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THE ELECTRONICS & COMPUTER MAGAZINE

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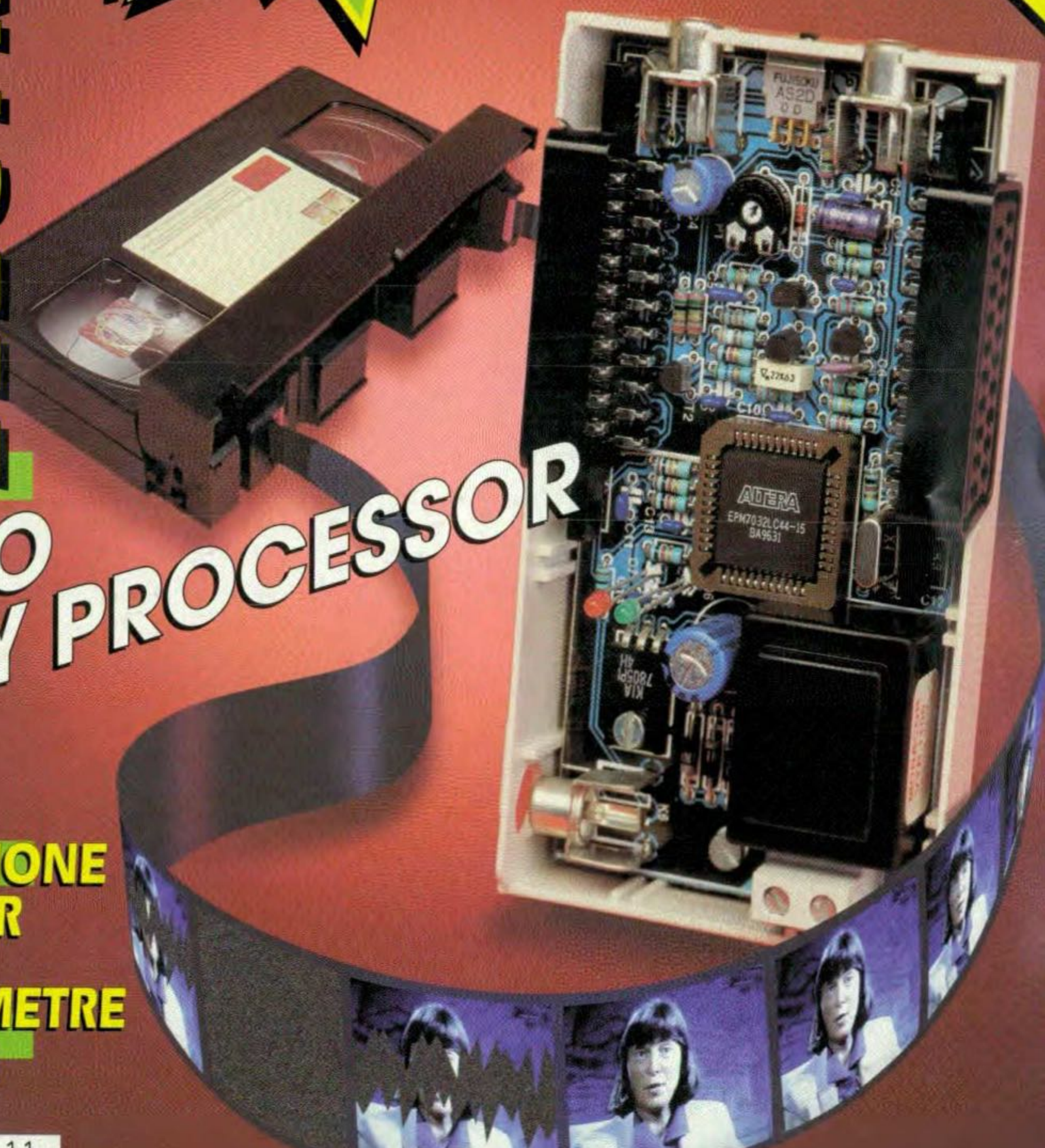

VIDEO COPY PROCESSOR

STEREO MICROPHONE AMPLIFIER

MINI 80-METRE RECEIVER

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ELEKTOR ELECTRONICS

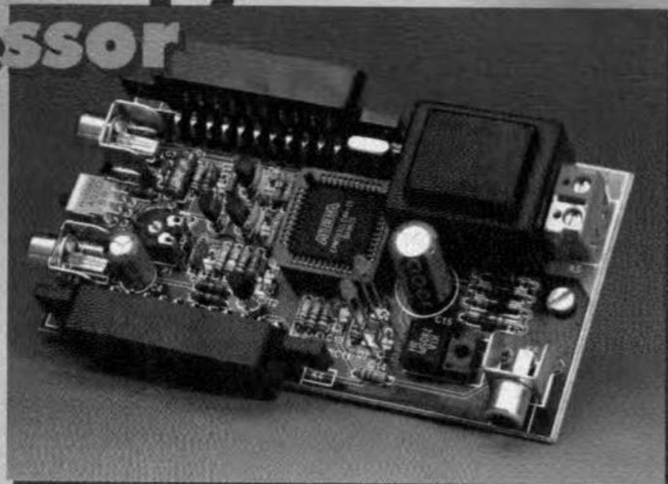
CONTENTS

November 1997 Volume 23
Number 260 ISSN 0268/4519

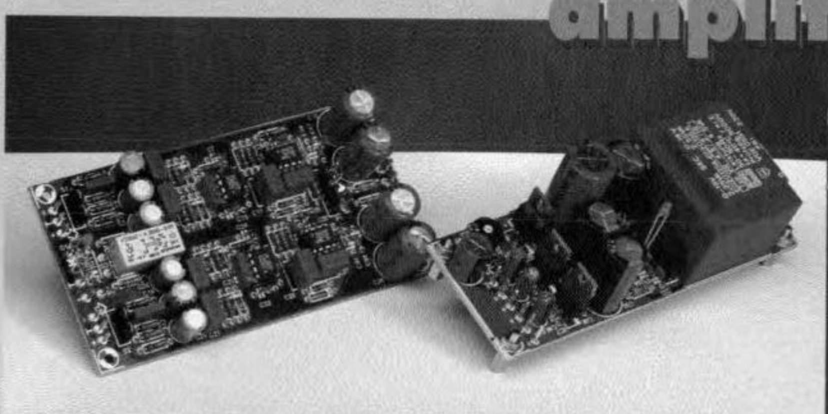
16-page SUPPLEMENT on Test & Measurement
Centrefold

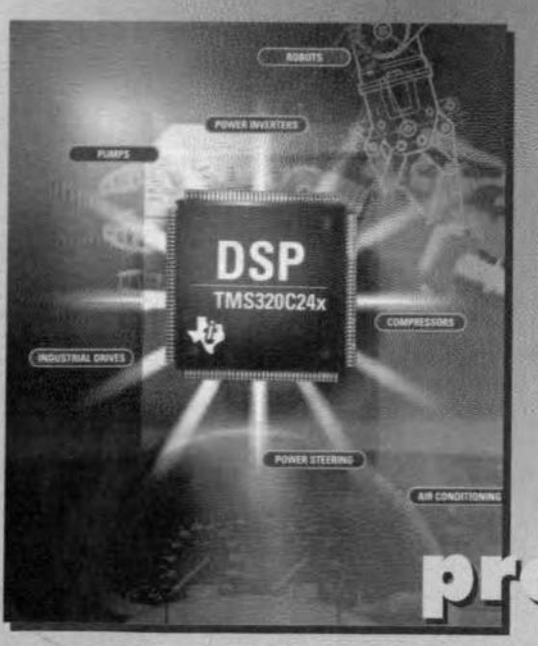
International Autumn/Winter Prize Competition
See pages 43 and 75

Video copy processor 28



Stereo microphone amplifier 22





48 Digital signal processing

* It is regretted that, owing to lack of space, the columns 'Electronics Now' and 'New Books' have had to be held over to next month.

ABC
AUDIT BUREAU OF CIRCULATIONS
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80-metres receiver

The circuit shown here will hopefully fill an important gap, so missing, in current DIY electronics. So, leave the computer alone for one night and let's build a short-wave radio. The 80-m band CW/SSB receiver to be described is simple to construct, yet is capable of high performance with a good aerial and, equally important, a good earth.

By Eric Edwards, GW8LJJ

In years gone by, all radio magazines produced 'blueprints'. These were circuit diagrams on a separate pull-out sheet that was blue and contained instructions to make a radio (wireless) set. They were the *plans*. Although the blueprints have long gone, and there is still a lot of interest in home constructing of radios, there does seem to be little information on building from the junk box. By contrast, there have been many articles where specialised components are required or coils have to be sourced commercially.

The radio we will build is of the regenerative type. The circuit is adjusted to the point of just oscillating when it is at its most sensitive. With a few minutes practice this control will become second nature in using whilst tuning the band.

The band I have chosen is eighty metres (3.5 MHz). This is a good frequency band to listen to local and DX (long-distance) radio amateurs. I have decided to use radio amateurs as our listening choice because on this band, and indeed all other short-wave bands, radio amateurs use single-side-band and CW (Morse code). This receiver is best suited for receiving these modes of transmissions.

To start with, you will need a box of some sort to house the receiver. This can be any size and type to suit your own personal taste. I used two boxes, one for the radio circuit and the other one for the loudspeaker. You can, of course, build the complete radio and loudspeaker into one box. This can be die-cast or any other type as long as it is strong. With any radio equipment, stability is most important. The radio is not to be built into a shoe box!

Design principle

As stated earlier, the radio is of the regenerative type. This principle was used in early radios employing a *reaction control*. The principle is, part of the output signal is sent back to the input and is amplified, this again is sent back and amplified even more. The amount of positive feedback is controlled by a regenerative potentiometer to the point of *just* oscillating. Since received signals present variable signal levels at the input of the receiver (albeit of the order of micro-volts), they actually control the oscillator activity. In this way we can receive SSB and CW signals at good sensitivity.

The *selectivity* can be controlled by either a very slow motion tuning drive or another variable tuning capacitor wired in parallel with the main tuning to provide what was called *bandspread* (or fine tuning). The tuning arrangement I used on one of my designs was from an old frequency wavemeter. This had a suitable tuning capacitor and a very slow tuning dial. Another of my designs used a 150-pF main tuning capacitor and a smaller tuning capacitor of about 30 pF as the bandspread control. A 150-pF or so can be made from a standard tuning capacitor by removing some of the plates. These are usually crimped onto the main shaft and can be extracted with pliers and brute force. Just imagine you are a dentist while tugging away at the capacitor vanes! Leaving about five or six vanes will be a useful main tuning control. Of course, you can buy one of these but they will be expensive as commercial radio manufactures do not use many of these nowadays. Look around on radio rallies or in junk stores for suitable tuning capacitors, or even old radios if they are not too expensive. A tuning type capacitor is also used in series with the aerial (it must be insulated from the front panel). This control is used to introduce a mismatch at the input and so attenuate strong signals to prevent overloading. You will soon realise the

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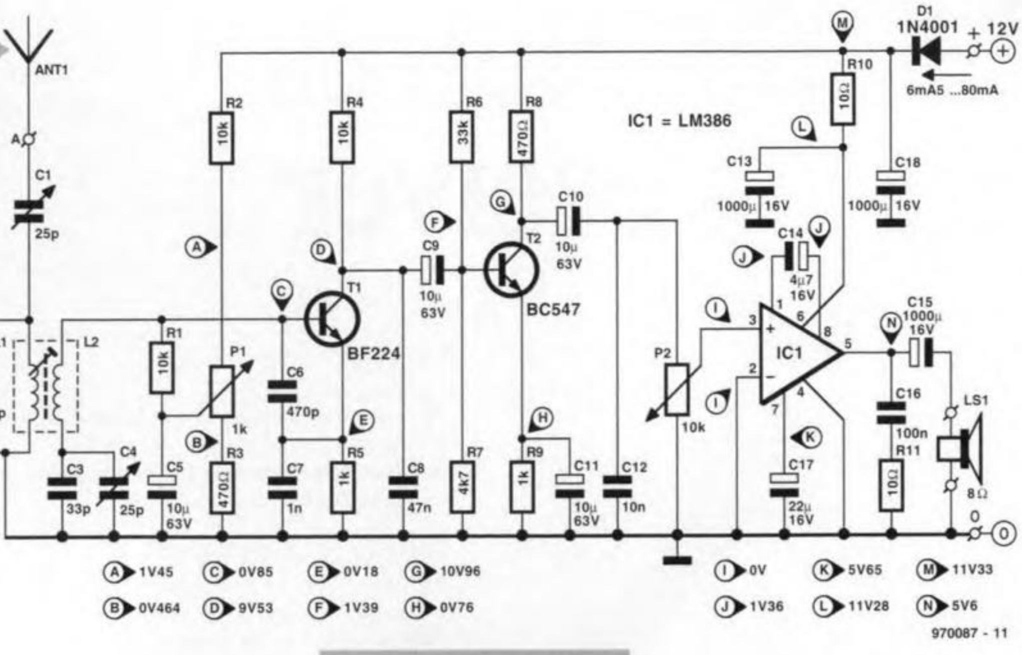


Figure 1. Circuit diagram of the CW/SSB regenerative receiver for the 80-m amateur radio band.

sensitivity of this little receiver, and you will be keeping the aerial at some attenuation.

Circuit description

The circuit as shown in **Figure 1** uses a Clapp oscillator in which the base bias of the oscillator transistor is adjusted by the 'reaction' potentiometer, P1. The tuning is by means of a variable capacitor or a trimmer in series with the bottom end of L2, the secondary winding on the coil former. As most of you will eventually be using some form of ex-radio tuning capacitor (with or without bandspread, see above), a single 25-pF trimmer, C4, and a parallel capacitor, C3, are designed into the circuit for the sake of convenience, and for quick testing.

Because of the very large gain variations produced by received signals, the collector voltage of the tuned oscillator in fact represents the demodulated signal. It is tapped via a capacitor, C9, and fed to T2, which acts as an audio preamplifier. Next comes a volume control, P2. An LM386, IC1, is used as the audio output stage. Although any output device will do here, I chose this IC because of its low cost, and I already had four of them in my junk box! The LM386 has the usual Boucherot network at its output (here, C16-R11) to prevent oscillation. Test voltages measured on our prototype of the receiver are shown at various important junctions in the circuit diagram.

In practice

The complete circuit is built on the custom designed printed circuit board shown in **Figure 2**. Construction of the receiver on this board is self explanatory as you only have to follow the parts list and the component overlay. Unfortunately the PCB is not available ready-made through our Readers Services.

The only component you have to make yourself is the combined inductor L1-L2. The components that make up the 10F1 inductor assembly (from Neosid) are illustrated in **Figure 3**. L1 consists of 15 close-wound turns of 0.1-mm dia. (SWG40) enamelled copper wire on the lower part of the former. Although the other winding, L2, is made with the same wire type, it differs from L1 in that it consists of two layers: the lower one has 40 turns, the upper one, 20. L2 is wound on the upper part of the 10F1 former. As the copper wire is quite thin, it is also brittle, so work carefully. Don't forget to remove the enamel layer from the wire ends, this is best done by carefully sandpapering or scratching (with a sharp hobby knife) the last 5 mm or so, and then pretinning with your solder iron. Check the continuity of the inductors at the base pins of the former, using the PCB component mounting plan and the circuit diagrams as guidance. Also make sure none of the base pins comes into contact with the metal cap which is pushed over the former once this is in place on the circuit board. Before mounting the finished inductor onto the board, check its looks against those of our prototype, shown in **Figure 4**.

If you are unable to produce the PCB shown in Figure 2, there is an alternative: veroboard construction. The method I have used is a little different in the sense that I use the veroboard up-side-down! I lay the components on the tracks and cut away the unwanted copper. The component leads do not go through the board as is the norm, but actually sit on the tracks. This keeps all component leads to an absolute minimum length. The un-coppered side is then stuck (with super-glue) to a copper ground plane. In RF circuits it is important to keep stray capacitance to an absolute minimum. One way of achieving this is by placing the complete circuit on an earthed plane such as copper clad board. The leads of components must be as short as possible, if not non-existent, as these can act as part of a tuned circuit especially at high frequencies. Although not as critical in the frequencies we are using here, it is still good practice to cut component leads very short.

The 12 Volts is applied, and the aerial with the all important earth is also attached. If you are using a bandset

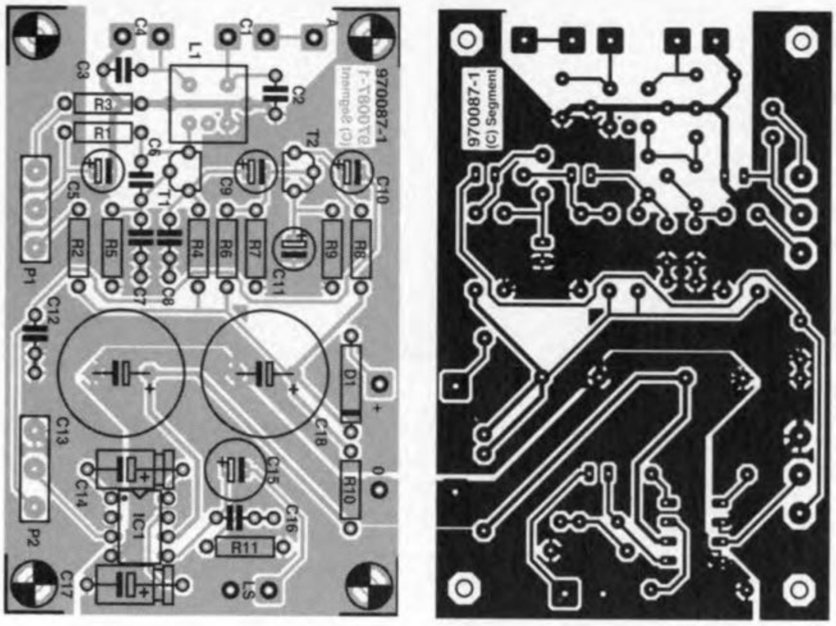
Back to Earth

Do not be tempted to use the mains earth, although good for what it is intended for in the household, it may not be the best for radio reception. Make an earth spike out of preferably copper, but almost any metal will do. Mine is made from an old copper pipe that I made into a point at one end for entering the garden, and a flat at the other end to which I drilled a small hole to fit a nut and bolt with a solder tag. The pipe is about ten inches long but this is not important[1], as any length will do depending how deep your garden goes. I soldered a length of good thick earth wire and fed it through a hole in my shack wall (Shack is a radio amateur's shed), which was then attached to a box with a couple of 4-mm sockets. The total length of the wire is about two meters. The shorter the wire and the heavier the gauge, the better the earth will be. Always keep it damp, no problem on wet days but you will need to throw a bucket of water over it on drier spells... when the water board is not looking.

[1] Note that the author lives practically at sea level. A much longer earthing rod may be required depending on your location and the soil condition. Tech. Editor.

2

Figure 2. Copper track layout and component mounting plan of the PCB designed for the receiver (board not available ready-made). The large copper plane provides very good decoupling of RF signals where necessary, and so helps to keep the receiver stable.



control as the main (coarse) tuning, then set this to about halfway (half mesh). The bandspread control is also set the same. If you have a good slow motion drive on the main tuning capacitor, then you will probably not need the bandspread capacitor (the bandspread capacitor is wired directly across the main tuning capacitor). Set the aerial tuning capacitor (here, trimmer C1) to half-mesh also.

If you have a signal generator or a grid dipper you can then align the coil to the middle of the band around 3.65 MHz, critical alignment being made with the coil and capacitor/s. Do not switch the receiver on. Couple the generator signal to the receiver input by means of a few turns of wire around the coil former. Set C4 to almost minimum capacitance, and peak the core for best reception of the test signal. The signal level may be measured by connecting a 10:1 (<15 pF load) oscilloscope probe to the hot side of C4. Carefully and alternately adjust the core and C4 until the signal level peaks. At this setting, L1 and C2 are tuned exactly to the centre of the band.

If you have a short-wave receiver already, you can find where you are by comparison. If you have neither of these, don't worry, you'll soon find what you want by experimenting. Remember, though, that due to radio-wave propagation the 80-m band does not come alive until after sunset, and that reception is generally better in winter time. Set the reaction pot also to about halfway. With all the necessary items attached slowly adjust the pot until a slight oscillation is heard. Turn the main tuning control whilst keeping the reaction in the state of just oscillating. When a signal is detected, adjust the bandspread (fine tuning, if fitted), aerial tuning and reaction until the signal is a pleasure to listen to. Obviously you will have adjusted the volume control accordingly. A little practice and you will soon realise how well this little circuit performs. I am listening to some continental amateurs as I am typing this text. I live quite low down by the coast and have a moderate wire aerial.

3

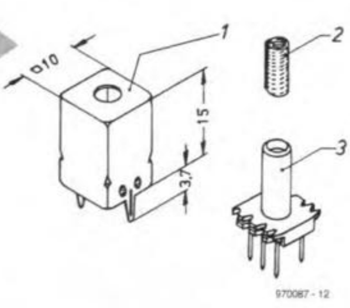


Figure 3. The components that make up the 10F1 assembly from Neosid. (1) screening can; (2) ferrite core; (3) coil former with base.

In conclusion

Enjoy building this little receiver and have hours of fun tuning around the band. The important factors are stability and short component leads. The rest is just patience to start with in getting top know this receiver. This really should not take more than a few minutes. You may experiment with the number of turns of the coils, but keep to the same ratio. Happy tuning!

(970087-1)

COMPONENTS LIST

Resistors:
R1,R2,R4 = 10kΩ
R3,R8 = 470Ω
R5,R9 = 1kΩ
R6 = 33kΩ
R7 = 4kΩ7
R10,R11 = 10Ω
P1 = 1kΩ linear
P2 = 10kΩ logarithmic

Capacitors:
C1,C4 = 25pF trimmer or tuning capacitor
C2 = 680pF
C3 = 33pF
C5,C9,C10,C11 = 10μF 63V
C6 = 470pF
C7 = 1nF
C8 = 47nF
C12 = 10nF
C13,C18 = 1000μF 16V radial
C14 = 4μF7 16V
C15 = 100μF 16V
C16 = 100nF
C17 = 22μF 16V

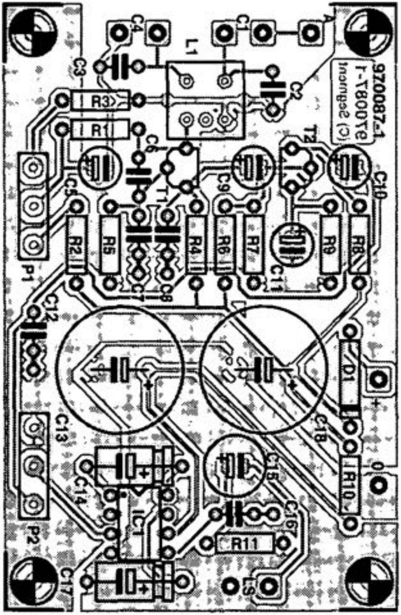
Inductors:
L1= Neosid 10F1 (brown dot on core)

Semiconductors:
D1 = 1N4001
T1 = BF224 or BF494
T2 = BC547B
IC1 = LM386

Miscellaneous:
LSP1 = 8 ohm 1 watt

4

Figure 4. Close-up of the finished input band filter, L1-L2, fitted in the circuit. Do not apply too much heat while soldering the enamelled copper wire, as the 10F1 base is prone to soften a little and the pins will easily come loose. Also note that we've used a few drops of candle wax to secure the windings.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(C) Segment
C1
C2
C3
C4
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C13
C14
C15
C16
C17
C18
R1
R2
R3
R4
R5
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R9
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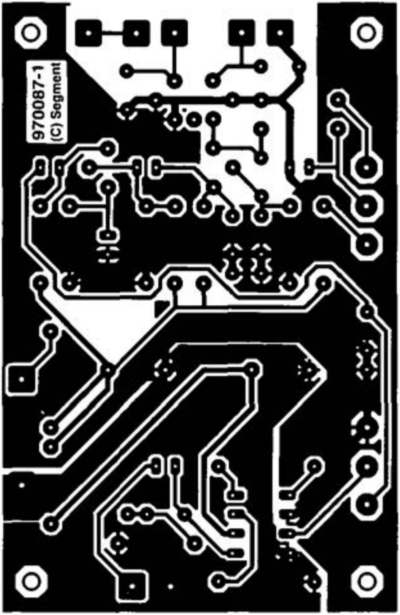
L1
T1
T2
D1
IC1
P1
P2
A
0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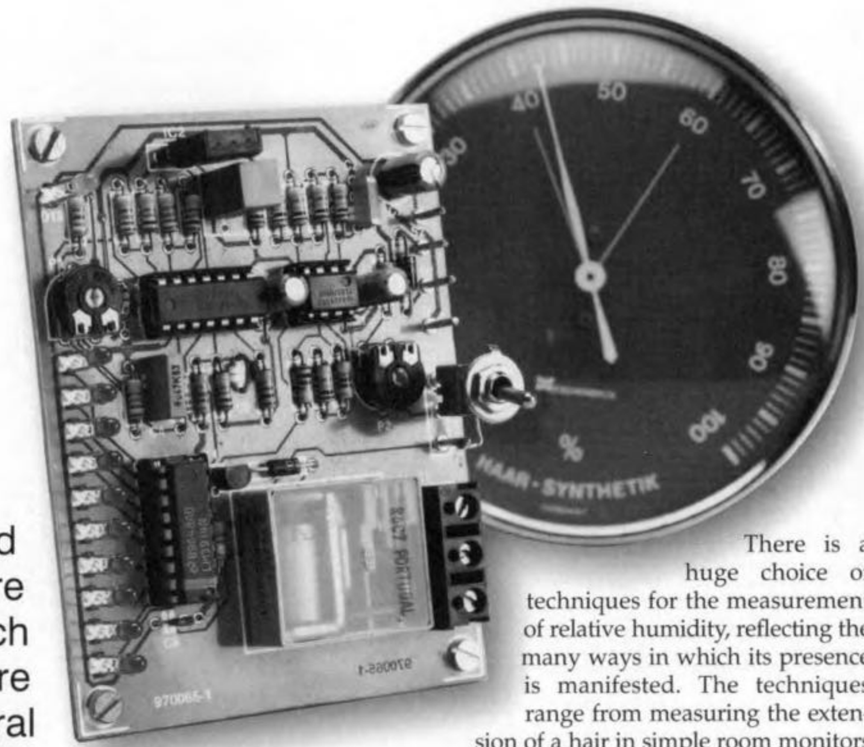
hygrometer

with LED display and switched output



The measurement and control of the moisture content of gases, such as the atmosphere around us, is an integral part of many industries, but is also of importance in the home and workshop. Although moisture, that is, water vapour, is not an ideal gas, for many hygrometry purposes it is sufficient to assume that it behaves ideally. In environmental applications the basic unit of moisture is relative humidity, which is the ratio in per cent of the actual vapour pressure in the atmosphere to the saturation vapour pressure of water at the same temperature. It is thus temperature dependent but independent of the atmospheric pressure.

Design by H. Bonekamp

There is a huge choice of techniques for the measurement of relative humidity, reflecting the many ways in which its presence is manifested. The techniques range from measuring the extension of a hair in simple room monitors to sophisticated electronic instruments.

The instrument proposed in this article is based on a special sensor of which a detailed description is given in the box further on in this article. It is especially intended for use indoors and has a range of 35–80 per cent RH (relative humidity). Its readout is made up of high efficiency light-emitting diodes (LEDs). Provision is made for driving an external fan or dehumidifier via an integral relay contact.

ELECTRICAL SENSORS

There are many substances whose electrical impedance changes with the surrounding moisture level. This change of impedance can be calibrated in terms of moisture concentration or relative humidity (RH).

Most of these electrical sensors consist of two electrodes separated by a material whose impedance changes in direct proportion to the relative humidity. The separating material may be a polymer, tantalum oxide, silicon oxide, lithium chloride mixed with plastic, or a host of others.

The sensor used in the present hygrometer is a Type NH-3 which may be considered a moisture-dependent potential divider. When a stable alternating voltage is applied across the divider, a potential is produced at the junction of the two divider branches which is directly proportional to the ambient moisture concentration. This voltage is rectified and differentiated, after which it is used to drive a bar of LEDs via a suitable display driver. The associated electronics circuits are simple.

CIRCUIT DESCRIPTION

In the circuit diagram in **Figure 1**, the moisture-dependent sensor is IC_2, while the display LEDs are driven by IC_4. Rectification and differentiation of the sensor output is effected by IC_{3a-c}.

To ensure that the moisture-dependent potential divider in the sensor functions satisfactorily, it needs to be powered by an accurately defined, symmetrical alternating voltage. This is provided by rectangular-wave generator IC_{1a}. This is a comparator with a totem pole (single-ended) output, in which the conditions for oscillation are met by the positive feedback provided by R_3. The time-constant, τ, is equal to R_4-C_2. The rectangular wave output is symmetrical (square wave) and has a frequency of 1 kHz. Its level of 3 V_{p-p} is stable since the generator is powered by a reference voltage, U_{REF}, of 3 V.

Since there should be no direct current through the sensor, the output of the generator, IC_{1a}, is coupled to the sensor, IC_2, via capacitor C_3.

The impedance of the sensor varies in direct proportion to the ambient relative humidity: from 2 kΩ at 90% RH to 10 MΩ at 30% RH. The consequent long time constants, and thus the sustained switching phenomena, are countered by resistor R_5 in parallel with the sensor.

A high dissipation in the sensor affects its linearity and the design therefore ensures that this remains below 1 mW over the entire measurement range.

The sensor output, taken from pin 2, is applied to op amp IC_{3b} via buffer IC_{3a}. The op amp not only raises the signal by ×2.7, but, since it operates asymmetrically, also rectifies it (half-wave). Moreover, in this stage the reference level is shifted from U_{REF} to earth.

The pulsating direct voltage at the output of IC_{3b} is differentiated by network R_{10}-C_4, after which the signal is buffered by IC_{3c}, and then applied to the input of IC_4 via calibration control P_1.

Circuit IC_4 is an accurate potential divider containing ten comparators each of which can drive an LED directly. The IC is therefore able to display variations in the applied direct voltage on a bar of ten LEDs. Its pin 9 enables a choice to be made between dot and bar

Figure 1. The circuit of the hygrometer consists of a sensor, rectifier/differentiator, display driver, bar of LEDs, and power supply.

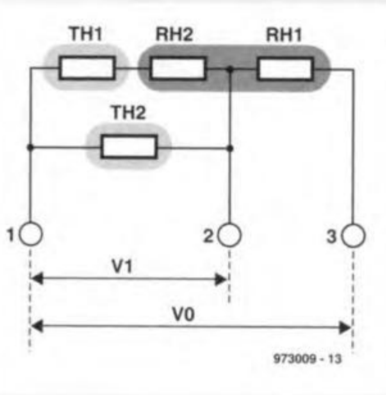
SENSOR

In the prototype, the moisture-dependent sensor is a Type NH3 from Figaro, but other suitable models may, of course, be used.

Basically, the sensor consists of a combination of a moisture absorbing material and a thermistor. The Type NH-3 is an ultra-linear model which uses two such combinations. The sensor proper is embedded in porous ceramic, which ensures stability as well as a long life. The electrodes are made from ruthenium oxide, RuO_2, which is also a porous material.

Both the sensor and thermistor are mounted on a ceramic substrate.

In the equivalent circuit shown, the resistors forming the potential divider are marked R_{H1} and R_{H2}, and the thermistors, TH_1 and TH_2. The applied alternating voltage is V_0, while the sensor output is V_1.



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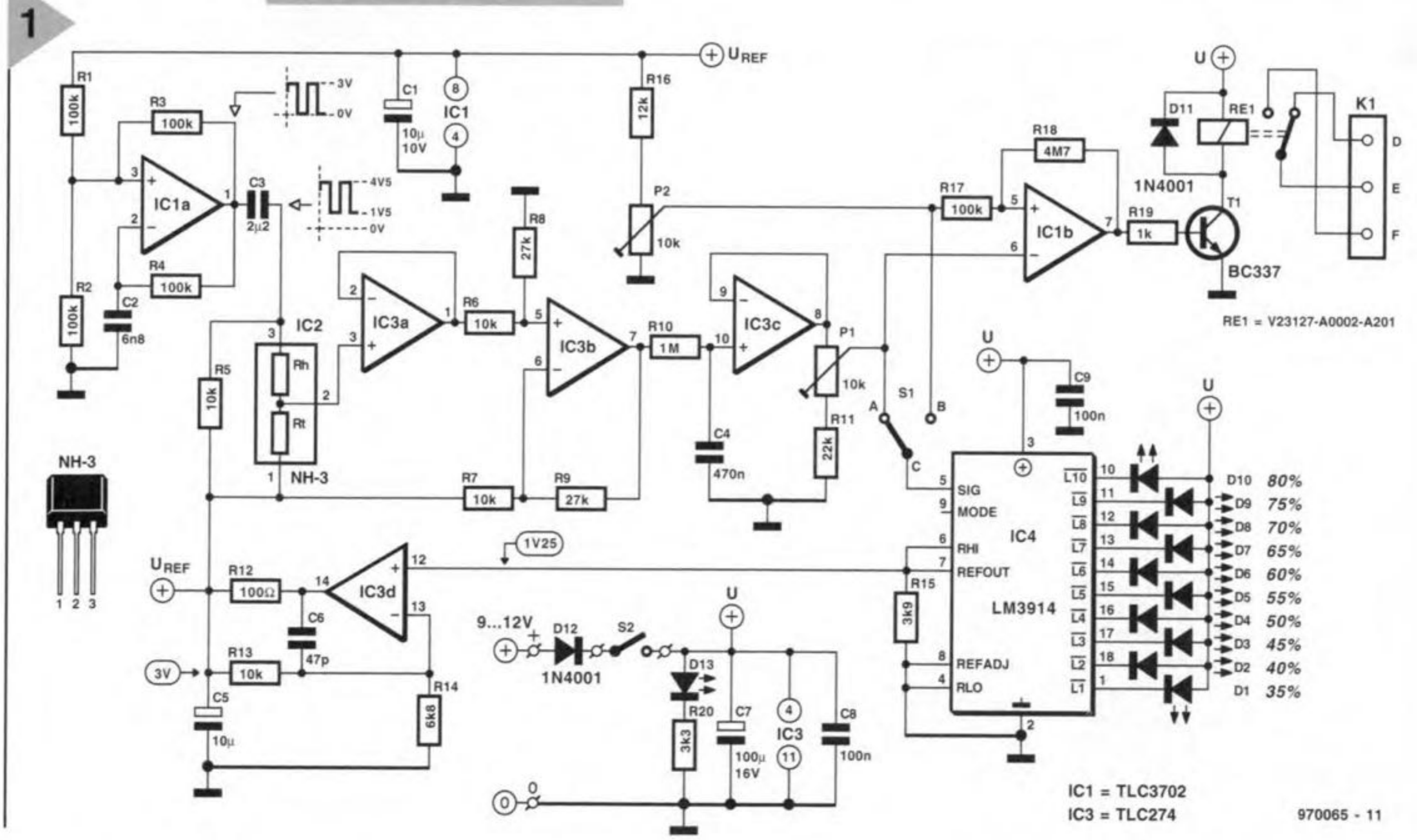

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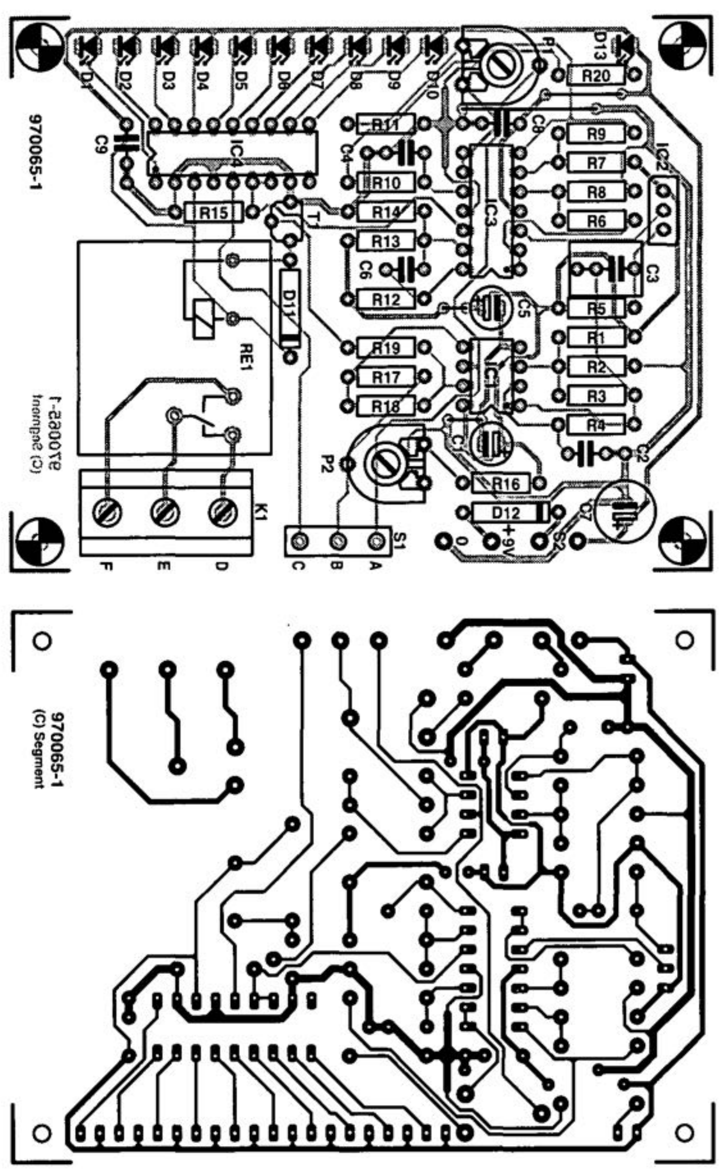

Figure 2. The printed-circuit board for the hygrometer holds all components.

Parts list

Resistors:
R_1–R_4, R_{17} = 100 kΩ
R_5–R_7, R_{13} = 10 kΩ
R_8, R_9 = 27 kΩ
R_8, R_9 = 27 kΩ
R_{10} = 1 MΩ
R_{11} = 22 kΩ
R_{12} = 100 Ω
R_{14} = 6.8 kΩ
R_{15} = 3.9 kΩ
R_{16} = 12 kΩ
R_{18} = 4.7 MΩ
R_{19} = 1 kΩ
R_{20} = 3.3 kΩ
P_1, P_2 = 10 kΩ preset

Capacitors:
C_1 = 10 µF, 10 V, radial electrolytic
C_2 = 0.068 µF, metallized polyester (MKT)
C_3 = 2.2 µF, metallized polyester (MKT)
C_4 = 0.047 µF, metallized polyester (MKT)
C_5 = 10 µF, 10 V, radial electrolytic
C_6 = 47 pF, ceramic
C_7 = 100 µF, 16 V, radial electrolytic
C_8, C_9 = 0.1 µF, high stability

Semiconductors:
D_1–D_{10} = LED, red, high efficiency
D_{11}, D_{12} = 1N4001
D_{13} = LED, green, high efficiency
T_1 = BC337

Integrated circuits:
IC_1 = TLC3702CP
IC_2 = NH-3 (Figaro) or equivalent
IC_3 = TLC274CN
IC_4 = LM3914N

Miscellaneous:
S_1 = change-over switch
S_2 = single-pole on/off switch (or push-button type – see text)
Re_1 = relay, 9–12 V, single change-over contact
K_1 = 3-way PCB terminal block, pitch 7.5 mm
PCB Order no. 970065 (see Readers Services towards the end of this issue)

modes (the prototype uses the dot mode).

Normally, the accurate reference voltage, U_{REF}, required for powering the sensor is obtained by a fairly complex arrangement. In the present circuit, it is, however, provided by display driver IC_4 and is available for external applications at pin 7. It is too low for the sensor, and is therefore raised by IC_{3d} to 3 V. The output of the amplifier is compensated for capacitive loads by R_{12} and C_6.

The switched output is driven by comparator IC_{1b}, which functions as a level detector. The comparator is provided with an hysteresis of about 4 per cent by R_{17} and R_{18} and drives relay Re_1 via transistor T_1. The change-over contact of the relay is linked to the outside world via connector K_1

The reference voltage applied to the non-inverting (+ve) input of IC_{1b} is derived from U_{REF} and set to the required level with P_2. This control thus sets the degree of relative humidity at which the relay is actuated.

The level setting is simplified by switch S_1. When this is set to position B, the sensor signal is interrupted, and in its stead the reference voltage is displayed on the bar of LEDs via IC_4.

CONSTRUCTION

Even less-experienced constructors will find the assembly of the hygrometer on the printed-circuit board shown in **Figure 2** a fairly easy affair, since there is no additional wiring and all components are housed on the board. As usual, mind the polarity of the diodes and electrolytic capacitors and fit the ICs in suitable sockets, although this is not strictly necessary.

The assembly may be fitted in any of a variety of suitable ABS (*acrylonitrile-butadiene-styrene*) cases. The really important thing is that the sensor is fully exposed to the ambient atmosphere, and this requires the drilling of a number of not too small holes in the enclosure.

There are two ways of obtaining the requisite power supply. When the instrument is used in a more or less permanent position, it is best to use a standard mains adaptor with an output of 9–12 V. Since the hygrometer draws a current of only 50 mA, such an adaptor need not be a heavy duty version.

For mobile use, it is, of course, necessary to use a 9 V battery instead of a mains adaptor. It is then, however, not possible to use the switched output. Transistor T_1, the relay, and most of the components associated with IC_{1b}, but not R_{18}, may then be omitted. Resistor R_{18} is necessary for the stability of the op amp.

Ignoring the relay, the circuit draws a current of about 14 mA. Although this is modest, do not forget to switch off the supply when the meter is not in use, otherwise the battery will soon be discharged. It is, therefore, better, if the meter is always used in mobile form, to replace S_2 by a spring-loaded (push button) on/off switch.

CALIBRATION

If the hygrometer is used (almost) exclusively to switch on a fan or dehumidifier in spaces suffering from damp conditions, the calibration is not terribly important: it then suffices to set P_1 to its centre position.

Where greater accuracy is desired, the instrument must be calibrated (by adjusting P_1) with the aid of a reference hygrometer.

Performance of the prototype compared reasonably well with that of some commercial hygrometers, at least over the range 40–80% RH. This is in accord with the manufacturer's sensitivity curve shown in **Figure 3**. It will be noted that the curve has a sharp(ish) bend below about 40% RH, but that its linearity over the remainder of the range is good.

The degree of dampness at which the ventilator or dehumidifier is switched on is set with P_2. This is best done by setting switch S_1 briefly to position B to display the reference voltage on the bar of LEDs.

[970065]

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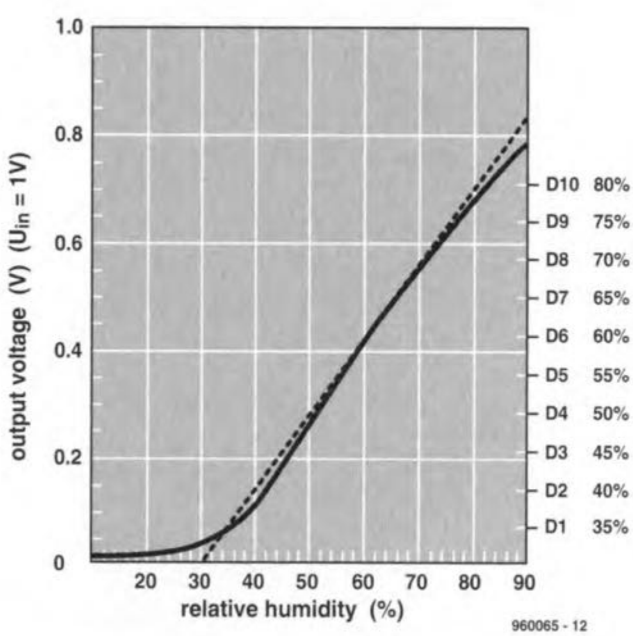
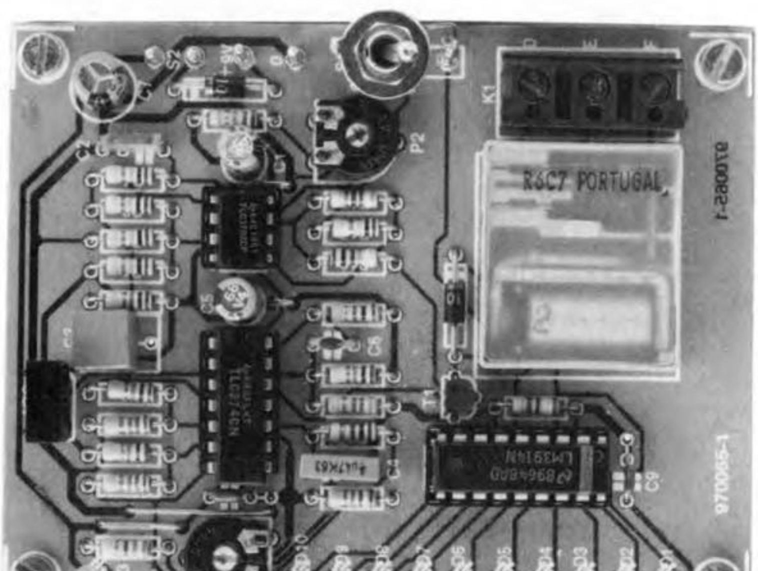


Figure 3. The sensitivity curve of the sensor is linear from 40% RH to 80% RH.



in passing ...

Another fine mess

About half a year ago my wife and I moved to a larger house. Detached, just 8 years old, and in a charming hamlet not too far from our respective work locations. Don't get me wrong, we are very satisfied with our choice (which we were able to make after half a year's intensive house-hunting), and the house is reasonably well built using good quality materials. It has every luxury I personally want (and that's not a lot), and it looks splendid after our most recent undertaking, all-round paint work!

To get to the point, the only trouble with the house is electrical. The previous owner claimed he had 'done a lot himself' to finish the construction of the house. Unfortunately, although reasonably versed in plumbing and other 'coarser' trades like masonry (the results of which far exceed my own capacities), he deserves a very low grade for electrical installation work. While the larger part of the electrical system has been installed by qualified people, the previous owner must have made the odd change and addition here and there after the electricity board inspector approved the system and walked off about 8 years ago.

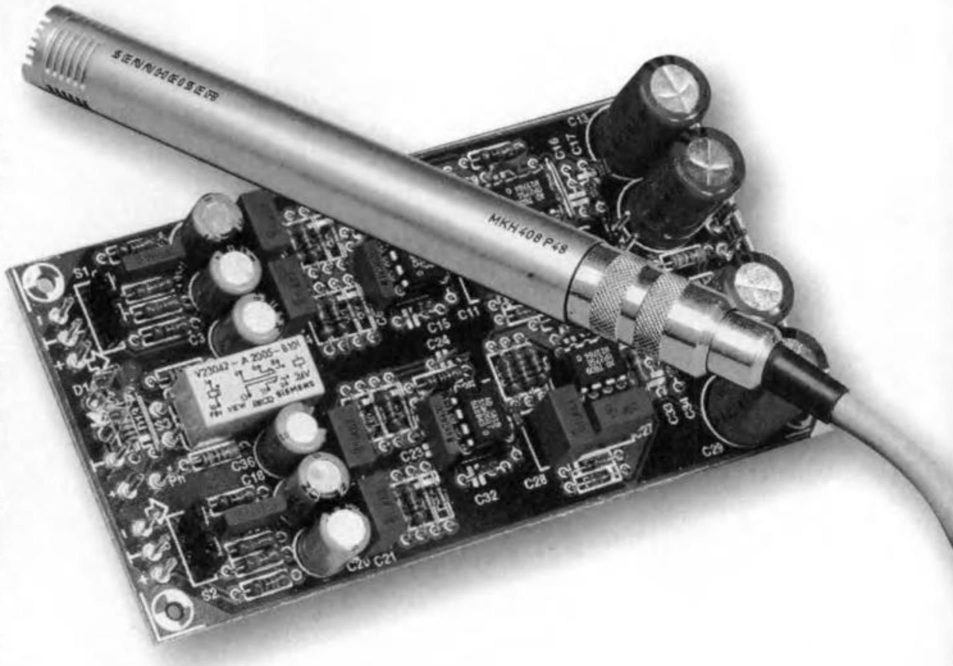
I now have a collection of A4 sheets on which I have attempted to unravel various attempts at layman's electrical installation. For example, I found that the connection of the central heating boiler to the thermostat fitted in our living room used half a dozen wires of assorted colours and a total length of more than 100 m, travelling up and down the house, either unprotected or through various pieces of plastic conduit. As I opened the old dial-type Honeywell thermostat, I was perplexed. The DIY CH control system was of a fearsome complexity, using a special 24-V transformer, a relay, a hidden switch and a very noisy valve (controlled by a mains extension cable running from the attic down to the scullery) to control the floor heating, and, to cap it all, an antediluvian switching clock which enabled the CH boiler from 7 am to 11 pm, with its vessel thermostat set to 75°C. Although all these components are of industrial quality (I wonder where he got them from), they are superfluous in my view. I removed all the wiring, the relay and the transformer, and installed a modern, intelligent, room thermostat which switches the gas valve using the 24-V supply also fitted inside the CH boiler. As far as I know, this is a standard, safe and approved connection based on just two medium-duty wires to the room thermostat. After a learning period of about one week in spring time, the heating system worked fine, and my efforts have been rewarded with a much reduced gas bill.

A few days ago I decided to install an infra-red (PID) controlled lamp in the front porch. No problem, I thought, until I opened the double switch in the hall and saw many more wires than I had expected, some just entering the switch case and leaving it again without any connection whatsoever. No assorted colours this time: just one: green!

Jan Buiting

stereo microphone amplifier

with balanced inputs and phantom power supply



The magnification of very weak audio-frequency signals is and remains a precarious affair. That is why a microphone amplifier good enough for professional applications must have balanced inputs and indisputably good specifications. Foremost, of course, is the faithful transducing of sound. Although this is a subjective aspect, it is confidently expected that the design of the amplifier described here will satisfy even professional audio engineers and technicians. It is a stereo model with integral phantom power supply for electret microphones.

Design by J Brangé

INTRODUCTION

There are not all that many good methods to amplify the very small a.f. output signals of a microphone without affecting the quality of the input sound. One of the best is undoubtedly that in which a balanced instrument amplifier, discrete or integrated, is used.

Today, the accent in design is on integrated circuits, ICs, rather than on discrete transistors. One of the many a.f. ICs in the gamut of Analog Devices is the SSM2017, which is eminently suitable for use in a balanced microphone amplifier. This operational amplifier has a small noise factor and low total harmonic distortion, combined with a large bandwidth and high slew rate. Also, its amplification may be set within wide limits.

CIRCUIT DESCRIPTION

From the foregoing, it is clear that basing the design of the microphone amplifier on an SSM2017 is a sound decision. The circuit diagram in **Figure 1** may look extensive, but it should be borne in mind that it concerns a stereo circuit. This means that there are two of every device and component with the exception of those in the power supply. The following description will be limited to the left-hand channel.

The microphone is connected to the balanced input amplifier, IC_1, via terminals L+ and L–. The facility for short-circuiting R_1 and R_2 is provided for cases in which the circuit is to be used as a line amplifier. The switches are then open to provide an attenuation of 10 dB. When the circuit is used as microphone amplifier, the resistors are short-circuited. So, if use as a line amplifier is not envisaged, the switches and R_1, R_2 may be replaced by a wire bridge.

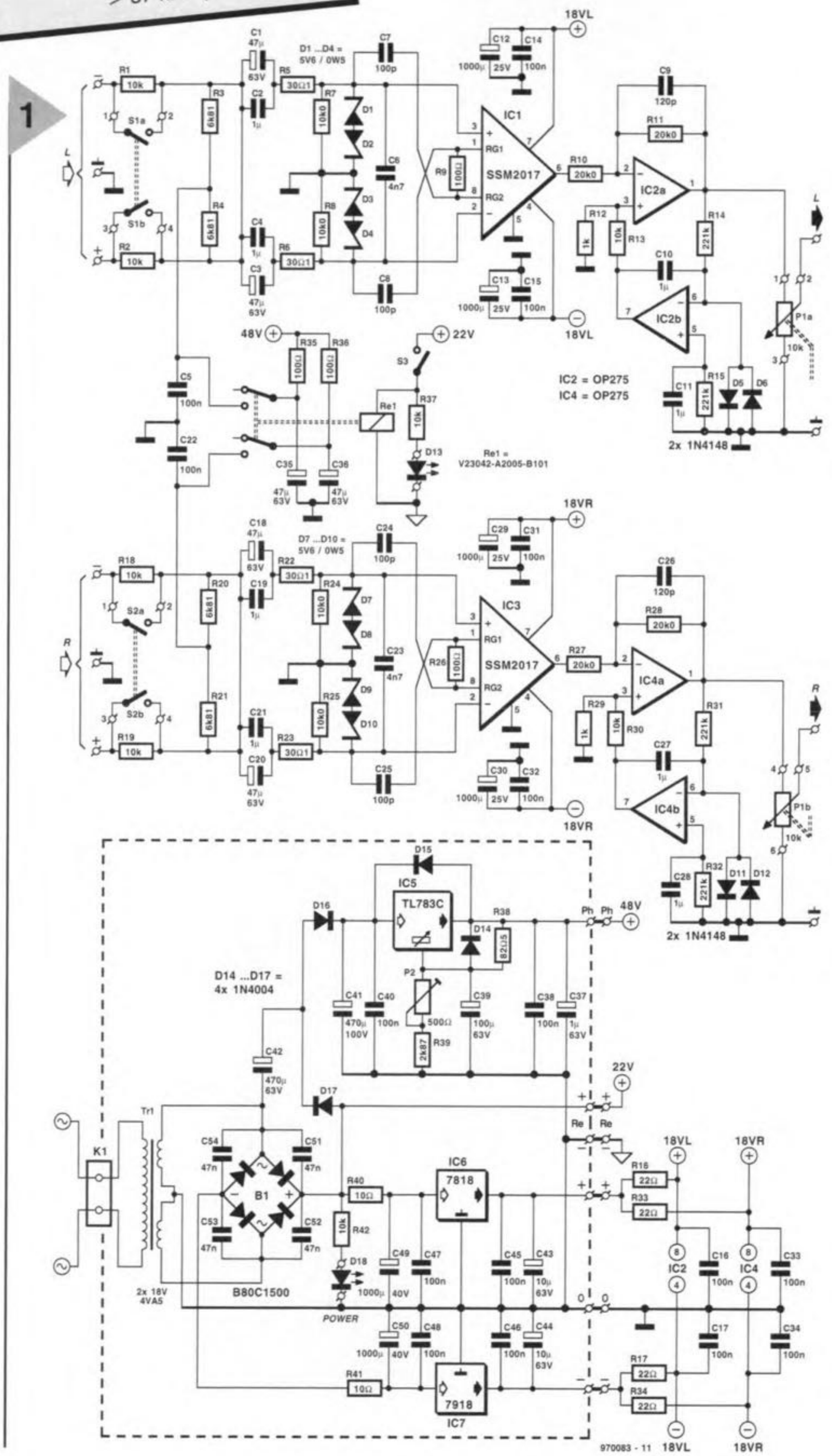
The supply voltage needed by electret microphones is provided via a phantom line. This means that the signal lines are used for carrying the supply voltage, which is, of course, not applied to the amplifier input. The volt-

Features

(measured with an input signal of 2 mV into 50 Ω)

Signal-to-noise ratio		>74 dB (BW 22 kHz linear)
Total harmonic distortion (20 Hz–20 kHz)		<0.03% (BW 80 kHz)
Channel separation	(20 Hz)	120 dB
	(1 kHz)	95 dB
	(20 kHz)	83 dB
Bandwidth		> 60 kHz (source 50 Ω)
		> 37 kHz (source 600 Ω)

Figure 1. Circuit diagram of the stereo microphone amplifier, which is based on an SSM2017 operational amplifier from Analog Devices.



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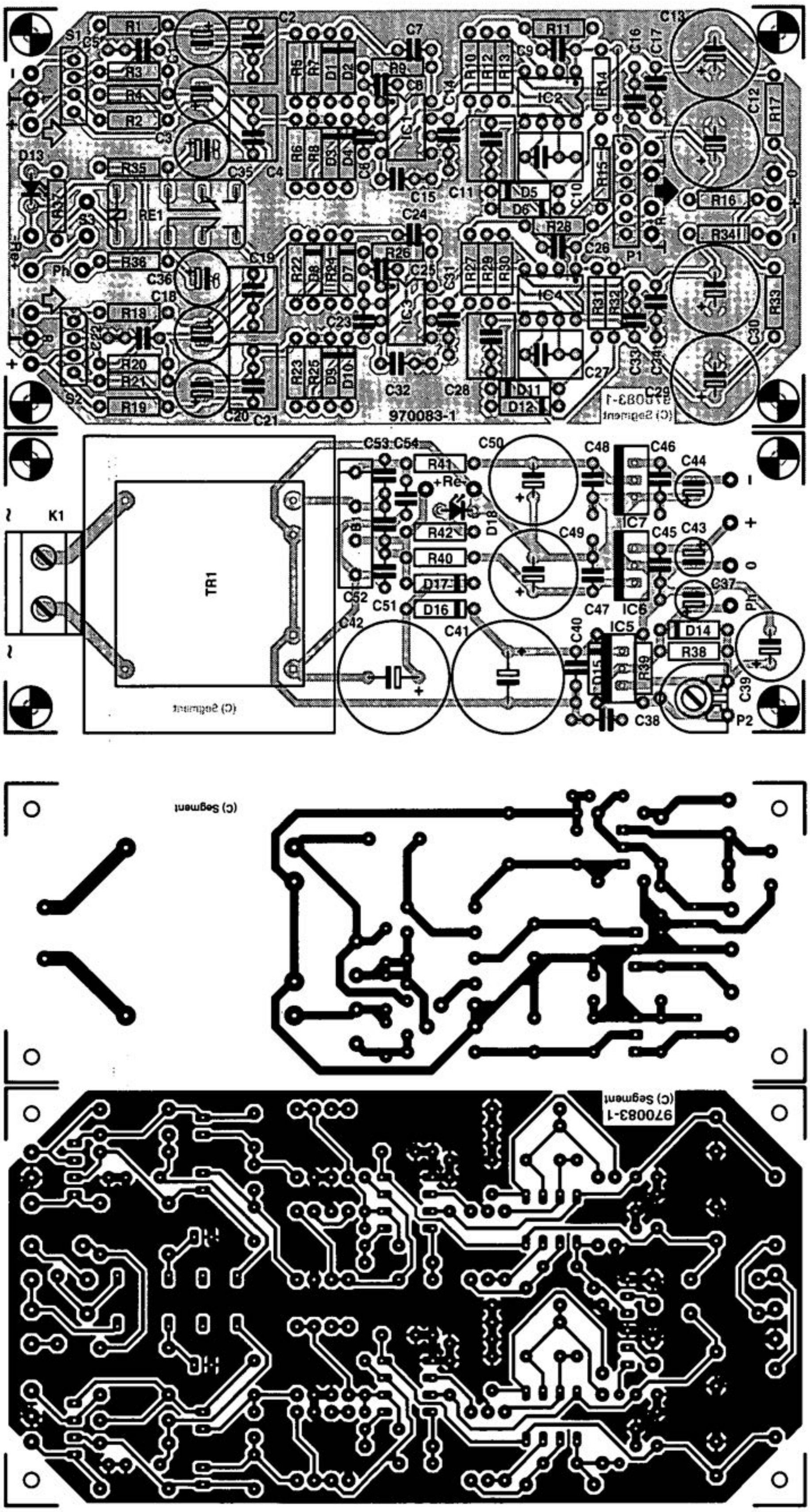

Figure 2. The printed-circuit board is in two parts to enable the amplifier and power sections to be kept isolated.

Parts list

Resistors:

R_1, R_2, R_7, R_8, R_{13}, R_{18}, R_{19}, R_{24}, R_{25}, R_{30} = 10.00 kΩ, 1%
R_3, R_4, R_{20}, R_{21} = 6.81 kΩ, 1%
R_5, R_6, R_{22}, R_{23} = 30.10 Ω, 1%
R_9, R_{26} = 100 Ω, 1%
R_{10}, R_{11}, R_{27}, R_{28} = 20.00 kΩ, 1%
R_{12}, R_{29} = 1.00 kΩ, 1%
R_{14}, R_{15}, R_{31}, R_{32} = 221 kΩ, 1%
R_{16}, R_{17}, R_{33}, R_{34} = 22 Ω
R_{35}, R_{36} = 100 Ω
R_{37}, R_{42} = 10 kΩ
R_{38} = 82.50 Ω, 1%
R_{39} = 2.87 kΩ, 1%
R_{40}, R_{41} = 10 Ω
P_1 = 10 kΩ, log, stereo potentiometer
P_2 = 500 Ω preset potentiometer

Capacitors:

C_1, C_3, C_{18}, C_{20}, C_{35}, C_{36} = 47 μF, 63 V, radial
C_2, C_4, C_{10}, C_{11}, C_{19}, C_{21}, C_{27}, C_{28} = 1 μF, metallized polyester (MKT), pitch 5 mm or 7.5 mm
C_5, C_{22} = 0.1 μF
C_6, C_{23} = o.0047 μF
C_7, C_8, C_{24}, C_{25} = 0.001 μF
C_9, C_{26} = 0.12 μF
C_{12}, C_{13}, C_{29}, C_{30} = 1000 μF, 25 V, radial
C_{14}–C_{17}, C_{31}–C_{34}, C_{38}, C_{45}–C_{48} = 0.1 μF, ceramic
C_{37} = 1 μF, 63 V, radial
C_{39} = 100 μF, 63 V, radial
C_{40} = 0.1 μF, metallized polyester (MKT), 100 V
C_{41} = 470 μF, 100 V, radial
C_{42} = 470 μF, 63 V, radial
C_{43}, C_{44} = 10 μF, 63 V, radial
C_{49}, C_{50} = 1000 μF, 40 V, radial
C_{51}–C_{54} = 0.047 μF, ceramic

Semiconductors:

D_1–D_4, D_7–D_{10} = zener 5.6 V, 500 mW
D_5, D_6, D_{11}, D_{12} = 1N4148
D_{13} = LED, high efficiency, red
D_{14}–D_{17} = 1N4004
D_{18} = LED, high efficiency, green

Integrated circuits:

IC_1, IC_3 = SSM2017 (Analog Devices)
IC_2, IC_4 = OP275G (Analog Devices)
IC_5 = TL783C (Texas Instruments)
IC_6 = 7818
IC_7 = 7918

Miscellaneous:

K_1 = two-pin terminal block, pitch 7.5 mm
S_1, S_2 = double-pole on switch (or wire bridge – see text)
S_3 = single-pole on switch
B_1 = B80C1500 bridge rectifier
Re_1 = 22 V relay, 2 contacts
Tr_1 = mains transformer, 2×18 V secondaries, 4.5 VA
PCB Order no 970083-1 (see Readers services elsewhere in this issue)

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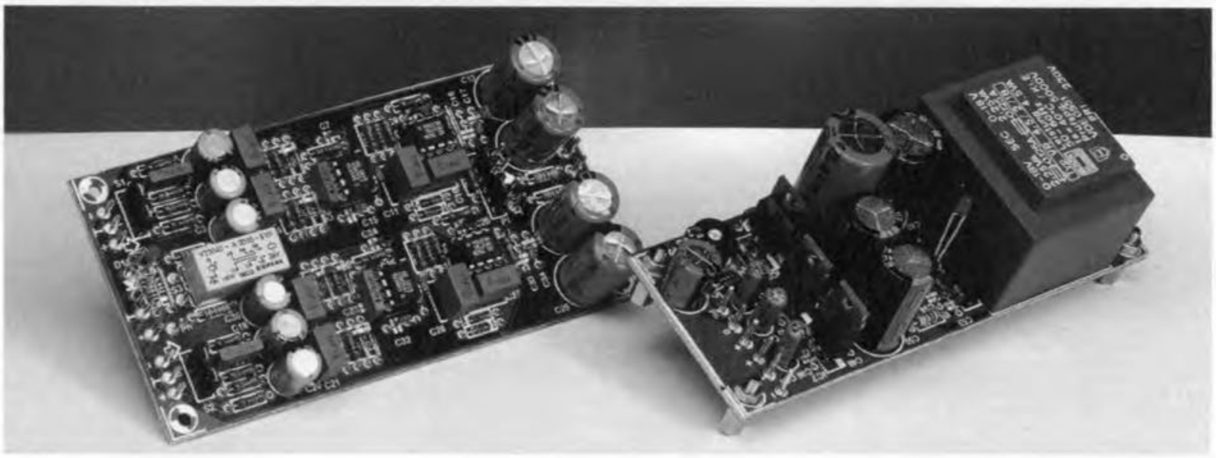

Figure 3. In the prototype, S_1, S_2 and P_1 have been omitted deliberately – see text.

age is 48 V and is linked to the input lines via relay Re_1, which is controlled by S_3. Network R_{35}-C_{35} provides additional smoothing of the supply line. The supply voltage is applied to the balanced microphone lines via potential divider R_3-R_4. RF decoupling is provided by C_5. The lighting of diode D_{13} indicates that the supply is on.

Capacitors C_1 and C_3 prevent any direct voltages from reaching the inputs of IC_1. They are bypassed for r.f. by C_2 and C_4.

Resistors R_5 and R_6, in conjunction with zener diodes D_1–D_4, suppress any peaks on the phantom supply lines that may be caused by the operation of S_3.

Resistors R_7 and R_8 form the input load of IC_1. As the pass-band of this IC is wide, it needs to be narrowed to reduce noise and distortion and also to largely suppress the effect of any r.f. radiation that may be present. The bandwidth is reduced by capacitors C_6–C_8, which must not be ceramic types.

One of the excellent features of the SSM2017 is that its voltage amplification is easily varied without affecting its input impedance and bandwidth. The amplification, α, is determined by the value of R_9 between pins 1 and 8. The relationship between these two quantities is given by

$$\alpha = (10^4/R_9) + 1.$$

With the value of 100 Ω specified for R_9 in the present circuit, this gives an amplification of ×100 (40 dB). For an amplification of ×31 (30 dB), the value of R_9 is 332 Ω, and for ×316 (50 dB), it is 31.6 Ω.

The output of IC_1 is applied to volume control P_{1A} via buffer amplifier IC_{2A}. This buffer is a special type of op amp, an OP275, which functions on the Butler amplifier principle. In this, a combination of bipolar and junction field-effect transistors is used to provide the low noise of the first and the speed and sound quality of the second. This arrangement results in a device with impressive specifications as regards noise contribution, distortion and slew rate.

The amplification of the buffer is set to unity with R_{10} and R_{11}.

The bandwidth of the buffer is limited by C_9.

To eliminate the fairly high offset (that is, imbalance) voltage produced by the SSM2017 (which may be as high as 200 mV, depending on the amplification), and since an electrolytic capacitor at the output was considered undesirable, integrator IC_{2B} is used. This op amp compares the potential at pin 1 of the buffer amplifier with that at its inverting input (pin 6). On the basis of this comparison, the integrator adjusts the input to the buffer in such a way that the overall offset voltage at the output is smaller than 1 mV at all times. Diodes D_5 and D_6 protect the integrator against high peak voltages.

POWER SUPPLY

The circuit needs three different supply voltages: a symmetrical one of ±18 V for IC_1–IC_4; a single one of 22 V for the relay; and a single one of 48 V for the microphone(s).

The ±18 V is provided by two voltage regulators, IC_6 and IC_7. The supply lines to the various ICs are bypassed for any noise and interference on the mains supply by resistors R_{16} and R_{17} and capacitors C_{12}–C_{17} (or R_{33} and R_{34}, and C_{29}–C_{34} in case of the right-hand channel).

The supply line for the relay is taken directly from the secondaries of the mains transformer since this need not be smoothed or regulated. Its level of 22 V is sufficient to drive the relay reliably.

Since the output of the mains transformer is not sufficient for directly generating the 48 V line for the microphone(s), a cascade network, D_{16}-D_{17}-C_{41}-C_{42} is used to boost the voltage across the bridge rectifier to about 70 V (open-circuit). This is reduced to +48 V by the combination of regulator IC_5 and preset P_2.

Diode D_{18} functions as on/off indicator.

CONSTRUCTION

It is obvious that in view of the very low a.f. input signal levels it is desirable to isolate the amplifier from the power supply, and this is why the printed-circuit board in **Figure 2** consists of two parts that should be separated before any construction work is commenced. In the final assembly, the two boards should be kept apart by at least 20 cm (8 in).

The construction itself is fairly straightforward, but a few points need to be borne in mind. Switches S_1 and S_2, if used, may be connected via 4-pin SIL headers.

In the same way, volume control P_1 may be connected via a 6-pin SIL header. Again, not everyone may deem this control necessary. When the amplification, and therefore the value of R_9 (and R_{26} in the right-hand channel), has been determined, the control is not really required, since the volume is set on the mixer panel or power amplifier. To enhance the signal quality it is then better to omit P_1 and, to protect the outputs, connect a 100 Ω resistor between pins 1 and 2 and 4 and 5 of the SIL connector.

It is advisable to house the power supply section in a well-insulated plastic (*Acrylonitrile-Butadiene-Styrene* – ABS) case.

The mains cable should be provided with a strain-relief.

The amplifier section is best housed in a metal case. The case should be connected to one of the earth connections on the amplifier board.

Finally, to keep any induced noise and interference to an absolute minimum, intertwine the three wires of the ±18 V line and earth, and the two wires of the +48 V line and earth.

[970083]

video copy processor

For obvious reasons, the rights owners of Hollywood 'block-busters' and other films released on video cassettes are doing everything they can to prevent financial losses due to unauthorised copying of their products. Most of these (huge) companies, including Disney, employ an anti-copy process called Macrovision™. In this article we will endeavour to unravel the operation of this technology, as well as describe a pretty intelligent processor* which defeats the pulses and other purposely introduced interference that prevents VCR-to-VCR copying of Macrovision™-protected VHS/PAL video tapes.

main features

- *removes copy interference signals from picture syncs on prerecorded video cassette tapes*
- *enables legal copying (VCR back-to-back for backup purposes only)*
- *enhances picture quality in playback mode*
- *future-proof thanks to programmable EPLD*
- *low component count*
- *drop-in unit with SCART or Cinch connectors for easy connection*
- *designed for VHS/PAL VCRs*

According to Macrovision Corp., their video cassette anticopy process is the most effective and widely used technology for preventing 'back-to-back' copying of videos using two VCRs. The technology, they say, has been applied to over 1.5 billion video cassettes worldwide, and is used by virtually every major Hollywood

Design by W. Foede

** The Editors expressly dissociate from any illegal use of this circuit when the copyright vested in video cassettes is violated by removing the prerecorded Macrovision components. If you are the rightful owner of a prerecorded video tape, you are entitled to make one copy of it for backup purposes only. Macrovision™ is a registered trademark of the Macrovision Corporation, Sunnyvale, CA.*

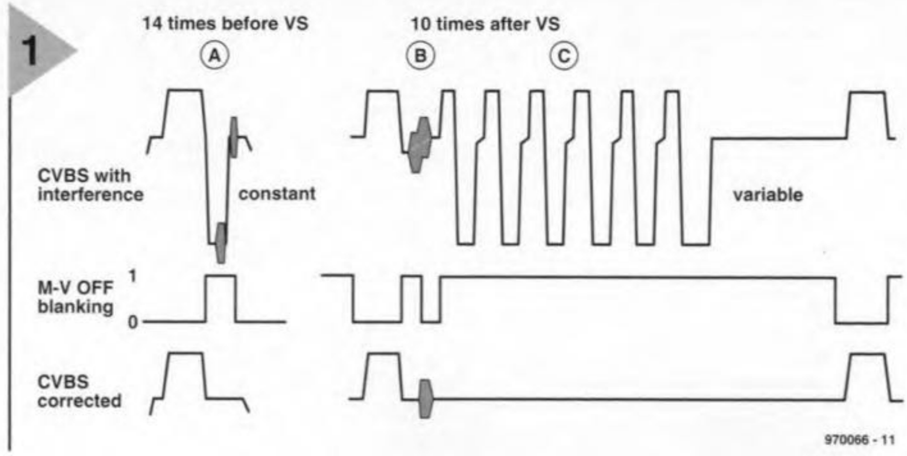

Figure 1. These drawings show the three types of interference (A, B and C) recorded on Macrovision-protected video tapes in order to make tape copying a waste of time. The Video Copy Processor described in this article detects all three types of interference, and kills it with the aid of a magic blanking pulse called M-V OFF. The drawings apply to VHS/PAL only.

studio, as well as by over 1,500 producers of special interest, corporate and educational programming.

Macrovision™ copy protection is available through 100 licensed duplicators throughout the world, including 40 licensed duplicators in the US and Canada.

Macrovision™ is applied to VHS tapes only, and seems to be more common in the US than in Europe. One of the best known videos protected by the system is the UK version of Disney's *Pocahontas*. Although the basic technology is the same, there exist slightly different NTSC and PAL versions of the system. **The processor described in this article was designed for use with VHS/PAL tapes only. It has not been tested with VHS/NTSC tapes.**

Like many other copy protection schemes and video encryption systems, Macrovision has been subject to a number of revisions aimed to keep the system compatible with evolving technology, in particular, of the video cassette recorder (VCR). Since an early (and very famous) publication in this magazine[1], the system has been improved considerably. The present processor has been designed to reflect all major changes made to the system up to the time of writing this article (August 1997).

What happens?

Suppose you copy a Macrovision protected tape to a second VCR. Because you are using a good quality tape in the recording VCR, and a fully-wired SCART-to-SCART link between the two machines, you expect the copy to be of reasonable quality. The disappointment comes when you play back the copy: the picture is useless because it is unstable. In some cases, there is colour flickering, or even a virtually dark picture. The sound may also disappear because the TV set detects an unintelligible picture and promptly mutes the audio circuitry. The Macrovision Corp. describes the effect as: 'copies made on most consumer VCRs are degraded to the extent they no longer have any entertainment value'. Neatly put, but obviously forgetting about the fun some electronics enthusiasts may have in defeating the system.

How is it possible that the TV set is not bothered by the Macrovision interference, and the recording VCR goes haywire? The answer is relatively simple. A TV set has a picture synchronisation circuit that employs a kind of flywheel to prevent small irregularities in the received signal upsetting the picture stability. Well, the purposely introduced Macrovision interference (more about it further on) is interpreted as a minor disturbance which most (but not all) TV sets have no problem in overcoming. The interference is definitely there, however. If you happen to have an older TV set available, try this: reduce the vertical picture height, or adjust the vertical sync so that the picture starts to roll slowly. With some luck, you may be able to see the Macrovision signals as bright vertical stripes near the top or bottom of the picture.

The recording signal circuitry in a VCR does not have a 'flywheel' function like the one implemented in a TV set. This function is not required and not appropriate because the recording method is based on FM (frequency modulation). All VCRs do, however, have an AGC (automatic gain control) whose operation is totally messed up by the interference bursts and other spurious pulses introduced by the Macrovision signal. In fact, the interference makes the VCR think that the picture is alternately far too bright and far too dark. In response to this fake information, the AGC reduces the video gain to virtually zero, which results in a darkened picture, and then pumps it again. The speed of the effect is such that the picture is 'murky' and unstable.

What happens? (gets technical)

For the following discussion it is assumed that you have a basic knowledge of the structure of a composite video signal.

The types of copy interference signal found on the latest pre-recorded video cassettes are shown in **Figure 1**. The basic thought is the same in all cases. All video recorders use AGC and black-level clamping for their recording circuits. These functions have their timing determined by the front edge of the line sync pulse (HS), and they start working on the rear porch (which starts well before the colour burst).

Interference type A

In a 'clean' signal, the instantaneous level of the rear porch is taken to represent the 'black' reference. Because it is not dependent on the picture (video) contents of the line, the level difference between the bottom of the sync pulse and the black level is an indication of the average video signal level. If the rear porch is pulled to the extreme white level as illustrated in Figure 1 (interference A), the process of establishing the video signal level is disturbed because a very large video signal is expected. As a result, the video gain is reduced to minimum. Because TV sets do not have a comparable gain control circuit in the video amplifier, the actual level is not reduced. None the less, the drive margin is changed considerably as a result of the 'fake' clamping level. As a result, the picture in the affected range becomes darker. A 'peak-white' burst (interference type A) is purposely inserted 14 times before each vertical sync pulse. The result is interference near the bottom of the picture.

The Video Copy Processor to be described further on detects the peak-white burst, and removes it from the rear porch.

Interference type B and C

As opposed to interference type A, types B and C are marked by levels which are varied periodically. This specific feature of the Macrovision system causes some parts of the picture to remain just about recognisable. The main interference is 'C' in the blanking period after the raster sync pulse, where the VideoText information nor-

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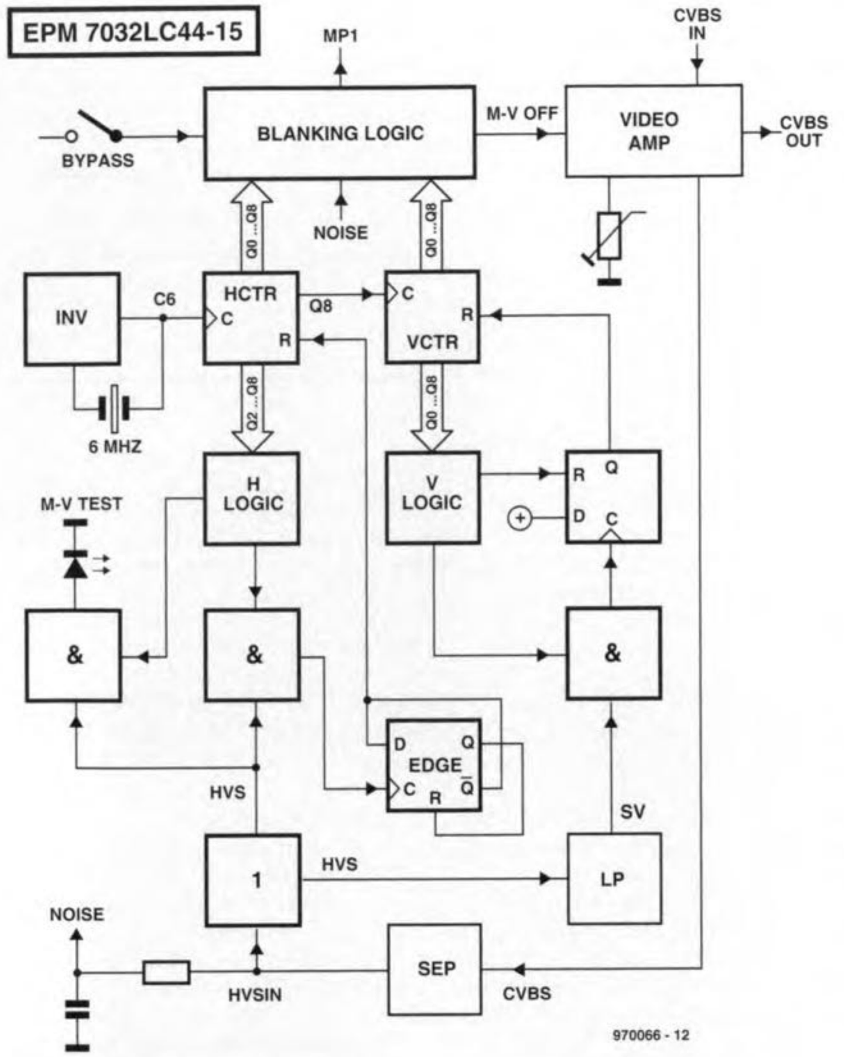

Figure 2. Combined block diagram of the circuit and the logic stashed away in the EPLD7032 chip.

mally resides. By inserting six new line sync pulses and six 'whitened' black references per line, the operation of the AGC circuits is totally disrupted, causing a muddled picture to be recorded: no brightness, no synchronisation! Because the muting circuit is activated, the sound channel with the film also disappears. The interference of type B is accurately matched to type C, so that TV sets show quasi-constant brightness in the upper range of the picture.

Older decoders remove the 'C' interference only, and forget about the 'A' and 'B' components. If only 'B' remains active, there will be a distinct flickering (rapid brightness variations) in the upper picture range. So, both have to be eliminated. Depending on the circuitry used in the TV set, interference type C may cause annoying horizontal picture tearing in the upper picture range when the tape is played back.

The present copy processor removes all three interference types from the composite video signal (CVBS) with the aid of its M-V OFF (Macrovision-Off) signal.

Circuit description

The heart of the circuit is an EPLD device type EPM7032 from Altera. This 'black box' contains a large number of digital functions which have been programmed into the chip. The IC is available ready-programmed from the Publishers (in combination with the PCB), or from kit suppliers. The main modules which are available in the EPLD, and their interaction, are shown in **Figure 2**. Mind you, not all of the blocks you see in the diagram are hidden inside the EPLD; some are external, tangible and of 100%-analogue design. Hence, our discussion of the operation of the circuit will cross-refer between the EPLD innards and the circuit diagram in **Figure 3**. An EPLD device was chosen here to reduce the component count and to defeat future changes to the Macrovision system (if and when they occur), simply by modifying the internal design of the chip.

Normally, an audio/video (AV) output supplies a positive CVBS (colour-video-blanking-synchronisation) signal with a level of about 1 V_{pp} into 75 Ω. So it is necessary for the video processor to provide a voltage gain of 6 dB (2×), and an output impedance of 75 Ω to ensure the same level is available at the output.

Transistors T4 and T3 form a complementary Darlington device whose gain depends on the ratio of (R3/R4) to R8. The polarity of the original signal is not changed by this amplifier. Using diode D1 the CVBS input signal is clamped at a level determined by voltage divider R6-P1-R5. Thanks to the preset, the circuit can be made to respond properly to input signals which are marginally out of spec (i.e., too weak or too strong). The result of the clamping operation is that the signal at the base of T4 has a stable bias level. By way of resistor R7, T4 can be disabled by the M-V OFF (Macrovision-Off) signal, which is digital and generated by the EPLD chip. The values of resistors R4, R3 and R7 are such that the 'off' level corresponds to 'black', while the amplified signal is not clipped at either side.

Emitter follower T2 acts as an output buffer. A simple level detector, T1,

Macrovision on the Internet

The Internet is a great source for further information on Macrovision. It's all there: technical, commercial, hearsay, hacker info, and totally uninformed stuff.

First of all, we should, of course, mention the site run by the Makers of the System. Surprise, surprise, this may be found at http://www.macrovision.com.

If you would like to participate in discussions about video-related subjects like the Macrovision anti-copy system, subscribe to the newsgroup rec.video. If you are new to this kind of communication, be sure to download, print and read the newsgroup's FAQ (frequently asked questions) before you start posting messages.

A separate, reasonably well informed FAQ (currently version 1.1) on Macrovision has been posted by Antti Paarlahti at http://www.cs.tut.fi. The particular strength of this FAQ lies in the sections which explain the operation of the Macrovision system in a technical as well as in a non-technical way. Interestingly, a number of technical references draw heavily on the 1988 Elektor Electronics Macrovision decoder/blanker.

extracts the horizontal and vertical sync pulses (HVS) from the CVBS signal. It also acts as an inverter, so that positive-going sync (S) pulses with a level of 5 V_{pp} are applied to pin 39 (HVSIN) of the EPLD chip.

Components R11 and C8 form a low-pass filter to block the colour burst and other high-frequency interference. By way of an edge detector, the leading edges of the S (sync) pulses start a 9-bit counter (HCTR) which is clocked at 6 MHz. This clock frequency is supplied by an oscillator inside the EPLD chip. The oscillator uses quartz crystal X1 as the frequency determining element. The programmed H-logic keeps the input blocked for about 63 μs to make sure that only the leading edge of the S pulse can start the counter at 64-μs intervals. Because the oscillator is a free-running type, the start instant may vary by 160 ns; not a problem here.

The HVS signal obtained from the sync separator is buffered and then applied to a two-stage low-pass filter, LP. In the actual circuit, the integrator is formed by external components C13-R20-C11-R19 which supply the vertical (raster) sync pulse, VS. Using a gate and the V-logic signal, VS starts another 9-bit counter. You guessed it, this one's called VCTR.

So far, so good. The complex analogue and digital circuits discussed so far, and in particular the HCTR and VCTR elements, allow any instant in the picture to be pinpointed. Using selected combinations of HCTR and VCTR output states, the module called 'blanking logic' determines when the Macrovision blanking pulses (M-V OFF) should be sent to T4 to eliminate the various types of interference. All bursts of type A which occur before VS are eliminated in this way. In this operation, the colour bursts in the 16 lines at the bottom of the picture are also lost. Fortunately, that is of little consequence because these lines are not normally visible. Starting at VS, the following picture components are blanked (i.e. pulled to black): the entire vertical blanking period, all 'B' and 'C' interference (*hurray*), and the contents of 8 picture lines before the VS instant, which we don't want anyway because additional 'rubbish' signals may be packed therein.

A double-pole change-over switch, S1, allows the processor to be made transparent. This is achieved by disabling the M-V OFF signal.

The power supply is conventional, being based around the familiar 7805 voltage regulator with a handful of 100-nF capacitors thrown in for decoupling at critical points.

The video copy processor may be connected using SCART cables or coax cables carrying the CVBS signal. More about this further on.

Socket K6 supplies trigger signals for your oscilloscope, and may be useful when it comes to analysing specific signals in the circuit or the input video signal.

Finally, there are two LED indicators in the circuit: one for the power supply (D7, red), and another which lights when a Macrovision signal is detected (D6, green).

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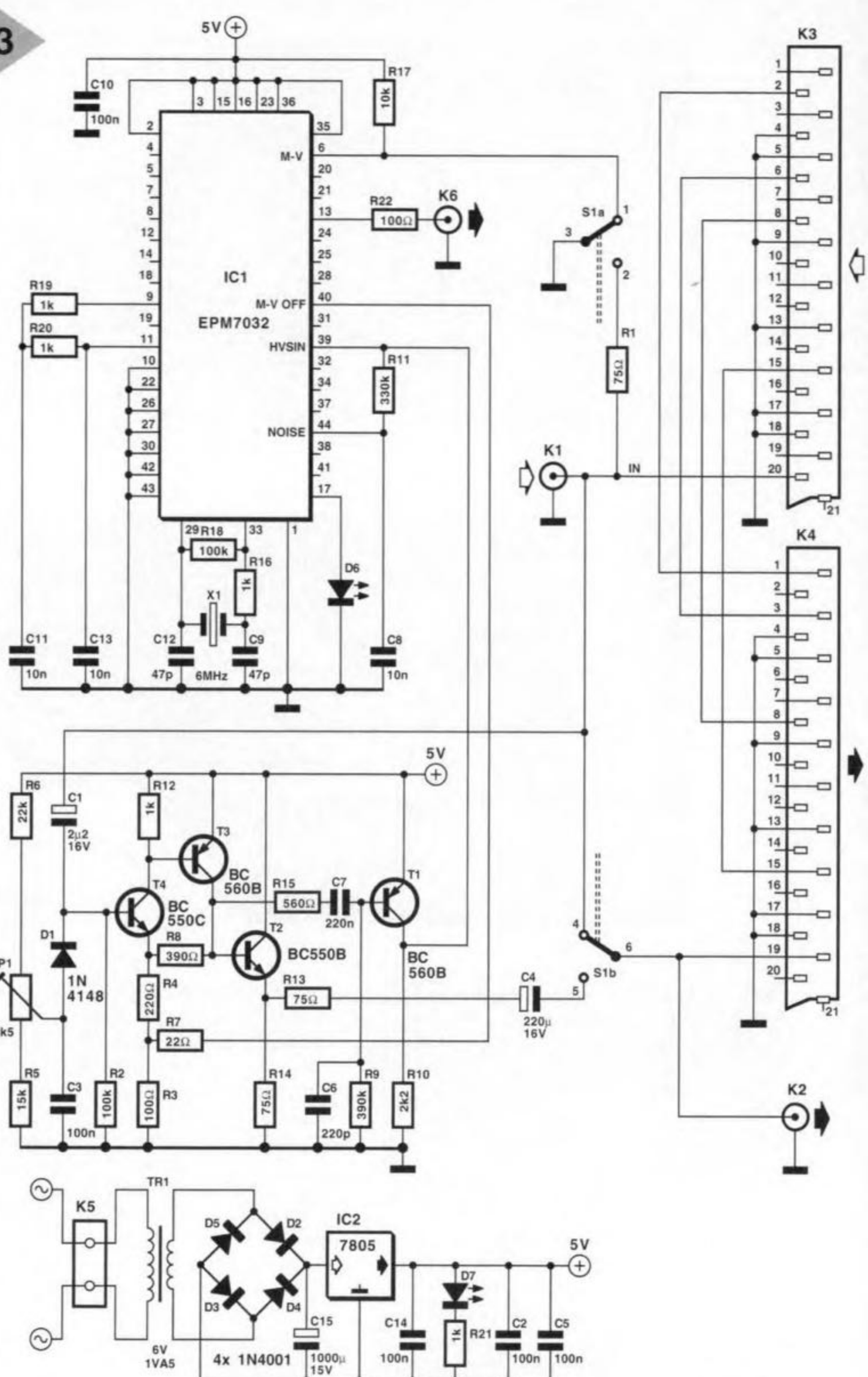

Figure 3. Circuit diagram of the Video Copy Processor for VHS/PAL. The heart of the circuit is a pre-programmed EPLD chip which combines a lot of logic functions. Sockets K3 and K4 may be omitted if you use cinch cable connections only (not recommended, though).

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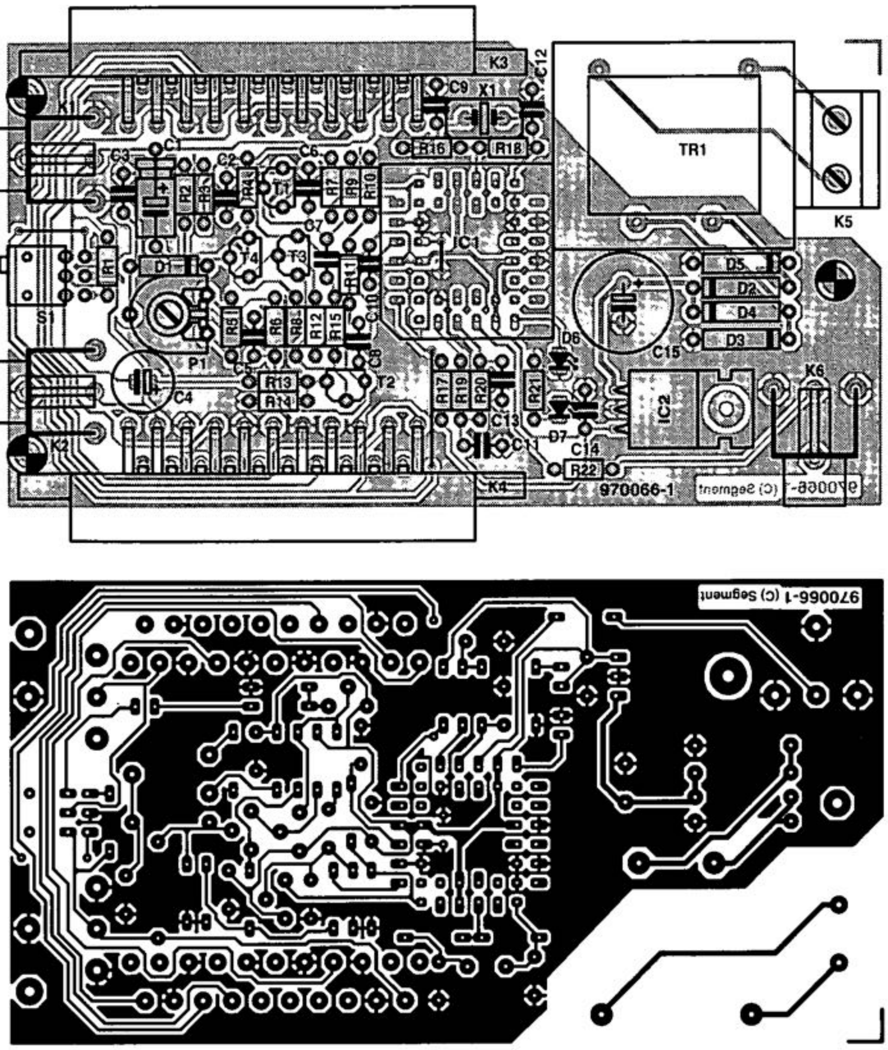

COMPONENTS LIST

Resistors:
R1,R13,R14 = 75Ω
R2,R18 = 100kΩ
R3,R22 = 100Ω
R4 = 220Ω
R5 = 15kΩ
R6 = 22kΩ
R7 = 22Ω
R8 = 390Ω
R9 = 390kΩ
R10 = 2kΩ2
R11 = 330kΩ
R12,R16,R19,R20,R21 = 1kΩ
R15 = 560Ω
R17 = 10kΩ
P1 = 2kΩ5 preset H

Capacitors:
C1 = 2μF2 16V
C2,C3,C5,C10,C14 = 100nF
C4 = 220μF 16V radial
C6 = 220pF
C7 = 220nF
C8,C11,C13 = 10nF
C9,C12 = 47pF
C15 = 1000μF 16V radial

Semiconductors:
D1 = 1N4148
D2-D5 = 1N4001
D6 = green LED, high efficiency
D7 = red LED, high efficiency
T1,T3 = BC560B
T2 = BC550B
T4 = BC550C
IC1 = EPM7032 z
IC2 = 7805

Miscellaneous:
K1,K2,K6 = cinch socket, angled, PCB mount
K3,K4 = SCART connector, angled, PCB mount
K5 = 2 way PCB terminal block, raster 7.5mm
S1 = double-pole changeover switch (PCB mount, angled, raster 2.54mm), e.g. Fujisoko AS2D, or Arrow D22-E subminiature, RS Components order code 149-276.
X1 = 6 MHz quartz crystal
TR1 = mains transformer, PCB mount, 6V 1.5VA (Monacor/Monarch VTR1106)
Case: type SD20 from Donau. Outside dimensions: approx. 125x70x40 mm.
PCB and programmed EPLD chip: order code 970066-C (see Readers Services page).

Figure 4. Copper track layout and component mounting plan of the printed circuit board designed for the Video Copy Processor. This board is available ready-made through our Readers Services, together with the programmed EPLD chip.

Construction

We've designed a very compact printed circuit board for the Video Copy Processor, and would advise anyone intending to build this circuit to use the ready-made board supplied through our Readers Services, or a kit supplier. For your reference, and for those of you who insist on making the board themselves, the artwork of the single-side printed circuit board is shown in **Figure 4**.

Mounting the parts specified in the components list is mostly plain sailing. There are three wire links on the board: one under the EPLD socket, one between switch S1 and connector K1, and between K4 and the EPLD socket. Before you do anything else, **locate and install these three wires**. Next, it is recommended to stick to this order when mounting the parts: resistors, capacitors, IC socket, semiconductors, connectors, transformer, crystal. The two LEDs should be mounted at a height that enables them to protrude from the top panel of the case. Watch the orientation of the *polarised* components. Yes, that includes the PLCC socket for the EPLD chip. Be warned, one error and you'll be spending hours on faultfinding, not even mentioning the cost of a new EPLD chip.

The dissipation of the 7805 on the board is quite low so that a heat-sink is not necessary. Use an M3 bolt, washer and nut to secure the regulator to the board.

For an exact fit, the size of the board and the position of the connectors demand the use of a type SD20 plastic case from Donau. Alas, there's no way you can avoid cutting, drilling and filing clearances for the SCART sockets, the LEDs, the cinch sockets, the decoder-bypass switch, and the mains input cord. Fortunately ABS plastic is an easy going material.

After a thorough visual check, the board is mounted in the case with the aid of three M3 screws (for which there are holes in the PCB), and 10-mm long PCB standoffs which are secured to the back plate of the case.

As a finishing touch, stick four rubber feet on the back panel to prevent scratches on your furniture.

Setting-up, connecting-up

Apart from preset P1, there are no adjustments in this circuit. Initially, the preset is set to mid-travel. If necessary, adjust it if the output level supplied by the 'playing' VCR is on the high or low side.

The Video Copy Processor is inserted in the AV (audio/video) link

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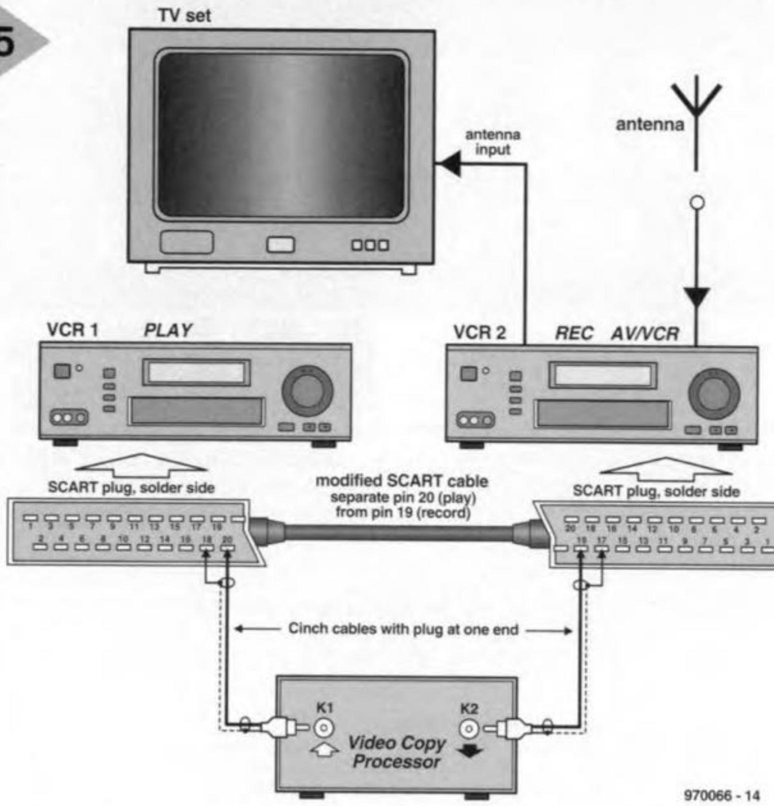

Figure 5. Use this alternative cabling scheme if for some reason you don't want to use two SCART cables to connect the Video Copy Processor between the two VCRs. The alternative consists of one SCART cable whose plugs are opened and modified to break one of the CVBS links. In this way you enable the CVBS signal to be passed through the Video Copy Processor using two coax cables fitted with cinch plugs. Each SCART plug on the cable so modified has to be marked to make sure it goes to the right VCR ('play' or 'record').

(cinch, BNC, DIN or SCART) between a **playing** and a **recording** VCR. Normally, two SCART cables will be used: one from the playing VCR to the input of the copy processor, and another, from the copy processor output to the recording VCR. This is the easiest solution, as the Copy Processor has two SCART sockets, K3 (input) and K4 (output) available for this purpose.

If you don't want to invest in two SCART cables, you may also make do with one by using the alternative connection scheme illustrated in **Figure 5**. You have to open up the two SCART plugs, though, and do some soldering inside. It is also necessary to mark the SCART plugs which go the different VCRs to prevent the cable being connected the wrong way around. Basically, the alternative connection boils down to breaking one of the two CVBS wires (pins 19 and 20), and giving it an alternative route through the Video Copy Processor by way of short pieces of coax cable terminated with cinch plugs.

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Reference:

1. Macrovision decoder/blanker, *Elektor Electronics* October 1988.

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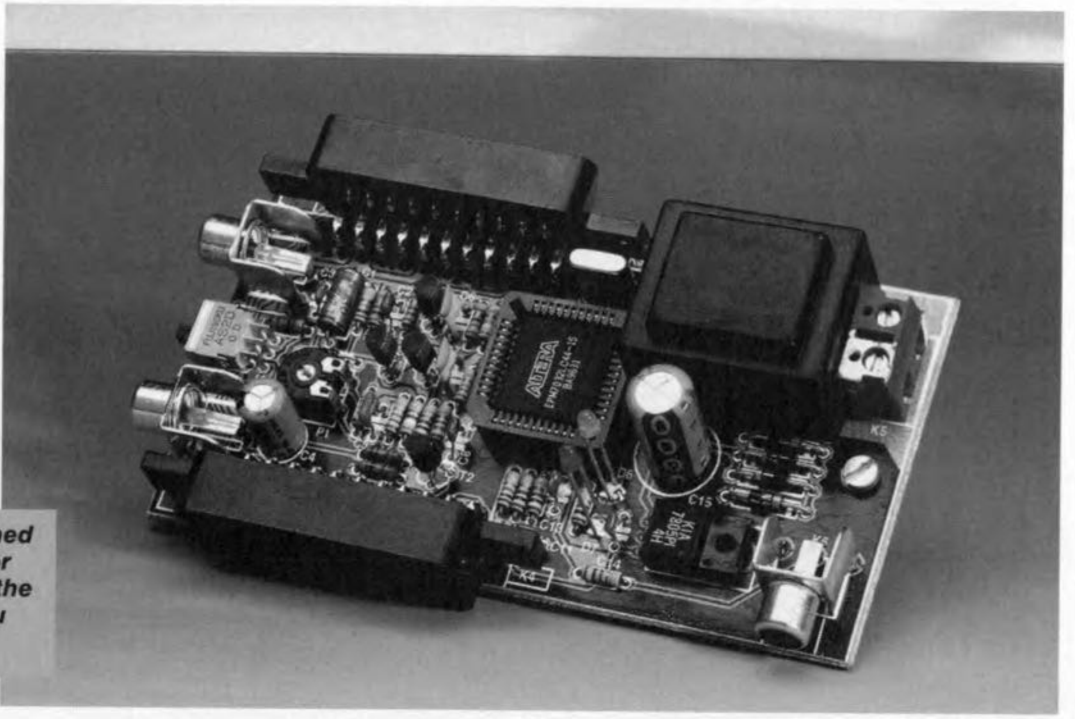


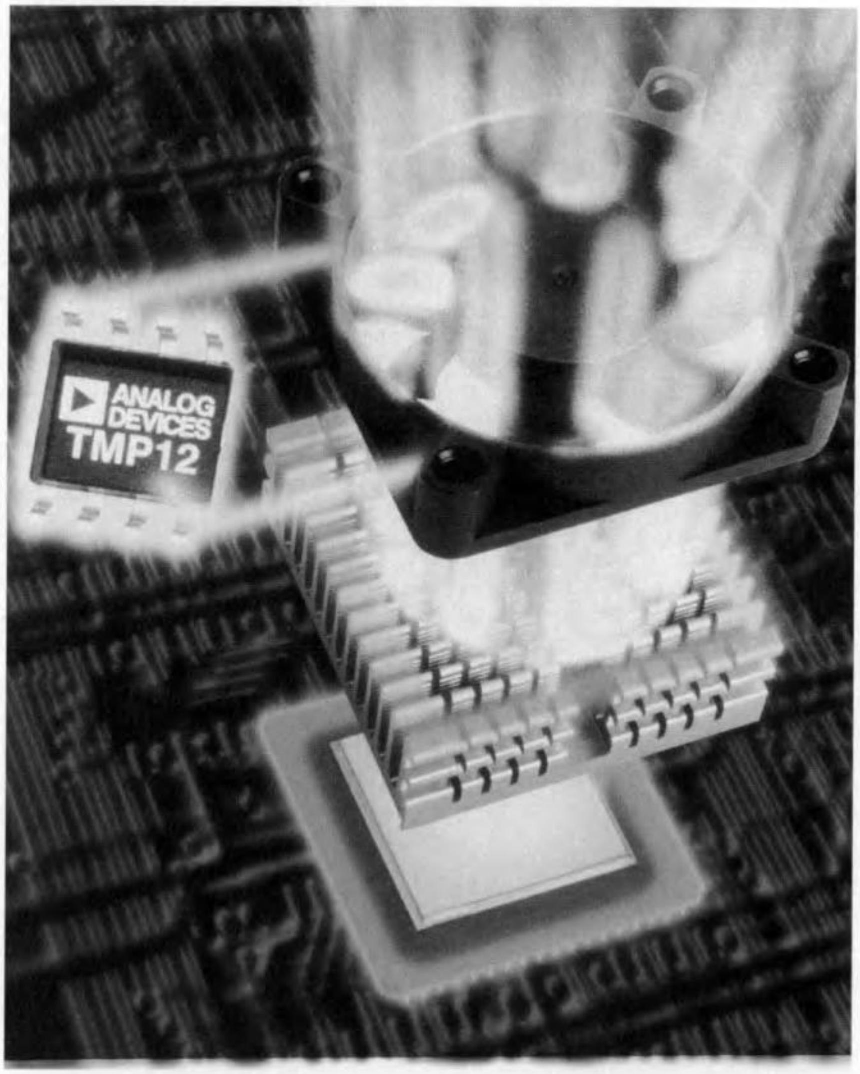
Figure 6. Finished board, ready for mounting into the compact Donau case.

APPLICATION NOTE

airflow and temperature sensor Type TMP12

effective monitoring of cooling systems

The TMP12 is a silicon-based airflow and temperature sensor designed to be placed in the same airstream as heat generating components that require cooling. Fan cooling may be required continuously, or during peak power demands, e.g., for a power supply, and if the cooling system fails, system reliability and/or safety may be impaired. By monitoring temperature while emulating a power IC, the TMP12 can provide a warning of cooling system failure.



The TMP generates an internal voltage that is linearly proportional to Celsius temperature, nominally +5 mV °C^{-1}. The linearized output is compared with voltages from an external resistive divider – R_1–R_3 in **Figure 1** – connected to the TMP's 2.5 V precision reference. The divider sets up one or two reference voltages, as required by the user, providing one or two temperature setpoints. Comparator outputs are open-collector transistors able to sink over 20 mA.

There is an on-board hysteresis generator provided to speed up the temperature-setpoint output transitions, which also reduces erratic output transitions in noisy environments. Hysteresis is programmed by the external resistor chain and is determined by the total current drawn from the 2.5 V reference.

An Analog Devices Application

Table 1. Specifications of the TMP12.

(V_S = +5 V, − 40 °C ≤ T_A ≤ +125 °C, sauf mention contraire)

Parameter	Symbol	Conditions	Min.	Typical	Max.	Units
ACCURACY						
Accuracy (high, low setpoints)		T_A = +25 °C		±2	±3	°C
		T_A = −40 °C to +100 °C		±3	±5	°C
Internal scale factor		T_A = −40 °C to +100 °C	+4.9	+5	±3.1	mV/°C
Power supply rejection ratio	PSRR	4.5 V ≤ +V_S ≤ 5.5 V		0.1	0.5	°C/V
Linearity		T_A = −40 °C to +125 °C		0.5		°C
Repeatability		T_A = −40 °C to +125 °C		0.3		°C
Long term stability		T_A = +125 °C for 1000 hrs		0.3		°C
SETPOINT INPUTS						
Offset voltage	V_{OS}			0.25		mV
Output voltage drift	TCV_{OS}			3		µV/°C
Input bias current current	I_B			25	100	nA
VREF OUTPUT						
Output voltage	VREF	T_A = +25 °C, no load	2.49	2.50	2.51	V
	VREF	T_A = −40 °C to +100 °C, no load		2.5±0.015		V
Output drift	TCV_{REF}			−10		ppm/°C
Output current, zero hysteresis	I_{VREF}			7		µA
Hysteresis current scale factor	SF_{HYS}			5		µA/°C
OPEN-COLLECTOR OUTPUTS						
Output low voltage	V_{OL}	I_{SINK} = 1.6 mA		0.25	0.4	V
		I_{SINK} = 20 mA		0.6		V
Output leakage current	I_{OH}	V_S = 12 V		1	100	µA
HEATER						
Resistance	R_H	T_A = +25 °C	97	100	103	Ω
Temperature coefficient		T_A = −40 °C to +125 °C		100		ppm/°C
Max. continuous current	I_H	Guaranteed but not tested			60	mA
POWER SUPPLY						
Supply range	+V_S		4.5		5.5	V
Supply current	I_{SY}	Unloaded at +5 V		400	600	µA
		Unloaded at +12 V		450		

The TMP12 airflow sensor also incorporates a precision, low-temperature coefficient 100 Ω heater resistor that may be connected directly to an external 5 V supply. When the heater is actuated, it raises the die temperature in the DIP package to about 20 °C above ambient (in still air).

FUNCTIONAL DESCRIPTION

The TMP12 temperature sensor section consists of a bandgap voltage reference which provides both a constant 2.5 V output and a voltage that is proportional to absolute temperature (VPTAT). The VPTAT has a precise temperature coefficient of 5 mV °C−1, and is 1.49 V (nominal) at +25 °C. The comparators liken VPTAT to the externally set temperature trip points and generate an open-collector output signal when one of their respective thresholds has been exceeded.

The heat source for the TMP12 is

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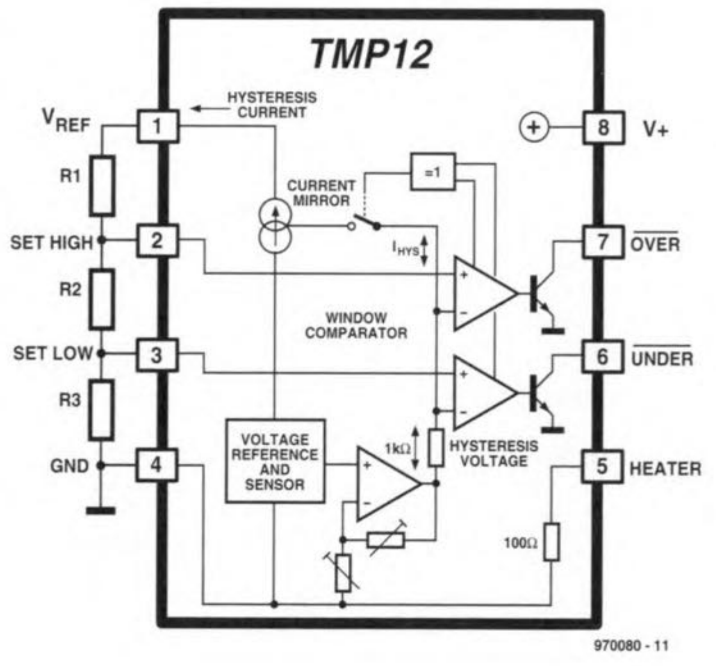

Figure 1. The main elements of the TMP12 are a temperature-dependent voltage source, two comparators, and a heater resistor.

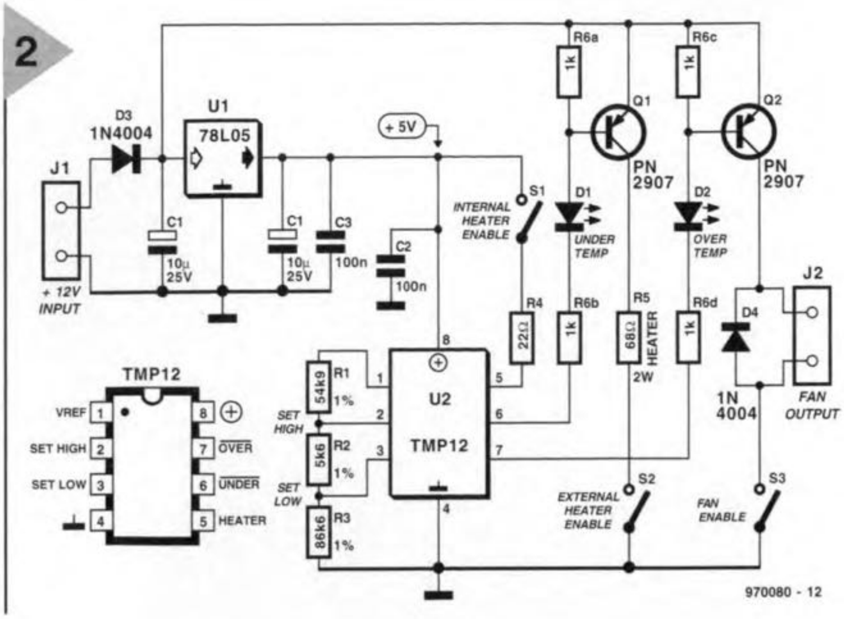

Figure 2. Typical basic application of the TMP12 (called Evaluation Board by Analog Devices).

an on-chip 100 Ω low temperature coefficient thin-film resistor. When this is connected to a 5 V source, it dissipates

$$P_D = V^2/R = 5^2/100 = 0.25 \text{ W},$$

which generates a temperature rise of about 32 °C in still air and about 22 °C in an airflow of 2.3 m s^{-1}. By selecting a temperature setpoint between these two values, the TMP12 can provide a logic-level indication of problems in the cooling system.

A proprietary, low temperature coefficient thin-film resistor, in conjunction with production laser trimming, enables the TMP12 to provide a temperature accuracy of ±3 °C (typical) over the rated temperature range.

The open-collector outputs are capable of sinking 20 mA, allowing the TMP12 to drive small control relays directly.

Operating from a single +5 V supply, the quiescent current is only 600 µA (max), without the heater resistor current.

PROGRAMMING THE TMP12

The basic thermal monitoring application requires only a simple three-resistor ladder voltage divider to set the high and low setpoints and the hysteresis. These resistors are programmed in the following sequence.

1. Select the desired hysteresis temperature.
2. Calculate the hysteresis current, I_{VREF}.
3. Select the desired setpoint temperatures.
4. Calculate the individual resistor divider ladder values needed to develop the desired comparator setpoint voltages at the Set High and Set Low inputs.

The hysteresis current is calculated from

$$I_{HYST} = I_{VREF} = 5\,\mu\text{A}\,°\text{C}^{-1} + 7\,\mu\text{A}.$$

For example, to produce 2 °C of hysteresis, I_{VREF} should be set to 17 µA.

The setpoint voltages, V_{SET} are calculated from

$$V_{SET} = (T_{SET} + 273.15)(5 \text{ mV } °\text{C}^{-1}).$$

This equation is used for both $V_{SETHIGH}$ and V_{SETLOW}. A simple three-resistor network, as shown in Figure 1, determines the setpoints and hysteresis value. The equations to calculate the resistors (in kΩ) are:

$$R_1 = (V_{REF}-V_{SETHIGH})/I_{VREF};$$

$$R_2 = (V_{SETHIGH}-V_{SETLOW})/I_{VREF};;$$

$$R_3 = V_{SETLOW}/I_{VREF}.$$

SELECTING SETPOINTS

Choosing the temperature setpoints for a given system is an empirical process, because of the wide variety of thermal issues in any practical design. The specific setpoints are dependent on factors such as airflow velocity in the system, adjacent component location and size, PCB thickness, location of copper ground planes, and thermal limits of the system.

TYPICAL APPLICATION

The circuit in **Figure 2** is a typical application of the TMP12 (it is called 'Evaluation Board' by Analog Devices). In this, the IC needs a single 12 V supply for operation, which is, however, required only for the fan and under-temperature heater. The 5 V line required by the TMP12 is produced by an on-board regulator, U_1.

Three switches are provided to (1) connect the internal heater; (2) disconnect the under-temperature heater; (3) disconnect the on-card fan.

The setpoint temperatures provided by the (unmodified) circuit are ≥ 40 °C (TMP12 junction temperature) for cooling fan ONm and ≤ 20 °C for heater ON. Thermal hysteresis is set to 2 °C. Different setpoint and hysteresis values may be set by changing the values of R_1–R_3 using the equations given earlier in this article.

The preset trip temperatures will cause the fan to cycle ON and OFF (in typical laboratory temperatures) with the TMP12's internal heater enabled.

Resistor R_4 is provided to adjust the internal power dissipation of the TMP12; this resistor should be removed when the pin-compatible low-power, programmable version of the IC, TMP01, is used.

The low-setpoint or heater function is used primarily for the TMP01 in heating/cooling control applications. However, the low-setpoint function is also included on the TMP12 airflow sensor as the function is easily changed to a second high-setpoint by inverting the logic output polarity (using an external gate). In that way, the TMP12 can provide actuation for primary and secondary cooling fans, or an auxiliary fan control plus over-temperature warning.

[970080]

radar-controlled door opener

This article provides suggestions and other useful data to help you design a door or lock opener system based on a ready-made X-band (10-GHz) radar module. Movement of any object or person within the radar's range is detected using the Doppler effect and a simple evaluating circuit which controls a relay.

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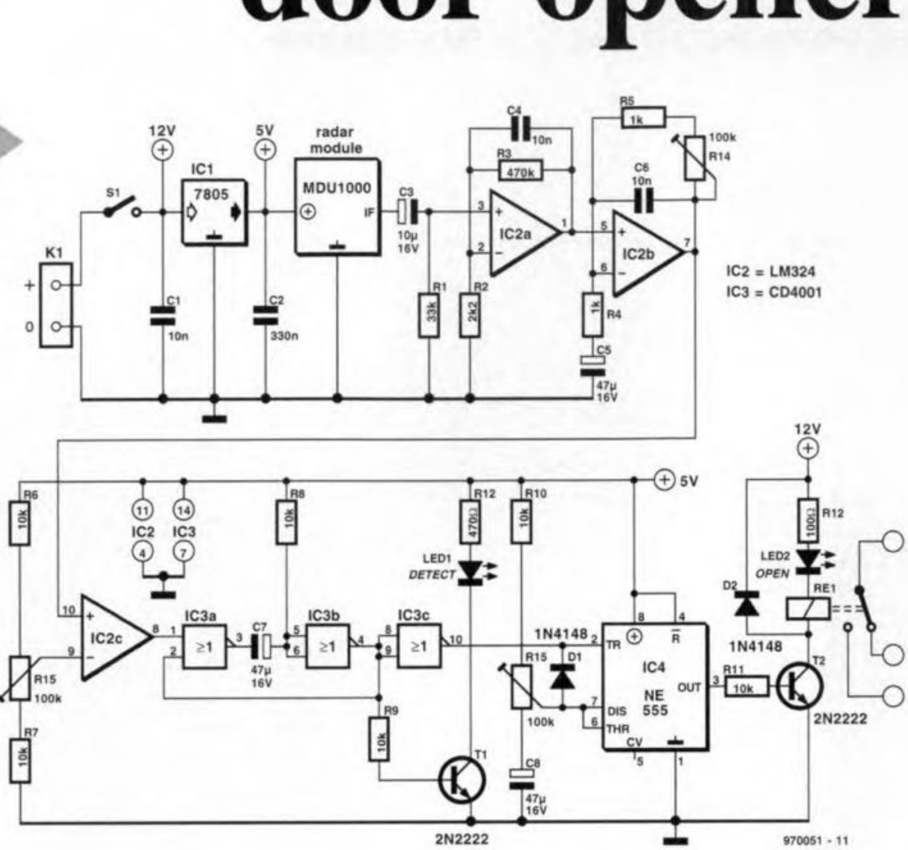

Figure 1. Circuit diagram of the radar-controlled door opener.

Principle of operation

The microwave detector which provides the switching information for the door opener is a volumetric radar designed to detect persons or objects within a specified area. On detecting anything that moves within its range, the system energizes a relay which, in turn, opens a door or provides access in another way.

The radar module used here is a Doppler type operating in the X-band (3cm, 10-GHz). In theory, the unit is capable of measuring object velocity between 50m/h (approx. 14cm/s) and 500km/h. Electronic evaluation of the (Doppler-shift) output frequency supplied by the radar module enables the system to reliably detect walking or running persons, or moving objects (like a car on the garage driveway).

In terms of sensitivity and reliability, the performance of this type of detector is superior over that of PID (passive infra-red) detectors because it is not in the least affected by fog, rain, darkness or intense sunlight. Also, thanks to the fact that microwaves traverse many materials like (plain) glass, wood and ceramics, without appreciable attenuation, the unit can be mounted out of sight, which is not normally possible with PIDs.

Circuit description

The complete circuit of the radar-controlled door opener is given in **Figure 1**. The heart of the circuit is the microwave radar module Type MDU1000. The Doppler-shift signal supplied by its IF (intermediate frequency) is first filtered by R1 and C3 which help to eliminate low-frequency components. Next, the signal is amplified by two opamps with respective gains of 210× (IC2a) and 1-100× (IC2b). The latter gain is adjusted with preset R14, and determines the detection range of the system. In the opamp amplifiers, capacitors C4 and C6 provide some bandwidth limiting by passing high-frequency components. An electrolytic capacitor, C5, in the second opamp stage serves to decouple the d.c. component.

The detection sensitivity of the system is adjusted with preset R15 which controls the bias at the – input of the third opamp in the circuit, IC2c, which is wired as a comparator.

If a significant Doppler shift occurs, the output of IC2c swing high and enables a monostable multivibrator (MMV) built from two of the four gates in IC3 to generate a 'person detected' pulse with a length of about 0.5 s. The delay, T, is expressed as

$$T = 1/(R8 \cdot C7) \quad [s]$$

The resulting control pulse at the output of gate IC3b causes driver transistor T1 to be switched on, so that the green 'detect' indicator LED1, lights.

The same pulse is inverted by IC3c, triggering a second (this time re-triggerable) MMV, IC4. The good old NE555 used in this position controls the 'on' time of relay Re1 via driver transistor T2. This time may be set to any value between 0.5 s (not very useful, we reckon, except of course to let the RoadRunner in, or a *very* fast cat) and about 5 s. The relay contact is used to switch on the door opener motor. A red indicator, LED2, lights when the motor is switched on.

The signal processing electronics

Design by S. Parizet

are powered from a regulated 5-volt rail provided by a 7805 voltage regulator. The relay is powered by the unregulated 12-V voltage.

The total power consumption of the system is about 80 mA.

Description of the radar module

The MDU1000 supplies a transmit power of about 13 decibel-milliwatt (dBm), or 20 mW. The IEEE C95.1-1991 standard specifies a maximum exposure level of 7 mW/cm^2 at a frequency of 10 GHz. At a distance of 1 m from the face of the MDU1000 radar module, the radiation level will about 0.72 μW/cm^2, or about 10,000 times lower than the norm.

The distribution of the transmitter power in the horizontal (H) and vertical (E, elevation) planes is shown in **Figure 2**. Note that the indicated power levels are relative, i.e. 0 dB in the graph corresponds to the maximum transmitter power of +13 dBm. The side lobes in the H-plane are normal for a microwave exciter designed for wide radiation in the X-band, i.e, without a WG16 waveguide flange. The –3-dB (half-power) angles are 36° and 72° for the H and E plane, respectively.

The detection threshold (sensitivity) of the receiver inside the MDU1000 is specified at –86 dBm by the manufacturer. This threshold (specified in dBm relative to an isotropic antenna) is shown as the straight line in **Figure 3**. Theoretically, a sensitivity of –86 dBm gives a radar range of 20 m. In practice, however, this figure will be derated to 15 m or so.

As shown in the mechanical drawing in **Figure 4**, the MDU1000 radar module is pretty flat and compact. There's nothing to adjust on the module, which has only three connections, IF out, +5 V and ground.

Final notes

According to the author, the MDU1000 Doppler radar module, as well as a complete kit of the system described here, are available from Saphir SARL, 9 Rue des Myosotis, F-94320 Thiais, France. Tel./fax (+33) 1.48.52.78.92.

A possible replacement for the MDU1000 is a microwave sensor from Conrad Electronics, order code 1081 11-99. Note that this replacement was not tested by the author.

Warning.
The installation and use of low-power X-band microwave radar systems for person detection like the one described here may be subject to radio and/or health regulations applicable in your country. In the UK, contact the DTI for further information.

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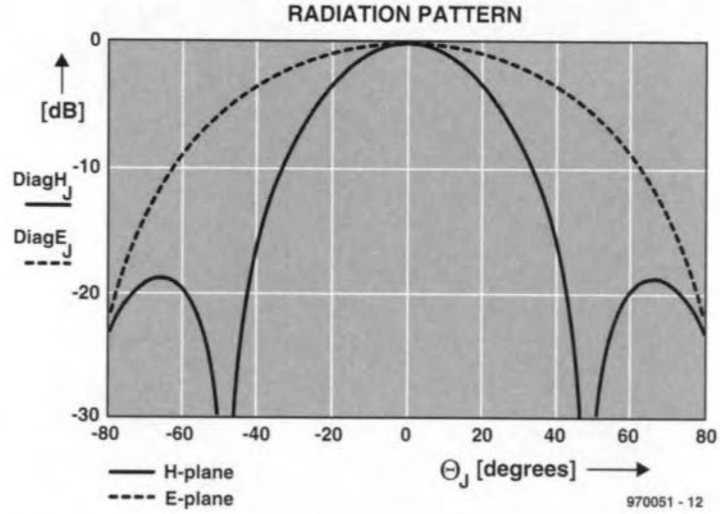

Figure 2. Radiation patterns of the MDU1000 in the horizontal (H) and elevation (E) planes.

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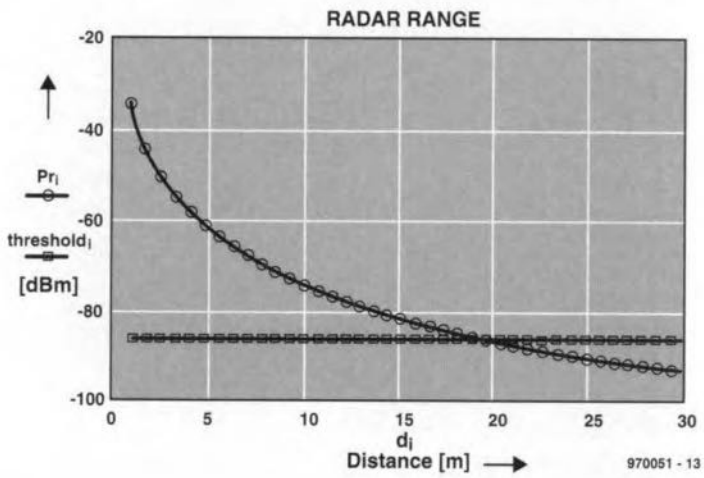

Figure 3. At a sensitivity of –86 dBm, the radar module has a theoretical range of 20 m.

Figure 4. Mechanical outline of the MDU1000 X-band radar module.

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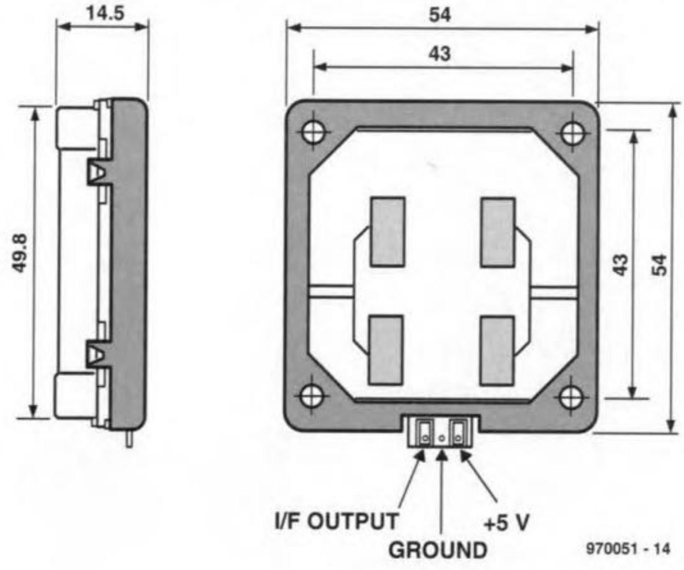
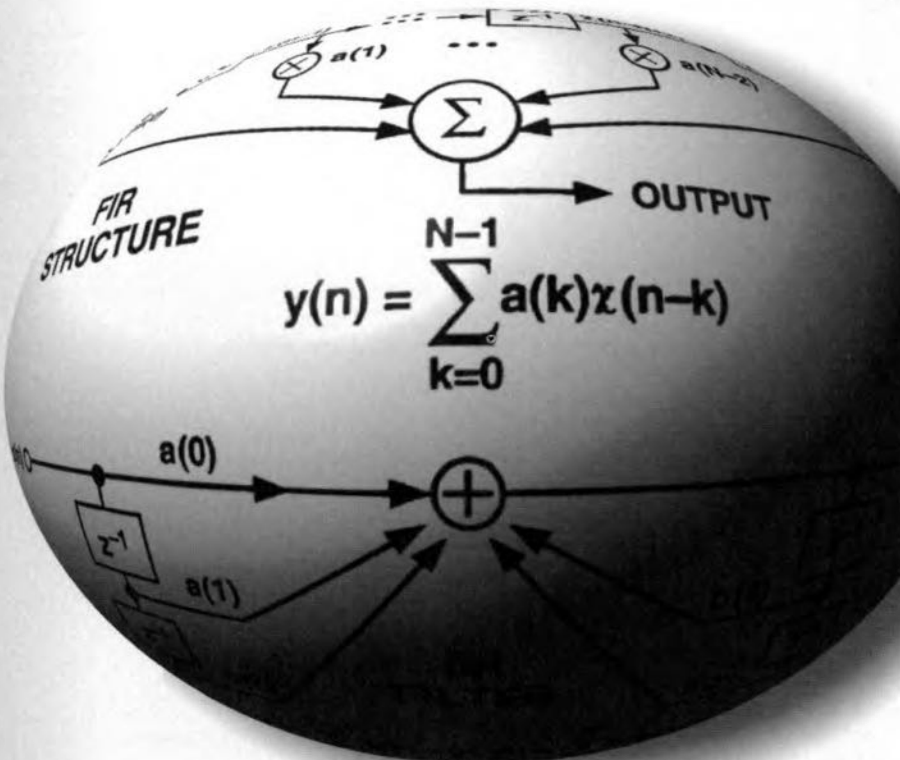


digital signal processing



Digital signal processors (DSPs) are used in many modern electronic systems to process analogue signals digitally. In essence, the hardware, software, and instruction sets of these devices are optimized for high-speed numeric processing tasks. In this article*, the design considerations for developing a DSP system are considered.

By our editorial staff

*This article is based on parts of 'Why use DSP' by D Skolnick and N Levne which originally appeared in *Analog Dialogue*, Vol. 31, Number 1, 1997, published by Analog Devices.

INTRODUCTION

To get an idea of the functioning and possible applications of a digital signal processor – DSP – it is instructive to compare an analogue and a digital circuit on the basis of a filter function.

An analogue filter is built from resistors, capacitors, inductors, and possibly active elements such as an operational amplifier – op amp. Such a filter is inexpensive and easily assem-

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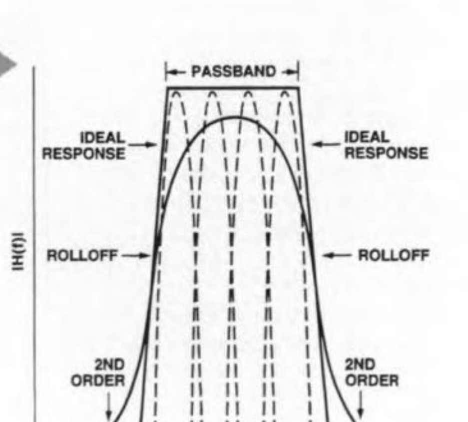

Figure 1. An ideal bandpass filter and second-order approximations.

bled but difficult to calibrate, modify, and maintain, and this difficulty increases exponentially with filter order. The use of a digital signal processor is therefore advisable in many instances. Such a processor enables a virtual filter to be effected by means of software. Such a filter is more flexible, easily repeated and more reliable than an analogue one. A further advantage is that if a higher order is desired, this affects the software only. In contrast to discretely built filters, the hardware remains unaltered.

As a practical example, consider an ideal bandpass filter. Such a filter has a frequency vs amplitude characteristic as shown in **Figure 1**. The response within the passband would be flat with zero phase shift, and in the stopbands there would be infinite attenuation. Useful additions would include passband tuning, bandwidth control, and stopband rolloff control.

It is clear from the characteristic that an analogue design would require a number of staggered 2nd-order high-Q sections. The tuning and adjusting of these would present great difficulties.

FILTER WITH DSP

With digital signal processing software, there are two basic approaches to filter design: the finite impulse response (FIR) and the infinite impulse response (IIR). The functional layouts of the two types of filter are shown in **Figure 2**.

The FIR filter's time response to an impulse is the straightforward weighted sum of the input signal and a finite number of previous samples. Since there is no feedback, the filter's response to a given sample ends when the sample reaches the 'end of the line'. An FIR filter's frequency response has no poles, only zeros.

By comparison, an IIR filter is called infinite because it is a recursive function, that is, its output is a weighted sum of the inputs and outputs. Since it is recursive, its response can continue indefinitely. An IIR filter's frequency response has both poles and zeros.

In Figure 2, x represents the input samples, y the output samples, a the input sample weighting, b the output sample weighting, n the present sample time, and M and N the number of samples programmed, that is the filter's order. Note that the arithmetic operations indicated for both types are simply sums and products. In fact, multiply-and-add is the case for many DSP algorithms that represent mathematical operations of great sophistication and complexity.

Approximating an ideal filter consists of applying a transfer function with appropriate coefficients and a

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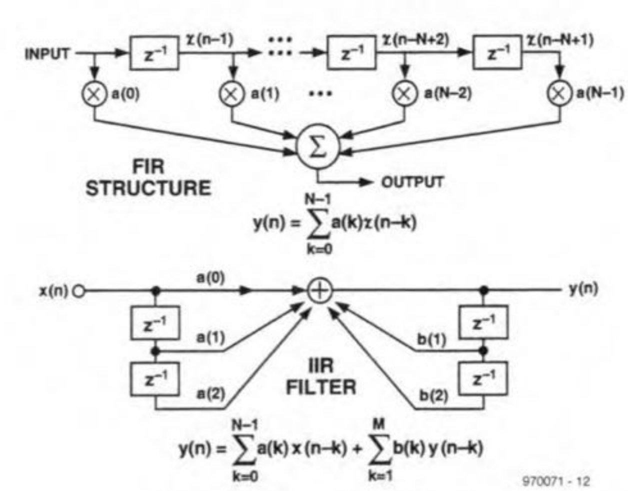

Figure 2. Filter (both FIR and IIR) equations and delay-line representation.

high enough order, or number of taps (considering the train of input samples as a tapped delay line).

Figure 3 shows the response of a 90-tap FIR filter compared with sharp-cutoff Chebyshev filters of various orders: 2, 4 and 6. The 90-tap example suggests how closely the filter can approach the ideal. Within a DSP system, programming a 90-tap FIR filter, like that in Figure 3, is not difficult. By comparison, attempting this level of approximation would be a very tedious affair. A further bonus of the use of a DSP to approximate the ideal filter is long-term stability.

REAL-WORLD SIGNALS

Real-world signals are analogue – the continuously changing energy levels of physical processes like sound, heat, light, electricity, magnetism. A transducer converts these levels into manageable electrical voltage and current signals, and an analogue-to-digital converter (ADC) samples and converts these signals to digital bits for processing. The conversion rate, or sampling frequency, of the ADC is critically important in the digital processing of real-world signals.

The sampling rate is determined by the amount of signal information that is needed for processing the signals adequately. In order for an ADC to provide enough samples to accurately describe the real-world signal, the sampling rate must be at least twice the highest-frequency component of the analogue signal. For example, to accurately describe an audio signal containing frequencies up to 20 kHz, the sam-

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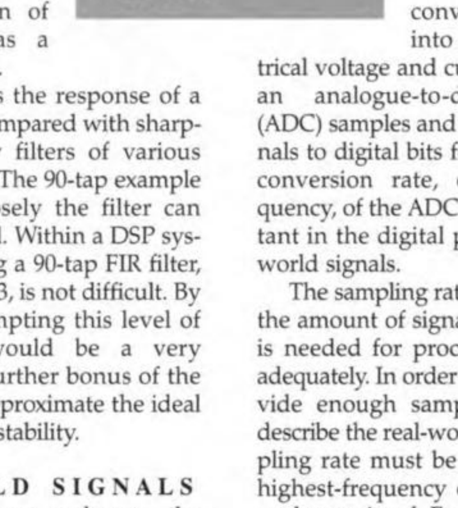

Figure 3. Filter response of sharp cutoff Chebyshev filters compared with that of a 90-tap FIR filter.

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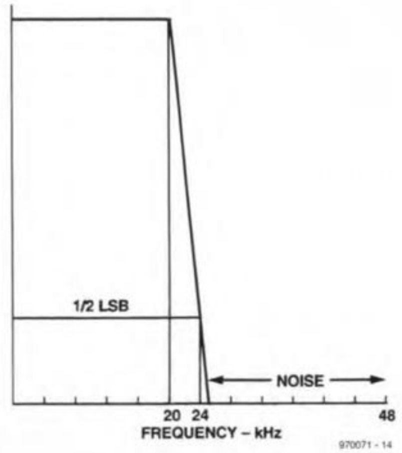

Figure 4. Response of an ideal anti-aliasing filter.

pling rate must be not less than 40 kHz. Since arriving signals can easily contain component frequencies above 20 kHz (including noise), they must be removed before sampling by feeding the signal through a low-pass filter ahead of the ADC. Since this filter, known as an anti-aliasing filter, has a finite frequency rolloff, additional bandwidth, say 2–4 kHz, must be allowed for the filter's transition band. This means that in practice the sampling rate must be higher by this bandwidth than the theoretical double value.

Figure 5. (a) Example of continuous processing of samples in a digital filter; (b) Example of batch processing of a block of data.

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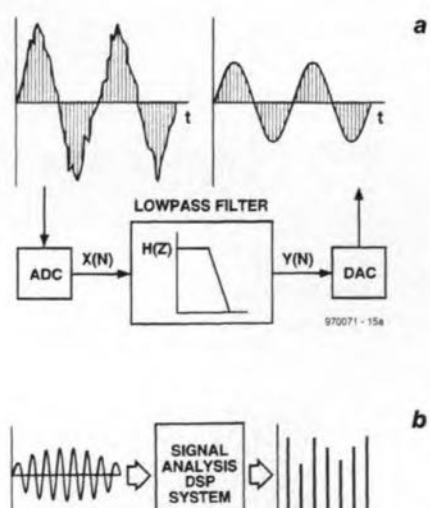

Figure 4 depicts the filter response needed to reject any signals with frequencies above half of a 48 kHz sampling rate. Rejection means attenuation to less than 1/2 least-significant bit (LSB) of the ADC's resolution. The curve in Figure 4 reaches this level at about 24 kHz. One way to achieve this level of rejection without a highly sophisticated analogue filter is to use an oversampling converter, such as a sigma-delta ADC. This typically obtains low-resolution (e.g., 1 bit) samples at megahertz rates – much faster than twice the highest frequency component – greatly easing the requirement for the analogue filter ahead of the converter. An internal digital filter restores the requisite resolution and frequency response.

PROCESSING REAL-WORLD SIGNALS

Once the system bandwidth requirements and the sampling rate have been established, a suitable DSP must be chosen. Since there are a number of these devices on the market, what are the necessary parameters?

Processing speed at a required sampling rate is influenced by algorithm complexity. As a rule, a DSP needs to finish all operations relating to the first sample before receiving the second sample. For instance, a 48 kHz sampling rate corresponds to a 20.833 μs sampling interval.

Next, consider the relation between the speed of the DSP and the complexity of the algorithm (that is, the software containing the transform or other set of numeric operations). Complex algorithms require more processing tasks. Because the time between samples is fixed, higher complexity calls for faster processing.

For example, suppose that the algorithm requires 50 processing operations to be performed between samples. Assuming a sampling rate of 48 kHz (sampling interval 20.833 μs), the minimum required speed of the DSP in millions of operations per second (MOPS) is $50/20.833 = 2.4$ MOPS.

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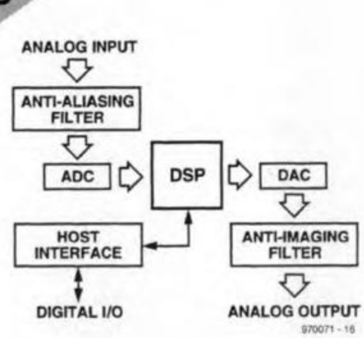

Figure 6. Block diagram of a DSP system. Its simplicity is due largely to the functionality of the programmable DSP.

Note that the two common ratings for DSPs, based on millions of operations per second (MOPS) and millions of instructions per second (MIPS), are not the same, since an operation may consist of several instructions.

SAMPLING REAL-WORLD SIGNALS

There are two basic ways to acquire data: one sample at a time or one frame at a time (continuous processing vs batch processing). Sample-based systems, like a digital filter, acquire data one sample at a time. As shown in **Figure 5a**, at each tick of the clock a sample comes into the system and a processed sample is output.

Frame-based systems, like a spectrum analyser, which determines the frequency components of a time-varying waveform, acquire a frame, or

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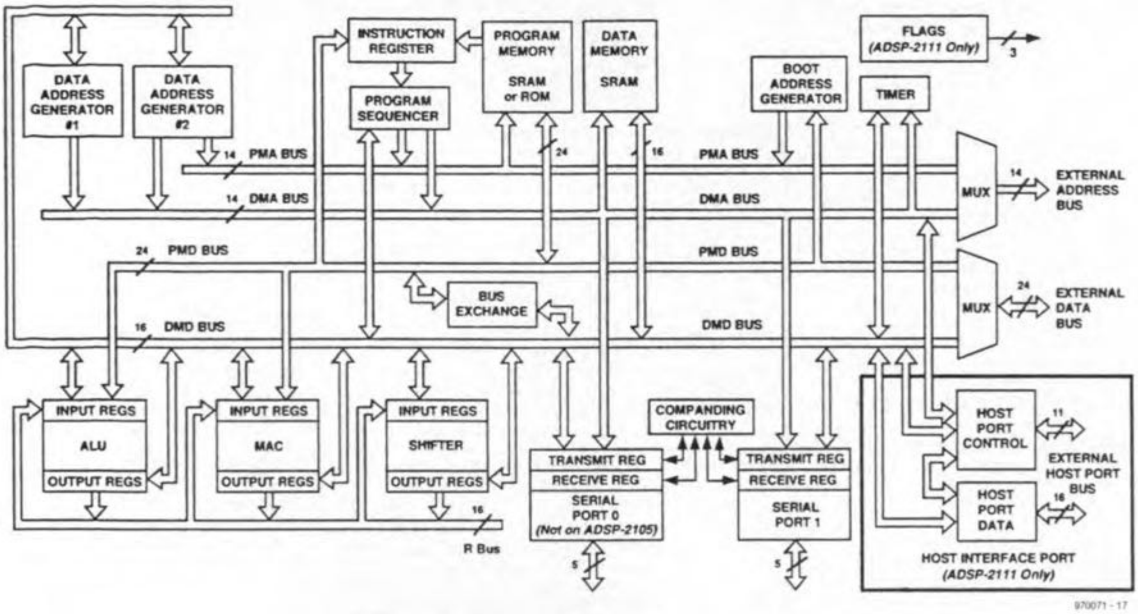


Figure 7. Internal block diagram of a modern DSP in the ADSP-21xx series from Analog Devices.

block, of samples. Processing occurs on the entire frame of data and results in a frame of transformed data as shown in **Figure 5b**.

For an audio sampling rate of 48 kHz, a processor working on a frame of 1024 samples has a frame acquisition interval of 21.33 ms (that is, $1024 \times 20.833\,\mu s = 21.33$ ms). Here, the DSP has 21.33 ms to complete all the required processing tasks for that frame of data. If the system handles signals in real time, it must not lose any data; so, while the DSP is processing the first frame, it must also be acquiring the second frame. Acquiring the data is one area where special architectural features of DSPs come into play: seamless data acquisition is facilitated by a processor's flexible data-addressing capabilities in conjunction with its direct memory-accessing (DMA) channels.

RESPONDING TO REAL-WORLD SIGNALS

It cannot be assumed that all the time between samples is available for the execution of processing instructions. In reality, time must be allowed for the processor to respond to external devices, controlling the flow of data in and out. Typically, an external device (such as an ADC) signals the processor by way of an interrupt. The DSP's response time to that interrupt, or interrupt latency, directly influences how much time remains for actual signal processing.

Interrupt latency(response delay) depends on several factors of which the most dominant is the DSP architecture's instruction pipelining. An instruction pipeline consists of the number of instruction cycles that occur between the time an interrupt is received and the time that program execution resumes. For example, if a processor has a 20 ns cycle time and requires 10 cycles to respond to an interrupt, 200 ns elapse before it executes any signal-processing instructions. In continuous processing, this 200 ns overhead will not hurt if the DSP finishes the processing of each sample before the next arrives.

In batch processing, however, an interrupted system wastes processor instruction cycles. For example, a system with a 200 ns interrupt response time running a frame-based algorithm, such as the FFT (Fast Fourier Transform), with a frame size of 1024 samples, would require 204.8 μs of overhead. That amounts to more than 10,000 instruction cycles wasted to latency. This waste is easy to avoid in DSPs having architectural features such as DMA and dual memory access: these let the DSP receive and store data without interrupting the processor.

DEVELOPING A DSP SYSTEM

Now the role of the processor, the ADC, the anti-aliasing filter, and the timing relationships between these components have been discussed, it is time to look at a complete DSP system. **Figure 6** shows the building blocks of a typical DSP system that could be used for data acquisition and control.

Note how few components make up the system, because so much of the system's functionality comes from the programmable DSP.

Converters, both analogue-to-digital (DAC) and digital-to-analogue (DAC), funnel data in and out of the DSP. The ADC timing is controlled by a precise sampling clock.

To simplify system design, many converter devices available today combine some or all of the following: an ADC, a DAC, a sampling clock, and filters for anti-aliasing and anti-imaging.

There are some other points that need to be borne in mind to arrive at a final design. For instance, the bandwidth of the anti-aliasing filter depends on the input signals used. Whether the interfacing between the ADC and DSP is serial or parallel depends on the desired speed. Parallel takes more space (both as regards board tracks and terminal pins), but is faster. Serial is slower, but much more compact and normally less expensive.

The incoming data is handled by the DSP's algorithm. When the processor completes the requisite calculations, it sends the result to the DAC. Because the signal processing is programmable, considerable flexibility is available in handling the data and improving system performance with incremental programming adjustments.

The DAC converts the DSP's digital output into the desired analogue output at the next sample clock. The converter's output is smoothed by a low-pass anti-imaging filter (also called a reconstruction filter) to produce the reconstructed analogue signal.

[970071]

Reference

Analog Interface and DSP Sourcebook by Alan Clements, McGrawHill (1993), ISBN 0 07 707694 x (reviewed in the May 1993 issue of this magazine).

make your own front panels

for a nice finish, the inexpensive way

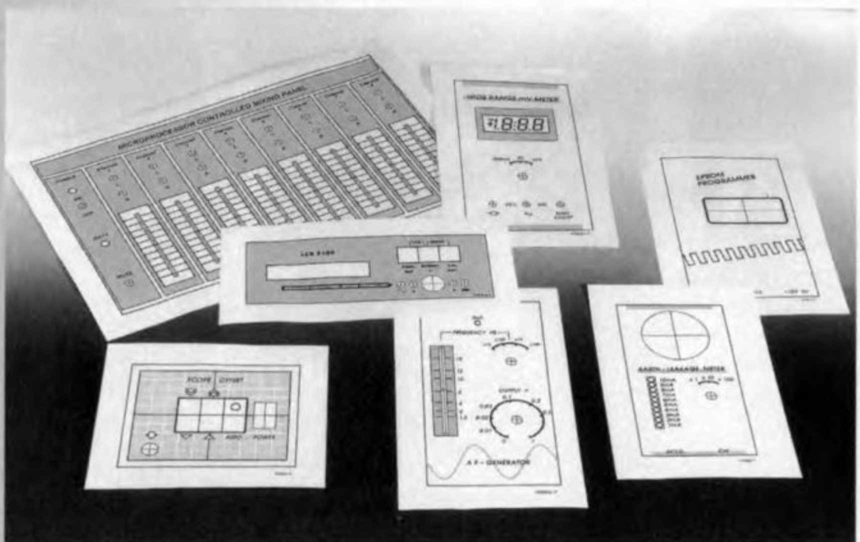


Figure 1. A modern inkjet printer enables you to produce a good looking front panel foil without too many problems.

When building your own electronic equipment, finishing the case with a smart looking front panel is often a problem. Because a ready-made front panel foil is rarely available through our Readers Services, most of you will be forced to produce one by yourselves. Fortunately, making your own front panels is relatively easy if you have a computer and a colour printer, or access to such a combination. In the Elektor Electronics laboratory, this technique is applied for finishing off prototypes which appear on photographs in the magazine.

It is only by means of a properly finished case and a suitable front panel that a home-brew project can be given the 'professional look'. These days, a variety of metal and ABS cases is available from electronics shops and by mail order, and finding a type which allows a published circuit to be housed will not be too difficult, especially if the outside dimensions and material are known from the relevant construction article. Designing and making the front panel, however, is much more of a problem, mainly because every project has its own requirements as regards the number of controls, type and position. An then we're not even mentioning scales to be printed around knobs, and other special treats.

By our laboratory staff

Unfortunately, designing, producing, storing and selling a ready-made front panel foil through our Readers Services is only viable if a certain minimum sales volume can be achieved. Hence, this product is now rarely seen in our product lists. The good news is that many of you have a computer and (access to) a colour printer of the inkjet type (e.g., an Epson, Canon or Hewlett-Packard printer). These two 'tools' form a perfect basis for making front panels with a professional look, and not expensive, either! Let's examine the production steps one by one.

Step 1. Use a drawing or photo retouching program (say, Corel Draw of Paintshop Pro) to design the front panel layout at true size. Where possible, use saturated colours, and no variable shades. Although modern inkjet printers offer quite acceptable results for screens in the backgrounds, actually printing these patterns remains difficult.

If you intend to use a certain enclosure for different applications, then a basic ('master') layout can be designed and saved as a kind of template for use later.

Step 2. Print the front panel design, using self-adhesive white foil sheets (suitable for the printer, of course), or glossy paper. Ordinary foil is not suitable in most cases because it does not absorb the ink properly. Allow the foil to dry completely after printing. Suitable foil or paper is available in relatively small quantities from stationers or computer shops (in some cases, by the sheet, else in boxes of 25 or 50 sheets).

Step 3. Glue the printed foil on a piece of solid paper (120-150 g). This is done to make the front panel more solid, and to be able to hide small surface irregularities. Glossy paper may be secured with two-component glue. If a

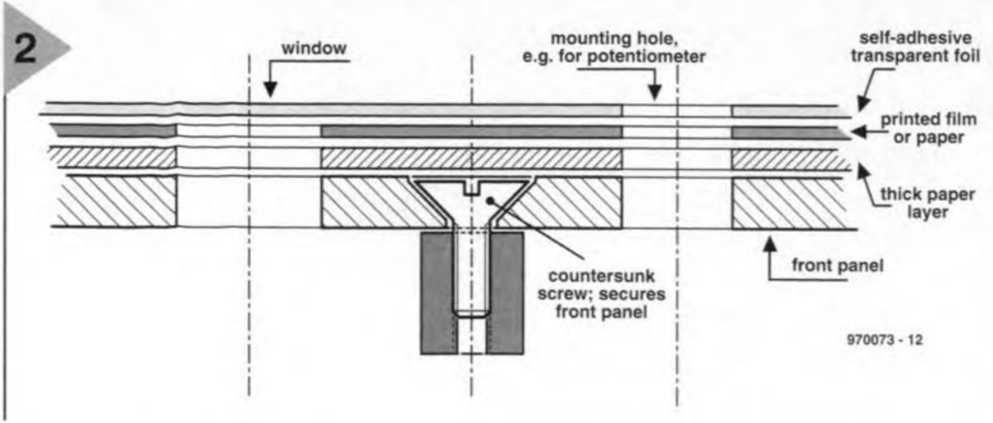


Figure 2. Basic construction of the front panel. On top, the self-adhesive transparent foil, then comes the front panel as printed on film or paper, and, finally, a layer of thick paper for extra strength, and to allow irregularities in the surface to be eliminated. The whole is glued on to the actual (metal or ABS) front panel supplied with the enclosure.

transparent window is required, this is the right moment to cut it out of the foil or paper layer using a sharp hobby knife. Next, cut the front panel exactly to size, so that it fits accurately on the (metal or plastic) front panel of the enclosure. This panel has to be completely drilled, filed and sandpapered before the front panel foil can be glued on. In this respect, it is important to carefully remove any burrs from the front panel. This is best done with a fine-grade sandpaper.

Step 4. Cover the front panel with a piece of transparent self-adhesive foil. Use a small paint roller (a clean one!) or a large pen to remove all bubbles from the surface. Cut off the foil around the panel, using a wide margin (at least 1 cm), and make all required holes. Next, wrap the excess foil around to the back of the front panel. If you can find a suitable glue, it is also possible to stick the panel, complete with the foil on it, directly on to the front panel. That does, however, require some experimenting!

The main function of the thick paper layer is to enable small irregularities in the front panel surface to be ironed out. If the surface is absolutely level, then the extra paper layer is not required.

Although the quality of the ink used in modern inkjet printers is constantly improving, discolouring does occur as a result of exposure to UV light. It is therefore useful to apply a transparent self-adhesive foil with a built-in UV filter. Ask your supplier about this specification. Obviously, when storing the equipment it is best to avoid a position where it is exposed to direct sunlight.

If you use a laser printer instead of an inkjet printer, you will have grey tones only. On the positive side, you don't have to worry about UV light affecting the durability of the front panel foil.

(970073-1)

the wall box

Brief parameters

Design	Two-way or three-way, vented
Woofer	WS13BF (Visaton)
Tweeter	DTW95NG (Visaton)
Wideband driver	FRWS5 (Visaton)
Frequency range	70 Hz – 20 kHz (–3 dB)
Crossover frequency	About 2 kHz
Music power	60 W
Nominal impedance	4 Ω
Volume	5 litres
Dimensions	300×418×138 mm (H×W×D)
Special aspects	Wideband unit provides sideways radiation

In these days of miniaturization, few people seem to feel a need for large loudspeakers in their living room. But, although they want small boxes, they still want good quality. The wall box presented in this article meets those requirements. It is a two-way system housed in a small, shallow enclosure that is inconspicuous in the living room.

Design by H. Baggen

There are many signs in the retail figures of the audio/hi-fi industry that the real hi-fi period of separates has peaked. These days, most customers specify a small or medium-sized system (mini towers are particularly popular right now) from which they expect exactly the same performance as from yesteryear's large installations. Loudspeakers are not excluded. Small, inconspicuous ones are wanted, but these must sound good and, more particularly, produce a good bass. Unfortunately, the unyielding laws of nature do not allow good bass reproduction from a small enclosure.

Nevertheless, if good drive units are used in a well-designed medium-sized box, good sound, small(ish) dimensions and affordability can go hand in hand, and this is what the present design is all about.

In the design, account was also taken of the fact that some people may use the present loudspeaker in a surround-sound system. Additional loudspeakers at right angles to the line connecting listener and main loudspeakers broaden the stereo reproduction.

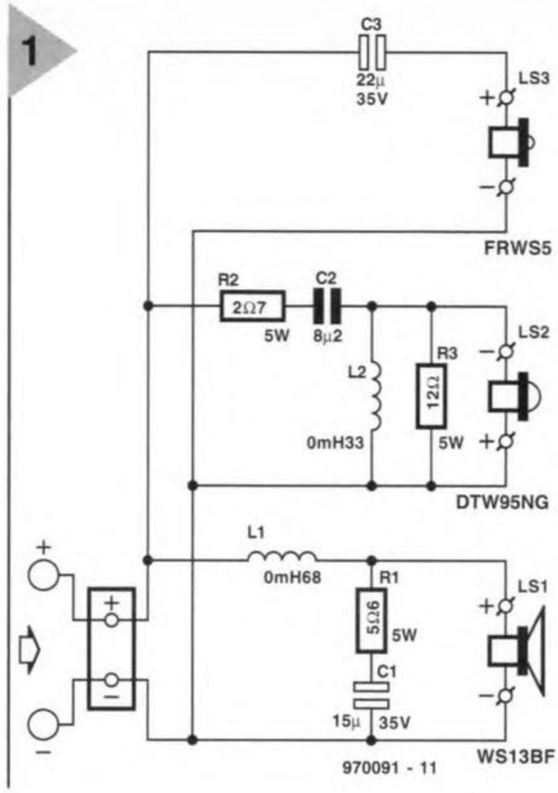
The enclosure is intended primarily to be hung from a wall: it is tuned for that purpose.

DESIGN CONSIDERATIONS

As mentioned earlier, the design specification includes reasonable cost, good sound reproduction that fits in well with a modern (stereo) TV receiver or hi-fi system, and is not too conspicuous in the living room. The result is a medium-sized, shallow box intended for wall mounting.

The requirement of providing good bass response from a medium-sized,

Figure 1. Circuit diagram of the crossover network for the wall box. The filter for the woofer and tweeter is a second-order one, while that for the wide-band unit is simply a capacitor.



not too expensive loudspeaker makes the choice of drive units highly critical. In the prototype, a 130 mm diameter bass/mid-range unit, a WS13BF, a 25 mm diameter tweeter, a DTW95NG, and a 50 mm diameter wideband unit, a FRWS5 are used. All three units come from the Visaton stable.

The Thiele/Small parameters of the WS13BF make it eminently suitable for use in a medium-sized box. It has a coated paper cone with a foam rubber surround and a 25 mm diameter voice coil.

The tweeter has a square mounting flange that goes well with the mid-range unit. It has an impregnated woven dome and heavily damped surround. The air gap is filled with magnetic ferrofluid, which provides good cooling and additional damping. Its frequency response is almost flat over the range 1.5–20 kHz.

The FRWS5 enhances the stereo effect. Strictly speaking, it may be omitted where this enhancement is not needed. However, in the prototype, it helped to produce a wide-angle stereo sound, which is particularly noticeable at the sides of the listening angle.

CROSSOVER NETWORK

The filter has also been kept as simple as possible without too many compromises. Its circuit diagram is shown in **Figure 1**. The network was designed with the aid of a computer simulation program that, based on the measured frequency and phase response of the drive units in the enclosure, calculates the overall response of the system including the crossover network. It also takes into account the positioning of the drive units and their radiation pattern.

The design of the network is a slightly modified second-order Butterworth filter with a few impedance corrections. This results in a second-order low-pass filter, L_1-C_1, for the woofer. Capacitor C_1 has another function as well: in conjunction with R_1 it provides the requisite impedance correction for the woofer, which is necessary for good filter performance.

The high-pass filter for the tweeter is formed by L_2-C_2, which, in conjunction with the response of the tweeter itself, results in a third-order rolloff. Resistors R_2 and R_3 lower the output of the tweeter by about 3 dB to match its output level to that of the woofer.

The cross-over frequency between woofer and tweeter is at about 2 kHz

The 'filter' for the wideband unit is simply a 22 μF bipolar electrolytic capacitor in parallel with the filters for the other two units. This cuts off the response of the wideband unit below about 800 Hz, which results in a satisfying additional spatial effect.

ENCLOSURE

The shallow box has a net volume of five litres and yet the 3 dB cut-off frequency is as low as about 68 Hz.

It will be seen from the introductory photograph and the construction plan in **Figure 3** that the woofer and tweeter are pointed slightly towards

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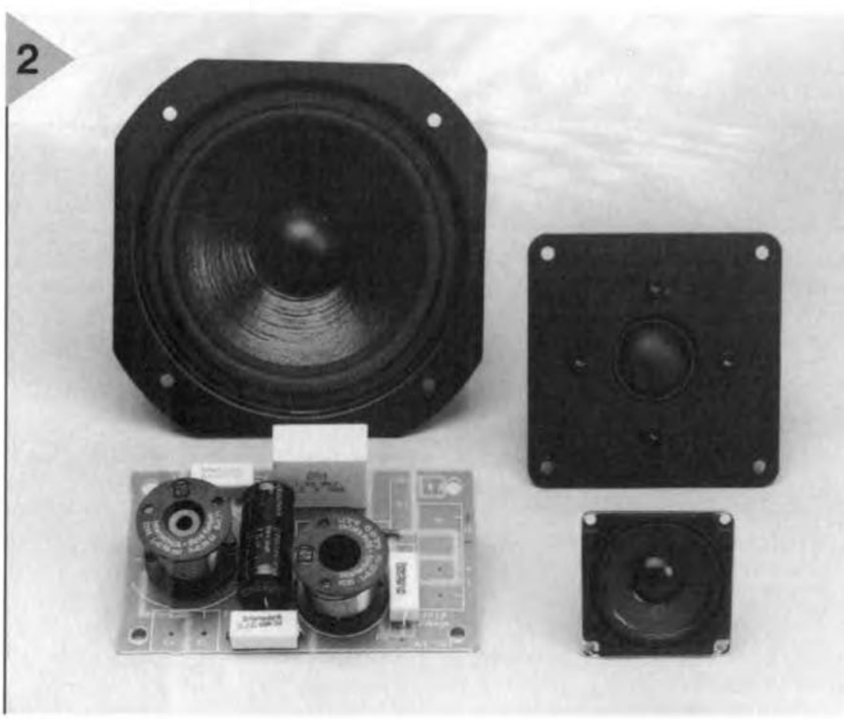

Figure 2. The drive units and crossover filter for the wall box (the electrolytic capacitor for the wideband unit is not shown).

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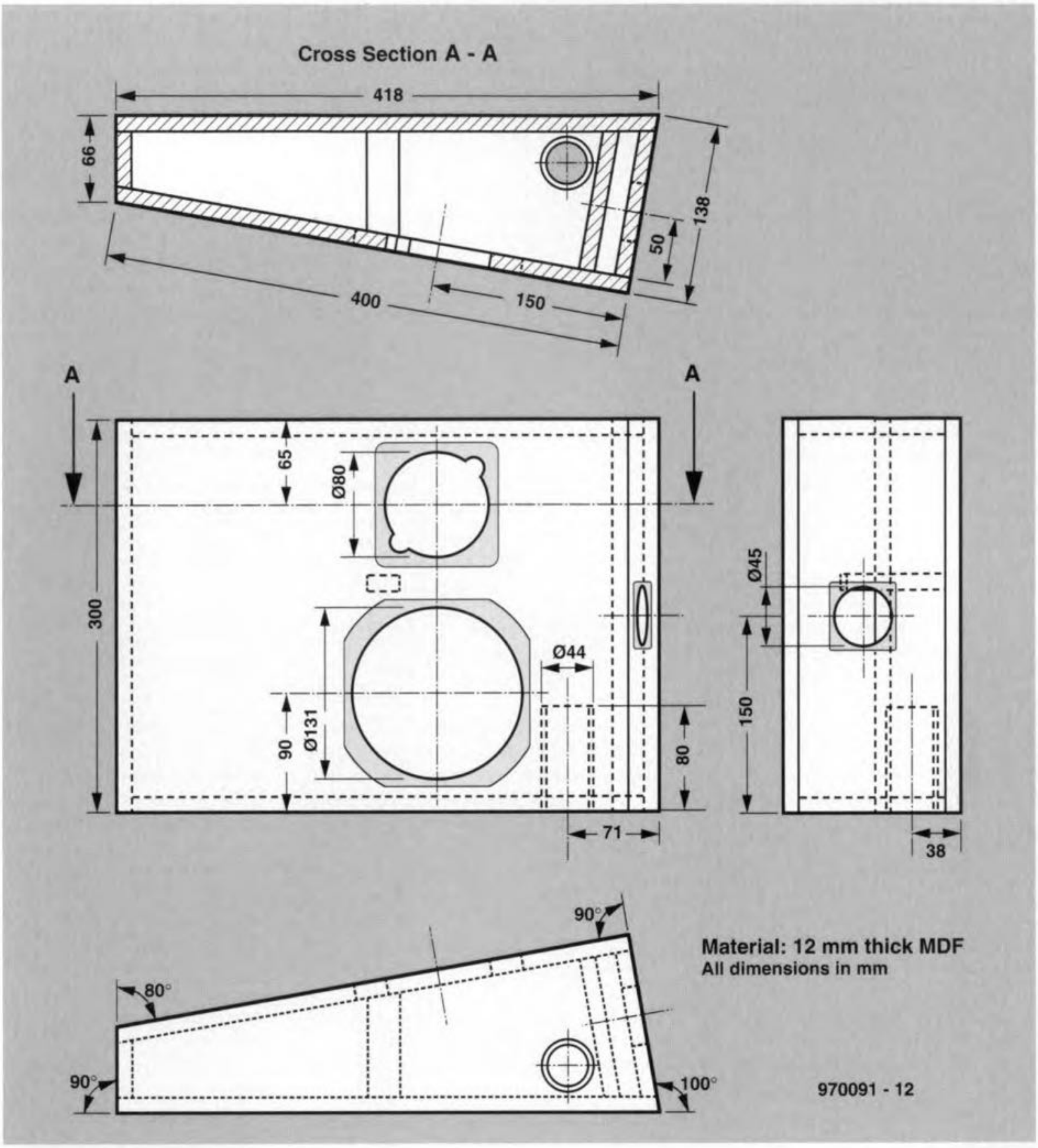

Figure 3. Construction plan for the box. Note that some of the panels are cut at 80° instead of the usual 90°.

the listener, whereas the wideband unit points slightly further forward. This makes it necessary for some of the constituent panels to be sawn at an angle of 80° instead of 90°. These angles make the enclosure look rather different from the usual rectangular box. Of course, if this angular construction is found too tedious, the box may be made rectangular with dimensions 300 × 160 × 160 mm.

Two holes are required in the front panel for the woofer and tweeter, and a single hole at the side for the wideband unit. The inside diameter of this latter hole should be enlarged with a suitable wood rasp or wood file to give the unit 'more air' at its rear.

A horizontal brace must be glued at the centre of the box between front and rear panel.

A partition at the side gives the wideband unit its own tiny chamber.

The bottom panel needs a suitable hole for the 44 mm diameter bass reflex port, which is 80 mm long. This port results in the the enclosure being tuned to about 50 Hz.

The crossover filter is best built on a piece of prototyping or similar board. Such boards can be bought from most specialist audio retailers. It is advisable to buy the components for the filters also from such a retailer, since electronics retailers normally do not stock them.

Fit the filter board in the box and feed the requisite connecting wires to the outside (those of the wideband unit through small holes in the partition). It is, of course, also possible to fit special (gold-plated) terminals to be underside of the box: this will look very pleasant.

Fit two small suspension plates at the top back of the box to enable it to be hung from the wall.

Fill the box with suitable wadding, wire up the drive units, and screw them into place. Mind the correct polarity, which is shown in the diagram (the tweeter should be in antiphase with the woofer).

The box may be finished to personal taste and in accord with the decorations in the living room.

POSITIONING AND USAGE

The finished loudspeakers should be hung from a wall at ear-height (that is, at a height from the floor of about 1.6 metres). Their position should be

Parts list (per box)

Resistors:
R_1 = 5.6 Ω, 5 W
R_2 = 2.7 Ω, 5 W
R_3 = 12 Ω, 5 W

Capacitors:
C_1 = 15 µF, 35 V, bipolar electrolytic
C_2 = 8.2 µF, metallized polyester (MKT)
C_3 = 22 µF, 35 V, bipolar electrolytic

Inductors:
L_1 = 680 µH air-cored or suitably cored, $R_i \leq 0.5\ \Omega$
L_2 = 330 µH, air-cored, $R_i \leq 0.5\ \Omega$

Drive units:
LS_1 = WS13BF, 8 Ω (Visaton)
LS_2 = DTW95NG, 8 Ω (Visaton)
LS_3 = FRWS5, 8 Ω (Visaton)

Miscellaneous:
PVC pipe, 44 mm outer diameter, 80 mm long
Wadding, about 40×25 cm
2 off gold-plated loudspeaker terminals

Medium-density fibreboard (MDF) panels 12 mm thick:
1 off 418×300 mm
1 off 402×300 mm
1 off 420×300 mm
1 off 114×300 mm
1 off 107×300 mm
2 off 392×42×378×112 mm

4

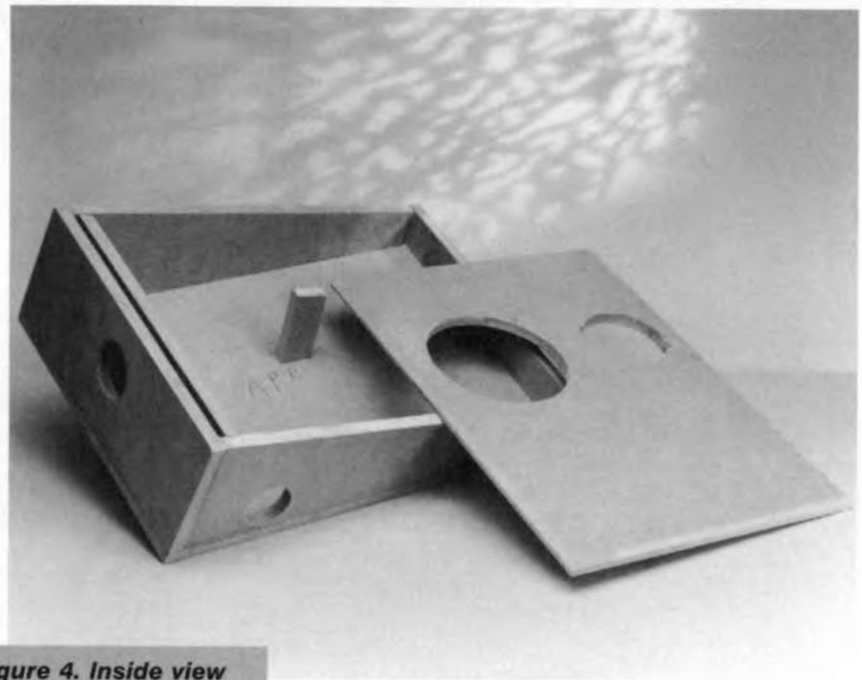

Figure 4. Inside view of the box before the top panel is put in place.

Figure 5. Frequency response curve and impedance curve of the wall box, incl. the wideband unit.

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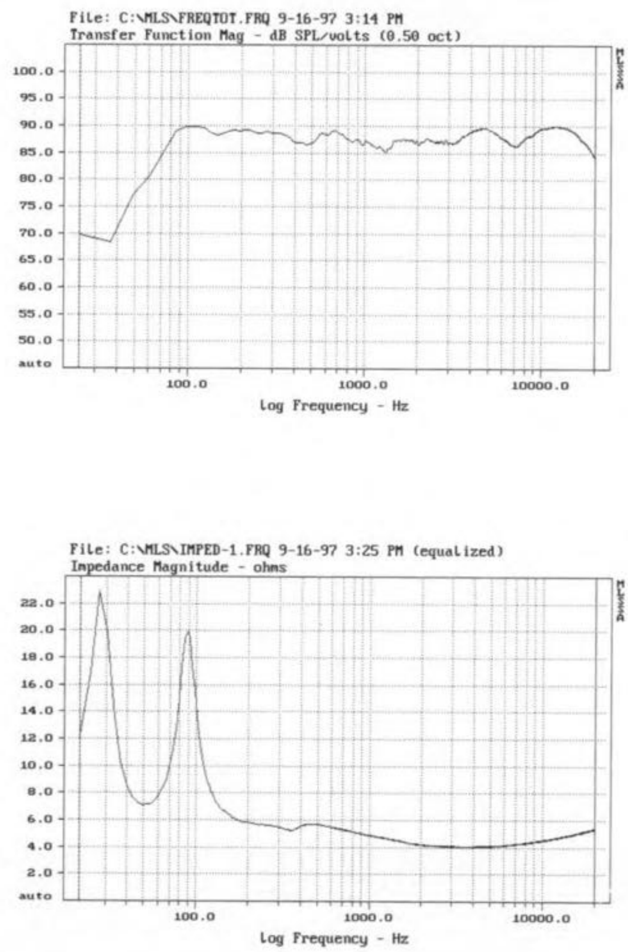


well away from corners so as not to lose the effect of the wideband unit. They should also not be placed too close (less than one metre) to a TV receiver, because the drivers are not shielded.

The frequency response and impedance characteristics are shown in **Figure 5**. Note that the minimum impedance is about 4 Ω at 10 kHz, which is a value that presents no risks to modern drive units.

A listening test on the prototype loudspeakers showed that, in spite of their modest dimensions, the sound quality and the bass reproduction are very good. Where top-class performance is required, they are best combined with a good-quality stereo system. The units are also very suitable for use as the front loudspeakers in a surround-sound installation.

[970091]

Corrections & Updates

The Wall Box

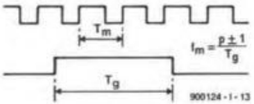
November 1997 - 970091

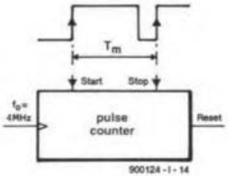
The outside diameter of the PVC tube for the bass reflex port should be 40mm, not 44mm.

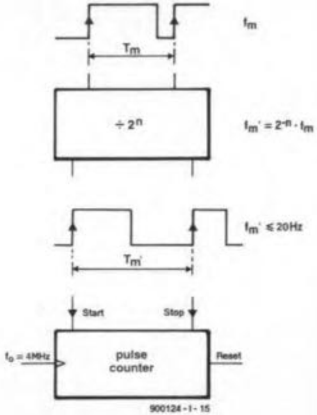
Frequency Meter and Event Counter Module

October 1997 - 970077

The drawings in the inset 'Principle of measurement' were mangled by our phototypesetting machine. The correct drawings are reproduced here.



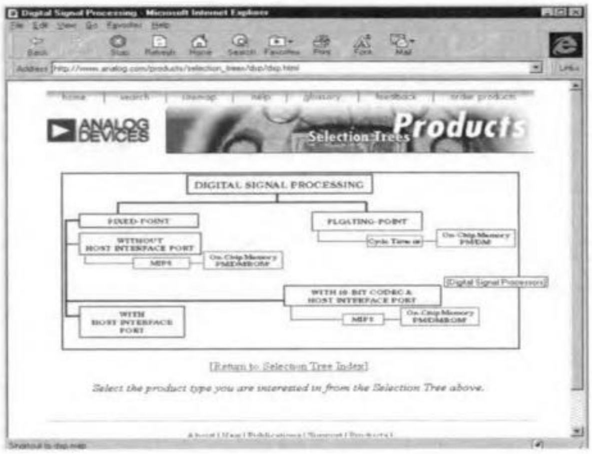


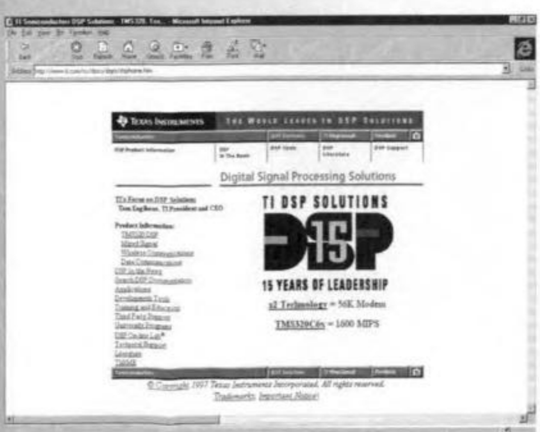


electronics on-line

DSPs on the Internet

If you can't wait to start working with digital signal processors (DSPs), a lot of relevant information may be found on the Internet. Datasheets and applications are published by this global network. On this page we hope to provide a number of starting points for a successful search.





DSPnet is a general forum which supplies information on practically any DSP. This site may be found at

http://www.dspnet.com or *http://www.technoline.com/dspnet/dspnet1.html*.

Both URLs point to the same location. Once at this site, the visitor is confronted with various forums, seminar schedules, datasheet files, tips and application notes. A complete overview of all information at this site may be found at the address

http://www.technoline/tol/dspnet/comp.html.

Interestingly, this site holds the specifications of all processors found in the standard supply program of all familiar DSP manufacturers. Using general qualifiers (like fixed or floating-point processor) or a specific requirement (say, regarding computing power), the processor having the desired features is quickly located.

Straight to the manufacturer

If you have already decided which DSP to use, you may also go straight to the desired site. In practice, three major manufacturers control the market for DSPs. Digital signal processors from **Motorola** are given an extensively presentation at the web address

http://www.motorola.com or *http://www.mot.com*.

The menus at these sites quickly guide you to the DSP main page at

http://www.mot.com/sps/dsp/products.

By virtue of the pointers you see on these pages, finding the necessary information on the DSP product line is easy.

Another large supplier of DSPs is **Texas Instruments** (*http://www.ti.com*) whose home page has a prominent location for everything related to DSPs. Choose

http://www.ti.com/sc/docs/dsps/dsphome.htm

for an excellent and extensive overview of TI's DSPs.

Last but certainly not least comes **Analog Devices** (*http://www.analog.com*) whose homepage seems to lack a direct reference to DSPs. A good idea is to use the search option and enter 'DSP' as the argument. The result is a screen full of references to technical information like datasheets and lots of other data covering the DSPs made by Analog Devices. Alternatively, choose a direct address. Simply type

http://www.analog.com/products/selection_trees/dsp/dsp.html,

and the main page about Digital Signal processing will be opened.

(975094-1)

Design kits for AC motor drives

International Rectifier has introduced a family of power conversion subsystem design kits for 230 V or 460 V ac motors from 370 W to 11 kW (0.5 to 15 hp)

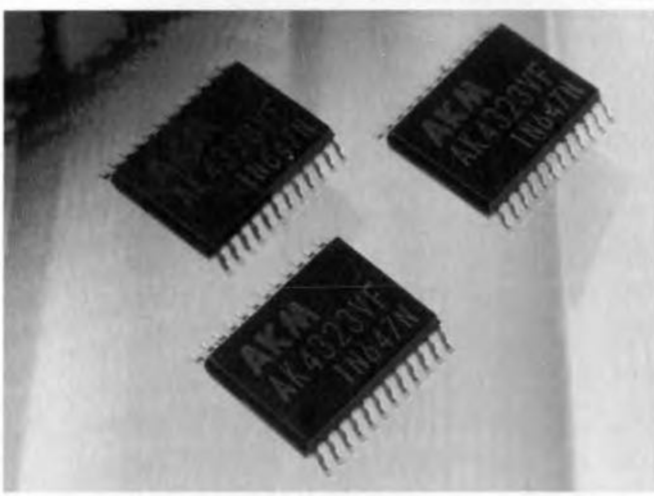
The PowIRTrain™ design kits combine all power conversion and power control functionality required for building state-of-the-art motor drives and controls, with or without braking, in modules or assemblies that are up to 75 percent more compact than other commercially available solutions.
The kits dramatically reduce the required design effort to translate power semiconductors into usable drive designs, eliminating component testing and device qualification, and taking into account thermal management and noise-free layout issues. They include both an integrated power stage and a fully optimised control stage, plus a bill of materials, complete schematics, Gerber photoplots for board layouts, and a comprehensive system specification.

(977199-1)

International Rectifier, Hurst Green, Oxted, Surrey RH8 9BB. Tel. (01883) 732020, fax (01883) 733410. Internet: www.irf.com.

Audio DAC with on-board PLL

New from DIP International is the AK4323 stereo D/A converter with an integral phase locked loop (PLL), designed for use in digital broadcast systems which use the 27MHz MPEG system clock.



The on-chip PLL reduces board area and saves cost as until now, separate external circuitry was required for the PLL. This part of the AK4323 provides selection of sampling clock frequencies locked to the recovered 27MHz MPEG clock, supports 2/3 division of master clock by some MPEG audio devices, and allows master clock frequencies including half rates to be chosen.
The DAC section of this CMOS device has 128x oversampling, a 20-bit 8x digital filter, combining high performance with low cost, making it ideal for high quality, cost sensitive applications. The built-in 2nd order SCF and 2nd order CTF minimises external filtering. There is digital deemphasis for 32, 44.1 and 48 kHz sampling, which makes the AK4323 compliant with most audio applications. The dynamic range of 100 dB supports true 16-bit performance. Supplied in a small 24-pin VSOP package, power supply requirements are 4.5 to 5.5 V.
Applications include digital set top boxes, television and radio broadcast systems and post production equipment that runs from a 27MHz system clock.

(977200-1)

DIP International Ltd., Sheraton House, Castle Park, Cambridge CB3 0AX. Tel. (01223) 462244, fax (01223) 467316. Email: 100343.304@compuserve.com

PICs with 8-channel ADC and 8k×14 program memory

Microchip's new PIC16C76 and 16C77 microcontrollers offer 8k×14 words of program memory, 376 bytes of data RAM and a very low power on-chip A/D converter.

The 5 MIPs EPROM-based microcontrollers contain an on-chip, 5-channel (PIC16C76) and 8-channel (PIC16C77), 8-bit A/D converter with sample and hold, accuracy of ±1LSB and an acquisition time of just 16ms. Available in small footprint 28-pin and 40-pin packages, these one-time programmable devices feature 200-ns microcycle times, 35 single-cycle instructions, and are ideal for embedded control applications such as air bag controllers, motor control, set top boxes and network switches. (977202-1)

Arizona Microchip Technology Ltd., Unit 6 The Courtyard, Meadowbank, Furlong Road, Bourne End, Bucks SL8 5AJ. Tel. (01628) 851077, fax 850259.

MAX187

Integrated Circuits

A-D Converters

ELEKTOR ELECTRONICS

DATASHEET 11/97

+5V, low-power, 12-bit serial ADC

General description

The MAX187 serial 12-bit analogue-to-digital converter (ADC) operates from a single +5V supply, and accepts a 0V to 5V analogue input. The device features an 8.5μs successive approximation ADC, a fast track/hold (1.5μs), an on-chip clock, and a high-speed 3-wire serial interface.

The MAX187 digitizes signals at 75ksps throughput rate. An external clock accesses data from the interface, which communicates without external hardware to most digital signal processors and microcontrollers. The interface is compatible with SPI™, QSPI™ and Microwire™.

The MAX187 has an on-chip buffered reference, and comes in space saving 8-pin DIP and 16-pin SO packages. Power consumption is 7.5 mW and reduces to only 10 μW in shutdown.

Manufacturer

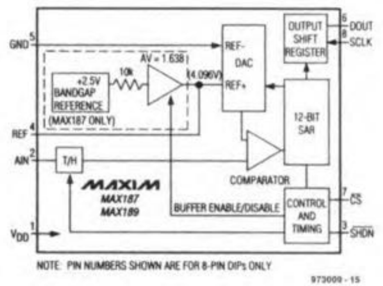
Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, U.S.A. Tel. (408) 737-7600, fax (408) 737-7194. Internet: www.mxim.com.

Application example

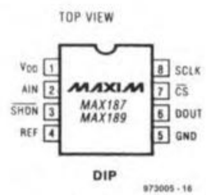
12-bit ADC Interface, *Elektor Electronics* October 1997.

Features

- 12-bit Resolution
- ±½ LSB Integral Nonlinearity (MAX187A)
- Internal Track/Hold, 75kHz Sampling Rate
- Single +5V Operation
- Low Power: 2μA Shutdown Current
  1.5mA Operating Current
- Internal 4.096V Buffered Reference
- 3-Wire Serial Interface, Compatible with SPI, QSPI, and Microwire
- Small Footprint 8-pin DIP and 16-pin SO



Functional diagram



Pin configuration, 8-pin DIP

Electrical characteristics	(V_{DD} = +5V ±5%;GND = 0V; unipolar input mode; 75ksps; f_{CLK} = 4.0MHz; external clock (50% duty cycle); internal reference: V_{REF} = 4.096V, 4.7μF capacitor at REF pin.					
Parameter	**Symbol**	**Conditions**	**Min**	**Typ**	**Max**	**Units**
DC ACCURACY (Note 1)						
Resolution					12	bits
Relative Accuracy (Note 2)		MAX187A			±½	LSB
		MAX187B			±1	
		MAX187C			±2	
Differential Nonlinearity	DNL	No missing codes over temperature			±1	LSB
Offset Error		MAX187A			±1½	LSB
		MAX187B/C			±3	
Gain Error (Note 3)		MAX187			±3	LSB
Gain Temperature Coefficient		External reference, 4.096V		±0.8		ppm/°C

NH-3

Sensors

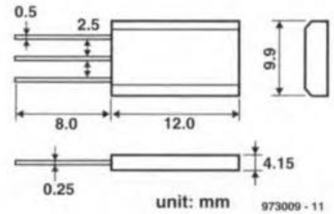
ELEKTOR ELECTRONICS

DATASHEET 11/97

The NH-3 is a **hybrid humidity sensor** whose output voltage to relative humidity is linearized. The sensitive element is a stable high polymer impregnated in porous ceramic, setting it apart from conventional high polymer type sensors.

Manufacturer

Figaro Engineering Inc., P.O. Box 357, 1000 Skokie Blvd, Room 575, Wilmette, IL 60091, U.S.A. Tel. (708) 256-3546, fax (708) 256-3884.



Structure and measurement

The NH-3 consists of two mounted relative humidity sensitive elements and two printed NTC thermistors for temperature compensation on an alumina ceramic substrate, allowing the output voltage to be linearized.

The AC supply voltage (V0) is applied to terminals 1 and 3. The output voltage is measured between terminals 1 and 2.

Features

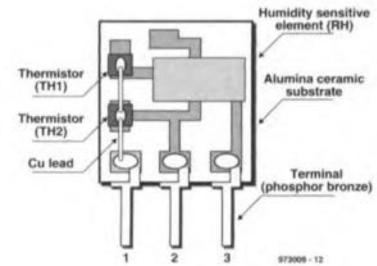
- Linear AC output voltage in range 30-90 %RH, at 10-40 °C
- Small temperature dependency
- High stability
- AC input
- Compact and light

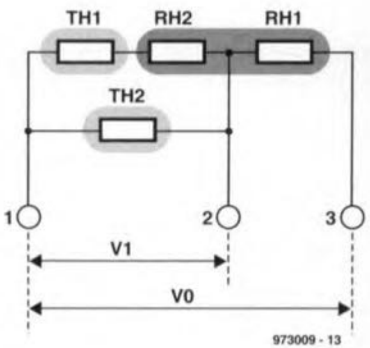
Applications

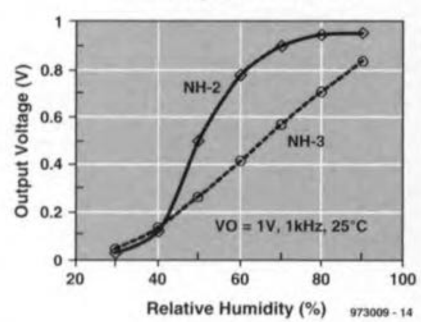
- Air conditioning systems
- Humidifiers, Dehumidifiers
- Dish or clothes dryers
- Copy or facsimile machines

Application example

Humidity Sensor, *Elektor Electronics* November 1997.







NH-3

Sensors

ELEKTOR ELECTRONICS

DATASHEET 11/97

Circuit & Operating Conditions	
Item	**Conditions**
Supply voltage	max. 2 V AC
Rated power:	1 mW (V0 = 1 V)
Operating temperature:	0 - 60 ºC
Operating frequency:	50 Hz - 1 kHz
Operating humidity:	20 - 90 %RH
Storage temperature:	−20 - 80 ºC (avoid continuous storage exceeding 100 hrs in 71-80 ºC)
Storage humidity:	0 - 95 %RH (at −20 - 40 ºC) 0 - 80 %RH (at 40 - 80 ºC, avoid continuous storage exceeding 100 hrs in 71 - 80 ºC, 80 %RH)

Electrical Characteristics					
Item	**Minimum**	**Typical**	**Maximum**	**Unit**	**Conditions**
Output voltage	0.34		0.47	VAC	AC 1.0V, 25ºC, 60%RH
Change ratio of voltage		0.015		V/%RH	AC 1.0V, 25ºC, 40-70%RH
Accuracy: (including temperature dependency)	−10		+10	%RH	AC 1.0V, 10-40ºC, 30-80%RH
Temperature dependency:	−5		+5	%RH	AC 1.0V, 10-40ºC, 30-80%RH
Response time:[1]		3	4	min	25ºC, from 30%RH to 80%RH, wind speed 2.5 m/s
Stabilizing time:[2]		100	150	msec	
Thermistor resistance		20kΩ ±5%			25 ±0.2ºC
B constant		4350K ±5%			

[1] Response time required to attain 90% of the change
[2] Time required for the unit to stabilize after powering up

MAX187

Integrated Circuits

A-D Converters

ELEKTOR ELECTRONICS

DATASHEET 11/97

Parameter	Symbol	Conditions	Min	Typ	Max	Units
DYNAMIC SPECIFICATIONS (10kHz sine wave input, 0V to 4.096V_{pp}, 75ksps)						
Signal-to-Noise plus Distortion Ratio	SINAD		70			dB
Total Harmonic Distortion (up to the 5th harmonic)	THD				−80	dB
Spurious-Free Dynamic Range	SFDR		80			dB
Small-Signal Bandwidth		Rolloff − 3dB		4.5		MHz
Full-Power Bandwidth				0.8		MHz
CONVERSION RATE						
Conversion Time	t_{CONV}		5.5		8.5	μs
Track/Hold Acquisition Time	t_{ACQ}		1.5			μs
Throughput Rate		External clock, 4MHz, 13 clocks			75	ksps
Aperture Delay	t_{APR}			10		ns
Aperture Jitter				<50		ps
ANALOGUE INPUT						
Input Voltage Range					0 to V_{REF}	V
Input Capacitance (Note 4)				16		pF
INTERNAL REFERENCE (reference buffer enabled)						
REF Output Voltage	V_{REF}	T_A = +25ºC	4.076	4.096	4.116	V
		$T_A = T_{MIN}$ to T_{MAX}, MAX187_C	4.060		4.132	
		$T_A = T_{MIN}$ to T_{MAX}, MAX187_E	4.050		4.140	
		$T_A = T_{MIN}$ to T_{MAX}, MAX187_M	4.040		4.150	
REF Tempco		MAX187AC/BC		±30	±50	ppm/ºC
		MAX187AE/BE		±30	±60	
		MAX187AM/BM		±30	±80	
REF Short-Circuit Current					30	mA
Load Regulation (Note 5)		0mA to 0.6mA output load		1		mV
DIGITAL OUTPUT (DOUT)						
Output Voltage Low	V_{OL}	I_{SINK} = 5 mA			0.4	V
		I_{SINK} = 16mA		0.3		
Output Voltage High	V_{OH}	I_{SOURCE} = 1mA	4			V
Three-State Leakage Current	I_L	$\overline{CS}$=5V			±10	μA
Three-State Output Capacitance	C_{OUT}	$\overline{CS}$=5V (Note 4)			15	pF
POWER REQUIREMENTS						
Supply Voltage	V_{DD}		4.75		5.25	V
Supply Current	I_{DD}	Operating mode		1.5	2.5	mA
		Power-down mode		2	10	μA
Power-Supply Rejection	PSR	V_{DD}=+5V, ±5%, external reference, 4.096V; full-scale input (Note 6)		±0.06	±0.5	V

Note 1: Tested at V_{DD} = +5V
Note 2: Relative accuracy is the deviation of the analogue value at any code from its theoretical value after the full-scale range has been calibrated.
Note 3: Internal reference, offset nulled. Excludes reference errors.
Note 4: Guaranteed by design. Not subject to production testing.
Note 5: External load should not change during conversion for specified ADC accuracy.
Note 6: DC test, measured at 4.75V and 5.25V only.

It is well known that a PC can be used to measure electronic quantities by making it behave as an oscilloscope through suitable software. Since a logic analyser is a variation of a digital oscilloscope, it stands to reason that a PC with its parallel inputs that can be scanned at high speed and its depth of memory to compute the measurement values is eminently suitable for use as a logic analyser. But where an oscilloscope displays real-time waveforms of signals, logic analysers store the signal for display in a non-real-time way.

Software design by Schröder

PC as 4-channel logic analyser

with or without electrical isolation

The number of input channels a logic analyser has defines the instrument's potential. If, for instance, the instrument is to be used to analyse an 8-bit microprocessor system with, say, a 16-bit address bus, an 8-bit data bus, and five or six control lines, the minimum number of input channels it needs is about 30. Unfortunately, a PC has normally only five parallel inputs so that the present logic analyser is restricted to four input channels.

The logic level at the four inputs is sampled at a rate of up to 1.2 MHz and saved in packets of 40 kbytes. Note that the sampling rate depends on the PC used: for instance, one that uses a 486 processor running at 33 MHz cannot go over 150 kHz.

The program makes three trigger modes available. There is also a 255 byte pre-trigger memory that determines the time interval before the trigger mode is set. The displayed diagrams may be printed or saved. There are numerous other facilities, such as scrolling, zooming, scanning of the display, and setting of the period and frequency. All four channels may be used for frequency measurement. Channel 1 may be used for the measurement of pulses, pulse width and pulse spacing. The results of measurements are shown in separate windows.

hardware

The only hardware needed for the logic analyser is for the protection of the Centronics interface and for preventing that this interface overloads weak signals. This is because the Centronics inputs draw a current that in certain circumstances may exceed the output capabilities of the signal source.

There are two circuits available, one with electrical isolation (**Figure 2**) and the other without (**Figure 1**).

In the circuit in Figure 1, resistors R_1–R_5 limit the input current to about 0.4 mA (from a supply line of 5 V). Circuit IC_1 is a transistor array with common emitter and open collector. The pull-down resistors, which are in parallel with the internal pull-up resistors and may, therefore, in some cases be omitted, prevent the bases of the transistors from functioning as antennas

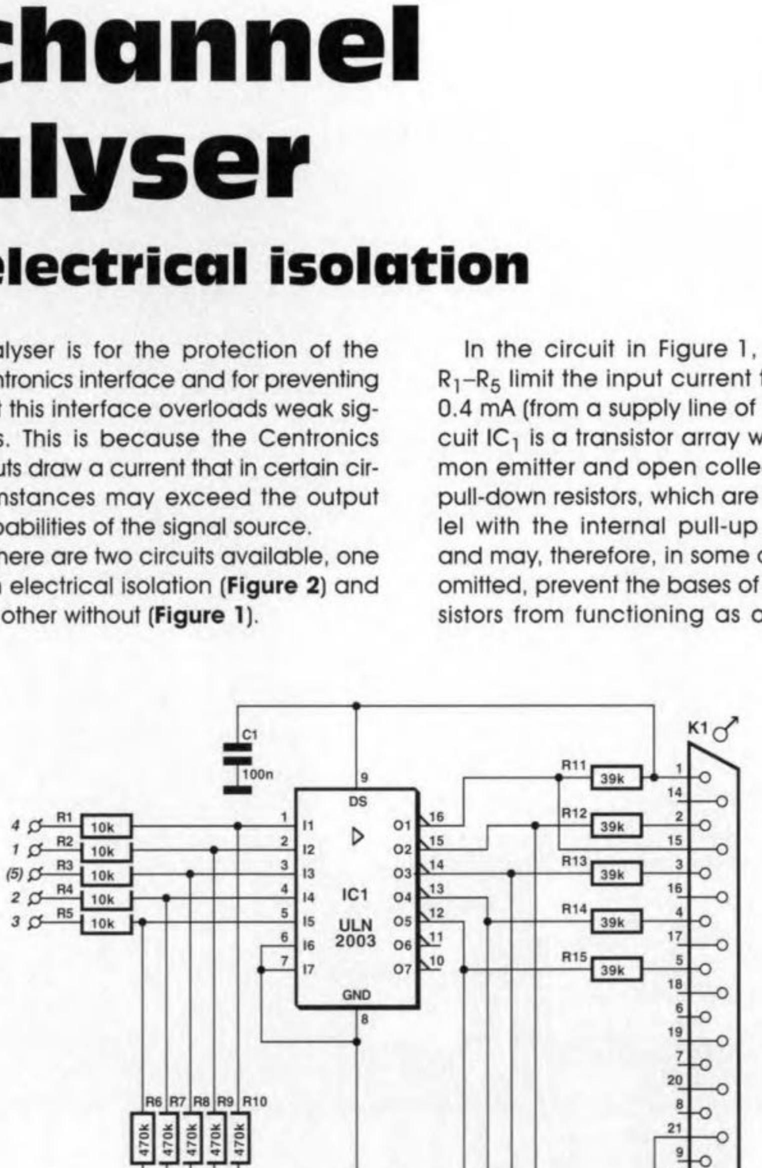


Figure 1. Circuit of the logic analyser hardware not providing electrical isolation.

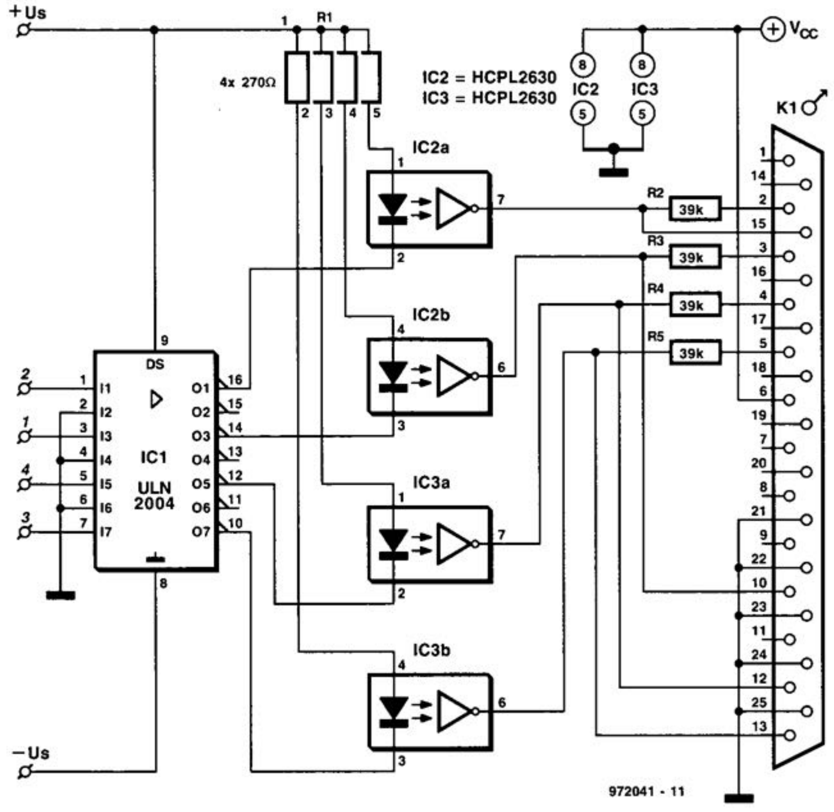


Figure 2. Circuit of the hardware for the logic analyser providing electrical isolation.

when the inputs are open-circuited. The pull-up resistors, which are linked to the data lines of the Centronics port, cause the collectors at the output to be logic high. Note that terminal (5) is not used.

In Figure 2, the current-limiting resistors are omitted since IC_1 has internal 10.2 kΩ series resistors. The output (at the collectors) of the internal transistors in the IC are linked to optoisolators in IC_2 and IC_3.

The optoisolators are not standard in that they have common ground and V_{CC} terminals and two open-collector outputs. To ensure 100% electrical isolation, the transistor array must be powered by the circuit on test via terminals $+U_S$ and $-U_S$.

When the input level is high, the outputs of IC_1 go low so that the LEDs in IC_2 and IC_3 light and the Schottky transistors at the outputs of these two ICs are on. The control lines, which are normally logic high owing to the pull-up resistors in SIL array R_1, are then pulled low.

The optoisolators have switching times of nanoseconds and are thus highly suitable for operation at high frequencies.

Figure 3 shows the printed-circuit board layouts for both circuits.

software

The logic analyser function is called "Logic_ An" and runs under DOS or in the DOS window of Windows 95. For satisfactory operation of the analyser, the computer should be an AT with VGA graphics and DOS 3.3 or better.

The program provides a three-part window as shown in **Figure 4**. On the right of this, the functions are represented by icons. When one of these is

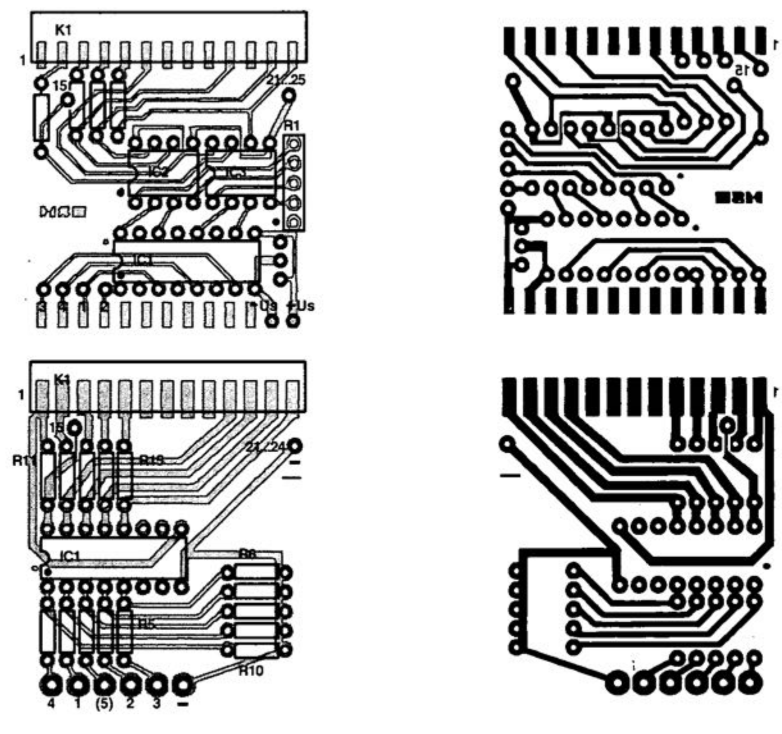


Figure 3. Printed-circuit boards for the hardware in Figures 1 and 2.

Parts list (Figure 1)

Resistors:
R_1–R_5 = 10 kΩ
R_6–R_{10} = 470 kΩ
R_{11}–R_{15} = 39 kΩ

Capacitors:
C_1 = 0.1 μF

Integrated circuits:
IC_1 = ULN2003 (Sprague)

Enclosure: to personal taste

Parts list (Figure 2)

Resistors:
R_1 = SIL array, 4×270 Ω
R_2–R_5 = 39 kΩ

Integrated circuits:
IC_1 = ULN2004 (Sprague)
IC_2, IC_3 = HCPL2630 (Hewlett Packard)

Enclosure: to personal taste

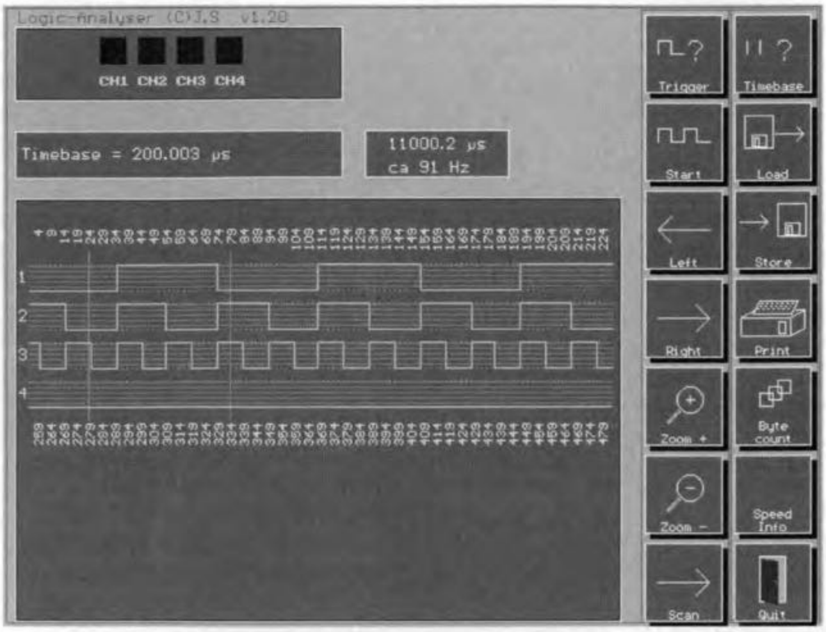


Figure 4. Main monitor screen of the logic analyser.

Hot keys

<key (n)>	function
<CTRL> <←>	left
<CTRL> <→>	right
<S>	scan (sample)
<+>	zoom +
<–>	zoom –
<CTRL> <Pos1>	start of plot
<ESC>	stop
<F1>	help
<F2>	frequency measurement in channels 1–4
<shift> <F2>	frequency measurement in channel 1
<F3>	pulse counting
<F4>	pulse interval

actuated, the desired action is enabled, or a menu is shown via which the wanted function can be selected or set. Selection is carried out with the aid of the arrow-keys and <enter> or by clicking the left-hand button on the mouse.

The integral help function can be selected with <F1>.

Triggering
There are three ways of starting the analyser: (1) by changing the instantaneous value of the sample bit; (2) by selecting a certain group of bits, and (3) by setting the status of a bit. In the case of (2) and (3), a small window is displayed in which the desired bit(s) can be selected with the relevant numeric key and set to low (grey) or high (yellow).

Start
This starts the plotting in the selected trigger mode. If a different one is selected, a window with the relevant trigger mask is shown. Start is stopped when any key is pressed.

Left/right
This shifts the window with the plotted data to the left or right as the case may be.

Zoom+/zoom–
This enlarges or diminishes the scale of the display of plotted data starting at the centre of the window.

Scan
This searches for the next position that is not in accord with the group of bits shown at the right-hand side of the window.

Time base
This sets the sampling value in microseconds, although this also depends on the ICs in the PC. Valid and exact values are ascertained empirically.

Load/save/print
This causes the plot to be loaded, saved or printed. The type of printer is set in the CFG (configuration) file.

Memory capacity
The largest number of bytes that can be saved is 40,000. The wanted, or expected lowest, number can be set as appropriate.

Speed info
This gives information on the length of time it takes for the important functions.INC and PORT IN to be executed, as well as on the approximate sampling frequency.

Stop
This terminates the program.

To determine the cycle time and the frequency, two help lines may be inserted into the measurement window with the mouse. This is done by pressing the left-hand mouse button at the start, hold it down and move the cursor to the stop position. As long as the mouse button is held down, a window is displayed with the cycle time and the corresponding frequency. Here again, the accuracy depends to a large extent on the specification of the PC.

auxiliary functions

The program also provides a number of auxiliary functions which can be selected with the relevant keys as follows.

<F1> switches the previously opened help window giving basic information and procedures to the program. When <enter> is pressed a second window with search masks is shown. The help window is disabled by pressing <ESC>.

<F2> starts the frequency measurement. It opens a window in which the frequencies in the four channels are shown online. <shift> <F2> does the same but only as far as the frequency in channel 1 is concerned. The measurement is ended by pressing <ESC>.

<F3> switches on the pulse counting function for channel 1. The pulses are summed until <ESC> is pressed.

<F4> switches on the pulse interval measurement for channel 1. The interval is displayed online until <ESC> is pressed.

further comments

In the configuration file LOGIC_AN. CFG there are program parameters that can be modified with any ASCII editor.

1 :Printer No.: [1, 2]
1 :Density of graphics print (1–3) (only in the case of the Epson ink-jet printer).
2 :[1]=Epson (9/24 ink-jet); [2]= HPPCL printer (e.g., Laserjet II).

The address of the status register (port address + 1) of the used printer interface must be entered into address file ADR.DAT. The relevant hexadecimal number must be preceded by the $ sign ($379 for LPT1 and $279 for LPT2).

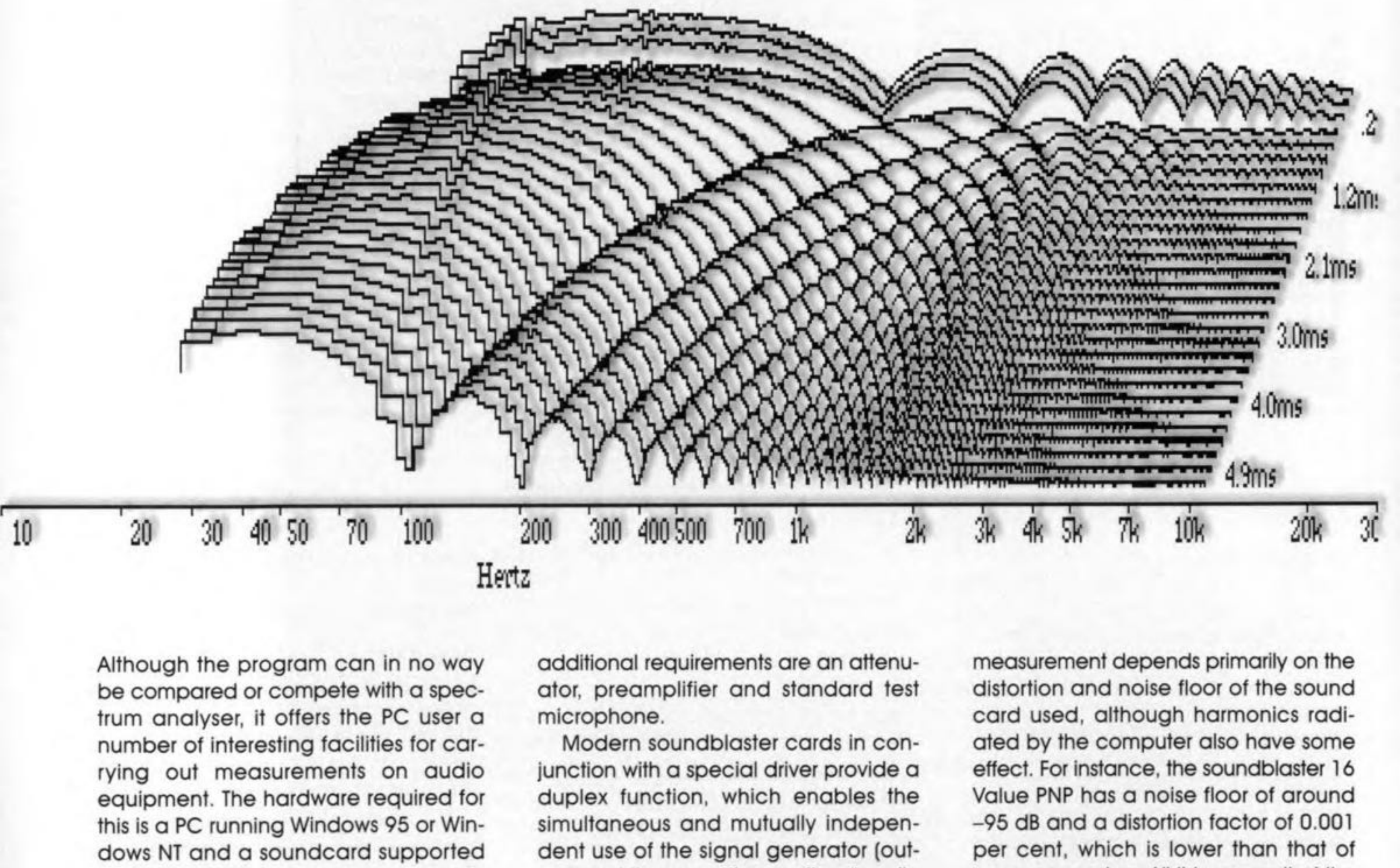
[972041]

"Cool Edit", a program from Syntrillium Software Corporation, was originally developed for the recording, processing and analysis of sound. It therefore contains a large number of assembly functions, filters and effects. In the course of time many other facilities have been added so that today the program is very suitable for carrying out measurements on sounds.

By Cees Ruytenberg

"cool edit"

software for sound processing and analysis



Although the program can in no way be compared or compete with a spectrum analyser, it offers the PC user a number of interesting facilities for carrying out measurements on audio equipment. The hardware required for this is a PC running Windows 95 or Windows NT and a soundcard supported by Windows. The properties and facilities of the program depend entirely on the type of soundcard.

The simplest setup is through the use of an inexpensive soundcard that is triggered by an external source, such as a test signal from a signal generator. Depending on whether the measurements are acoustical or electrical, additional requirements are an attenuator, preamplifier and standard test microphone.

Modern soundblaster cards in conjunction with a special driver provide a duplex function, which enables the simultaneous and mutually independent use of the signal generator (output) and the recording section (input).

To obtain a better signal-to-noise ratio, it is possible (with Windows 95) to use two soundcards that operate on different I/O (input/output) addresses and DMA (direct memory access) channels. In this setup, one card is used as generator at all times and the other as measuring card. The accuracy of the measurement depends primarily on the distortion and noise floor of the sound card used, although harmonics radiated by the computer also have some effect. For instance, the soundblaster 16 Value PNP has a noise floor of around –95 dB and a distortion factor of 0.001 per cent, which is lower than that of many generators. All this means that the properties of the soundcard do not set a limit to many applications.

noise suppression and spectrum analysis

The additional facilities of the program are particularly valuable in the case of very precise measurements. A typical example is the process of noise suppression. In this, the statistical content of a recorded signal can be ascertained, and stored or erased. This may concern a recording of an instrument amplifier with short-circuited input, or a recording of the quiet just prior to the arrival of a test signal. The information obtained in both cases may be superimposed on the test signal and subsequently subtracted from it so that a clean test signal without spurious elements remains. This may result in a noise reduction of up to 20 dB without loss of quality. In the case of weak signals, this may make it possible to measure well below the noise floor (down to −120 dB). The associated window is shown in **Figure 1.**

The spectrum analysis function offers several interesting possibilities. The amplitude scale (in decibels) is divided according to a manual setting. The frequency scale is determined by the selected sampling rate and may be displayed linearly or logarithmically, as required. The number of FFT (fast Fourier transform) points may be set between 2^7 (128) and 2^{16} (65536).

When the program starts for the first time, the spectrum analysis is displayed in a window the size of about a quarter of the monitor screen. The window may be enlarged to full screen size in the usual manner with the cursor and the left-hand kob of the mouse to make reading the scale more accurate.

Figure 2 shows an example of an analysis of the output signal of an external sine wave generator at a frequency of 1000 Hz (FFT= 8192 points). The second and third harmonic at −90 dB and −84 dB are clearly visible. After the data has been accepted, downsampling and vertical scale adaptation are carried out. When the mouse pointer is placed within the window that shows the spectral content, the frequency and amplitude for that point are displayed. If part of the recording is selected with the editor, the analyser measures and displays the spectral content at the centre of the selected section.

All frequencies occurring within the entire selected section are measured with the function 'scan'. In the presence of noise, this results in a levelling of the curve and gives a good display of the transition in the case of measurements with a sweep signal. 'Scan' is, in fact, the differentiation of a number of spectra that fall within the selected time interval.

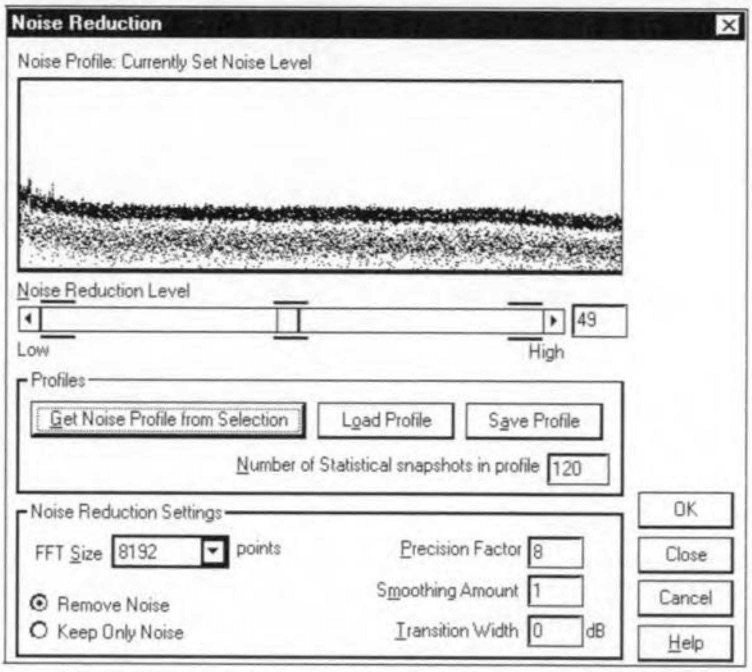


Figure 1. This window provides the noise reduction facilities. The profiles may be saved for later use.

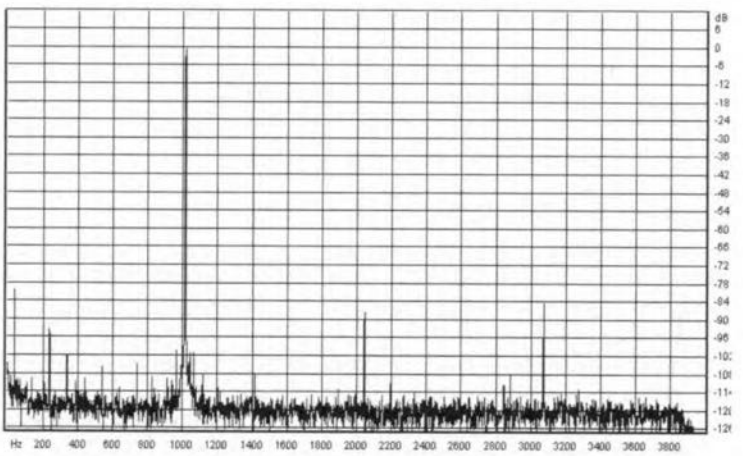


Figure 2. Harmonic distortion of the 1000 Hz output signal of an external generator (FFT=8192 points).

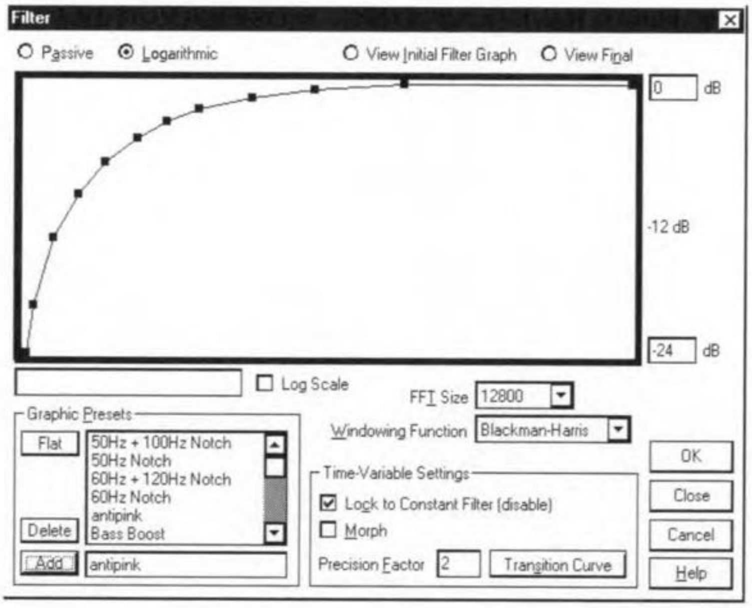


Figure 3. Correction curve to linearize the descending characteristic of pink noise (saved as antipink).

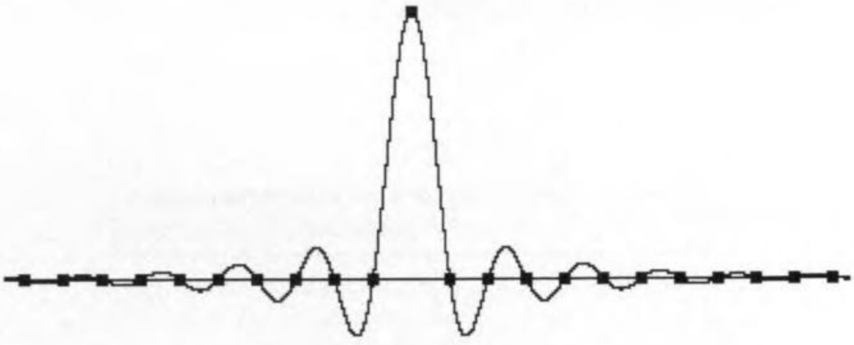


Figure 4. Approximation of a sin(x)/x function. The sampling points are indicated by squares.

filter functions

Specific interference, such as hum, can be eliminated with the filter function. This function enables the accurate setting of the rolloff and the frequency. The number of FFT points on the filter response characteristic may be chosen on the basis of the wanted accuracy. Unusual rolloffs and a number of filter types, such as pass-band, stop-band, notch, peak, low-pass, high-pass, or combinations of these, are available with the graphics interface.

Spectrum analysis will show a descending characteristic when measurements using pink noise are carried out. This results from the decreasing energy contained in the signal with rising frequency. This aspect should be borne in in mind when loudspeaker systems are being tested. Fortunately, with "cool edit" it is possible to correct the slope of the characteristic after the measurements have been taken. The relevant correction curve is shown in **Figure 3**; it may be stored permanently and recalled at all times.

Curves for linearizing the measurement system may be obtained in a similar manner.

generator function

The generator function enables almost any kind of waveform to be produced.

In the first place, quiet may be generated for any length of time depending on the available space on the hard disk and on the RAM.

Signals such as a burst of sin(x)/x pulses, or a constant tone – rectangular, sinusoidal, sweep – may be generated for a given time. When such a signal is replayed, it is available at the output of the soundcard. It may be replayed continuously by replaying it in the loop mode. Therefore, signals need not be generated for longer than a second. By setting the duration intelligently or by editing the ends of the waveform suitably, a seamless, endless signal may be obtained. This is also possible with unusual waveforms.

As an example, see the graphical representation in **Figure 4** of approximating a sin(x)/x pulse whose first pole lies above the audio range. For this, a sinusoidal signal at a frequency of 10 kHz is selected for one-half period, that is, for 50 μs. In the case of a burst of pulses (see **Figure 5**) a certain frequency applies to one or more cycles inserted into a quiet of 1 s.

Depending on the measurement, normal replay or loop mode may be selected for the reproduction. Other choices are brown noise, white noise, and pink noise, for which the same processing options may be used.

When the generator function is used at the same time as the soundcard is used for recording (duplex operation), the sampling rates must be the same. Owing to the limited mixing facilities of some soundcards, it is advisable to generate via the right-hand channel and to use a mono configuration during the recording, which automatically causes the recording to be done via the left-hand channel. This arrangement ensures that crosstalk is minimized.

When very clean measurements are needed, it is advisable to use a second card, an external generator, or a second computer with suitable soundcard.

After the desired waveforms have been generated, samples may be picked up with the mouse and placed elsewhere (see **Figure 6**). This gives a wide choice of possibilities to manipulate waveforms or create new signals.

downsampling

Downsampling, that is, 'convert sample type', enables the resolution of spectral measurements to be enhanced greatly.

Measurements taken at a high sampling rates (44.1 kHz or 48 kHz) contain much unnecessary detail if analyses of low frequencies are required (1 cycle at 100 Hz takes about 400 samples). Unfortunately, in FFT analysis the range is divided linearly so that a measurement of 1024 at the lower end of the range leads to coarse steps of about 20 Hz. When the measurements points are then linked, the resultant curve, if plotted on a logarithmic scale, will show a number of 'kinks'.

True, the number of FFT points may be increased to 2^{16} (65536) resulting in frequency steps of ony 0.3 Hz, but this has the disadvantage of causing long computation times when a slow computer is used, or the 'scan' function is selected.

The use of downsampling results after anti-aliasing and conversion of the samples in a new waveform from which all high frequencies, except those lower than half the sampling rate, have been

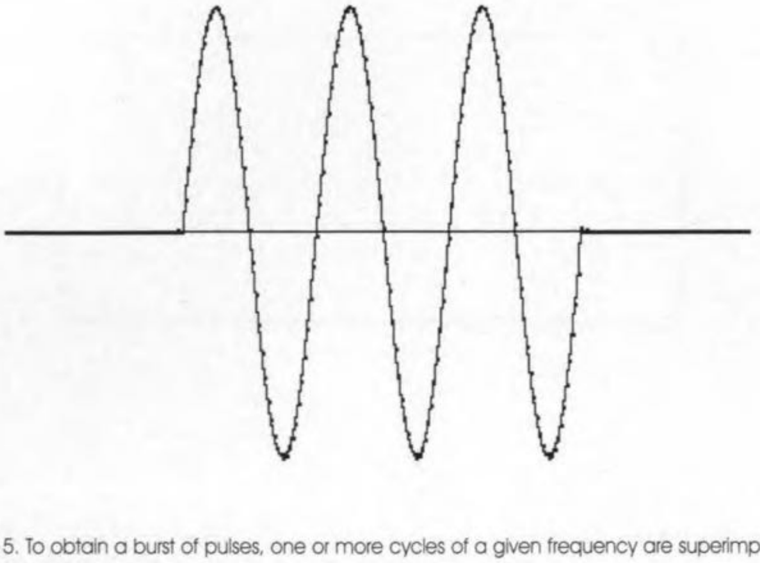


Figure 5. To obtain a burst of pulses, one or more cycles of a given frequency are superimposed on to a quiet of 1 second.

removed. The scale of the spectrum analyser adapts itself to this, so that low frequencies are more clearly displayed.

In this way, a sampling rate of 6000 Hz results in a frequency scale of not more than 3000 Hz, so that in case of an FFT of 1024 points the resolution is 2.9 Hz. In downsampling, there is no obligation to use the sampling rates allocated by the manufacturer, because the soundcard is not really used for the measurements. These are just arithmetical operations that after computation can be represented by a waveform or range of frequencies. Therefore, if downsampling down to 400 Hz or even lower is selected, low frequencies can be read accurately (as required, for instance, during the measurement of a woofer in its enclosure).

transfer measurements with a logarithmic scale

The logarithmic scale enables the user to carry out transfer measurements of which, in contrast to the linear scale, all frequencies can be read clearly. However, the discrete measurement points at low frequencies can be seen, so that for a satisfactory measurement not fewer than 2^{12} (4096) test points are needed. When a very high resolution at low frequencies is wanted, downsampling may be used.

As an illustration, consider an experiment in which three different test signals are applied to a 17 cm woofer (see **Figure 7**). The inputs are measured with the aid of the duplex function of the soundcard, combined with the generator functions in the software, and stored in a file.

All signals were applied to the system at such a level that the linearity was not jeopardized. The soundcard was driven to just below the saturation level (up to 16 bits). The pink noise was converted to white noise via an electronic correction network so as to obtain a real measurement (this would also have been possible with filter setting 'antipink'). After this, the 'scan' function was used for the selected 2 seconds.

In the second case, a linear sweep was applied for half a second, after which the 'scan' function was used to obtain the transfer characteristic.

In the third case, a sin(x)/x pulse was applied. Since only a single pulse was used, a quiet of a few seconds was measured (test signal removed from the loudspeaker), after which the test signal was applied to the woofer. Additionally, a record was made of the quiet, which was then saved with the functions of 'noise suppression'.

Subsequently, the program computed the statistical content of this signal. The resultant signal was then superimposed

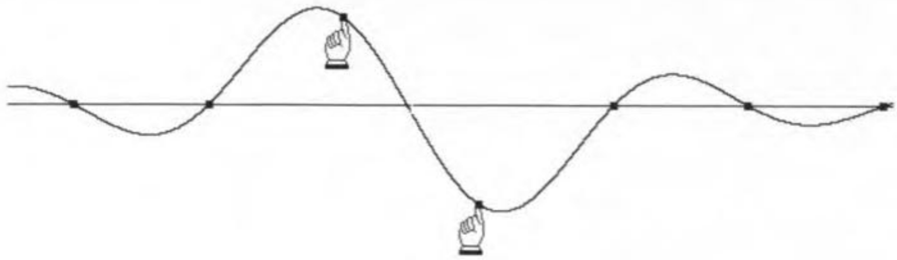


Figure 6. The separate samples may be picked up and relocated with the mouse.

on to the test signal and subtracted to yield a clean test signal. (This method is much faster than that of other systems in which typically 100 measurement results are superimposed on to one another and then differentiated). Afterwards, the time window was superimposed on to the test signal and closed before echoes from the measurement range could distort the measurement result.

As will be seen in Figure 7, the energy content (signal-to-noise ratio) of the pink noise measurement is clearly the highest of the three, followed by the linear sweep. The sin(x)/x method is the worst as far as energy is concerned. However, this gives more information about the system being tested since spurious reflections from the measurement range are removed (the time window determines the purity of the measurement and the lowest frequency to be measured).

This signal may also be used to analyse the pulse response.

finally

In taking measurements, the knowledge, experience and insight of the operator are just as important as the quality of the hardware: the soundcard, preamplifier, and test microphone.

The facilities of 'cool edit' will, no doubt, be extended and enhanced in fulture versions. Therefore, any feedback from present users to the manufacturer is of great value. Internet users may obtain more information and an evaluation copy from
http://www.syntrillium.com

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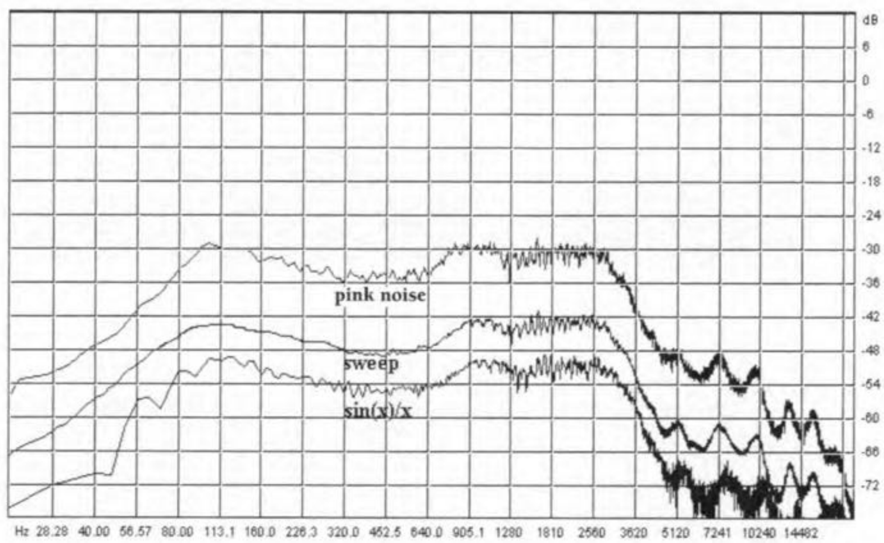

Figure 7. The result of three measurements on a 17 cm woofer with different test signals.

Until not so long ago, triggering was used only in oscilloscopes to make certain parts of a signal visible. With the advent of new measuring instruments, new trigger techniques have had to be developed. Today, apart from in oscilloscopes, special trigger techniques are used in counters, scopemeters, and graphical multimeters. This article discusses several trigger techniques used in modern measuring instruments from Fluke*.

Fluke (UK) Limited

trigger techniques

no longer confined to oscilloscopes

The function of a trigger circuit in a measuring instrument is simply to find a reference point in a repetitive signal at which the time base (that is, the measurement) can be started. The user may add a number of peripheral conditions. **Figure 1** shows basically how triggering is achieved in an analogue oscilloscope. Not only is the voltage level at which a trigger pulse must be output variable, but the polarity of the signal edges, rising or falling, is selectable. If desired, a hold time may be observed after the trigger point has been reached.

There are variants of the basic trigger circuit that take the type of input signal into account. Even analogue oscillo-

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scopes are equipped with TVL(ine), TVF(rame), LF-rej(ect), HF-rej(ect) facilities on the trigger control, which enable certain frequencies to be filtered from the signal, so that triggering at other, wanted, frequencies is more readily achieved.

Additional functions are AC and DC coupling with which the direct voltage component in the input signal may be passed or blocked. Because of these functions, the user is able to determine exactly where the time base of an oscilloscope is to be triggered.

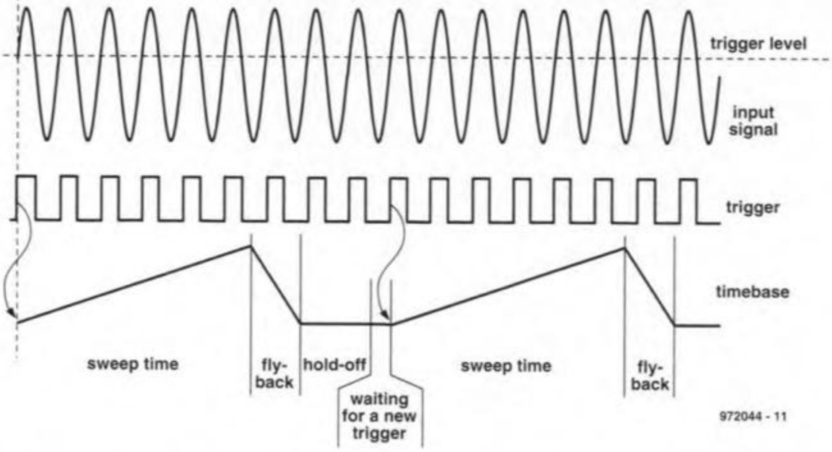


Figure 1. Triggering is a central function in an oscilloscope. Today, trigger circuits are also found in some multimeters and counters.

digital oscilloscopes

Most digital oscilloscopes feature pre- and post-trigger delay to enable a signal to be analysed at both sides of the trigger point, but the Fluke ScopeMeter has two additional delay modes. The first is delay by events and the second is delay by *n* cycles.

Consider the latter first. When using an oscilloscope, you will run across waveforms that just will not trigger well. At times, they appear stable, but at other times they jitter across the screen. A typical example of such a signal is shown in **Figure 2**. If triggering would take place at every zero crossing, the result would be a jittery signal in which various waveforms would be superimposed on to one another. In this case, triggering with a delay of *n* cycles is used. With this, an instruction is given to wait for *n* cycles before another trigger pulse is given. If in Figure 2 $n = 3$, triggering will take place at identical pulses.

Meters that are used for the analysis of automotive electronics systems often have to cope with complex repetitive signals in which, for instance, each 4th, 5th, 6th or 7th pulse is missing. Here also, the *n*-cycle trigger mode comes to the aid of the user, because there is a risk that automatic triggering (autoset) will lead to chaos. The delay with this mode renders the image stable again. For instance, if the 4th pulse is missing, choose a delay of 3 cycles.

In the event mode, use is made of two input signals: the master trigger and the event trigger, which is used to increment an event counter. In this mode, it is, of course, the intention that the two signals are in some way related to one another. A rather similar situation occurs when the armature voltage of a motor is connected via an attenuator box to a generator.

Therefore, the measuring instrument can be informed that the master trigger is enabled only after a certain number of events has taken place. This complex and relatively tedious way of triggering may be used in exceptional cases to obtain a stable image on the screen.

peak detector

The input voltage in a digital oscilloscope is sampled by an analogue-to-digital converter (ADC), which at predetermined intervals measures and quantizes the value of a voltage. Provided that the sampling rate is high enough, it is possible to reconstruct a signal. Normally, at least 5–10 samples per input cycle are needed to obtain an acceptable representation of the input waveform.

The only problem that remains is that spikes (very short pulses) remain unde-

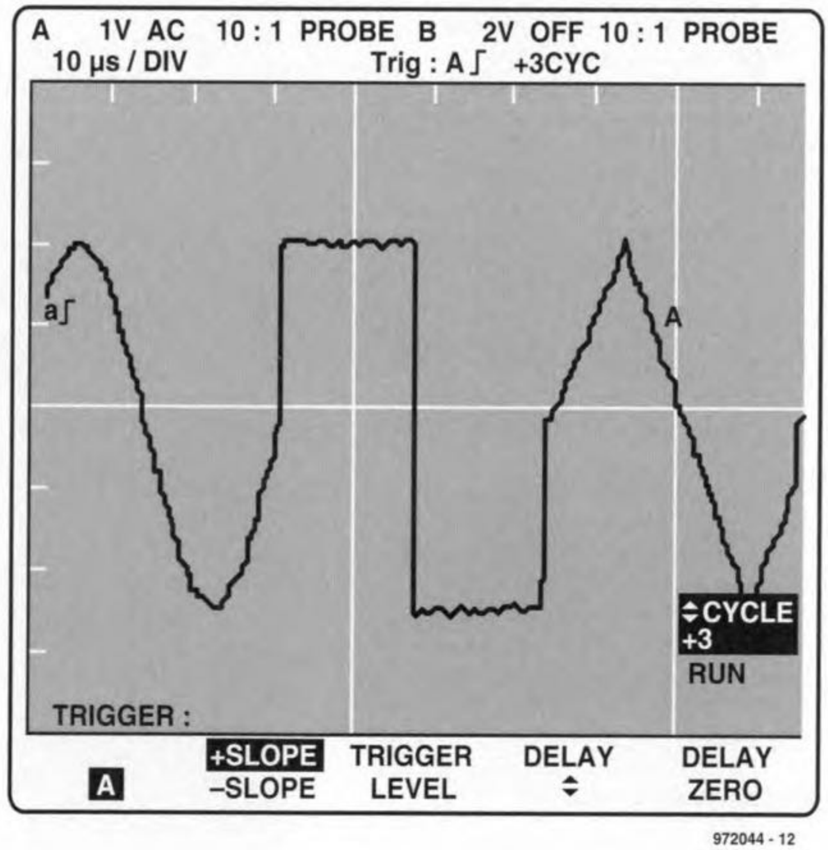


Figure 2. The n-cycle function enables the display of complex, repetitive signals on the screen. Spurious signals are ignored.

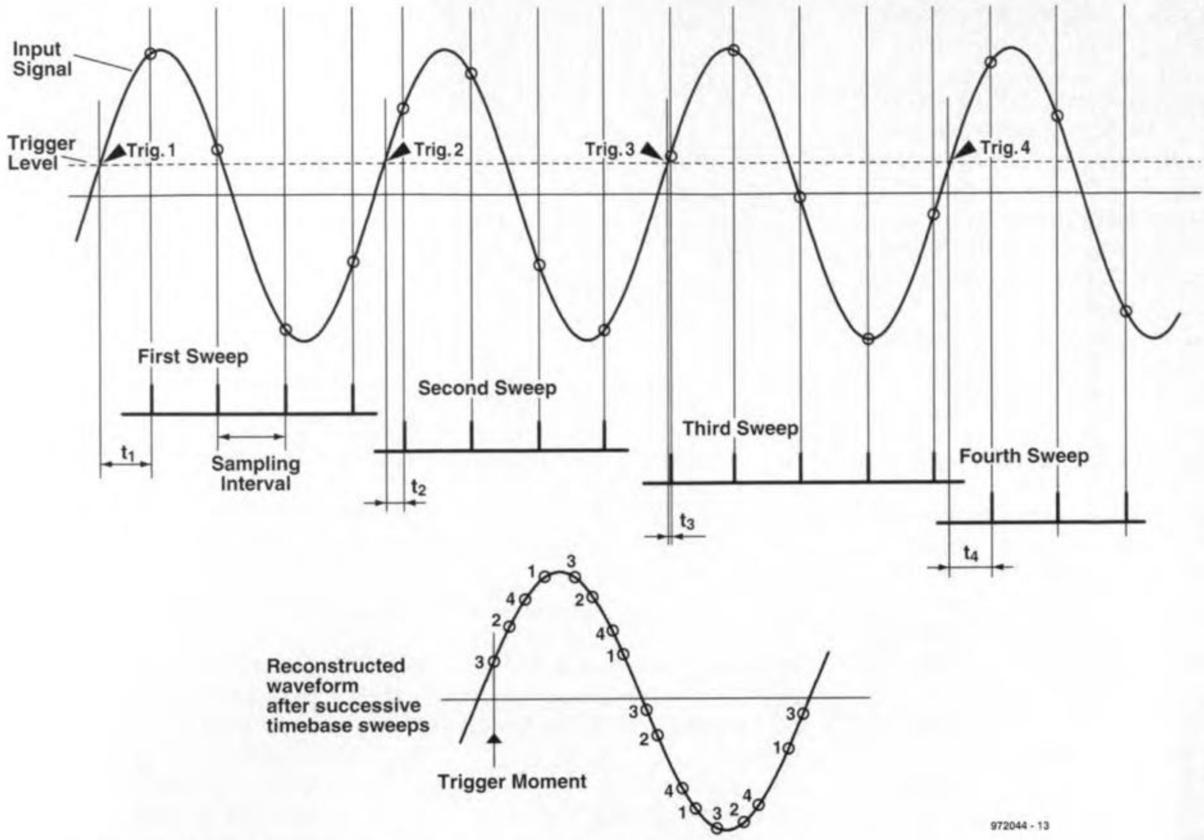


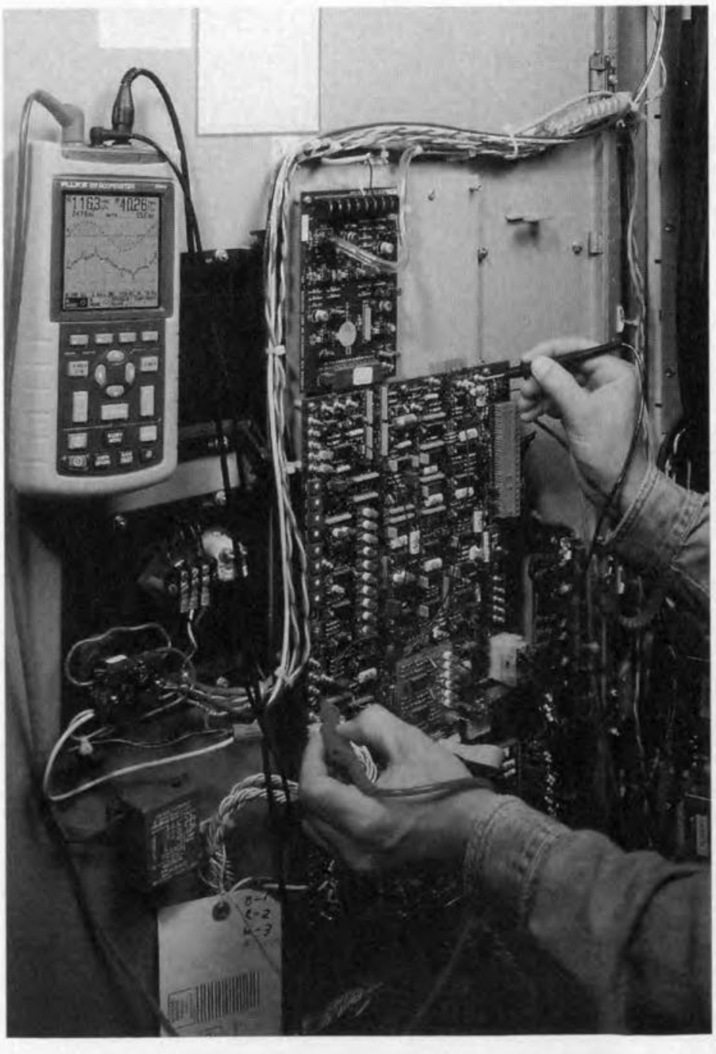
Figure 3. Repetitive sampling of a periodic quantity makes high sampling rates possible.

tected. If these occur between the samples taken, they will not be visible on the screen. This problem is obviated in a modern oscilloscope by a good-quality digital peak detector. This circuit samples the signal at a high rate, say, 200 megasamples s^{-1}, which gives a resolution of 5 ns.

Using repetitive techniques on a stationary signal in which additional measuring points are gathered, an equivalent or effective sampling rate of one or several gigasamples s^{-1} is possible (see **Figure 3**). This means that the time resolution is of the order of 1 ns. The limited memory depth in which samples are stored means that this high resolution is only possible at the fastest timebase setting. At lower timebase settings, the sampling rate is reduced to a value at which the memory can store all the samples corresponding to that time base setting. A limitation of repetitive sampling is therefore that signals with a low duty factor, particularly narrow pulses, are difficult to analyse, unless the earlier mentioned peak detector is used.

counters, a case apart

Today, oscilloscope functions are integrated not only in graphic multimeters, but also in multifunction counters. Such a counter measures multiple time inter-



vals, all running from a unique start trigger point of the signal to several stop trigger points at various trigger levels. The stop trigger point is incrementally scanned over the input signal. The multifunction counter then measures the successive time intervals.

Interesting examples of such counters are contained in the Fluke 160 Series of MultiFunction Counters. Where in an oscilloscope a test signal is sampled horizontally, that is, in the time domain, in these counters sampling takes place vertically, which is called voltage sequential sampling. In this, the time intervals between different voltage levels are measured. The resolution of the measured time interval is always <1 ns; not just at the fastest timebase setting as with traditional DSOs. The magic here is that vertical sampling inherently produces a sampling density over time that depends on the need to capture signal variations.

See **Figure 4** for a comparison of the two modes of sampling.

The designers of the counter have turned the world through 90° to better digitize timing signals. Although this technique appears expensive since use cannot be made of standard components, part of the extra cost of the requisite trigger and sampling circuits is compensated by the much simpler design of the display driver circuits.

In vertical sampling, a large number of trigger levels are used and measurements taken of the time intervals when a given trigger level is exceeded or not met. For this it is necessary to use a repetitive signal so that multiple measurements, each at a different trigger level, can be taken.

The interesting point about this technique is that by definition time intervals are measured that match the parameters of the input signals. If no signal variations occur, for instance, at the flat top of a pulse, no sampling takes place. In other words, samples are taken and stored only if they contain factual information.

A dot joiner is used to connect two measurement points on the screen to render the constant voltage visible.

The output signal in vertical sampling is a two-dimensional array that contains a number of measurement points that describe the waveform. Each point in the array contains information as to time (*x* coordinate) and as to the signal level (*y* coordinate).

The resolution of the voltage measurements. depending on the setting of the internal attenuator, is 1, 10, or 100, which is more than sufficient for most applications.

It is clear from the foregoing that this trigger and sampling technique is eminently suitable for the detection and measurement in an inexpensive manner of narrow pulses and glitches in repetitive analogue and digital input signals, such as in a counter. Another application where this is of great use is in the calibration of radar pulses. Such pulses are 1 μs wide and have a PRF (pulse repetition frequency) of 1 ms, resulting in a duty factor of 1:1000.

connect and view

The 'autoset' function in modern analogue and digital oscilloscopes has simplified triggering immensely. When this function is selected, the incoming signal is analysed and disassembled into its constituent components. On the basis of this, the optimum setting is selected automatically.

In the latest generation of Fluke's scopemeters, this function has been enhanced by the addition of 'connect and view'. This automatically ensures that as soon as a signal is applied to the input terminals an autoset instruction is given. Any change in the input signal automatically actuates a new autoset instruction.

This facility enables the triggering of signals at frequencies down to 1 Hz. Its chief advantage is that no control needs to be touched to obtain good plots of the signal on the screen. This is especially useful when measurements are taken in difficult conditions.

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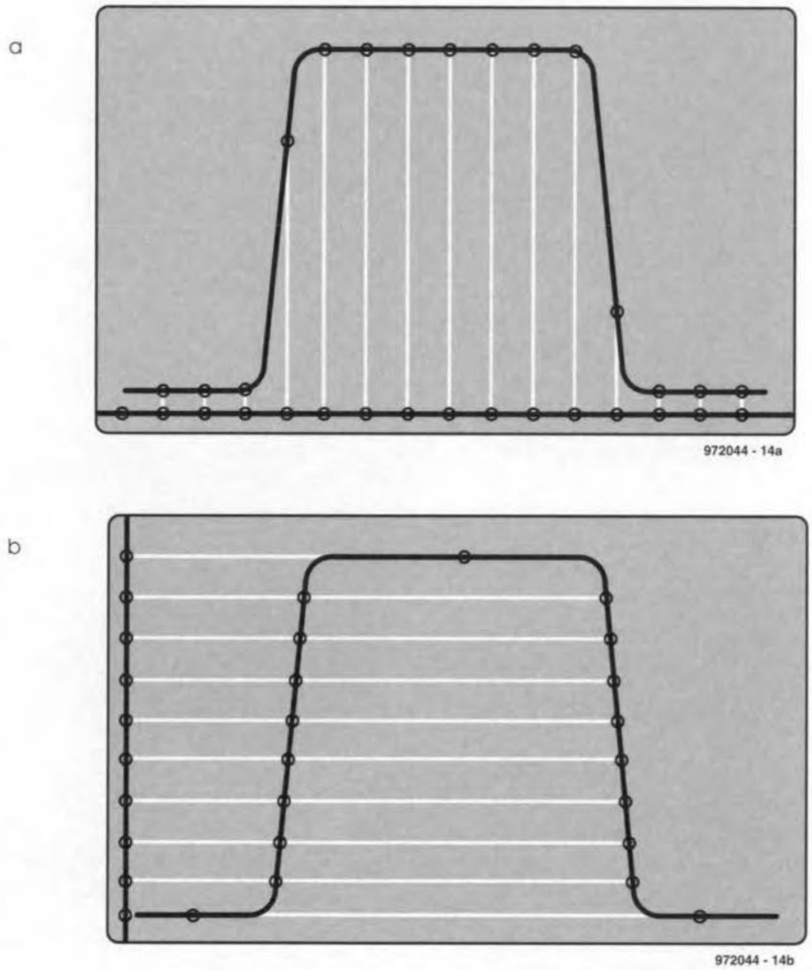


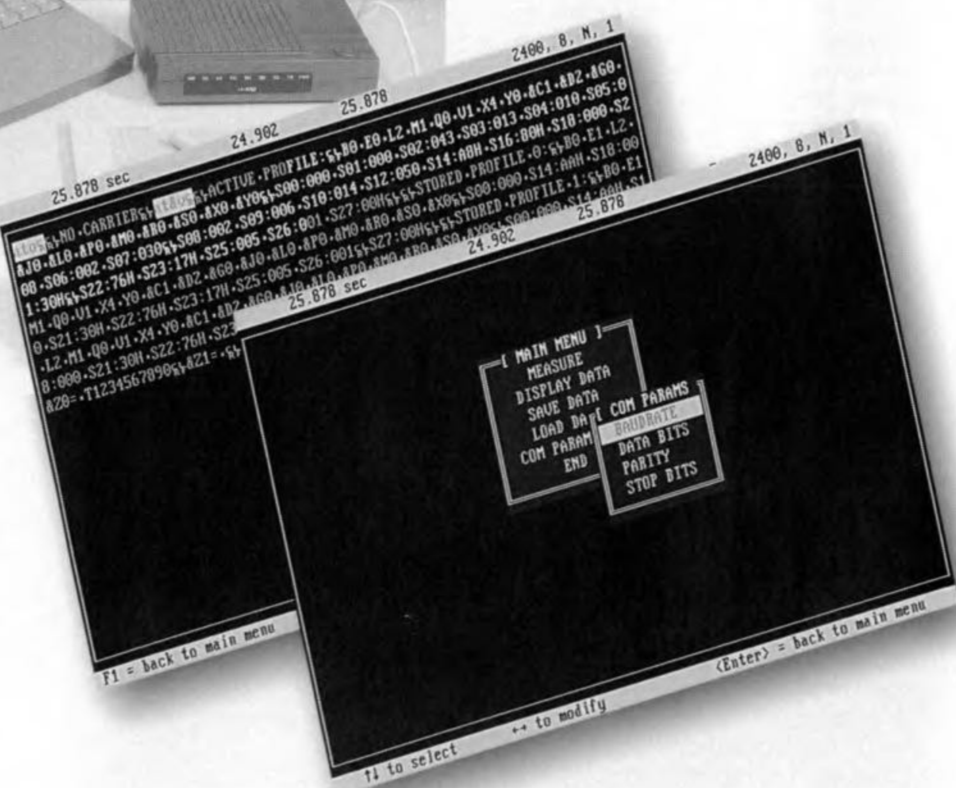
Figure 4. Comparison of (a) conventional horizontal sampling, and (b) vertical sampling.

"ASCOPE" is a computer program that enables the transfer of asynchronous signals between two pieces of equipment to be monitored and, if required, saved for later analysis. In this, use is made of the standard serial port(s) COM_1 and COM_2 of the PC. Other than a simple test cable, no specific hardware is needed.

By J Clerx

"ASCOPE"

measuring analogue signals with a PC



There are not many simple aids available for monitoring digital traffic over serial links, such as an RS232 link. In this article, an IBM compatible PC is considered of which one serial port is connected to the unit whose signals are to be controlled or inspected. If a second serial port is available, this may also be used to enable the traffic in both directions to be monitored.

The test cable is a very simple adjunct (see **Figure 2**).

ASCOPE (acronym for asynchronous oscilloscope) is a DOS program specially written for this application. It arranges the correct test settings and the display of the measured signals on the screen. The program is written largely in Turbo C from Borland. Only a few, time-critical, routines are in assembler (MASM).

properties of the program

Facilities provided by, and properties of, the program are as follows.

• Simple operation via selection of

Figure 1. The "ASCOPE" program measures asynchronous (RS232) signals and displays them on the screen of the monitor. Afterwards, they may be saved as a data file.

menu-listed items.

- During the measurement there is an indication of the time, in seconds, elapsed since the onset of the measurement cycle.
- Simultaneous measurement in two directions. The direction is indicated by 'normal' or 'negative' colour display on the screen.
- For each character, in both directions, the time is registered with a resolution of 1 ms.
- Control characters, such as STX (start of text), ETX (end of text), LF (line feed), etc., are displayed by means of an adapted character set.
- The measured data may be analysed after the measurement has been completed. The cursor may be moved freely through the measured data with the aid of PgUp, PgDn, and the arrow keys. The time of registration is displayed at all times with a resolution of 1 ms.
  Cyclical buffers ensure that the last measured characters are retained for further analysis.
  Data may be saved under an arbitrarily chosen name.
- Start and stop markers may be placed to enable only the important parts of information to be saved.
- Earlier saved data are available for analysis at all times. Markers may be placed on these to enable only those of interest to be recalled for a future analysis.
- Parameters such as baudrate, parity, number of databits and number of stopbits may be set via the menu.
- Status lines at the top and bottom show a number of data, such as available keys.

operation

The program is started simply by the instruction

ASCOPE <enter>,

whereupon the first window with general information appears. Pressing a key causes the main menu to be displayed. The arrow keys enable the desired selection to be made and executed via <enter>. The choices then available are as follows.

Measurement

When 'measurement' is selected, the measurement is started. However, if there are data in the buffer that have not yet been saved, a message is displayed asking whether these data should be saved for later use.

When the measurement is started, each character that passes the measurement line is immediately displayed on the screen. The direction of the data stream determines the kind of display (normal or high intensity).

The correct setting of the parameters is described in COM parameters later on. The appropriate value of the parameters is shown at the right-hand side of the upper status bar.

New data are constantly added at the bottom, while older ones are removed from the screen when this is full by scrolling.

Measurements are ended when the F1 key is pressed, whereupon the main menu is displayed again.

Display data

When display data is selected, the data that are in the buffer (via Measurement or Load data) are displayed. The cursor may be moved through these with PgUp, PgDn and the arrow keys.

The upper status line shows the time, in seconds and milliseconds, of the first character in the buffer, the last character in the buffer, and the character on which the cursor is placed. This may be very useful when it is desired to measure the exact response times in the communication between the two

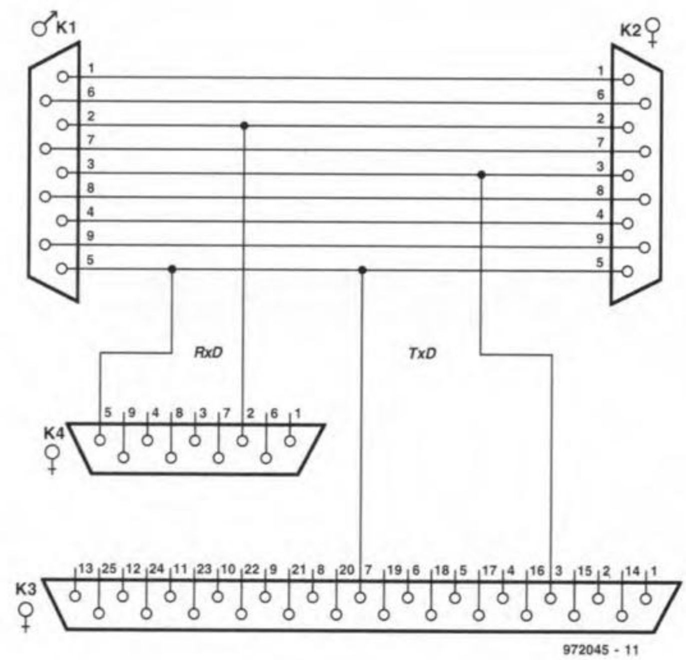


Figure 2. Construction of the test cable. See Figure 3 for other types of connector.

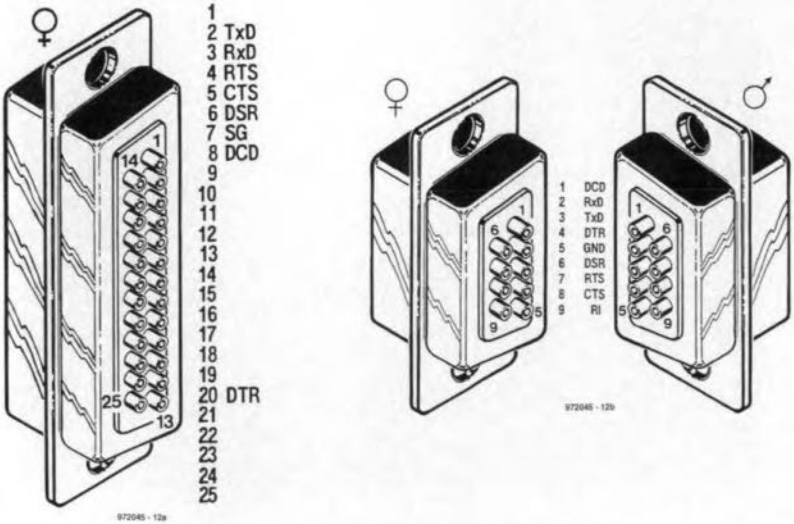


Figure 3. Pinouts of a 9-pole and a 25-pole RS232 connector.

pieces of equipment.

A certain portion of the data may be marked. To do this, place the cursor at the beginning of the portion to be marked and press F6. Then, place the cursor at the end of the portion of the data to be marked and press F7. The portion so marked is shown markedly different. When these data are saved (see 'save data'), only the marked portion will be saved.

Pressing F1 reverts to the main menu.

Save data

With 'save data' selected, it becomes possible to save information contained in the buffer. The name, and if desired, the name of the disk and name of the pad, have to be entered. If a portion of the information has been marked, only that portion will be saved.

Load data

When 'load data' is selected, data already on the disk will be loaded anew. Apart from the name of the data, the name of the pad and that of the disk on which the data is stored may be entered.

COM parameters

When 'COM parameters' is selected, a submenu will be displayed. This contains the various parameters that are to be set in accordance with the settings of the signals on the link which are to be measured. This means that the baudrate, the number of data bits and stopbits, as well as the parity setting must be known. Selection is made with the keys marked with the up and down arrows; modifications are made with the keys with left-hand and right-hand arrows. The selected value is shown at the upper status bar. The submenu is replaced by the main menu with <enter>.

End

When 'end' is selected, the program is terminated.

hardware needed

To use the datascope, a PC and a suitably adapted test cable terminated in Sub-D connectors are required.

When a PC has two serial ports, one of these normally has a 9-pole connector and the other a 25-pole connector. It is obviously advisable to use one type of connector for the two measurement links (both female). The interconnecting cable is provided with a male and a female connector, so that it can always be used as an extension cable.

It is also possible to use an adaptor of which a variety of different kinds is available. The adaptor is inserted between one of the pieces of equipment and the cable. The link to the measurement PC is via this adaptor. The adaptor may be 9-pole-to-25-pole or vice versa to adapt the cable for the kind of measurement to be carried out.

The "ASCOPE" program can be found on the CD-ROM *Electronics software 96-97*, Order no. 976003, available through our Readers Services found towards the end of this issue. The relevant files are found in NL/23.

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