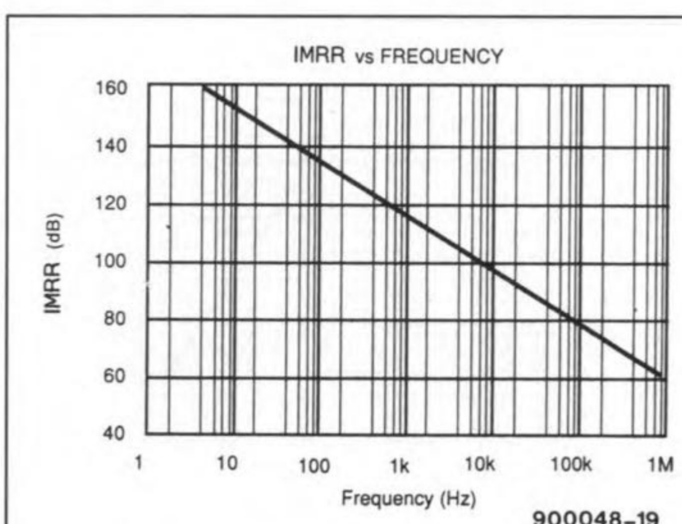


Fig. 9. To work with high isolating frequencies, the circuit must be designed for smaller maximum isolating peak voltages.

has a pronounced irregular behaviour. For example, a 10-volt, 800 kHz input signal results in an attenuated (–30 dB) 800 kHz output signal (this may be read from the ampliciation characteristic and the left-hand y-axis). At the same time, a 200 kHz interference signal is produced (which may be read from the triangular characteristic and the right-hand y-axis). At the 200 kHz point on the x-axis it will be seen that the interference signal is attenuated by a further 10 dB. Expressed in figures, this means that the output consists of a 800 kHz signal at a level of 316 mV and one of 200 kHz at a level of 100 mV. If the interfering signal is eliminated with the aid of a filter, the isolating amplifier can work with signal frequencies that are higher than the carrier frequency.

The same applies to the isolating voltage, to which the Nyquist theorem is, of course, equally applicable. To eliminate an interfering signal caused by the isolating voltage from the output, the IMMR as a function of frequency should be added to Fig. 8. For example, when V_{ISO}=1000 V at 800 kHz, the IMMR≈ –62 dB as may be seen from Fig. 9. Part of the output signal, viz. 794 mV at 800 kHz, was already seen to be an interfering signal. From Fig. 8 it is seen that the output additionally contains a 200 kHz signal at a level 30 dB below that of the first interfering signal, that is, 92 dB below 1000 V or 25 mV.

In this connection, Figures 10 and 11 should also be taken into account. The maximum permissible isolating voltage decreases with rising frequency. Furthermore, when V_{ISO} rises and the rise time exceeds 1000 V/μs, the triggering of the sense amplifier may go awry with the result that the condition of energy-less signal transfer is no longer met, and a com-

SPECIFICATIONS

At T_A = 25°C and V_S1 = V_S2 = ±15V unless otherwise noted.

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
ISOLATION					
Voltage Rated Continuous AC 60Hz		1500			VAC
100% Test 1	1s, 5pc PD	2400			VAC
Isolation Mode Rejection			140		dB
Barrier Impedance			10^{14} ‖ 2		Ω ‖ pF
Leakage Current at 60Hz	V_{ISO} = 240Vrms		0.18	0.5	μArms
GAIN	V_O = ±10V				
Nominal Gain			1		V/V
Gain Error			±.05	±.30	%FSR
Gain vs Temperature			±10		ppm/°C
Nonlinearity			±.008	±.015	%FSR
INPUT OFFSET VOLTAGE					
Initial Offset			±5	±50	mV
vs Temperature			±200		μV/°C
vs Supply			±2		mV/V
Noise			4		μV/√Hz
INPUT					
Voltage Range		±10			V
Resistance			200		kΩ
OUTPUT					
Voltage Range		±10	±12		V
Current Drive		±5	±15		mA
Capacitive Load Drive			1000		pF
Ripple Voltage(2)			10		mVp-p
FREQUENCY RESPONSE					
Small Signal Bandwidth			50		kHz
Slew Rate			1.5		V/μs
Settling Time	V_O = ±10V				
0.1%			50		μs
0.01%			150		μs
Overload Recover Time			150		μs
POWER SUPPLIES					
Rated Voltage			15		V
Voltage Range		±4.5		±18	V
Quiescent Current: V_{S1}			±4.5	±6.5	mA
V_{S2}			±4.5	±6.5	mA
TEMPERATURE RANGE					
Specification		0		70	°C
Operating		–25		85	°C
Storage		–25		85	°C
θ_{JA}			100		°C/W

NOTES: (1) Tested at 1.4 X rated, fail on 5pC partial discharge leakage current on five successive pulses. (2) Ripple frequency is at carrier frequency (500kHz).

Table 1. Technical specification for the ISO 122P at T_A = 25 °C; V_{S1} = V_{S2} =±15 V and R_L = 2 kΩ.

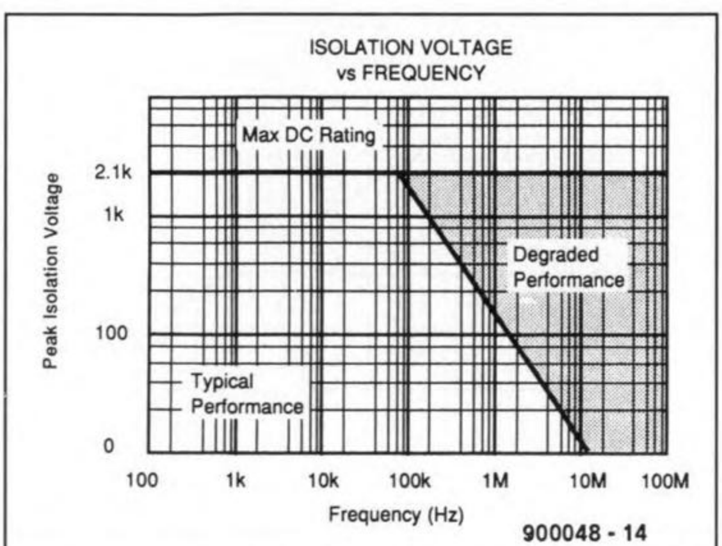


Fig. 10. Isolating voltages at high frequencies cause degrading of the output signal.

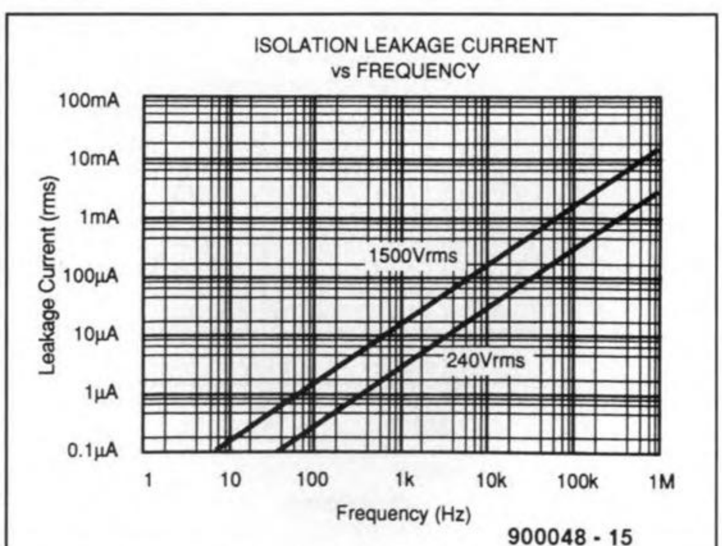


Fig. 11. At high frequency isolating voltages there is no longer en energy-free transfer across the barrier.

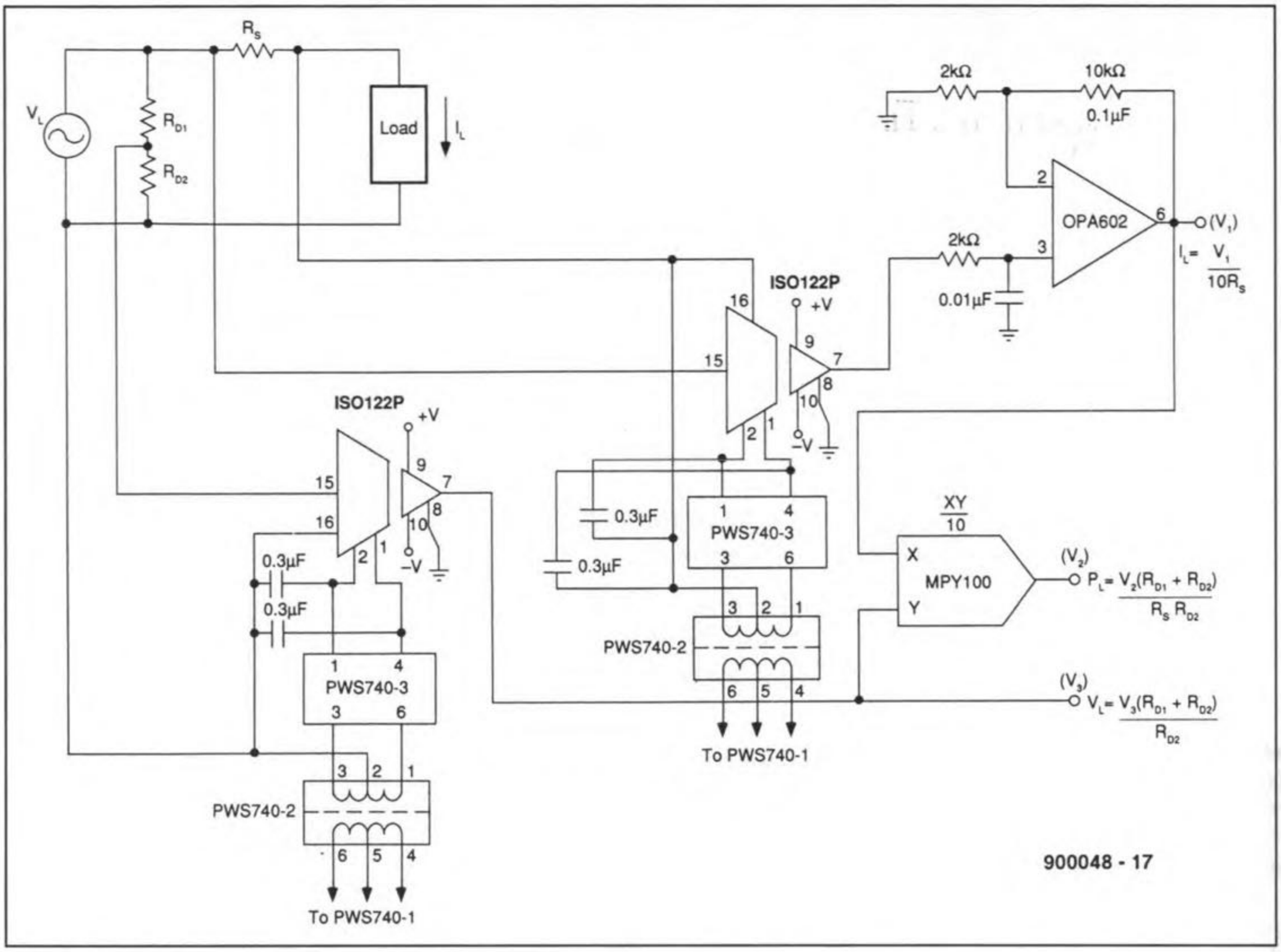


Fig. 12. Typical application of an isolation amplifier: a measuring instrument with current, voltage and power indication.

mon mode current flows across the barrier. It should be noted that supply voltages below ±15 V reduce the maximum permissible slew rate by about half.

A typical application: a power measuring instrument

Finally, as an example of what kinds of application may be satisfied with a typical isolating amplifier such as the ISO 122P (although other types may also be used, of course), we have chosen a power measuring instrument, whose circuit is shown in Fig. 12. The instrument can indicate the load current, the source voltage and the resulting power dissipated in the load.

One of the isolating amplifiers is used to evaluate the source voltage with the aid of potential divider R_{D1}-R_{D2}, and the other to measure the load current with the aid of current sensor R_S. Both amplifiers are connected in an identical manner.

The input earth that serves as the reference for the input potential at pin 15 is not grounded. Only the output sections are referred to ground potential.

The Type PWS740-1, PWS740-2 and PWS740-3 devices are Distributed Multichannel Isolated DC-to-DC Converters from Burr-Brown. These converters are able to produce up to eight ±7–20 V supply voltages, which are galvanically isolated from one another, from a single 7–20 V direct voltage. Currents of up to 60 mA may be drawn from each of the resulting supplies.

Although this is not a cheap way of producing power supplies, it guarantees that no interference will be transferred from the mains to the isolating amplifiers.

The MPY100 is a four-quadrant multiplier-divider that, apart from multiplication, performs analogue square-root and division without the bother of external amplifiers or potentiometers. Here it is used to compute the power, $P = UI / 10$.

Since the multiplier always divides by 10, the output of the voltage-indicating amplifier may be used directly as the multiplicand, whereas that of the current-indicating amplifier must first be amplified by 10 before it can be so used..

The Type OPA602 high-speed precision operational amplifier, also from Burr-Brown, is used as an active low-pass filter and impedance converter. Other types of high-speed precision operational amplifier may, of course, also be used. ■

Further information on all devices discussed in this article may be obtained from

Burr-Brown Worldwide Headquarters
P O Box 11400
TUCSON, AZ 87734
USA
Telephone (602) 746 1111
Fax (602) 889 1510
or
Burr-Brown International Ltd
1 Millfield House
Woodshots Meadow
WATFORD WD1 8YX
England
Telephone (0923) 33837
Fax (0923) 33979.

IN QUEST OF A PANGRAM – PART 2

g00103-II

by Lee C.F. Sallows

Combinatorial explosion

At long last the program was finished and started. Roughly a million combinations had already been tested during the development period. The trouble with previous versions had been their hopelessly slow speed. Even the latest program could only test something like ten new combinations per second. This was still sluggish, but bearing in mind the hefty letter crunching involved (16×16 additions in calculating the SUMPROFILE alone, for example), I thought it probably couldn't be greatly improved upon. Vaguely I wondered how long it would take before a solution popped up. Being a greedy consumer of valuable processor-time, the program ran at nights as a low-priority 'batch-job' on the Computing Centre's VAX 11/780 machine. Every morning I would hasten to call up the job file, running my eye swiftly down the screen in search of 'EUREKA', which would precede a printed record of the magic combination of number-words. As day succeeded day without result, the question of how long it would be before all possibilities had been exhausted gradually assumed importance. It was a matter I had never given any serious attention. 10^7 cases had already been examined. Let's see, how many would there be altogether ...?

The calculation is an absurdly simple one and even now I blush to recall first seeing what the result implied. Programmatically, the ten totals in each of the sixteen ranges are cycled exactly like the 0–9 digits on the rotating number disks of the familiar tape-counter or odometer. Advancing this software counter a single step results in the next combination of totals being clicked into position, ready for the pangram test. The all-zero state will correspond to the first or lowest set of number-words: the bottom row of scale numbers in Fig. 2. Just as the mechanical counter begins at 0 and steps in turn through every number (that is, every possible digit sequence) up to the highest, so the program runs through all possible combinations up to that coinciding with the top row in Fig. 2. In effect, we are systematically examining every single histogram that can be plotted. About halfway through the process, the example shown for the near-solution totals will come up for testing. How many such graphs can be drawn in Fig. 2? The answer is clearly the same as that number displayed on our sixteen-digit odometer after stepping through all possible positions: a string of sixteen 9s (plus one for the zero position) = 10^{16}. Is there a golden vein running through the ten-deep strata? A milky nipple crowning the Gaussian breast? At a speed of ten combinations per second, to find out is going to take $10^{16}/10$ seconds. A pocket calculator soon converts this to more intelligible units.

In searching for an autogram, my computer program could only test something like ten new combinations per second. How long would it be before all possibilities were exhausted?

There seemed to be something wrong with the one I was using. Every time I worked it out, the answer was ridiculous: *31.7 million years!*

I was so unprepared for the blow contained in this revelation that initially I could hardly take it in. The whole object of turning to a computer in the first place had been to canvass huge numbers of combinations fast. Now that the truth had dawned, I began cursing my naivety in ever embarking on such a fool's errand. True, I was an electronics engineer, not a professional programmer. However, the more I contemplated the kinds of speed at which a realistic program would have to run, the more preposterous the whole computer venture appeared. Conceivably, a somewhat faster program could be written. But even checking at a rate of one million combinations per second, it would take three hundred and seventeen years to run through the ten-deep range of possibilities!

Yet, thoughts of millions of combinations per second put me in mind of *megahertz*. And megahertz brought my thoughts back to Electronics. This in turn prompted an idea, a fanciful notion, for the first few days no more than an idle phrase repeated in the head, a good title perhaps for a science-fiction story: *The Pangram Machine*.

Initially, I didn't take the thought seriously. I was disconsolate after the embarrassing failure of the computer project, and the absurd expression 'pangram machine' mocked hollowly at the back of consciousness. Yet suddenly the vague intuition began to crystallize; in a flash I saw how a central process in the program could be simulated electronically. Taking this mechanism as a starting point, I tried translating other aspects of the algorithm into hardware. It worked; it was easy. A few hours later, I was amazed and thrilled to find the broad outlines of an actual design already clear in my mind.

The Phoenix now emerging from the ashes of the Pangram Quest soared serenely to the sky, smoothly circled, swiftly swooped, and soon bore me off, a helpless prisoner in its relentless talons. For the next three months I would be pouring all my energy into the development and construction of a high-speed electronic Pangram Machine.

The Pangram Machine

How seriously should a word puzzle be taken? Though only the size of a smallish suitcase, the apparatus to emerge from three months' intense activity packed more than two thousand components on to thirteen specially designed printed-circuit cards. More than a hundred of these were integrated circuits, each containing on the average something like fifty transistors. Foresight of this complexity might have dissuaded me from starting. In the event, the completed machine turned out to involve a good deal more electronics than originally planned.

At the heart of the device is the electronic equivalent of a continuously-stepped sixteen-digit odometer: a clock-driven cascade of sixteen Johnson counters: see Fig. 3 for all that follows. The clock is a simple 1 MHz square-wave generator producing a continuous train of 10^6 pulses every second. As mentioned above, however, even checking at this rate, ten-deep ranges would take 317 years to explore. A reduction was therefore demanded, the choice of new range-length being determined primarily by the availability of standard 8-output devices. Each counter is thus a circuit with eight outputs that become actuated consecutively by successive pulses presented to its single input. Before the clock is started, a RESET button on the control panel (see photograph in Part 1 on page 57) enables all counters to be initialized or 'zeroed', meaning that all '0' out-

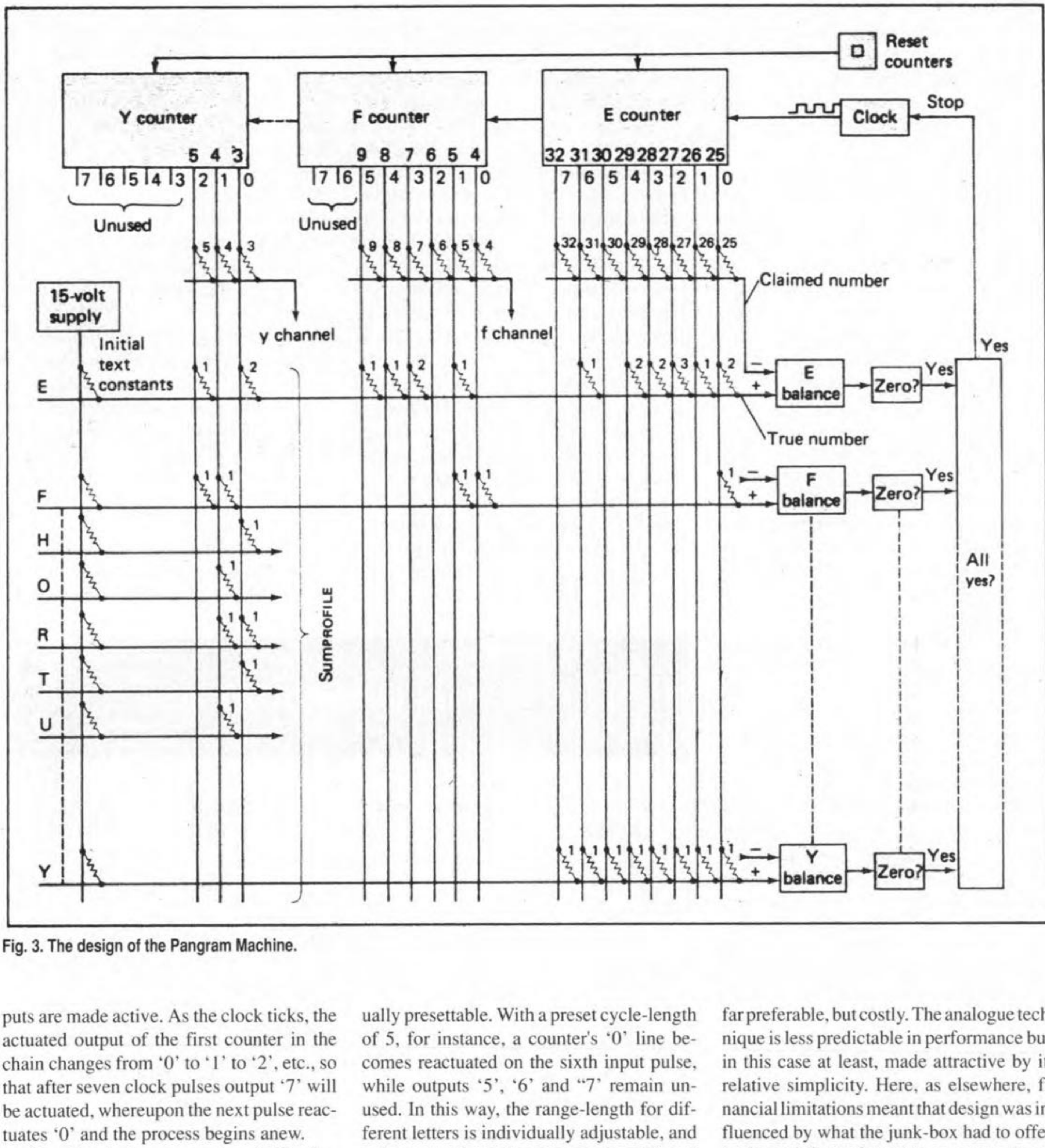


Fig. 3. The design of the Pangram Machine.

puts are made active. As the clock ticks, the actuated output of the first counter in the chain changes from '0' to '1' to '2', etc., so that after seven clock pulses output '7' will be actuated, whereupon the next pulse reactuates '0' and the process begins anew.

Coupling between counters is like between odometer disks in that, after completing one cycle, it is arranged for a single pulse to be sent to the input of the following counter in the cascade. Eight cycles of the first are thus needed to step the second counter through one. In this way, every new clock pulse results in actuating a unique combination of sixteen output lines. After 8^{16} pulses, all combinations will have been run through and, unless halted, the entire process will begin again.

Even so, calculation shows that running time must still be measured in *years* unless some further limitations are introduced. In fact, the cycle-length of counters is individually presettable. With a preset cycle-length of 5, for instance, a counter's '0' line becomes reactuated on the sixth input pulse, while outputs '5', '6' and "7' remain unused. In this way, the range-length for different letters is individually adjustable, and a shorter total running time can be achieved (at the price of narrower ranges). Figure 3 shows that the *y*-counter's cycle-length has been reduced to 3, for example. Later we shall turn our attention to the actual set of ranges used.

Now, just as in the computer program, the object of actuating different combinations of output lines is to call up sets of PROFILES whose corresponding elements will be added together so as to form a SUMPROFILE (as discussed above; I leave the initial text constants temporarily out of account). Electronically, the instantiation and addition of PROFILES may be achieved by the use of digital or analogue techniques. The former is far preferable, but costly. The analogue technique is less predictable in performance but, in this case at least, made attractive by its relative simplicity. Here, as elsewhere, financial limitations meant that design was influenced by what the junk-box had to offer. In the end, I was forced to use an analogue approach but, since other parts of the circuitry are digital (the counters, for example), the overall design is really a hybrid.

Accordingly, the PROFILES 'called up' by actuated counter outputs take the form of resistor fan-outs feeding specific patterns or profiles of discrete current levels into sixteen common lines representing the SUMPROFILE. Every counter output is associated with a predetermined number-word (shown in counter-boxes). An actuated output is one transistor connected to a 15-volt supply and thus able to deliver current; non-actuated outputs are simply left unconnected (these are so-called open-collector outputs). The

PROFILE of each number-word is implemented as a set of resistors connecting the counter output to appropriate SUMPROFILE lines. These are the horizontal lines *E*, *F*, ...*Y* shown in the diagram. (Sixteen 0.5 Ω resistors, not shown but electrically important, connect each of these to ground or 0 V).

Current drawn from actuated outputs thus divides into a number of resistor-adjusted streams and is distributed over the *E*, *F*, ... *Y* lines of the SUMPROFILE so as to represent the contribution of each PROFILE-number. PROFILE summing is thereby achieved almost without doing anything: the current produced in each SUMPROFILE line (and hence the voltage across its 0.5 Ω resistor) is simply the aggregate of the subcurrents injected into it via the resistors in the presently actuated set of PROFILES.

The number and value of the resistors used in each case depend entirely on the PROFILE being simulated. Choosing an arbitrary unit of current to represent one letter, double this value will stand for two, and so on. In fact, with the exception of *seventeen* which alone contains four *e*'s, values in the PROFILES are always 0, 1, 2 or 3. Since 0 is indicated by no current = no connection, all PROFILES (excepting that for *seventeen*) can be implemented by resistor sets built up from just three discrete values of resistance: x Ω, $x/2$ Ω, and $x/3$ Ω, yielding current levels of 1, 2, and 3 units, respectively. (In reality, $x = 3920$ Ω, a high value relative to the 0.5 Ω resistor over which the sum voltage falls: this is important for achieving summing linearity). A concrete example is shown for the *y*-counter's *three* and *four*. The small diagonal zigzags are the resistors. The numbers printed alongside represent not their resistance but the number of current units (15 V/3920 Ω = 3.82 mA) they pass into the SUM PROFILE line: *three* = 2 *e*'s, 1 *h*, 1 *r*,1 *t*; *four* = 1 *f*, 1 *o*, 1 *r*, 1 *u*.

So far, so good: the current entering each + input of the boxes marked BALANCE is a measure of the number of *e*'s, *f*s, etc. actually occurring in the present set of sixteen actuated number-words; every microsecond a new set is switched in . But the SUMPROFILE is incomplete without the initial text constants—themselves comprising no more than a special PROFILE and thus representable as a set of fixed-bias currents. Hence, a further array of sixteen resistors is connected permanently from the 15 V supply to each SUMPROFILE (see Fig. 3).

Now, in the program SUMPROFILE, totals (representing true letter frequencies) are compared with the labels of the PROFILES (the numbers corresponding to the number-words) to check for pangramhood. These label-numbers are simulated by an extra resistor-determined current derived from each counter output (top row of resistors). *E*-label currents are fed to the – input of the *E* BALANCE box, and so on. Comparison of SUMPROFILE and label currents takes place in the BALANCE boxes; each box is a differential amplifier whose output voltage is a fixed multiple (the amplification factor) of the difference between its two input currents (or voltages, depending on how you look at it). In this way, SUMPROFILE and label-numbers are weighed against each other in the BALANCE; only if they are equal will the output voltage be zero, or close to zero, volts. Of course, all sixteen pairs are weighed simultaneously.

The rest ought to be obvious. The 'Zero?' boxes are window detectors: circuits signalling a logic 1 ('yes') if their input voltage lies within a predetermined voltage range or 'window'. The window in this case is a narrow one (±50 mV) centred on 0 V. All window-detector outputs go to a sixteen-input AND gate. If sixteen zeros turn up together, the AND gate will fire, thereby stopping the clock, freezing the counters, and turning on an inessential but comforting EUREKA! lamp mounted on the control panel. The magic set of number-words sought will now be represented by the frozen combination of actuated outputs. In order to signal which these are, counter positions are indicated (in binary code) in the form of sixteen groups of three LEDs visibile through a red perspex front panel. When a table is used to translate LED patterns into number-words, it will remain only to double-check the result by hand and, if it is correct, ring for the champagne.

Though all very well on paper, in reality the analogue techniques used in the machine are messy. Circuit capacitance and amplifier settling times set a practical limit to the speed of operation. When the clock ticks and switches in a new set of PROFILES, electronic havoc breaks loose as overshoots, oscillations, glitches, and gremlins conspire to drive window-detectors into palsied indecision. After a while, electrons begin to simmer down and circuits settle into a new steady state. For this reason, rather than

Ranges of Number-Words

LETTER	NEAR-SOLUTION TOTAL	RANGE	RANGE LENGTH	INITIAL CONSTANT
E	27	25–32	8	7
F	6	4–9	6	2
G	3	2–7	6	2
H	5	3–8	6	2
I	11	8–14	7	4
L	2	1–4	4	1
N	20	17–23	7	10
O	14	12–17	6	11
R	6	3–8	6	2
S	28	24–30	7	21
T	21	18–24	7	7
U	3	1–6	6	1
V	6	3–8	6	2
W	10	7–13	7	5
X	4	1–6	6	1
Y	5	3–5	3	1

Fig. 4. Ranges of values of number-words as actually built into the Pangram Machine.

going straight to the STOP input of the clock as shown in Fig. 3, the AND gate output is actually sampled some 900 nanoseconds after clock pulse onset—that is, at the last moment of the clock cycle, only 100 ns before the next pulse arrives. This idea, among others, was due to Willie van Schaijk, without whose friendly and expert assistance the machine might never have left the ground. Using the (TTL) technology at my disposal, a clock frequency of 1 MHz was the highest I was able to achieve in these circumstances. Given more funds, it would probably not be difficult to improve on this by a factor of 10. Digital techniques bring their own problems: I am not convinced that a worthwhile gain in speed could be won for the large investment needed.

Although all 16 counters have eight outputs each, it is impossible to exploit these unrestrictedly, since to examine all possible combinations at a clock rate of 1 MHz would still take $8^{16}/10^6$ seconds = 8.9 years. Range lengths were therefore tailored to each letter so as to retain a reasonable chance of finding the pangram while bringing the running time down to about one month. Flexibility was maintained by providing printed-circuit cards with easily altered solder links allowing preadjustment of each counter's cycle length. Selection of the ranges to be used was a ticklish business, involving careful analysis of letter frequencies in number-words. Those finally settled upon may be seen in Fig. 4 (numbers under RANGE stand for number-words).

Notice that *e*, having a high frequency and being therefore less predictable than other letters, receives the maximum range length of 8. On the other hand, *y*, occurring exactly once in every number-word from *twenty* upwards but in no others, can appear only 3, 4, or 5 times in the pangram given the ranges for *e*, *n*, *s*, and *t*. This is hardly a trivial insight: were *y*'s range length increased to 4, ten days would be added to the running time. As it is, to run through the combinations generated by the ranges in Fig. 4 will take $[8\times6\times6\times6\times7\times4\times7\times6\times6\times7\times\times7\times6\times6\times7\times6\times3]/10^6$ seconds = 31.36 days. Anything longer would have been unendurable.

In the program, the PROFILE for *one* contained –1 in the *s*-position to cancel what would otherwise be an *s* too many in the initial *s*-constant. However, minus values are not resistor-representable in the machine. As may be seen in Fig. 4, there are only three letters (*l*, *u*, *x*) in whose ranges *one* occurs. To deal with these cases, after reducing the initial *s*-constant by 3, an *s* is added to the PROFILES of number-words higher than *one* in their ranges. The range for *l* thus becomes: *one*, *two* + *s*, *three* + *s*, *four* + *s*; in other words, number-words above *one* bring their plural *s* with them. There is no reason that this couldn't be done for every number-word in every range (with corresponding reduction in the *s*-constant), but it would mean a lot of extra resistors.

Failure

After twelve weeks' concentrated effort, the machine drew near to completion. As a prototype, it had posed a host of technical problems to be faced and overcome. First, there had been a pilot phase to investigate the feasibility of an analogue implementation.

After twelve weeks' concentrated effort, the world's first pangram machine drew near to completion, and was ready for testing

How fast could the critical summing and balance circuitry perform? Despite normal pessimistic expectations, small-scale trials yielded promising results. The only way to discover whether the full-scale version would function satisfactorily would be to build it. At length the long program of design and construction culminated on the day the machine stood ready for a crucial test: would it successfully identify and halt at a magic combination?

To find out, I introduced deliberate changes in the resistor-represented initial text constants; by feeding the machine with false data about letter frequencies in the introductory text, I could trick it into halting at a prearranged pseudo-magic combination. Subtracting *o* and adding an *i* and *n* should cause it to stop at that combination of real totals represented in the previously discussed hand-produced solution:'twenty-one', the true number of *t*'s, then replacing 'twenty-nine'. Using the 'manual clock' and 'select counter' controls to preadvance the five highest or 'most significant' counters in the odometer chain (*u*, *v*, *w*, *x*, *y*) to their appropriate totals (3, 6, 10, 4, 5), it would take only a few minutes for the faster-cycling counters to reach the remaining numbers in the magic combination. Starting the clock, I watched anxiously as the changing pattern of binary-coded LED displays reported the steady increment of counter positions.

Suddenly and soundlessly the counters locked, the EUREKA! lamp came on, and the correlation monitor confirmed 16 hits in a row. This was it; the machine had passed the acid test. With the correct text constants loaded and a few other loose ends tied up, one week later all was ready for the launching of this singular rocket on its 32-day voyage into the unexplored regions of logological space.

Lift-off came almost eight months after the publication of Rudy Kousbroek's audacious challenge. Cees Wegman, a spiritual godfather to the project who had watched sympathetically through the long months as I gracelessly declined from suave insouciance to crazed intensity, came along to perform the deed of honour. A bottle of wine was broached, and three of us sat with glasses raised as he ceremoniously clicked the starting switch to RUN (it was a fitting tableau for some quixotic latter-day Velasquez, I couldn't help musing).

The ensuing period found me hovering nervously over the machine. Among other things, there was the nagging worry of machine reliability; what guarantee was there of faultless operation over so long a period? The answer, of course, was *none*. All I could do was maintain sporadic surveillance with an oscilloscope, and halt the machine at three-day intervals to perform checks with the pseudo-magic combination. After a while the suspense became nerve-racking. Mornings were worst. On waking, the first thought in consciousness would be *has it halted?* It took nerves of iron to go patiently through the morning's ablutions before tensely descending to the living room where the machine was installed on my desk. Opening the door with great deliberation, I would quickly go in and transfix the machine with a questioning gaze. And there would be the flickering LEDs as the counters slowly switched their way through the 2.71×10^{12} combinations. One million a second for 31.366 days. It was a torturing experience. The novelty of watching the machine soon wore off and the edge of expectation blunted, but a single second's distracted attention was accompanied by the thought that another million chances had already elapsed, so perhaps NOW??? ... and my glance would be wrenched back to the twinkling array of lights. After months of frenzied activity in building the machine, this period of enfored waiting was a cruel contrast of frustrated inertia and protracted disappointment.

But it was highly conducive to thinking up means for shortening that time. Before long, I saw that by halting the machine at key points in its travel and limiting the cycle-length of certain counters through calculated intervals, redundant checks on predictably invalid blocks of combinations could be obviated. Temporarily truncating the t-counter's range to exclude *eighteen* and *nineteen*, for instance, meant that all values of t contained a y so that y could occur only four or five times. Testing cases for which $y = three$ could thus be skipped during such a phase. Using dodges of this kind, I was able to slice nearly ten days off the originally estimated running time.

Meanwhile, the grains of sand—and of hope—were running out inexorably. Day succeeded day and week succeeded week with no sign of an EUREKA! Twenty-two days after launching, the machine had checked out every (undisqualified) combination of number-words within its capacity without finding the magic pangram. Since oscilloscope monitoring and a subsequent test with the modified initial text constants showed the machine to be functioning properly, I was not in any serious doubt about this negative result.

The crushing truth was that there never had been a needle in the haystack; the Quest for the Pangram had failed.

The third part of this article, reproduced by kind permission of Springer-Verlag, Heidelberg and New York, will appear in the October issue of *Elektor Electronics*.