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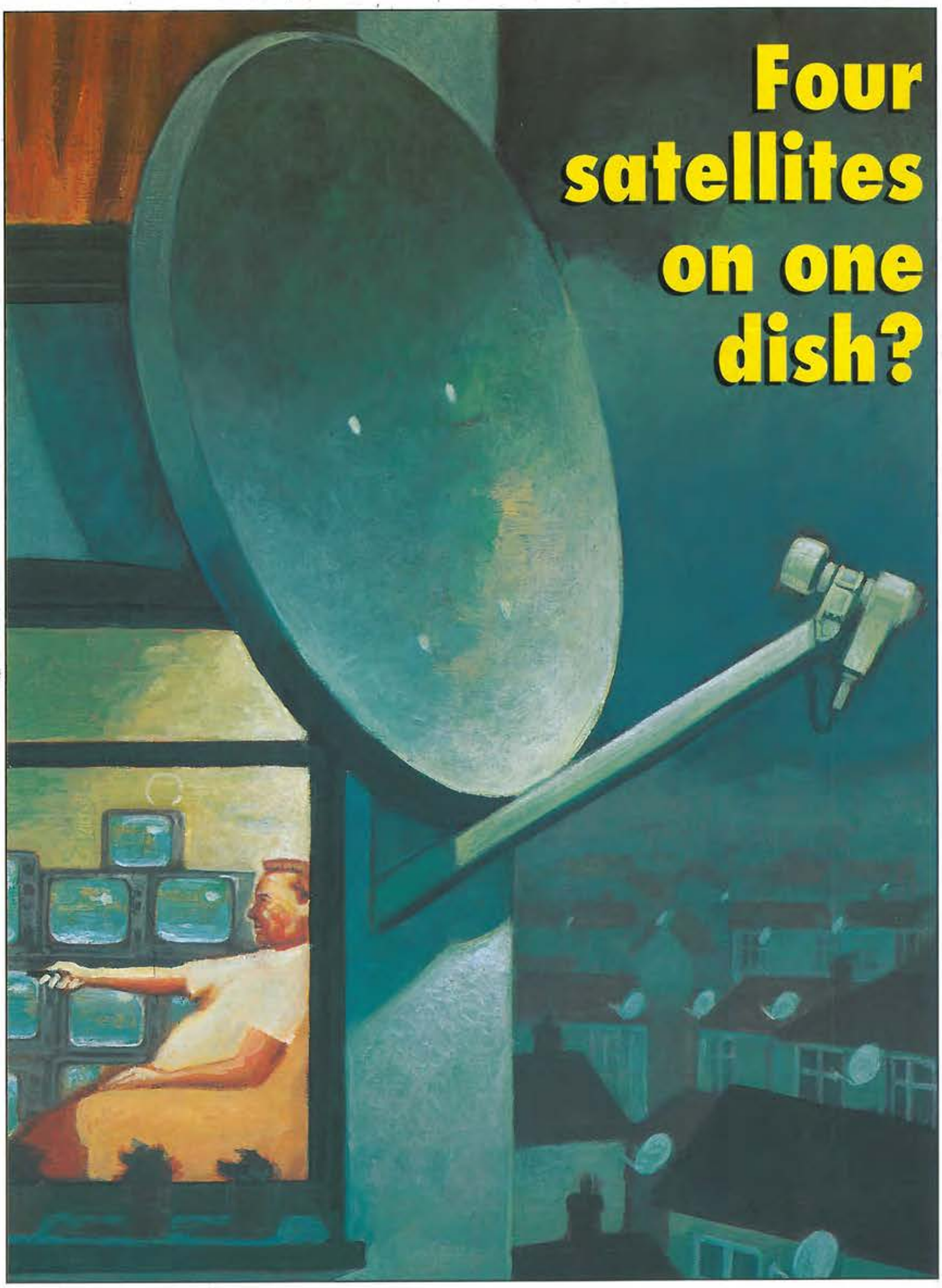

Four satellites on one dish?

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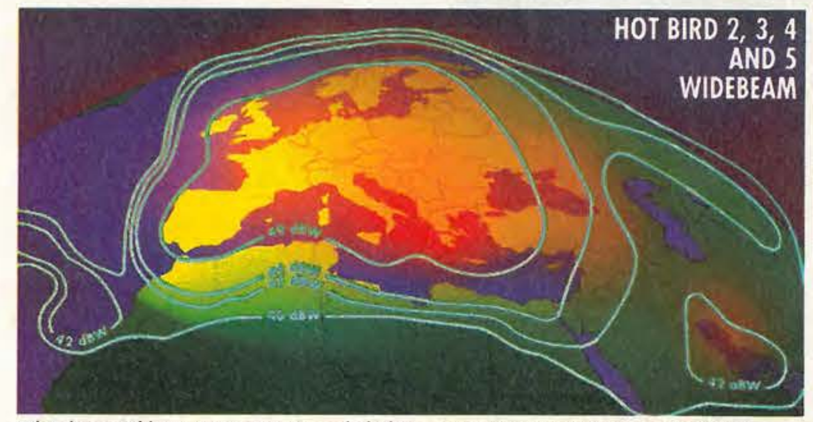
Contents

Cover – Hashim Akib

Reader offer

Electronics World readers: obtain a 25% discount on this sturdy dual read-out bench power supply. *Turn to page 224.*

Load this month's free CD-ROM into your PC or Mac to find a wealth of dsp information, including not only data, but also applications and software tools (sorry ROW – UK readers only).



Thinking of buying a motorised dish to increase you viewing options? Take a look at Cyril Bateman's alternative first.

Regulars

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REED BUSINESS INFORMATION

Devices for increasing the range

I sometimes wonder how many original *Wireless World* readers are still around these days. Plenty I hope, like me who enjoy the current 'combination magazine', but still remember the original *WW* as an inseparable part of the engineering lab scene of the sixties and seventies.

It had a great blend of wisdom and a caustic – sometimes irreverent – wit, along with superior test gear projects for enthusiasts. Test gear that actually worked and could be built in the lab as 'homers'.

So on the bench in my loft I still have a working *WW* signal generator with frequency meter and an audio millivoltmeter. Testimony to my misdirected – from my Employer's viewpoint – zeal and the need in those days for making what you definitely could not afford to buy. Products from Global low-cost markets had not yet arrived.

As for wit and wisdom, they frequently came in equal measure from the prolific pen of 'Cathode Ray', who was actually MG Scroggie, the celebrated author of classics such as the 'Foundations of Wireless' and the 'Radio Laboratory handbook'. I found a particularly thought-provoking paragraph on the first page of the former, entitled, 'What Wireless Does', as follows.

> *'People who remark on the wonders of wireless seldom seem to consider the fact that most people can broadcast speech and song merely by using their voices. We can instantly communicate our thoughts to others, without wires or any other visible lines of communication and without even any sending or receiving apparatus outside of ourselves. If anything is wonderful, that is. Wireless, or radio, is merely a device for increasing the range...'*

Blimey! What a lot of juice in that lemon. For a start, my staff and I – and anybody else for that matter – on the same office site can jolly well relearn to talk to each other face to face, rather than using electronic mail. They will probably get a real message across quicker!

We can have better communication in general. Managers to their staff , people to people. We might even become interesting and entertaining speakers again if we rediscover this 'wonder' and stop being electronic 'paper tigers'.

Secondly, and on the other hand, what an incredible saga it has been – and still is – for the Electronics industries that have dedicated themselves to 'merely increasing the range.'

From the late twenties to the late fifties, every major country had its own radio industry: true engineering and style setting giants: major employers who dominated their semi captive markets.

The arrival of the transistor with its 'nomadic', one per person end equipment possibilities and world wide compatible battery supplies created an overnight global radio market with victory going to the lowest-cost, highest-quality designs. Europe and America fell to the East. Where are the great names now?

Almost forty years later and turning to the 'new age' consumer (nee communications) market, Europe bit back and has taken the world lead in digital hand-held phones and indeed in digital video and audio communications amongst others.

So somewhere along the road we have learnt to do things differently – and better – on the European scale. But how about the UK? Well, the great UK radio industry which had all but fizzled out in the sixties was largely replaced by generously funded defence work. And this was replaced by... what? Well. Ahem... By inspired take-over and investment from the 'colonists', both end-equipment and components.

The first wave came from the USA, then Japan, Korea, Germany and Scandinavia. What's their secret and what qualities have they brought to bear that we couldn't manage?

Having worked over thirty years in electronics and semiconductors for UK, American and Japanese companies I can hazard a few guesses. How about a pioneering spirit to develop new products for new markets? A monastic dedication to getting things right in detail and quality approach to keeping them right? A pragmatic, non hierarchical type of organisation that gives everybody something to go for, and rewards them for getting there? Detailed senior management involvement in the business to give knowledge and confidence to invest for the long haul?

There's a perennial 'stuck bit' in the UK electronics psyche. Anyway, lets be pleased that we are now a vigorous part of a multicultural, UK, European and world-wide electronics business which is constantly striving to develop new 'devices for increasing the range'. And let *Electronics World* maintain a critical watch that we succeed!

***Jim Duckworth**, Executive General Manager – Electronic Components Group, Hitachi Europe Ltd and Deputy Managing Director – Hitachi Europe Ltd.*

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UP DATE

Mobile phones – safe or not?

Mobile phone manufacturers and governmental research agencies have been criticised by a UK researcher for their inadequate response to the growing – but as yet unproven – belief of the health risks of mobile phones.

Roger Coghill, who runs a bio-electromagnetics laboratory in Pontypool, believes the public should be warned explicitly about possible health risks from mobile phones in light of recent research.

Last week, Australian doctor Andrew Davidson published study results in the Medical Journal of Australia which show a rise of more than 50% in brain tumour cases in Western Australia between 1982 and 1992. Davidson believes this may be due to growth in analogue mobile phone usage during the period.

"Davidson has put his finger on something important,"said Coghill. "There is no doubt that there are some significant bio-effects from overuse of these devices. There are seven million mobile phone users in the UK all putting the most radiative electrical appliance ever devised next to the most sensitive organ in the human body."

So far, the European Commission is yet to carry out a research programme into the health risk of mobile phones because it has still to decide how it will be funded. Leo Koolen, of the EC's telecommunications directorate, told EW that the programme is likely to come under the European Union's Fifth Framework R&D programme, which will not begin until next year.

Ben Greenebaum of the World Health Organisation said that its international agency for research on cancer is planning a major epidemiological study, but that, "you're talking a number of years" before it is completed.

Meanwhile, Coghill has launched an independent court action against a local mobile phone distributor for contravening the 1987 Consumer Protection Act. "The law says that you may not sell a product that is unsafe unless [the manufacturer] warns the public of the risk," said Coghill, who wants to see warning labels put on all mobile phones.

62 000 components and products to go on-line

Catalogue distributor CPC is trialling an on-line 'Internet-related' catalogue service as part of a £10m expansion in its Preston-based business. The move is a response to growth in the company's consumer spares distribution business.

"There are exciting plans for new routes to market as well as new markets to look at in 1998," CPC's managing director Chris Haworth commented.

The company is believed to be acting as a pilot site for a BT on-line sales service. Haworth would not give any more details of the service except to say it would be introduced later this year.

Haworth decided some time ago not to introduce a cd-rom version of the company's catalogue – as other distributors had done. Instead, CPC would move straight to on-line transaction services.

As part of its six year expansion plan CPC has purchased land where it will build a 56 000 sq. ft warehouse, with fully automated order handling, this year. The six acre site, next to CPC's existing site at Fulwood, will have potential for further expansion of the company's warehousing capacity. A second 60 000 sq. ft warehouse is scheduled to be built after the year 2000.

"We have enjoyed a compound rate of expansion of almost 20 per cent for the past ten years," said Haworth.

In brief

New ferroelectric ram backer

Ramtron has added to the list of licensees for its patented ferroelectric ram, or FRAM.

Asahi Chemical Industry joins Fujitsu, Hitachi, Samsung, SGS-Thomson and Toshiba in licensing FRAM. The two companies will develop jointly the technology for Asahi to produce in densities up to 64kbit. Asahi's semiconductor business currently produces custom devices for mobile telecoms.

Products suited to use the memory include mobile phones, handheld pcs, industrial control systems and medical electronics.

The FRAM arena is forecast to be worth $10bn by 2000.

Lion's share for satellite tv

Digital satellite television will for a short while eclipse digital terrestrial television (dtt), forecasts analysts Datamonitor.

Even though there will be up to 1.4m European households with dtt by the year 2002, the satellite broadcasters will have control of the digital market with a 65% share and 13m digital subscribers.

Digital tv on a pc?

Hitachi and Intel are collaborating on developing technologies for pcs to receive terrestrial based digital television signals.

MMIC business acquired

TriQuint Semiconductor has signed an agreement to acquire the monolithic microwave integrated circuit (MMIC) operations of the defence systems and electronics group of Raytheon for $39m. TriQuint is making the acquisition to address such applications as satellite, local multipoint distribution systems and point-to-point digital radio.

FTC unsettles Intel and DEC

The US Federal Trade Commission is investigating the $700m patent settlement between Intel and Digital Equipment because of possible antitrust implications. The move could delay the deal.

First V-Chip television

Canadian firm Tri-Vision Electronics and Samsung claim to have developed the world's first V-Chip enabled television. The V-Chip TV allows the the content of television programs viewed in the home to be controlled..

New logic

"Boolean logic is inefficient," claims scientist Karl Fant. Is he serious? Very. His company is taking an asynchronous approach to circuit design which uses a logic structure called null convention technology, reports Richard Ball.

A US company is challenging one of the cornerstones of modern electronic design – Boolean logic. The firm, Theseus Logic, proposes an asynchronous logic approach to circuit design that does not use Boolean logic.

Theseus promises lower power and better electromagnetic performance from its circuits, much like existing asynchronous techniques, such as those from UK start-up Cogency, or Manchester University's Amulet project.

However, Theseus' designs use an alternative logic structure called null convention logic, or NCL. This describes different gate structures from conventional logic.

"Boolean logic is inefficient," claims Karl Fant, chief scientist at Theseus and driving force behind NCL. "Asynchronous circuits using boolean logic have local delay sensitive tracks." This could lead to races and glitches in the logic. The reason for this is that Boolean logic, "is not symbolically complete", says Fant. This is due to the function of a circuit still depending on time, represented by the clock.

For a logic circuit to be symbolically complete, it should have no dependence on time and therefore require no clock. Also, propagation delays of individual elements and the wires linking them should have no effect on the circuit.

Attempts have been made for several decades to reduce the time dependency of Boolean logic circuits and produce effective asynchronous circuits.

Many of the projects being done today, including Amulet, the asynchronous ARM processor, use the concept of Sutherland's micropipelines. These link blocks of Boolean logic and each block waits for the result from the previous one. Data 'ripples' through the circuit.

Fant argues that this is the wrong approach. His NCL technique for time-independent circuits uses a new class of logic function. Basically, NCL adds the null class to the logic to indicate to a function that no data is present on that wire or signal. To make NCL symbolically complete, it needs to be a four value system – true, false, intermediate and null.

In a commercial electronic circuit, four values are impracticable. There are two data states, one represented by a 0V level, the other by 5V or whatever the supply is.

In NCL, one voltage has to be null to tell the function that no data is present yet. The other voltage represents valid data. So to indicate true or false as well, there must be two wires for each signal, one for true, and one for false. Only one can show data at a time, to indicate true or false. The diagram of the full adder in NCL shows how each input and output to the function, in this case A, B, Z, carry in or carry out, has two wires.

This adds an overhead in terms of tracks or wires, but is compensated for by fewer logic functions.

When a particular logic function, the adder for example, has yet to complete its task, then both outputs from that function, Z and C_{OUT}, will indicate a null value on both their wires. When the function is complete, one wire from each of the outputs indicates a true value, and the next logic block in the chain can start its function.

The simplest NCL gate, **Fig. 1**, has a number of inputs and an output. The output only indicates a true data value when the threshold count is reached. In the case shown, the output is set when three or more inputs show true.

For complex reasons, in order for NCL to work in real circuits, the output must be held if any, but not all, of the inputs go null. Feedback provides this hysteresis, **Fig. 2**.

As soon as four inputs show data, the output is set, and three more inputs are also set. Now, the output only becomes null when all the inputs are null.

The NCL circuit symbol for the feedback function is shown on the right of Fig. 2.

Circuits can be built up using blocks of NCL gates linked together with the two wire signals. The links between logic blocks are less prone to races and glitches than are conventional asynchronous circuits.

In order to prove the ideas, Theseus has manufactured test circuits. The latest, a 0.5μm cmos Asic, performs a two-dimensional discrete cosine transform using 170 000 transistors.

The chip was designed via industry standard tools. "It ran at 30MHz," said Fant. "We expected about 70MHz, but the router was not used effectively."

Tests on the circuits showed that supply voltages could be altered while the circuits were in operation, from 0.5 to 9V. This allows both speed and power consumption to be adjusted.

Unlike other asynchronous circuit techniques, NCL can be prototyped using programmable logic, much like synchronous circuits. "The general opinion in the industry is that you can't do asynchronous circuits on fpgas," said Fant.

To prove his point, Fant implemented an FFT butterfly function on a *Xilinx XC4010*. Although slow, it ran first time managing a few megahertz.

Fant says these tests prove that NCL can be used with existing design methods and tools, including synthesis.

The company is now looking to license the technology and carry out custom designs. To help further prove its feasibility, Theseus is working with Sanders – a Lockheed Martin company – to develop NCL circuits for the F-22 fighter aircraft testbed. ■

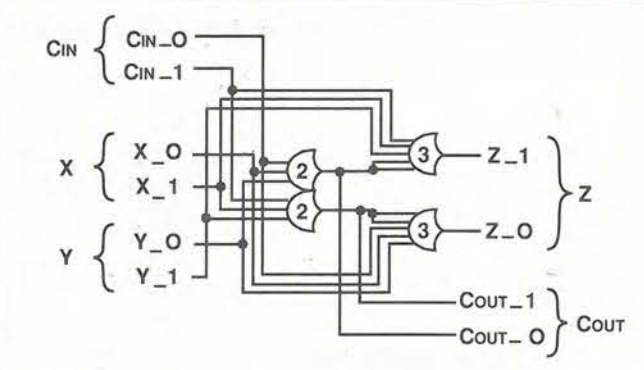
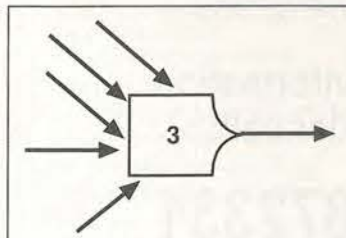
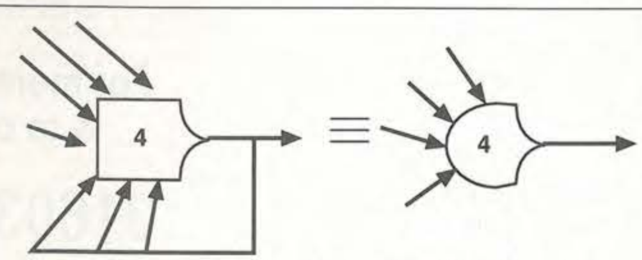


Fig. 3. A full adder is implemented easily in NCL – and the representation is simpler than with the Boolean alternative.



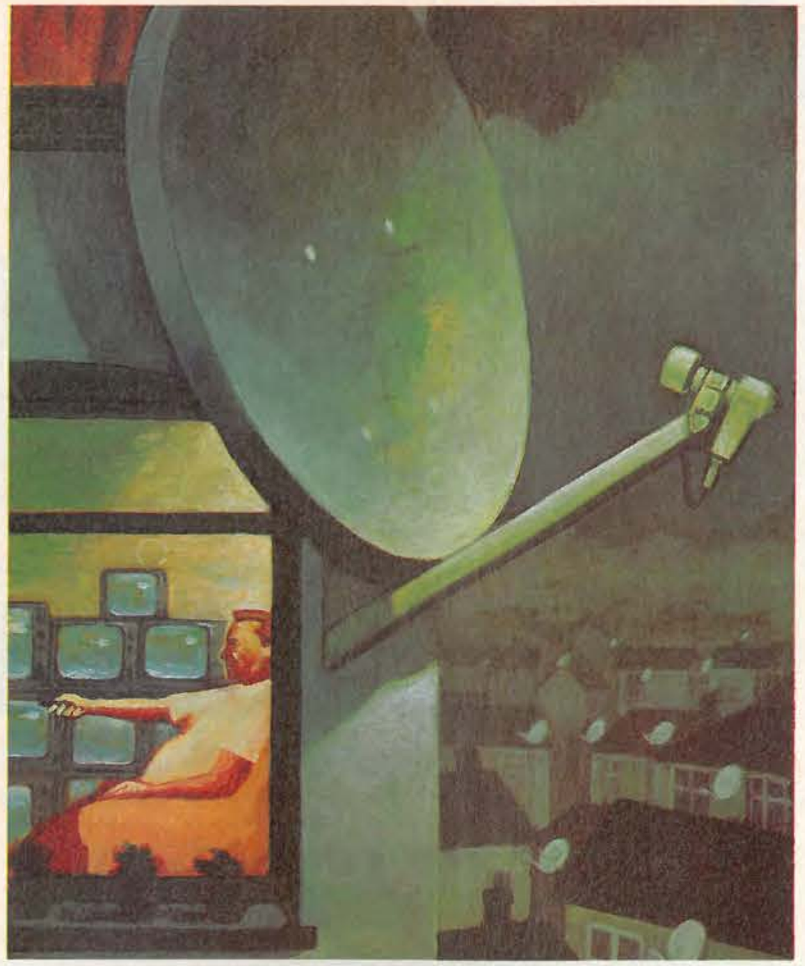
Threshold gate

Fig. 1. Simplest null convention logic gate – essentially a threshold gate. Output is true when three or more inputs are true.



Threshold gate with weighted feedback

Fig. 2. How NCL is implemented in real world circuits with feedback to produce hysteresis. Output becomes true when all four inputs are true. But it will only revert to null when all four inputs are null.



Motorised dishes are versatile, but using multiple lnbs on one dish instead has three advantages. It's cheaper, it allows viewing from more than one satellite simultaneously and switching between satellites is instantaneous. As a bonus, there's no noisy motor to reduce system reliability. Cyril Bateman explains how it's done.

Four birds on one dish?

Publicity for the UK's long promised upgrade to digital television via satellite, suggests that in addition to a new set-top decoder box – costing perhaps a subsidised £200 – viewers may also need to upgrade to a universal lnb, if they already have an old satellite receiving system.

Less well publicised is the fact that this UK digital service from Sky is planned to be supplied from a totally new satellite.[1] Due to be launched in January 1998, this satellite is Astra 2, located at 28.2° east.

UK satellite viewers have receiving dishes directed to the present Astra 1 range of seven satellites, located at 19.2° east. Other non-Sky digital services are planned or already transmitted from the Astra high frequency satellites 1E, 1F and 1G.

In the UK, a typical Sky dish is 60cm in diameter and has a half power beamwidth less than 3°. This is too narrow to receive signals from both Astra 1 and 2 satellites. The larger 80cm dish, common in northern counties, has a beamwidth of only 2°. In the UK only one dish per domestic property is allowed without planning permission.

These Astra 1 satellites comprise only a small proportion of satellites that can be received in UK. In direct competition with Astra, Eutelsat has eight satellites broadcasting television programs.[2] The most popular of these, called Hot Bird, are located at 13° east, but others can be found at 16° east, 10° east and 7° east.

A number of transponders located near 1° west, are popular with enthusiastic viewers. In general programmes from 1° west, are targeted to the Scandinavian countries. Much of the movie film content is transmitted in its original English soundtrack, but many programs require a D2Mac decoder for reception.

In total some 250 satellites are located in geo-stationary orbit in the Clarke belt, of which 30 can be received in the UK.

Multi-satellite reception

How then can both the existing Astra 1 and the imminent Astra 2 transmissions, be received, together with say 13° east and 1° west?

The obvious choice is a steerable or motorised dish assembly. While this would enable all 'visible' satellites to be received, it requires a more expensive receiver, positioner electronics and a motorised dish mount. It would receive only one satellite at a time, taking perhaps one minute to rotate the dish to a program from a different satellite.

At less cost, many viewers have already installed a second lnb and receive two satellites on one dish, usually Astra 1 at 19.2° east and Hot Bird at 13° east. This method has two advantages, instantaneous satellite switching and with two receivers, simultaneous viewing or taping of two programmes, from the same or different satellite. I have used just such a system[3] ever since I installed my dish December 1990, **Fig. 1**.

Since these satellites are distanced by more than 6°, with a dish half power beamwidth less than 3°, just how is this reception from two satellites possible ?

Satellite dishes are simply parabolic reflectors. In comparison, consider that most common parabolic reflector, found in the car headlamp. Two bulb filaments are used. The one at the reflector focus generates the main beam. A second filament offset above this focus point, generates a deflected or dipped beam.

In reverse fashion, this reflector can be used to focus light from the sun, into a tiny intense point of light. Aimed a little off-line from the sun, another less distinct, displaced focus will be seen.

Radio waves reflect from metal surfaces in a similar manner. A satellite dish concentrates the in-line received signals to a point at the entrance to the lnb. Slightly off-line signals are focused away from the lnb feed horn, so are not received.

Satellite dish forms

Two main forms of satellite dish are used. The concentric-fed style dish, visibly shaped like a parabola, is used mostly for motorised systems. While providing the highest possible sig-

Fig. 1. One of the earliest dual-lnb satellite installations in the UK, as installed in December 1990. Both the original Marconi 1.3dB noise lnbs continue in daily use.

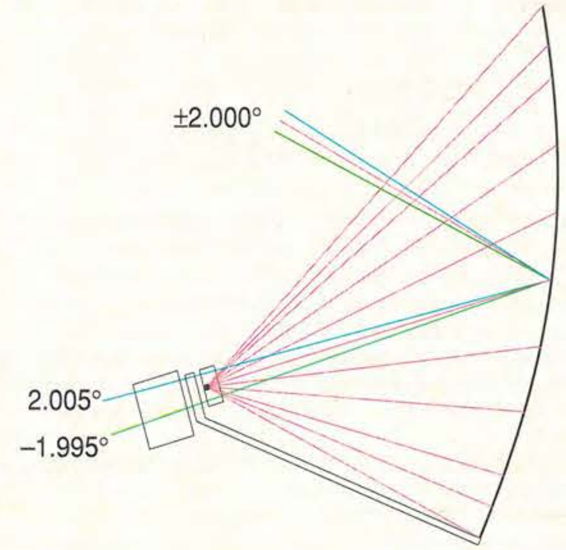


Fig. 2. Ray traced, scale side view of Technisat dish, as used for my dual-lnb installation. With 2° of vertical deflection, beam deviation is negligible.

The new technique in a nutshell

Traditionally, a satellite dish is first roughly aligned using a compass bearing for azimuth, then with an inclinometer for elevation. Final alignment is used to attain maximum signal on a spectrum analyser or field strength meter.

This compass bearing must allow for the Earth's local magnetic deviation, and the compass reading can be disturbed by adjacent metalwork.

Compared to alignment of a terrestrial uhf television antenna, which receives some signal almost regardless of its orientation, a satellite dish receives no signal at all until it is within a couple of degrees of its true alignment, in both azimuth and elevation.

The required azimuth and elevation can be obtained using published tables, or computer programs. Alternatively it can be manually calculated using,

$$elevation = \arctan\frac{\cos X - 0.151269}{\sin X}$$

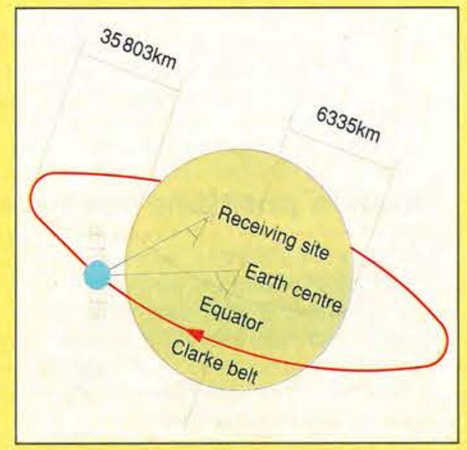
Here, X=arccos(cos$Y$cos$Z$) where Y is the satellite longitude minus receiver longitude and Z is the receiver latitude.

$$azimuth = \arctan\frac{\tan Y}{\sin Z}$$

Don't forget to add magnetic deviation to this.

With a concentric-fed dish, setting true elevation is straightforward, by comparing line-of-sight to the inclinometer. But this is not so for the more commonly used offset-fed dishes. A few offset-fed dishes have approximate elevation angles inscribed, but establishing the optimum elevation angle for such dishes relies almost exclusively on trial and error adjustment, while keeping an eye on the field strength meter. Unless the azimuth has first been set correctly, the meter measures nothing at all.

A starting location for a second lnb can be calculated using engineering formulae, but these require knowledge of the dish dimensions and assume it has a true parabolic shape. Final positioning relies on trial and error location using a field strength meter.



A more precise method, needing neither calculation nor specialist equipments, can be based on observation of the dish reflecting an image of the sun. While for true elevations this method can only be applied for a few days each spring and autumn, true azimuth can be assured throughout the year, whenever the sun shines. This method applies to both concentric and offset dishes, and regardless of variations of dish shape or size.

This article explores both methods.

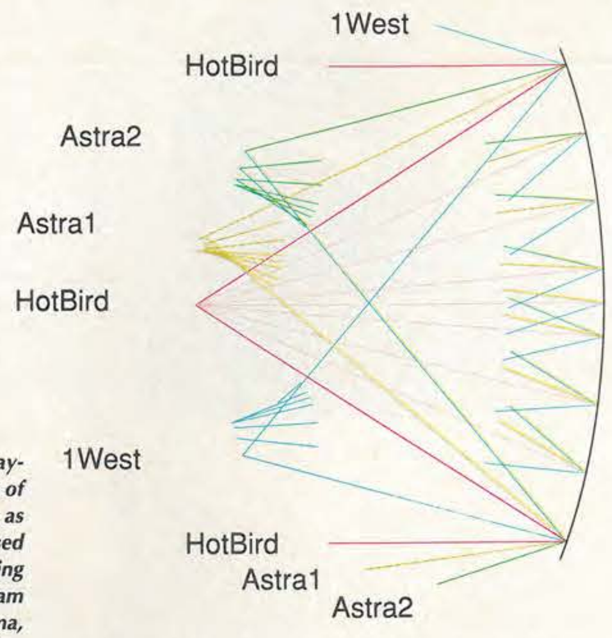


Fig. 3. Accurate ray-traced top view of Technisat dish, as measured and used for tests. Increasing effects of beam deviation and coma, with deflection angle, are clearly visible.

nal gains, the metal work of the lnb intrudes into the received signal path, forming a shadow on the dish surface.

The offset-fed dish is used for low-cost fixed Sky dishes. It is effectively a segment taken from a true parabola, of focal length roughly equal to the shortest distance from lnb to dish. At first glance it looks quite unlike any normal parabolic reflector. Its main advantage is that the lnb does not intrude into the received signal path, so there is no dish shadowing, **Fig. 2**.

Typical signal gains of 33dB for 45/50cm, 35dB for 60cm, 37dB for 75cm, 39dB for 90cm, 40dB for 1m and 41dB for 1.2m offset dishes are obtained.[4] These provide acceptable reception with signal contours at 51dBW for 60cm, 49dBW for 75cm, 47dBW for 90cm and 45dBW for 1.2 metre dishes.

Multi-lnb dish behaviour

With a car headlamp, light from its focus is reflected as an in-line, parallel beam. Light originating slightly away from this focus becomes a slightly off-line, diverging dipped beam.

Due to the reflector's curved surfaces, this dipped beam is bent through a less than expected angle. The degree that these angles differ, called the beam deviation factor, depends on the ratio of the reflector's focal length to its aperture or diameter. This factor increases with deflection, **Fig. 3**.

Used to receive light from a slightly off-line source, the reflector's curved surfaces now exaggerate the angle the focus is deflected by the reciprocal of the beam deviation factor. The focal point becomes slightly diffused, due to coma effects, reducing gain. Radio waves are reflected in the same way as these light waves.

Most good radio-engineering handbooks provide equations or graphs, which allow many factors to be estimated.[5] But all require known and accurate measurements of the dish and many offset dishes are not true parabolas. With an already erected satellite dish, taking measurements might be extremely inconvenient – especially if its location requires using one hand for safety, on a ladder.

Aligning the dish

Traditionally, satellite dishes are first roughly aligned using compass and inclinometers to approximately the correct angles. Compass reading must be compensated for magnetic deviation and can be affected by local metalwork, especially window frames, lintels or the dish assembly. With an offset dish, the required elevation angles are not easily assured. Consequently final alignment using signal strength meters is inevitable.

While this traditional method works well when aligning to just one satellite, finding the location for a second offset lnb requires allowance for the beam deviation factor together with some trial and error searching using a signal-strength meter.

My present dish was originally aligned using these methods. While professional installers can justify the needed equipment, many interested experimenters cannot.

An alternative alignment method

To prevent the sun's reflected rays damaging the lnb, satellite dishes are deliberately coated with a mat, optically non-reflecting paint. If the sun's reflection from the dish could cause damage, might it also be used to align, or measure a dish's behaviour, for off-line reception?

Satellite positions easily received in the UK

While the Astra satellite, owned by SES, is the best known to UK viewers, its seven satellites at 19.2° east, providing some 90 analogue and 220 digital tv channels, are just the tip of an iceberg. Over 30 satellites exist above Europe, eight being owned by Eutelsat.

SES is currently building three satellites, Astra 2A and 2B destined to locate at 28.2° east and Astra 1H at 19.2° east.

Eutelsat has seven satellites building, some intended to replace and upgrade existing satellites at 16, 13, 10 and 7° east. A major carrier of telecommunications and link transmissions, not all Eutelsat capacity provides tv channels. Its three satellites, which comprise the Hot Bird cluster at 13° east, serve over 60 million home, cable and collective households, making Eutelsat one of the largest operators world wide.

Many satellites are positioned or aligned such as to require larger than normal satellite dishes. Others, allowing for local trees or buildings, will be below the usable horizon. Just how many of these 30 satellites can be easily received in the UK using low cost dishes ?

Obviously the Astra 19.2° location can be received throughout the UK. At my home in Norfolk, I could receive from 28.5° east to 27.5° west, encompassing all 30 satellites, but some require a larger dish or non-standard lnbs.

In practice, using my present dish and lnb suitably re-aligned, viewable strength signals can be received from 28.5° east, 19.2° east, 16° east, 13° east, 10° east, 3° east, 1° west, 5° west, 8° west, 18° west. Many of these require decoders or decoder cards not available in UK. Some are Secam only transmissions while others are completely non-English language. This leaves the new Astra at 28.2° east, existing Astra at 19.2° east, Eutelsat at 13° east, and Intelsat/Thor 2 at 1° west as the most desirable.

Eutelsat at 16° east and 10° east, being mostly clear PAL transmissions, might interest those of you with linguistic abilities. However these intermediate 3° locations impose practical difficulty when using a multi lnb fixed dish for reception, due to the physical sizes of the lnb feedhorn needed and the signals' focii.

For a few days each spring and autumn, the sun's elevation coincides with, and shines through, the Clarke Belt, causing a temporary loss of satellite television reception. It is possible to predict the time when sun and satellite azimuths coincide and it is possible to make the dish temporarily optically reflective by washing it with soapy water. For these few days, this is an elegant, practical solution.

Many offset-fed satellite dishes may not be true parabolas, so observation of the sun's reflection provides the most accurate and simple solution. Precise lnb locations simply plotted onto a card and used to pre-assemble suitable lnb mounts result in little final adjustment. This method can be used to align one or multiple satellites.

You need only know the time and date when the sun will reach the required azimuth and elevation. For convenience I have listed these for Manchester, this city being a median site for the UK. The exact times for your location will differ from Manchester time by perhaps a minute, causing a small, consistent error for each satellite.

For convenience, these times can also be applied a couple of days before and after the stated dates, **Table 1**.

An all year round solution?

The Earth traverses the sun in 23 hours 56 minutes, so each degree of azimuth takes approximately four minutes. Old time sundials depend on the sun being near due south at midday throughout the year. So like the sundial, can we also use the sun for all year measurements, not just for a few days?

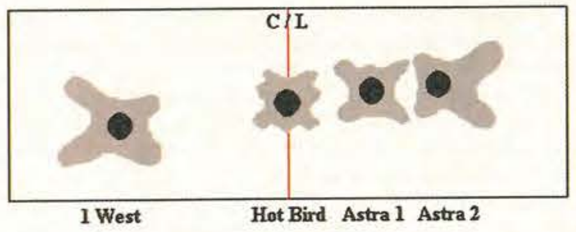


Fig. 4. Sketched Sun reflections, 10 October, 1997 found using Technisat dish. The dark centre hotspot is most important, but the effects of coma are most obvious.

To facilitate observations of the sun's reflections, I assembled a simple lightweight open-fronted viewing box. It was 50cm wide by 15cm tall and 11cm deep at its centre, having a curved back of radius 10% less than the distance from lnb to dish. To this curved back I attached some white card. The whole box was temporarily supported, immediately in front of the lnb cap.

All dishes used to receive off-line signals exhibit a diffused focus due to coma. Most of this originates from the dish extremities. By off-line I mean focussed, but with an offset lnb, as opposed to on-line, where the lnb is at the centre of a symmetrical parabolic dish. Hot Bird in Fig. 2 is on line.

Sensitivity to these diffused signals is reduced since they are on the limits of the lnb's viewing angle. As a result, the focus of the rf waves seen by the lnb is defined more sharply than the visual image. Plotting both the hot spot and the coma is useful in that it helps you guess what improvement you are likely to get from adjusting the lnb on test.

Table 1. Solar azimuth and elevation with satellite alignment time. Times based on GMT at Manchester.

Satellite - location	Astra 2 - 28.2°E		Astra 1 - 19.2°E		Hot Bird - 13°E		Thor 2 - 0.8°W	
Date	Time	Azimuth	Time	Azimuth	Time	Azimuth	Time	Azimuth
1 March 1998	10:07	36.6°E	10:47	26.1°E	11:14	18.7°E	12:16	1.6°E
12 October 1998	9:41	36.6°E	10:22	26.1°E	10:49	18.7°E	11:50	1.6°E

What is a low-noise block?

The low-noise block, or lnb, mounted on the satellite dish, comprises two main parts. These are a mechanical feed horn and waveguide for frequencies from 10.7 to 12.75GHz, and the first down frequency converter electronics of a double superheterodyne receiver. The first local oscillator in the lnb runs at a fixed frequency while that of the second local oscillator in the receiver varies.

Astra analogue television channels use horizontal and vertical polarisation alternately. Inside the lnb waveguide, two fixed 'pickup' probes couple the relevant polarised rf signals to low-noise rf amplifiers. A nominal 13V or 17V power supply, fed via the coaxial output cable, determines which signal polarisation is used. Signals from all either vertically or horizontally polarised transmissions are sent to the receiver, but not both simultaneously.

Depending on the satellite's azimuth with respect to your location, some beam rotation occurs in transit. This ranges from near –20° for Astra 2 to +2° for 1° west. In practice, the easiest way to accommodate this is to adjust the lnb for best reception while tuned to one polarity signal, then switch the receiver polarity and rotate the lnb for minimum signal.

Low-noise blocks intended for offset dishes receive over a 70° cone of reception, sensitivity then being typically 10dB down, to minimise unwanted side pickup. This angle giving an *F/D* typically 0.8 vertically and 0.65 horizontally.

The original lnb oscillators ran at 10GHz. Receivers covered 950MHz to 1750MHz, permitting reception from 10.95 to 11.75GHz. When the Astra 1D satellite was launched, enhanced lnbs containing a 9.75GHz local oscillator were introduced, together with receivers featuring an extended tuning range. This satellite transmitted frequencies down to 10.7GHz,

The Astra 1E, 1F and 1G satellites using the 'high-band' frequencies saw the introduction of the 'universal' lnb having dual first local oscillator frequencies of 9.75 and 10.6GHz. Oscillator frequency is selected by the presence or absence of a 22kHz switching tone which is output by the receiver and superimposed on the lnb supply voltage. Consequently the first intermediate frequency sent by the lnb to the receiver covers frequencies from 950 to 2150MHz. This demands high quality double-screened air-spaced 75Ω coaxial cable.

These high-band frequencies are for digital rather than analogue transmissions. Many are already in use, but targeted to mainland Europe – not the UK. For its UK viewers, Sky has reserved 14 of the 32 transponders on Astra 2A, each capable of up to 10 television channels using digital compression techniques.

While most satellite transmissions use horizontal or vertical polarisation – some including the original BSB satellite – use circular polarisation. This is mostly right handed.

Following the BSB/Sky merger, the original BSB satellite was renamed Thor, and moved to 0.8° west to serve Scandinavian countries. Circularly polarised transmissions cannot be received using a commercial vertical/horizontal voltage-switched lnb.

Fig. 5. Swedish Microwave software shows a simple representation of dual lnb positions. All the software's simulations adopt similar, simple, graphical presentations of results.

I mentioned earlier that twice each year, for a few days, the sun tracks the Clarke Belt. Its reflection from a dish indicates precisely the changing elevation needed for each satellite. With my dish aligned on 13° east, the sun's reflections for our four target satellite positions on 10 October 1997 indicated a change in lnb height of some –18mm for 1° west to +22mm for 28.2° east, **Fig. 4**.

Compared to the azimuth changes of ±18°, elevation changes are small – less than the receiving area of an lnb. With allowance for beam deviation, the tabled elevations could be converted into lnb height changes with sufficient accuracy. Simple trigonometric calculations based on the measured distance from the lnb to the dish centre and some final adjustments on test, could suffice.

Your main task then is to accurately determine the time when the sun's azimuth matches that of the required satellite, for any day of the year. For any given day, the easiest way to calculate satellite azimuth from your location and the sun's time to this azimuth is to use a dedicated satellite program. Two such programs can be downloaded from Internet.

Free software tool for dish alignment

Swedish Microwave[6] provides a freeware package called *SMWLink* that can be used to quickly calculate antenna alignments for any needed satellite or any two satellites, in a doubled-up system. Results are demonstrated graphically to eliminate confusion between positive and negative azimuth values, **Fig. 5**.

A piece of software called *Satmaster Pro*[7] can be used to calculate solar times and all other information needed to align your dish. Its tabular results are less user friendly than the Swedish Microwave package. All equations and calculations used are fully explained in the book Guide to Satellite TV.[4]

Alternatively, the sun's time to an azimuth could be manually calculated using solar tables. Remember though that a satellite's claimed location is maintained with respect to the centre of the earth. Azimuth and elevation for a dish vary according to your local latitude and longitude. Requiring three dimensional trigonometry, I much prefer the simpler software approach.

For convenience, the Table 1 and **Table 2** optimum timings for several dates were calculated using *Satmaster Pro*. Based on Manchester and GMT, they are acceptably accurate throughout the UK, and remain so for a couple of days either side of the given dates in Table 2.

Table 2. Satellite alignment times for 1998, solar azimuth only. Based on GMT at Manchester.

Satellite	Astra 2	Astra 1	Hot Bird	Thor 2
Location	28.2°E	19.2°E	13°E	0.8°W
5 April	10:18	10:53	11:16	12:07
3 May	10:27	10:58	11:18	12:02
7 June	10:41	11:08	11:26	12:04
5 July	10:47	11:14	11:31	12:10
2 Aug.	10:40	11:09	11:29	12:11
6 Sept.	10:14	10:49	11:11	12:03

Putting it into practice

Having determined the mounting positions for the extra lnbs, how does reception of more than one satellite work in practice?

Modern satellite receivers have selectable inputs for two lnbs and recent designs have an inbuilt system called DiSEqC, designed to control motors and external switches.

This Digital Satellite Equipment Control system was devised by Eutelsat[2] as a means of controlling all necessary satellite dish ancillaries using pulsed 22kHz tones, fed along the coaxial downlead. This removes the need to run multiple coaxial leads, power and control cables from a receiver/positioner to the dish, simplifying installation.

Older receivers can be suitably equipped using switches made by specialist suppliers. Receivers unable to provide the 22kHz tone needed to switch a universal lnb between its low and high bands can be equipped with external tone generators. Switches and tone generators are produced by Global Communications[8] and SFM Engineering.[9]

Satellite transmissions have a footprint. A receiving dish in the outer regions of this footprint needs to be bigger in order to deliver the same signal strength. Maps indicating power contours, available from all satellite operators, are included in both software packages, **Fig. 6**.

Off-line reception will be reduced depending on the degree of mis-alignment. For convenience, the graphs and equations[5] needed have been combined into a table of losses for various dish combinations and angles. Used with the relevant contour map, these indicate the expected signal level to be received, and hence predict the quality of reception, **Table 3**.

A noisy but watchable analogue picture will be attained with signals 3dB lower than recommended – even to 6dB using the best low noise receivers and lnbs. Digital signals are more critical. With signals above the minimum, a noise free picture results. With a lesser signal all reception ceases, so it is essential to ensure digital signals have sufficient strength, allowing for the effects of rain and cloud.

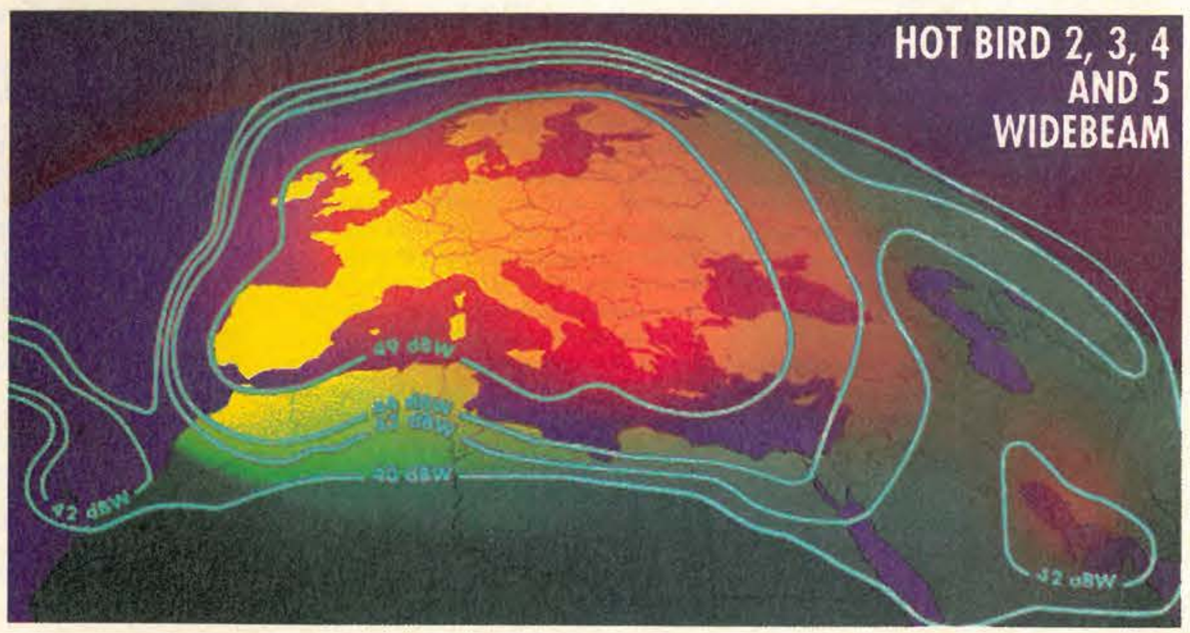


Fig. 6. One of many signal strength contour maps available from the satellite operators. Eutelsat's Hot Bird family reaches an extremely large potential audience.

Table 3. Beam deviation at a nominal 10° deflection angle, also Gain / Loss predictions for Offset Satellite Reception, both relating to Focal Length/Dish Diameter.

Dish	Beam deviation factor		Beam offsets versus decibel losses by f/d ratios				
F/D	Transmit	Receive	1dB	2dB	3dB	4dB	5dB
0.45	0.85	1.18	3.90°	6.50°	7.90°	9.30°	10.20°
0.50	0.86	1.16	5.30°	7.80°	9.50°	11.20°	12.30°
0.55	0.88	1.14	6.30°	9.30°	11.30°	13.30°	14.60°
0.60	0.89	1.12	7.40°	10.90°	13.30°	15.60°	17.20°
0.65	0.91	1.10	8.60°	12.70°	15.40°	18.10°	19.90°
0.70	0.92	1.09	9.90°	14.60°	17.70°	20.80°	22.90°
0.75	0.93	1.08	11.30°	16.60°	20.10°	23.70°	26.10°

Table 4. Satellite transmissions used for dish signal strength tests.

Satellite	Channel	Language	Frequency	Polarisation	Elevation
1 west - Thor 2	Kanal 5	Swedish - English	11.341GHz	Vertical	28.91°
1 west - Thor 2	TV Norge	Norwegian - English	11.421GHz	Horizontal	28.91°
13 east -Hot Bird	RAI UNO	Italian	11.366GHz	Vertical	27.33°
13 east -Hot Bird	Eurosport	German	11.390GHz	Horizontal	27.33°
19.2 E - Astra 1	Eurosport	English	11.258GHz	Vertical	25.82°
19.2 E - Astra 1	TNT	English	11.023GHz	Horizontal	25.82°

In reality

Just how do these calculations compare with actual measurements ?

Using my old *Technisat* dish, original Marconi lnb with a 70mm diameter horn and an *HS101* signal strength meter, I carefully re-aligned the dish to Astra 1 and set the meter sensitivity to read +1dB. I then re-aligned the dish to 13° east, and with the meter sensitivity unchanged, manually moved the lnb to maximise Astra 1 signals. The meter indicated a loss of 0.8dB, **Table 4**.

While centred on 13° east, I measured this optimised physical lnb displacement used to receive Astra from 19.2° east, a satellite alignment difference of 6.2°, as 73mm horizontal, 12mm vertically higher.

Similarly with the dish aligned to 1° west, I measured the loss of Astra 1 signal as 3dB, confirming the theoretical predictions, Table 3.

With the dish aligned to 13° east, the new Grundig lnb secured to an adjustable mount set to receive 1 west, I measured a reduction in strength of 4dB. I believe this is in part due to its smaller, 56mm feed horn collecting less of the diffused coma signals than did the larger feed horn. It would seem advisable to use larger lnb horns for such extreme deflection angles, with smaller horns used for small deflection angles.

Even so with a semi-clouded sky, the unencrypted Pal signals from Sweden 5 and TV Norge were noise free, while all signals from Astra are completely noiseless except under the most adverse weather conditions.

In the short term, re-aligning my dish on 10° east would benefit the 1° west signals by some 0.5dB while leaving the Astra 1 signals almost unchanged. For this my lnb mounts would have to be rebuilt.

Astra 2 planned signal contours are intended to permit reception using a 45-50cm dish. This equates to a signal some 2 or 3dB stronger than direct reception of Astra 1. Hence Astra 2 should be receivable on a second offset lnb on your existing dish which remains aligned on Astra 1. A minimum sized Astra 1 dish cannot provide good reception of Hot Bird transmissions.

Receiving both analogue and digital satellite television

While the straightforward method of receiving both analogue and digital television is to use a steerable dish, this has the disadvantage it can receive only one satellite at a time. Astra viewers must choose whether to receive either analogue from 19.2° east or digital from 28.2° east. This prevents any possibility of viewing one source while taping the other, or parents watching one choice with children the other.

Installing two dishes in the UK, requires planning permission[10] as does an extra large dish.

The objective of this article was to examine the possibility of receiving all major satellite positions of interest on one fixed dish. The voltage and 22kHz switching arrangements used mean that it is not practicable to supply both analogue and digital receivers from one low-noise block. But all interesting satellites other than the new Sky Astra 2, support both analogue and digital transmissions.

Perhaps the most straightforward and flexible method is to use the twin-output, low-noise and digital-ready lnbs now available. These provide essentially two totally independent lnbs, each with selectable horizontal or vertical polarisation and low or high band. These are now available in both small and large feed-horn versions.

For this investigation I bought a Grundig[11] *AUN10T*, which is a universal, twin-output, 56mm diameter small feed horn model, recently introduced. It has a 0.7dB noise figure.

Grundig assures me that leaving one output awaiting a digital receiver and hence for now unconnected is perfectly acceptable. This unconnected output must however be covered to prevent water ingress.

Note that two different lnb types are available, both having two outputs. The dual output versions look identical, but provide H channels only from one output with V channels only from the other, so are less suitable for the options described in this article.

Aligned to 13° east, a one size larger dish is needed to provide noise free signals from both Astra 1, Astra 2 and 13° east. Depending on your location, this might also be possible with the dish aligned to 10° east. Both alignments would permit acceptable reception up to 1° west, using a fourth lnb.

From my results it seems feasible to receive all four desirable satellites using offset lnb's mounted on a dish at least one size larger, but preferably two sizes, than the minimum size Astra 1 dish needed for your location. Unfortunately, until Astra 2 is actually working, its reception cannot be proven experimentally.

Any dish system aligned to receive satisfactory signals from 13° east, would also supply signals from 16° east and 10° east. The main problem is physically positioning the lnbs, rather than one of ensuring sufficient signal strength. An increase of dish diameter to, say, 90cm, with its resultant increased lnb displacements, would facilitate reception of these satellites while increasing signal level for all channels.

In principle, all satellite dish installations are subject to planning control, particularly should you live in a National Park, one of the conservation areas or a listed building. At present though, a 70cm dish size is permitted without planning permission, increased to 90cm for western and Northern counties. Your local planning office will supply details applicable to your local area.

In France, which almost invariably receives much stronger signals, a 1m dish is permitted. Negotiations between Eutelsat and the last Government were expected to relax the above size restraints, but at the time of writing, the June 1995 restrictions are still in force.[10]

Having performed this investigation on 1 March, I now plan to install a fourth lnb on my dish and align this system using the sun's reflection. ■

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Stand-alone data logger

You can leave this small, cheap, autonomous and battery-powered data acquisition box almost anywhere, quietly gathering data ready for later retrieval and post processing on a pc. Pei An and Pinhua Xie explain its how it works.

In environmental monitoring applications, parameters such as temperature, humidity, or water level or pollution need to be monitored continuously over long periods.

A conventional pc-based data acquisition system can be used for such an application, but it may not be ideal. Firstly, such a system involves a computer and a data logger, making it expensive. Secondly, the physical size is large. Thirdly, power assumption will be high, and this implies that a powerful battery pack is required in applications where there is no mains supply.

A stand-alone data logger is a useful device for such an application. Firstly, it is dedicated. Its only task is to acquire data and save the data into its memory. It can be connected to a computer any time to allow its collected data to be transferred and analysed.

Such a data logger can be made small in size and with ultra-low power consumption. Such a stand-alone logger's small size allows it to be placed in any location. It can collect data continuously over a long period of time without having its battery changed.

This article describes a design of such a data logger. It has one analogue input channel with an input range of 0 to 2.5V and the a-to-d conversion accuracy is 12-bit. It has an on-board memory capable of holding 1000 a-to-d conversion results.

When data is to be downloaded, the logger connects to a pc via an RS232 serial port. When driven by a lithium PP3-sized battery it could capture data for a month or so unattended.

It is possible to build this logger into an enclosure the size of a small calculator. The complete system is illustrated in **Fig. 1**. Bear in mind that different sensors and signal conditioning and amplification circuits may be needed for different applications.

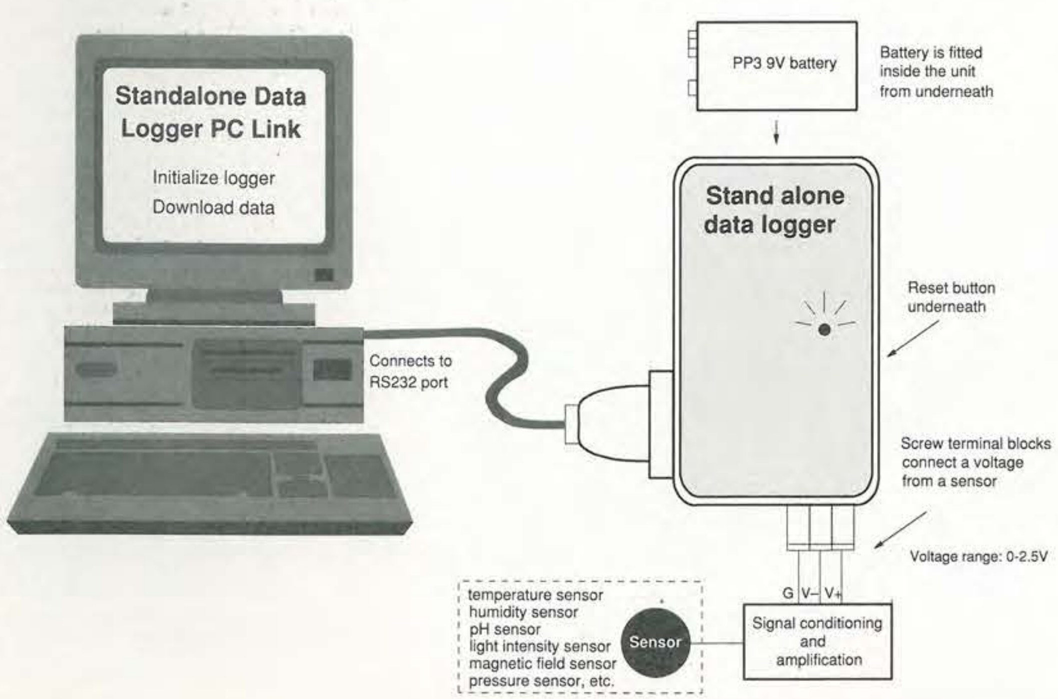


Fig. 1. A stand-alone data logger is used to acquire analogue data from the external world. The logger connects to a host computer via its RS232 port for initial configuration. After that it can be disconnected from the computer and placed in a designated location to acquire data. After the data acquisition session is completed, it is connected to the host computer again for data downloading. The logger only measures analogue voltage. A sensor, amplification and signal conditioning circuits are required to complete the system.

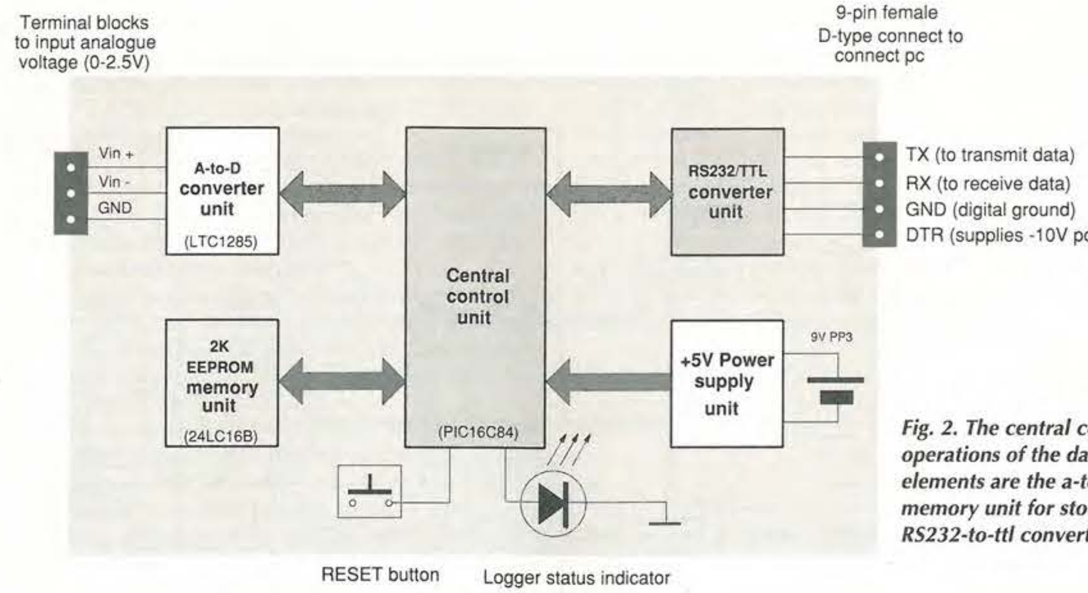


Fig. 2. The central control unit controls all operations of the data logger. Three main elements are the a-to-d converter unit, the memory unit for storing the data and the RS232-to-ttl converter.

How data logger works

The data logger has three operating modes. These are the initialisation mode, the data logging mode and the data downloading mode.

In initialisation mode, the user specifies the start time of data logging and scanning interval – i.e. the period between two consecutive data loggings. This is done by plugging the data logger to the RS232 port of a host computer.

After the initialisation, the logger enters data logging mode. It can now be disconnected from the computer and placed to a designated location. The data logger converts analogue signal into digital data at a fixed interval and stores the data into its memory.

Data logging is terminated either by pressing the reset button on the logger or when the total number of stored data exceeds 1000. At this point, the logger is connected to the host computer once more for data downloading. During downloading, the data stored in the data logger are transferred into the computer.

Hardware details

Figure 2 shows the logger's block diagram. The system comprises five units. They are,

- Central controller based on the *PIC16C84*
- *LTC1285* analogue-to-digital converter
- *24LC16* memory unit ,
- RS232-to-ttl converter
- power supply

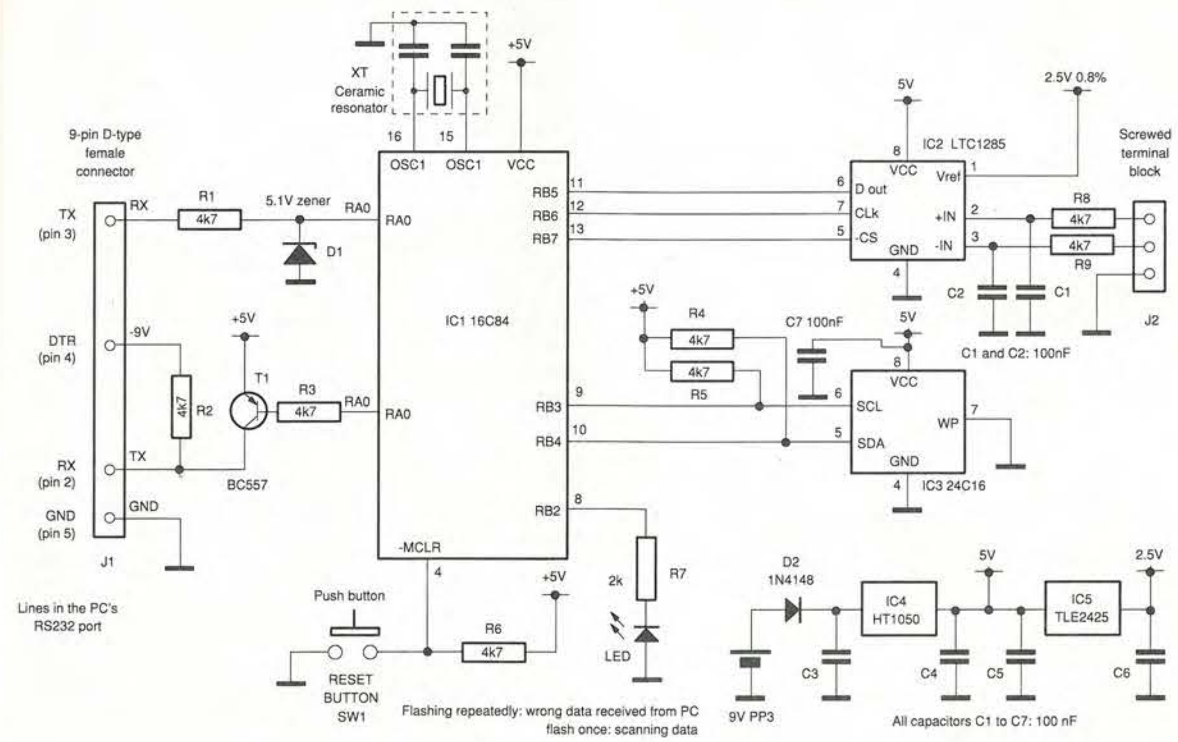
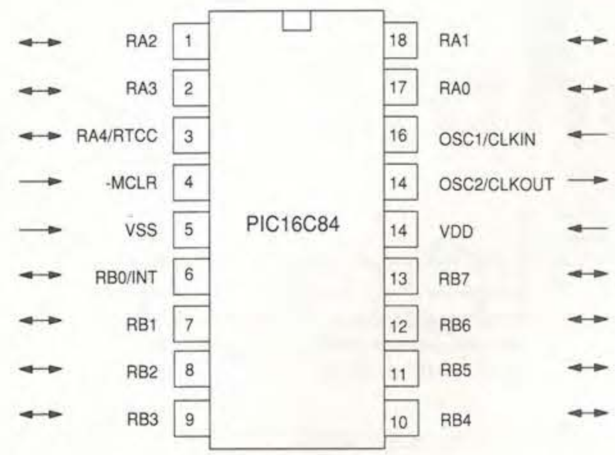


Fig. 3. Complete circuit diagram of the stand-alone data logger. A PIC16C84, an LTC1285 a-to-d converter and a 24LC16B 2Kbyte electrically-erasable prom are used in the circuit, which may be constructed on a single-sided board.



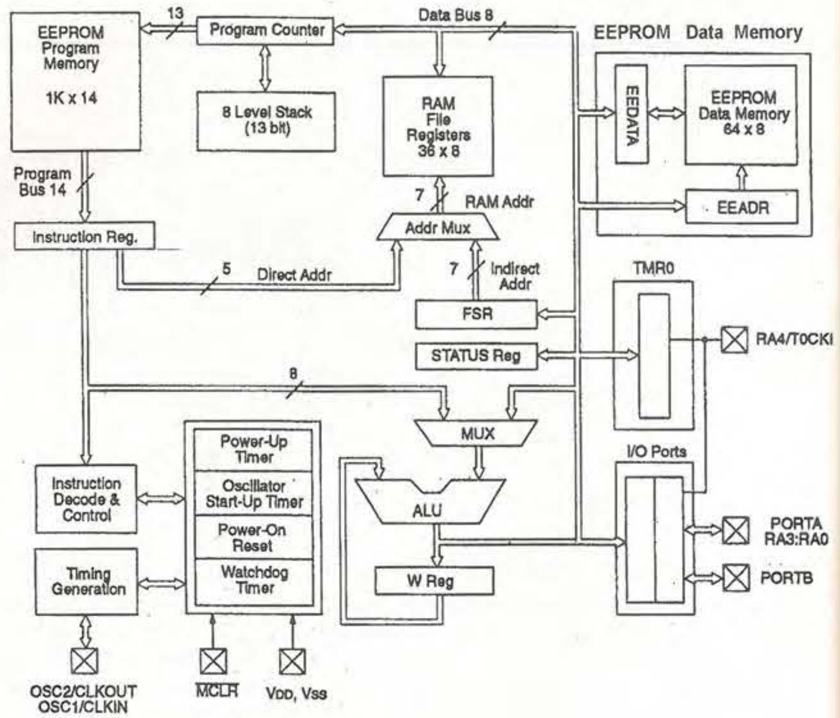


Fig. 4. Pin-out and the internal block diagram of the PIC16C84 microcontroller. This is an 18-pin DIL device with only 35 instructions, making its programming easy to learn.

Fig. 5. Pin-out and internals of the LTC1285 12-bit a-to-d converter. It has a three-wire serial i/o bus, so hardware design is easy.

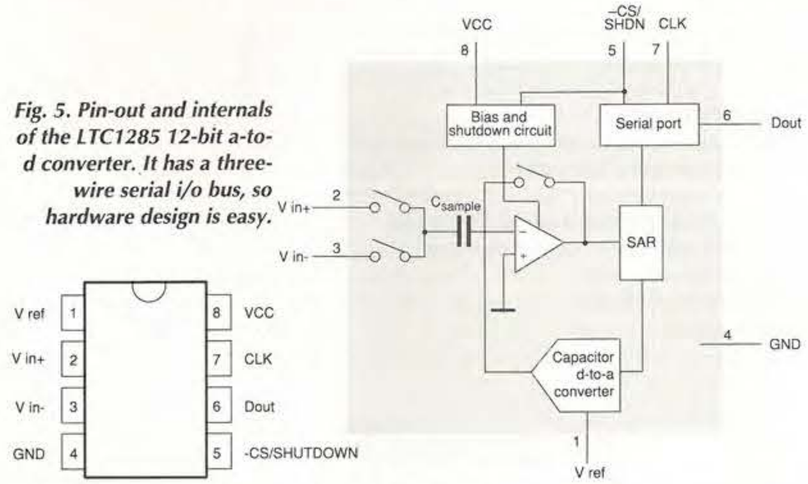


Table 1. In this application, lines of the PIC controller are designated as follows.

Line	Pin	Description
Port A		
RA_0	18	Serial data input to the PIC (connected to Tx of the PC's RS232 port)
RA_1	17	Serial data output from the PIC (connected to Rx of the PC's RS232 port)
RA_2	1	Not used
RA_3	2	Not used
RA_4	3	Not used
Port B		
RB_0	6	Not used
RB_1	7	Not used
RB_2	8	Control of the logger status led (output)
RB_3	9	Serial clock of the I²C bus (SCL) for the 24LC16B (output)
RB_4	10	Serial data of the I²C bus (SDA) for the 24LC16B (input and output)
RB_5	11	Serial data output (D_{out}) from the LTC1285 (input)
RB_6	12	Serial clock (CLK) of the LTC1285 (output)
RB_7	13	enable (–CS) of the LTC1285 (output)

The complete circuit is given in **Fig. 3**. The system utilises only three key ICs, namely the controller, the a-to-d converter and the memory. The *LTC1285* a-to-d converter has a serial peripheral interface, or SPI, for all i/o operations. The 2K eeprom has a 2Kbyte capacity. It communicates with external devices using an I²C bus. The *PIC16C84* manage the gathering of data from the a-to-d converter and the storage of it in the eeprom. It also looks after communication with the host computer via the RS232 port.

Central control unit. The central control unit is based on a Microchip *PIC16C84* peripheral interface controller. This device is a relatively recent addition to Microchip's range. It has has an electrically erasable memory to store program, making it particular useful for product development. This is why we adopted it for this application.

The pin-out and internal block diagram of the *16C84* is shown in **Fig. 4**. Pins 14 and 5 connect to the positive and negative rails of a power supply. The supply voltage range is 2 to 6V. Power supply current is typically 2mA at 5V and 4MHz clock frequency. This drops to several tens of microamps when the IC is in standby mode.

Pin 4 is the master clear. It must be held high in normal operation. Pin 15 and 16 connect to a crystal or ceramic resonator up to 4MHz. The *16C84* has a 1024 word 14-bit wide electrically erasable prom to store instructions and a 64 byte eeprom to store data. There are 15 special function registers and 36 byte-wide general purpose registers.

There are two i/o ports. Port A is brought out on $RA_{0\text{-}4}$ while port B is on $RB_{0\text{-}7}$. Each individual line can be configured as an input or output. As an output, any line is able to source 20mA and sink 25mA. Line RA_4 has a secondary function. It is used for the timer/counter modules. Also, RB_0 doubles as an external interrupt input.

The PIC has only 35 single-word instructions, which makes programming it easy to learn. In this application, the PIC works in the crystal-oscillator mode. A 4MHz ceramic resonantor – a three pin device – is used, Fig. 3. Input/output lines of the PIC are committed in as in **Table 1**.

Analogue to digital converter unit. The analogue-to-digital converter core is a Linear Technology *LTC1285CN8* 12-bit a-to-d converter using successive approximation conversion, **Fig. 5**. It requires a power supply of 2.7V to 6V on pins 4 and 8 and a reference voltage on pin 1.

Typical supply current is 260μA at a 6.6kHz sampling rate and with a 2.7V power rail. When in standby mode, the supply current drops to several nanoamps. The *1285* has a differential analogue input on pin 2 and 3 and the analogue input leakage current is typically 1μA.

The converter communicates with other circuitry through a three-wire SPI serial interface. These three wires are –CS/SHDN, CLK and D_{out}. On going low, pin 5 selects the chip and initiates data transfer. If the pin goes high, the converter enters standby mode.

Pin 7 is the clock input. It synchronises the serial data transfer and determines conversion speed. At the falling edge of the clock signal, each bit of the 12 bits of an a-to-d conversion result is sent out from D_{out} pin.

The operating sequence of the *1285* is shown in **Fig 6**. Data transfer is initiated at the falling edge of the chip select, pin 1. Following chip select's falling edge, the second clock pulse enables data output D_{out}. A null bit (logic 0) appears first on D_{out}, pin 6. At the next 12 falling edge of the clock, the 12 bits of the a-to-d conversion result appear on D_{out} one by one.

In the present circuit, –CS, D_{out} and CLK connect to RB_7, RB_5 and RB_6 of the PIC. The PIC sets RB_7 (–CS) and RB_6 (CLK) as output lines. Line RB_5 is set as an input.

Memory unit. The memory unit uses a *24LC16B* 2Kbyte eeprom from Microchip. The memory is organised is 2Kbyte memory locations. It is possible to erase and write to the memory up to a million times.

The chip requires a 2.5V to 5V power supply with a typical current consumption of 1mA in active mode and 10μA in standby mode. It has an I²C bus for data transfer operations and it operates as a slave device on the bus, **Fig. 7**.

Lines $A_{0\text{-}2}$ have no function and can be left open. Pin WP is for write protection and is normally tied to the ground to enable write operation. Lines SCL and SDA are the clock and data lines of the I²C bus.

Data can be written to and read from the rom via the I²C bus. The write operation has two modes – byte-write mode and page-write mode. The former writes a single byte to a memory location. The latter writes 256 bytes to a block in one go. The read operation has a current-address-read mode and a random-read mode. Byte-write mode and the random-read mode are used in this application. Their timing sequence is described below, **Fig. 8**.

Following a start condition on the I²C bus, an eight-bit slave address byte is clocked into the memory from the controller. The slave address from bits 7 to 0 is:

1, 0, 1, 0, B_2, B_1, B_0 and R/–W.

Bits 7 to bit 4 are the permanent address of the *24LC16* memory. Bits $B_{2\text{-}0}$ specify one of the four memory blocks. When R/–W is high, the operation is a read operation, otherwise it is a write operation.

After the slave address bits are transferred into memory, an address byte is transmitted to it which specifies a particular memory location in the selected memory block. This address is written to the address pointer of the *24LC16* and its value ranges from 0 to 255.

If the operation is a write operation, the eight bits of data are sent to it next. In random-read mode, after writing to the address pointer, a start condition is generated again and it is followed by sending slave address bits with the R/–W bit set to 1, to signify reading. Now, the data stored in the memory is sent out bit by bit.

List 1. These nine bytes are sent to the PIC controller in the logger from the pc immediately after initialisation.

```
byte 1: data logging launch time, year (0-99 decimal)
byte 2: data logging launch time, month (1-12 decimal)
byte 3: data logging launch time, day (1-31 decimal)
byte 4: data logging launch time, hour (1-24 decimal)
byte 5: data logging launch time, minute (1-59 decimal)
byte 6: scanning rate (1=1 second, 2=1minute, 3=1 hour)
byte 7: delay number high byte, Dh
byte 8: delay number mid byte, Dm
byte 9: delay number low byte, Dl
```

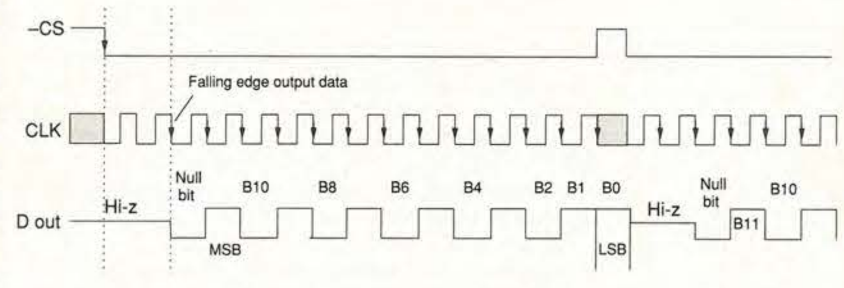


Fig. 6. Timing sequence of the LTC1285. After –CS falls, the converter enters the data conversion stage. The falling edge of the third clock pulse causes D_{out} to output bit 11 of the conversion result. The following 11 clock pulses causes D_{out} to output bit 10 to bit 0.

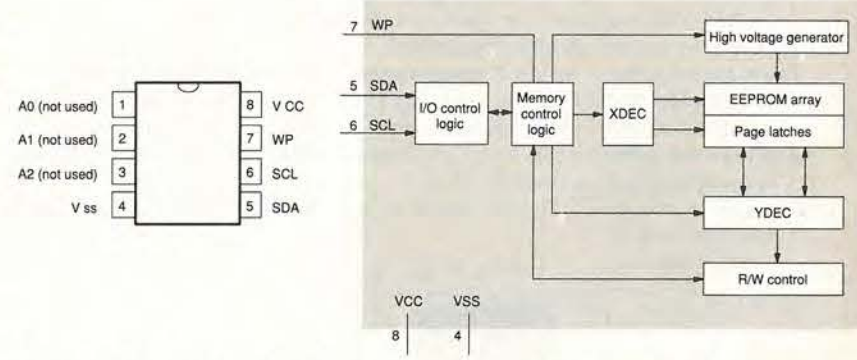
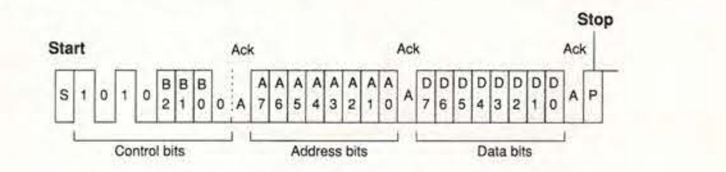
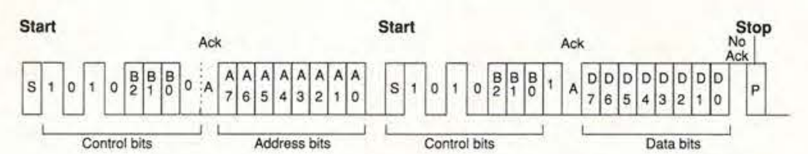


Fig. 7. Details of the 24LC16 2kbyte eeprom. It has an I²C bus comprising a clock line, SCL, and a data line, SDA.



(a) Byte write operation



(b) Random read operation

Fig. 8. Timing sequence of the 24LC16 eeprom. See details of the I²C bus operation in the panel.

List 2. When the logger is to off-load its data, it send this sequence to the pc first, followed by the stored data.

```
byte 1: data logging start time, year (0-99 decimal)
byte 2: data logging start time, month (1-12 decimal)
byte 3: data logging start time, day (1-31 decimal)
byte 4: data logging start time, hour (1-24 decimal)
byte 5: data logging start time, minute (1-59 decimal)
byte 6: scanning rate (1=1 second, 2=1minute, 3=1 hour)
byte 7: total number of data logged, lower 8 bits, Dl
byte 8: total number of data logged, upper 8 bits, Dh
```

In the present circuit, SCL and SDA are controlled by the PIC via $RB_{3,4}$. Both lines are pulled high by $R_{4,5}$ to form an I²C bus. The PIC permanently sets RB_3, the SCL line, as an output line. Depending on the I²C operation in progress, SDA on RB_4 is set as an input or an output.

RS232/ttl translator unit. The function of this unit is to perform voltage conversions between RS232 and ttl logic levels. From the circuit diagram, you will see that the Rx line – i.e. the line from which the logger receives data, RS232 voltage level – is converted into a ttl voltage level using a simple clamp based on R_1 and zener diode D_1. This converter does not have an inverting action.

The Tx signal – the signal output from the logger, RS232 voltage level – is generated by a circuit consisting R_2, R_3 and Tr_1. The circuit requires a positive and a negative power supplies. The former is from the +5V power supply of the data logger board. The latter is 'stolen' from the RS232 port of the computer. The DTR, or data-terminal ready, line in the pc's RS232 port is set low which outputs a –10V level. The pin-out of the pc's RS232 port connector and its functions are given in **Fig. 9**.

Supplying power

As Fig. 3 shows, the power supply is a PP3 9V battery, regulated to +5V using an *HT1050* regulator. This is a 5V fixed voltage regulator with a maximum supply current 30mA. It offers a very low dropout voltage of 100 mV and a quiescent current of 3.5μA.

The 5V supply is converted into 2.5V by the *TLE2425* 2.5V voltage reference for use by the a-to-d converter.

Our stand-alone data logger implementation is constructed on a single-sided pcb and is housed in a slim ABS box.

Software for the PIC

The PIC software divides into three main procedures. The first is the initialisation procedure, the second is the data logging procedure and the third is the data downloading procedure, **Fig. 10**. Their functions are described briefly below:

After pressing the reset button, the initialisation procedure is activated once the PIC detects a serial byte AA_{16} sent by the host computer at its RA_1 pin. After this, the procedure receives nine initialisation bytes, as described in **List 1**.

Delay period in second is calculated using the following,

$$2.6\times(256\times256\times D_h+256\times D_m+D_l)$$

To carry out the initialisation, the host computer must send AA_{16} and the nine bytes through its RS232 port.

After the PIC receives the ninth byte, it automatically enters the data logging mode. Firstly, the data logger is in the sleep mode until the launch time of data logging is reached. While the PIC is not logging data, the PIC, a-to-d converter and the memory are all in sleep mode.

The PIC wakes up and makes the *TLC1285* converter perform an a-to-d conversion. The resulting bits are read into the PIC serially. After reading 12 data bits, the PIC writes the value into the *24LC16B* eeprom.

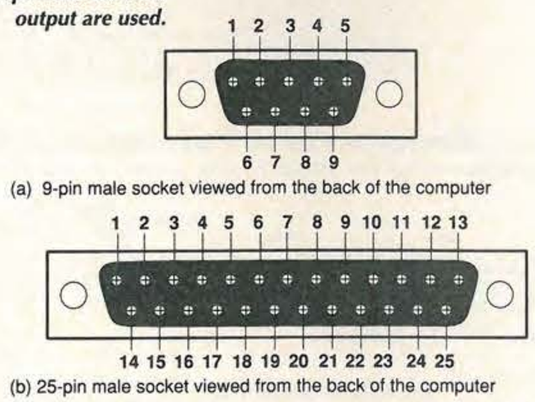
Next, the PIC goes back to sleep. It waits for a time period as specified by the scanning interval and then starts another data logging cycle. There are two ways to terminate the data logging. One is to press the reset button anytime. The other is that the number of data stored in memory exceeds 1000.

Data downloading to the pc is activated once the PIC detects an RS232 serial byte 55_{16} sent by the host computer from its RB_1 line, after the reset button is pressed. Following this, the data logger begins to output data. The first eight bytes are data headers and the logged data follows, **List 2**. The total number of data stored by the data logged is calculated using $D_h\times256+D_l$.

The bytes succeeding those are data bytes. Two bytes rep-

Continued over page

Fig. 9. Pin-out of the RS232 port on IBM compatibles. In this application, only the Tx transmit output from the PC, the Rx input and the DTR output are used.



Pin functions of the RS232 connectors

25 pin	9 pin	Name	Direction (for pcs)	Description
1		**Prot**	-	Protective ground
2	3	**TD**	Output	Transmit data
3	2	**RD**	Input	Receive data
4	7	**RTS**	Output	Request to send
5	8	**CTS**	Input	Clear to send
6	6	**DSR**	Input	Data set ready
7	5	**GND**	-	Signal ground (common)
8	1	**DCD**	Input	Data carrier detedt
20	4	**DTR**	Output	Data terminal ready
22	9	**RI**	Input	Ring indicator
23		**DSRD**	I/O	Data signal rate detector

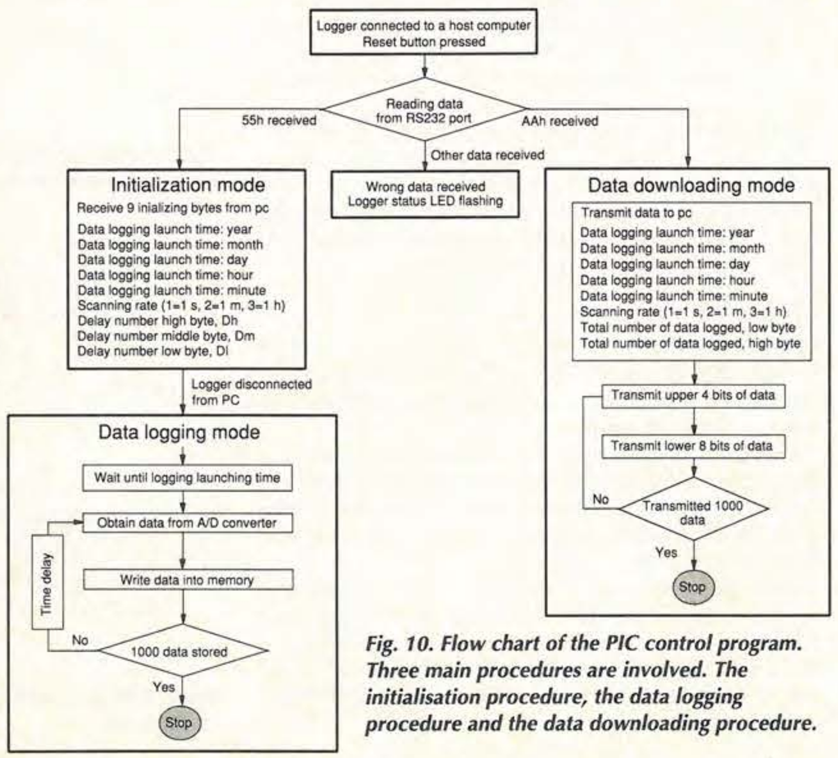


Fig. 10. Flow chart of the PIC control program. Three main procedures are involved. The initialisation procedure, the data logging procedure and the data downloading procedure.

What is I²C bus?

Devised by Phillips, I²C stands for inter-IC-communication. It is a data bus that allows integrated circuits or modules to communicate with each other.

The bus allows data and instructions to be exchanged between devices via only *two* wires. This greatly simplifies the design of a complex electronic circuits. There is a family of I²C compatible devices available for various applications. They include i/o expansion, a-to-d and d-to-a conversion, time keeping, memory and frequency synthesis, etc..

Principle of the I²C bus. The I²C bus consists of two lines: a bi-directional data line called SDA and a clock line called SCL. Both are pulled up to the positive power supply via resistors. An I²C bus system is shown in **Fig. A**.

A device generating a message is a 'transmitter' while a device receiving a message is the 'receiver'. The device controlling the bus operation is the 'master' and devices controlled by the master are 'slaves'.

The following communication protocol is defined:

- a data transfer may be initiated only when the bus is not busy
- during the data transfer, the data line must remain stable whenever the clock line is high.

Changes in the data line while the clock line is high is interpreted as control signals. The following bus conditions are defined, **Fig. B**.

- **Bus not busy**: both data and clock lines remain *high*
- **Start data transfer**: a change in the state of the data line *from high to low* while the clock is *high*, defines the start condition
- **Stop data transfer**: A change in the state of the data line *from low to high* while the clock is *high* defines the stop condition.
- **Data valid**: The state of the data line represents valid data after a start condition. The data line is *stable* for the duration of the *high period of the clock signal*. The data on the line may be changed during the low period of the clock signal. There is one clock pulse per bit data. Each data transfer is initiated with a start condition and terminated with a stop condition. The number of data bytes transferred between the start and stop conditions is not limited. The information is transmitted byte-wise and the receiver acknowledges with a ninth bit.
- **Acknowledge bit**: Each byte is followed by an acknowledge bit. The acknowledge bit is a high level put on the bus by the transmitter whereas the master generates an extra acknowledge related clock pulse. The acknowledge bit is a low level put on the bus by the receiver. A slave receiver which is addressed is obliged to generate an acknowledge bit after the reception of each byte.

The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse in such a way that the SDA line is at a stable low state during the high period of the acknowledge related clock pulse. A master receiver must signal an end to the slave transmitter by not generating an acknowledge on the last byte that has been clocked out of the slave.

How the bus operates. Before any data is transmitted on the bus, the device which should respond is addressed first. This is carried out with the seven-bit address byte plus R/–W bit transmitted after a start condition. A typical address byte has the following format:

```
Fixed Address bits + Programmable address
bits + R/-W bit (in total 8 bits)
```

The fixed address depends on the IC and it can not be changed.* The programmable address bits can be set using the address pins on the chip. The last bit is the read/write bit which indicates the direction of data flow. The byte following the address byte is the control byte which depends on the IC used. Following the control byte are the data bytes. The serial data has the format shown in Fig. 8.

*Although some I²C devices have inputs that can modify the address depending on their logic state, allowing more than one of the same ic to be used on the same bus – Ed.

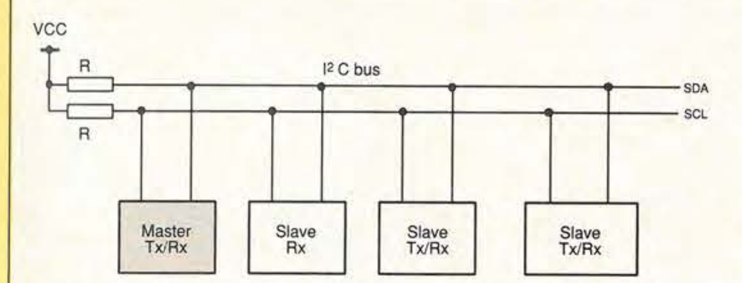


Fig. A. An I²C bus consists of only two data lines: serial clock, SCL, and serial data, SDA. I²C compatible devices connect to the bus using these two wires, making hardware design simple.

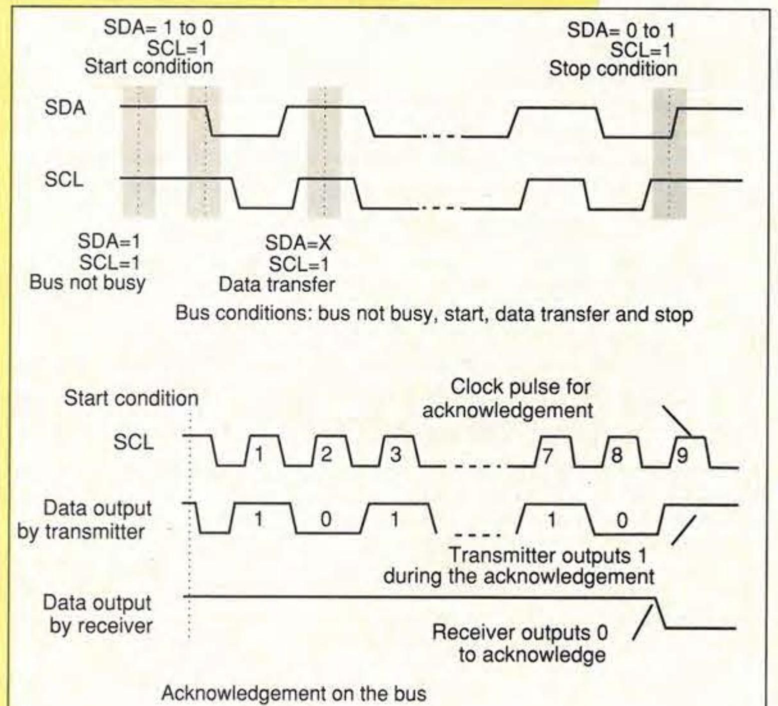


Fig. B. Timing sequences for Bus Not Busy, Start, Stop and Acknowledgement.

List 3. Outline of how to send initialisation data to the logger using Turbo Pascal 6.

```
Procedure init_logger
{initialize data logger}
begin
    ...
    ; {send AA=10*16+10 byte to start initialization};
    Port[RS232_address]:=10*16+10
    delay(1000); {a short delay}
    Port[RS232_address]:=start_year;      delay(1000);
    Port[RS232_address]:=start_month;     delay(1000);
    Port[RS232_address]:=start_day;       delay(1000);
    Port[RS232_address]:=start_hour;      delay(1000);
    Port[RS232_address]:=start_minute;    delay(1000);
    Port[RS232_address]:=scanning_inteval;    delay(1000);
    Port[RS232_address]:=delay_h;         delay(1000);
    Port[RS232_address]:=delay_m;         delay(1000);
    Port[RS232_address]:=delay_l;         delay(1000);
    ...
End'
```

List 4. Routine for downloading data from the data logger into the pc.

```
Procedure readdata;
...
Function data:byte;
{to read data from COM port with valid-data-received detection}
begin
    ; {check if a new valid data is received}
    repeat until (Port[RS232_address+2] and 1) =0
    {check if a valid serial data is received by the COM port}
    data:=port[RS232_address]; {read the received data}
end;
begin
    port[RS232_address]:= 5*16+5; {to start data downloading pro-
cedure}
    yearx:=data;
    monthx:=data;
    dayx:=data;
    hourx:=data;
    minutex:=data;
    scan_intervalx:=data;
    number_lowbyte:=data;
    number_highbyte:=data;
    Total_number:=number_lowbyte+number_highbyte*256;
    for i:=1 to total_number do
    begin
        d1:=data;
        d2:=data;
        data_from_logger[i]:=(d1*256 + d2)* 2.50/4096;
    end;
```

List 5.

```
Procedure Write_interrupt_enable(RS232_address,
Output_byte: integer);
{to enable interrupt identification register on certain
conditions
output_byte=1, to generate an interrupt flag when a valid
serial data is received}
begin
   Port[RS232_address+1]:=Output_byte;
end;
```

resent a 12-bit a-to-d conversion. The upper four bits are sent first, then the lower eight bits.

The procedure causes the PIC to output 2000 bytes – which is in effect 1000 data words. While the PIC is doing so, it does not generate any handshake signals. The host computer must be able to detect each valid received byte and to read it. This is easily achieved on modern computers.

The program list of the PIC control software is lengthy and it is not possible to include it here, but it is available from the authors, as described in the Technical Support panel.

PC link software

This section describes how a personal computer controls the data logger and presents some hands-on programming examples. Turbo Pascal 6 programming language is used.

The following Pascal procedure shows how to send initialisation data to the data logger via its RS232 port. A Pascal command

```
"Port[RS232_address]:=DATA"
```

is used to output the variable DATA from the RS232 port. The RS232 port address (RS232_address) should be supplied to this command.

There are various ways of finding the RS232 address. You will notice that initialisation starts by sending a AA_{16} byte to the stand-alone data logger. A short delay is need between each data transmission, **List 3**.

The following procedure downloads data from the data logger. The download procedure starts with a Pascal command ,

```
"Port[RS232_address]:=55h"
```

Next the pc reads data from the data logger, **List 4**.

When the computer reads serial data from the RS232 port, it must be able to detect when a valid serial data is transmitted from the data logger to the computer. This is achieved by enabling the selected COM port to generate a valid-data-received identification.

After a valid data transmission is completed, bit 0 of the interrupt-identification register of the selected COM port goes low. The register has an address of: RS232_address+2. The way to do the checking is shown in 'Function data:byte' in List 4.

To enable the COM port to generate a valid-data-received identification, the procedure in **List 5** is used before calling the above procedure. The variable output_byte should be 1.

After the program reads all the data from the data logger, It saves the data into a dos text file. The data can be analysed by spreadsheet packages such as the Microsoft excel.

The complete program list of the pc link software is lengthy. It is available from the authors if required. ■

Technical support

Designers' kits containing all the necessary components to construct a complete stand-alone data logger are available from the authors. This includes a pre-programmed PIC. Source code for the PIC and the computer linker program are also available. Please make your enquiry to Dr Pei An, 11 Sandpiper Driver, Stockport, Manchester SK3 8UL, UK. Tel/Fax/Answer:+44-(0)161-477-9583. Alternatively, e-mail to pan@fs1.eng.man.ac.uk.

```
    port[RS232_address]:= 5*16+5; (to start data downloading pro-
cedure}
    yearx:=data;
    monthx:=data;
    dayx:=data;
    hourx:=data;
    minutex:=data;
    scan_intervalx:=data;
    number_lowbyte:=data;
    number_highbyte:=data;
    Total_number:=number_lowbyte+number_highbyte*256;
    for i:=1 to total_number do
    begin
        d1:=data;
        d2:=data;
        data_from_logger[i]:=(d1*256 + d2)* 2.50/4096;
    end;
```

logger. A short delay is need between each data transmission, **List 3**.

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List 5.

```
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    Port[RS232_address+1]:=Output_byte;
end;
```

To enable the COM port to generate a valid-data-received identification, the procedure in **List 5** is used before calling the above procedure. The variable output_byte should be 1.

After the program reads all the data from the data logger, It saves the data into a dos text file. The data can be analysed by spreadsheet packages such as the Microsoft excel.

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ADC-42 WINNER

Open line break finder

Time-domain reflectometry or capacitance measurement are, in most circumstances, successful methods of finding breaks in cable pairs. But, if the lines are loaded at intervals with inductors to improve frequency response, those methods are often useless at the frequencies used.

This problem is overcome in this arrangement by the use of a very low-frequency capacitance measurement, in practice a long CR with a capacitance multiplier for magnification. A 15s CR eliminates the line inductances and the multiplier confers better accuracy on an expanded scale.

Figure 1 shows the multiplier using a Miller circuit, where the effective value of the capacitor is increased by the gain of the transistor. There is, however, leakage to cope with and the **Fig. 2** circuit is a bridge arrangement to balance out dc leakages.

Figure 3 is the practical circuit. Discharge the line capacitance and adjust VR_1 until voltmeter D is reading just below the knee of the curve in Fig. 2a), point Q, logging the reading which is the balance point to be made before any further readings. You may like to practice this procedure using a 0.1μF capacitor.

With S_2 set to 'balance', set the voltmeter to point Q. Switching to 'measure', temporarily closing S_1 zeros the meter and discharges the cable; releasing the switch, the fixed offset of the bridge causes the meter reading to rise slowly as long as the cable is charging. When the meter stops moving, log the reading.

Ranges are 3, 10 and 30km, corresponding to 1.5, 5 and 15s to reach full scale with a 30mV offset and a constant 1mA current. Accuracy is better than 5%, which is sufficient to place the fault near enough for a tdr method to be used, if necessary.

As regards components, K_1 is a normally-open dual reed relay; VR_3 adjusts the op-amp gain to make full-scale reading match analogue voltmeter scale on calibration; $VR_{4,5,6}$ are for scale adjustment; and the *2N3906* is in a constant-current circuit to measure charging time using a low-leakage 18μF capacitor.

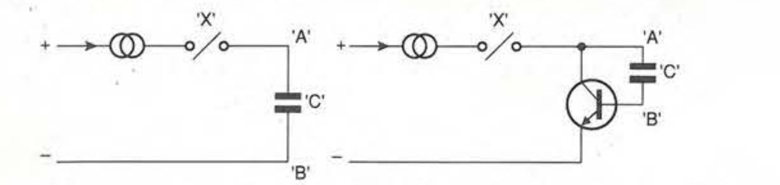


Fig. 1. Measuring the rise time of a long CR *avoids the problem of inductance found in high-frequency methods of detecting line breaks. The Miller circuit magnifies the capacitance.*

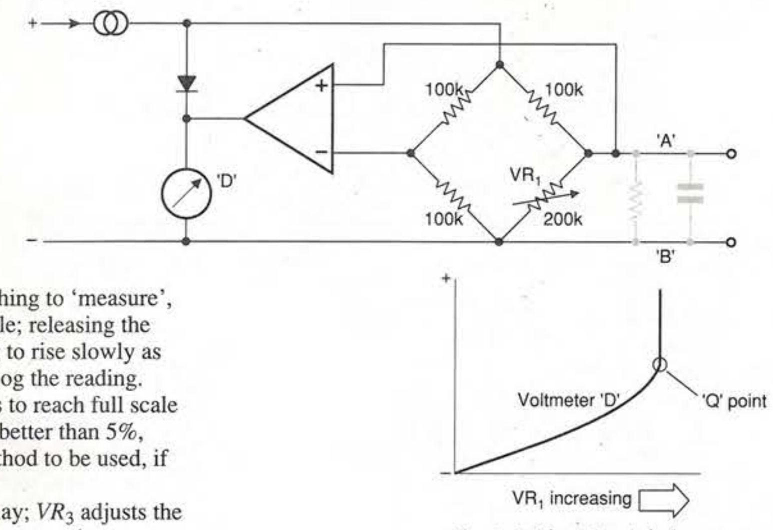
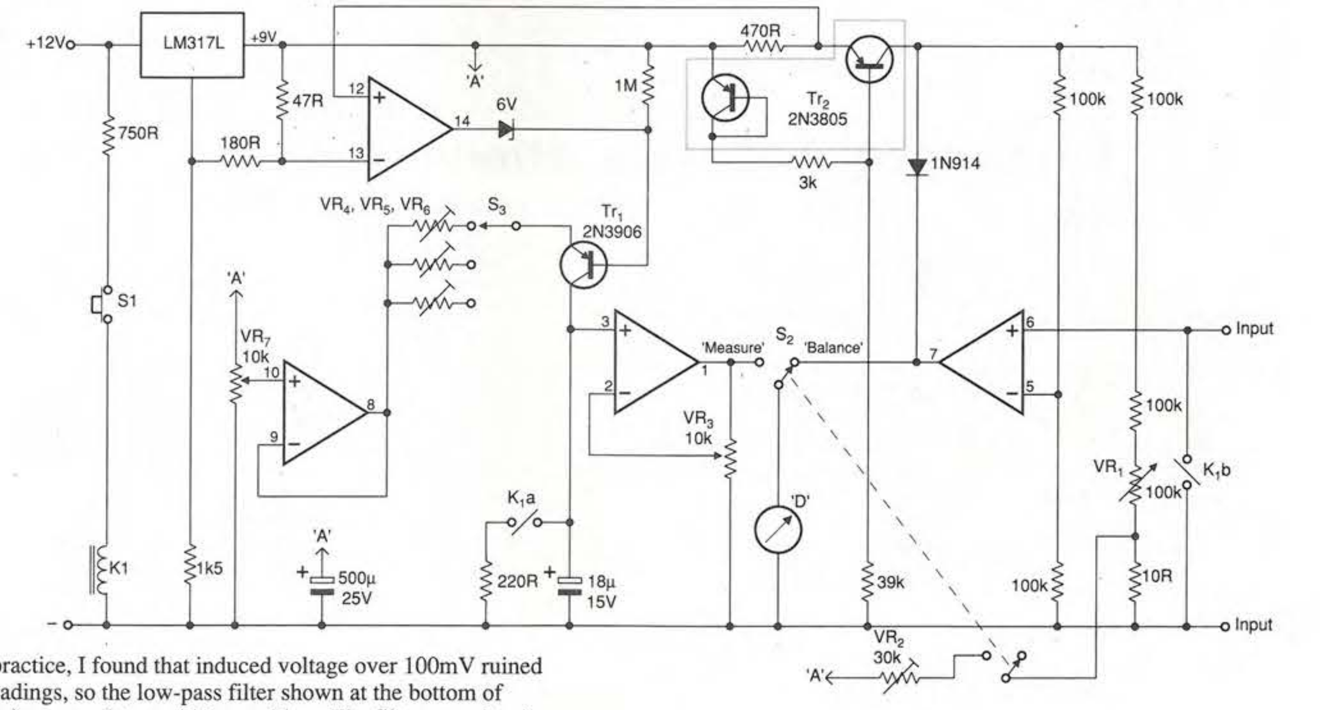


Fig. 2. Bridge circuit balances out leakages in line capacitance. At (a), the meter characteristic; point Q is the meter reading point.



In practice, I found that induced voltage over 100mV ruined the readings, so the low-pass filter shown at the bottom of Fig. 3a) was used to avoid the problem. The filter capacitor is equal to one scale unit, with the result that one unit must be subtracted from readings.

J H Knox
Northwest Territories
Canada

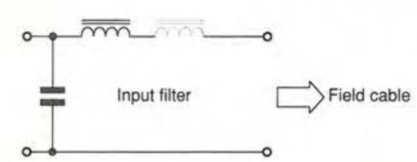


Fig. 3. Practical circuit and filter to reduce induced noise.

Positive feedback linearises temperature measurement

Temperature-dependent resistor *Pt-100* is in common use for temperature measurement, presenting 100Ω at 0°C and 138.5Ω at 100°C. The plot shows its characteristic, in which is implicit an error of 0.7°C at 50°C. This circuit reduces the error to about 0.05°C.

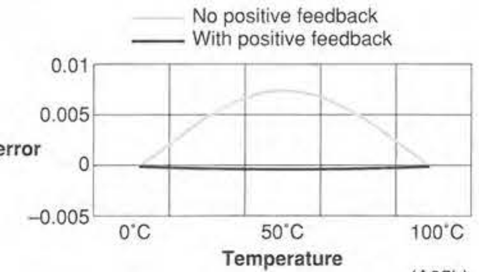
Excitation current for R_t is determined by R_1 and V_{ref} and is, in this case, 0.83mA, low enough to prevent the sensor self-heating. Negative feedback to the main amplifier, op-amp IC_{1b}, comes via R_f, while $R_{2,3,4}$ form the positive feedback path.

When the sensor is cool, the positive feedback is less than the negative feedback, so that the output is low and stays low until the sensor warms up and the two feedback voltages become equal. Voltage divider $R_{3,4}$ sets the amount of positive feedback, R_4 being set initially to 15kΩ.

To adjust the circuit, calibrate at 0°C and 100°C, at which points the output of the prototype was –0.3334768 and –1.5008071. Calculate the output required at 50°C (–0.9166916 in the prototype) and adjust R_4 to obtain that voltage. These adjustments are a little interactive.

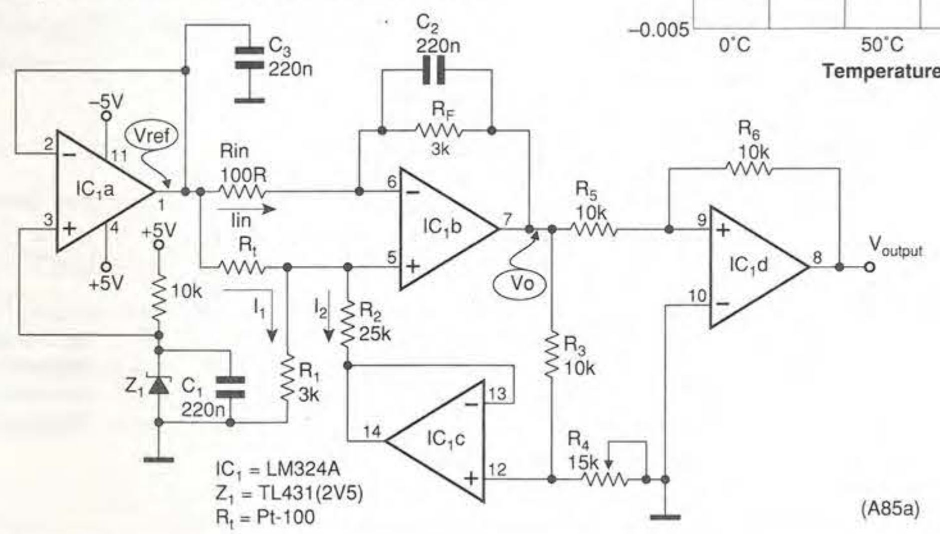
The plot shows the improvement in linearity from around 0.007% to a barely discernible 0.0005%, and gain is much increased.

Fung Tak Sang
Singapore
(A85)



(A85b)

Temperature dependent positive feedback reduces non-linearity and augments gain in resistive temperature measurement.



(A85a)

Economical 1dB/step attenuator

Step attenuators, commonly used in gain or loss measurement, come in several forms, each with its own pros and cons: bridged-tee types are economical, but have a tendency to introduce errors at higher frequencies; those using tee or pi pads are complicated to switch; ladder attenuators, such as the one shown here, are simple and economical, although they do impose a fixed minimum attenuation.

This is a 75Ω design with 1dB per step; this allows the use of fewer resistors than would a 10dB/step type.

In the 0dB position, attenuation is 20dB, provided by the input pad, the 1dB steps being incremental above the 20dB to a maximum of 30dB. Because of the input pad, input resistance of the attenuator remains within 0.5Ω of the 75Ω characteristic impedance in all positions and the output impedance stays within 1Ω of 75Ω.

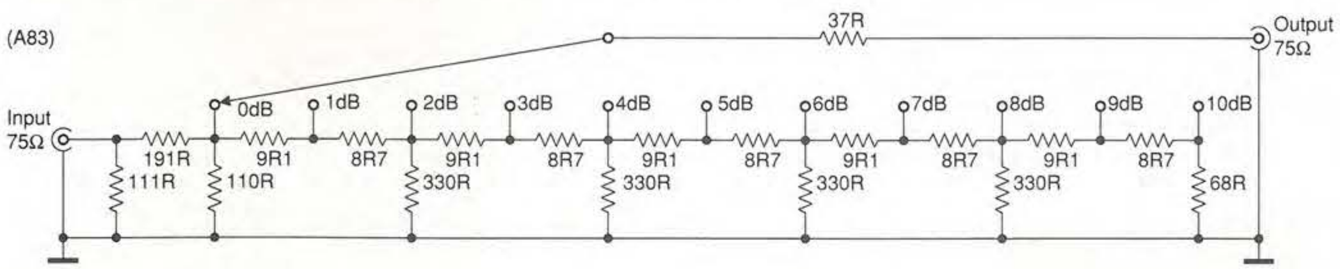
With the resistor values shown, attenuation is within 0.03dB of nominal; most are standard values and others can be made up from parallel resistors. Metal-film types of 1% tolerance are suitable and E96 values would give negligible error.

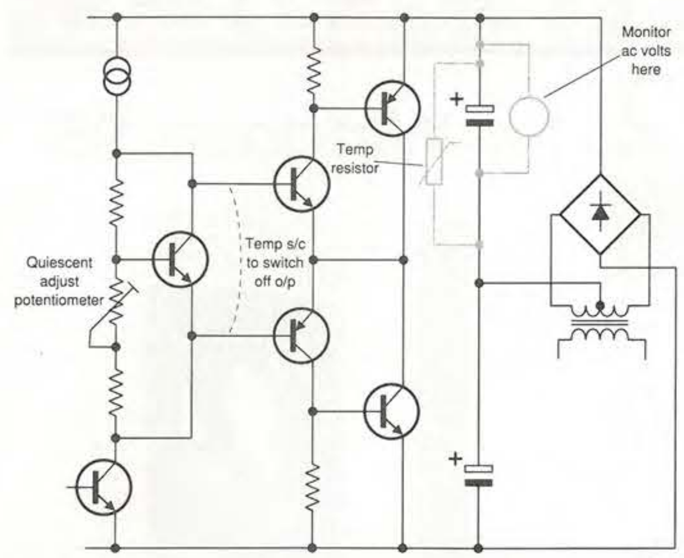
To obtain 50Ω or 600Ω attenuators, scale the values shown to two-thirds or eight times. If an increased variation in input resistance with switch position can be tolerated, the loss in the input pad can be reduced to 10dB.

Ben Sullivan
Waterlooville
Hampshire
(A83)

Ladder step attenuator is economical, accurate and remains accurate at higher frequencies. There is a fixed minimum of 20dB, with 1dB steps above that.

(A83)





Setting up Class AB quiescent current without dismantling the amplifier. Reservoir ripple current is a measure of the current drawn and may be measured instead.

Class AB set-up

Setting up the quiescent current in a Class AB output stage – particularly when no monitoring point is provided, usually requires unsoldering joints to get the meter in series. If the power supply to the amplifier is a conventional mains type with rectifier and reservoir capacitor, this method avoids all that. Ac ripple voltage on the capacitor is proportional to the current drawn and may therefore be used to indicate the quiescent current setting.

To do this, reduce the current in the output stage to zero by setting the potentiometer or shorting the bias network. Connect a resistor across the reservoir of a value to draw a current equal to the required quiescent current, measuring the ripple voltage with an ac millivoltmeter. Now disconnect the shorting resistor and the short on the bias network if you used one; set the potentiometer to give the same reading on the millivoltmeter as before.

S J Kearley
Address not known
(A87)

Infrared remote remote control

A leg-saving circuit, allowing volume on extension speakers to be adjusted from the same room.

When infrared remote control for an audio system is not remote enough, for example when you wish to adjust volume from extension speakers in another room, this circuit detects the ir signal from the controller and transmits it via a twisted pair to another transmitter led near the amplifier.

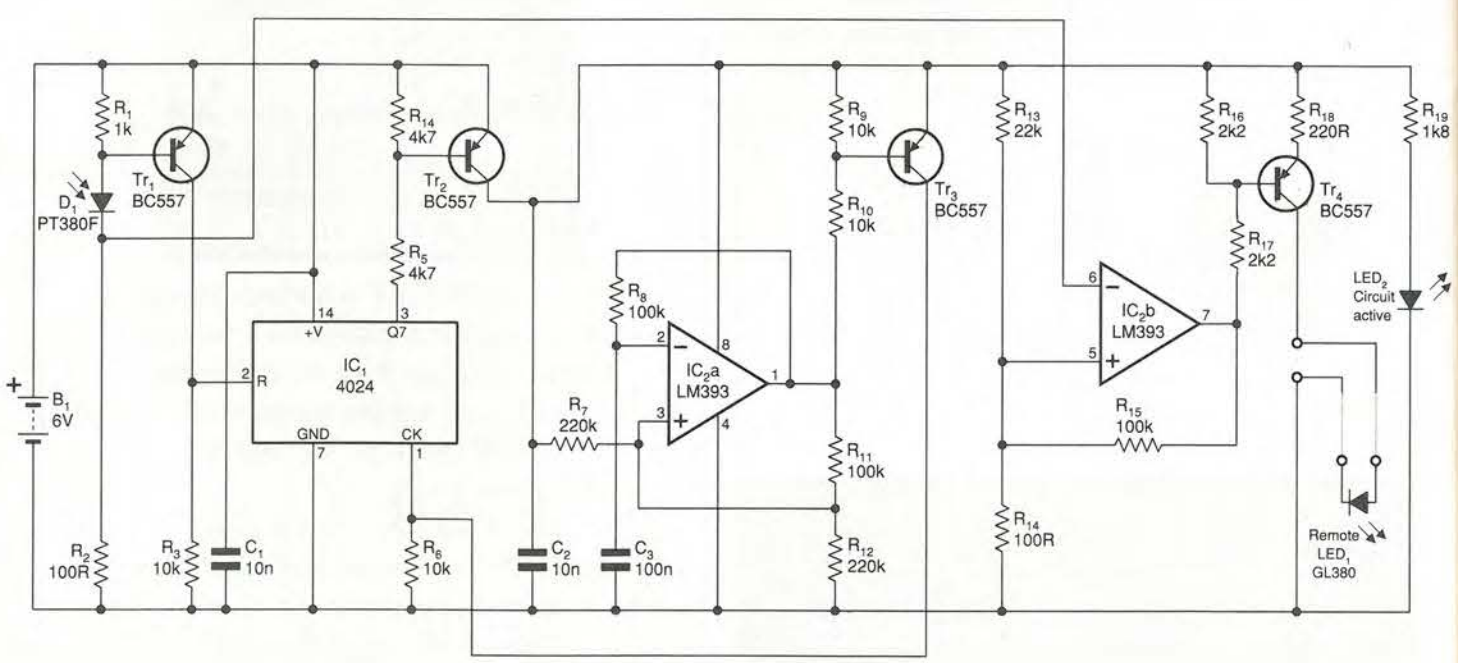
When quiescent, output Q_7 of IC_1 is high and Tr_2 is off. When the photosensor D_1 detects the ir signal from the remote controller, Tr_1 resets IC_1 to turn Tr_2 on. Current through the sensor develops a signal voltage across R_2, which is amplified by the op-amp and Tr_4, modulated current being driven into the 7/0.2 twisted pair into Led_1 near the amplifier.

After about 1s after the controller pulses have finished, the oscillator formed by the first op-amp has applied enough pulses to the counter ic to drive its Q_7 pin high, which switches off Tr_2 and returns the circuit to its quiescent state. Led_2 confirms the operation.

Diode D_1 and Led_1 are made by Sharp and are obtainable from RS Components. They are used in this circuit because the sensor diode has a filter to prevent ambient light causing current to flow and drain the battery.

Using the Sharp devices allows only 3μA to flow in daylight. If these devices are not used, place the sensor in a position to reduce the ambient light falling on it.

S J Kearley
(A86)

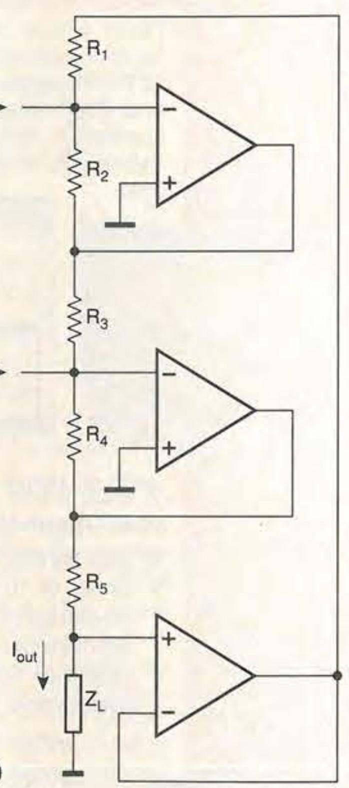


Current-controlled current source

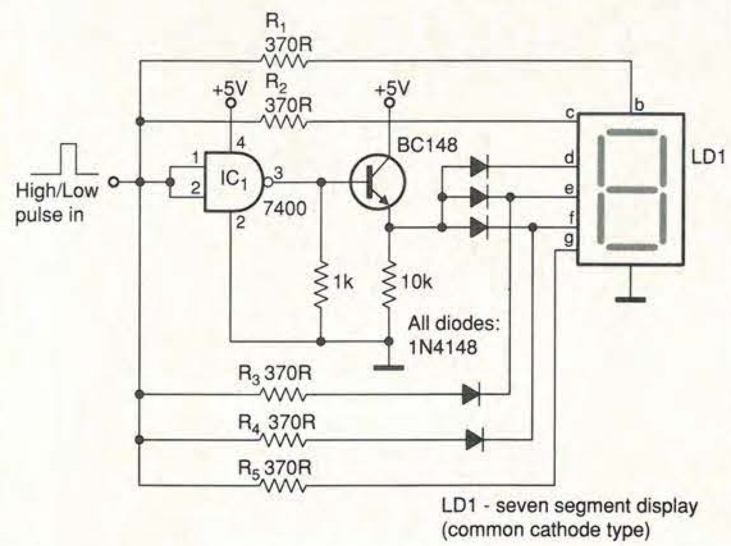
To maintain a high output impedance, the current source uses local positive feedback and a single resistor controls the gain I_{out}/I_a-I_b. If $R_1=R_4=R$ and $R_2=R_3$, then,

$$I_{out}=\frac{R}{R_5}(I_A-I_B)$$

L Szymanski
Stamford
Lincolnshire
(A89)



Current source is controlled by one resistor and provides a high output impedance.



Pulse polarity indicator

To show whether a pulse is positive or negative going, a common-cathode, seven-segment display indicates H or L.

For positive-going pulses, segments b, c, e, f, and g are driven through R_{1-5} to illuminate the segments to give H. A negative-going pulse is inverted by the *CD7400* to drive segments d, e and f for an L, the bottom two diodes isolating the e and f segments.

Raj K Gorkhali
Kathmandu
Nepal

Electrifying farm animals

Spark generator using a car ignition coil, which can be used for many purposes; here it deters farm animals from wandering.

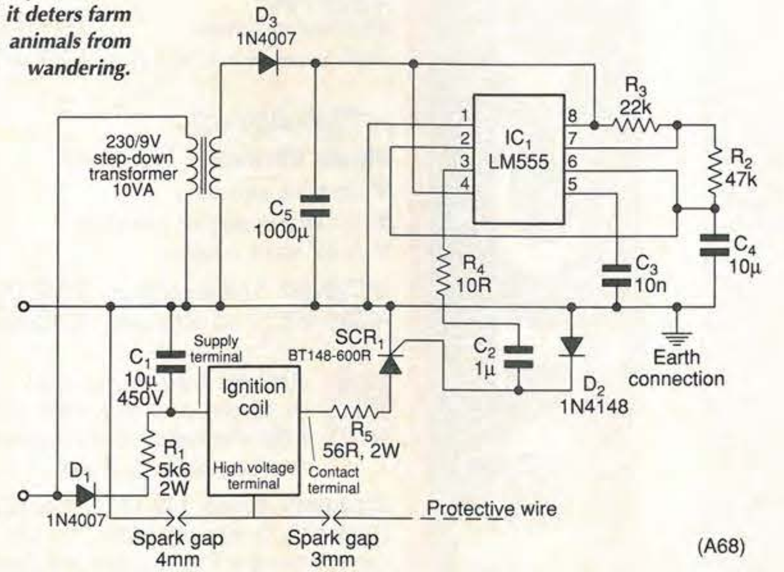
This high-voltage spark generator has various uses, but this description is of a barrier for farm animals; it keeps them in, somewhat surprised but with damage only to their dignity. The generator supplies pulses at 2Hz, which flash over to a barrier wire insulated from earth when an animal touches it.

A high-voltage capacitor discharges through a car ignition coil to produce the h-v pulses. The 230V ac input is rectified by diode D_1 and charges capacitor C_1 through R_1. Counter/timer IC_1, a *555*, fires SCR_1 through R_4 and C_2, discharging C_1 through the transformer primary, D_2 clamping the negative-going pulses to ground. Varying the value of R_1 alters the energy of the spark.

Spark gaps of 3mm and 4mm are placed in series to the barrier wire and ground respectively, so that normally the spark flashes across to ground but goes across to the barrier wire when it is touched.

Circuit ground must go to an earth terminal and, since this could interfere with other installations, an isolation transformer should be used to provide the 230V ac input.

C W W Palihawadana
Dehiwala
Sri Lanka
(A88)

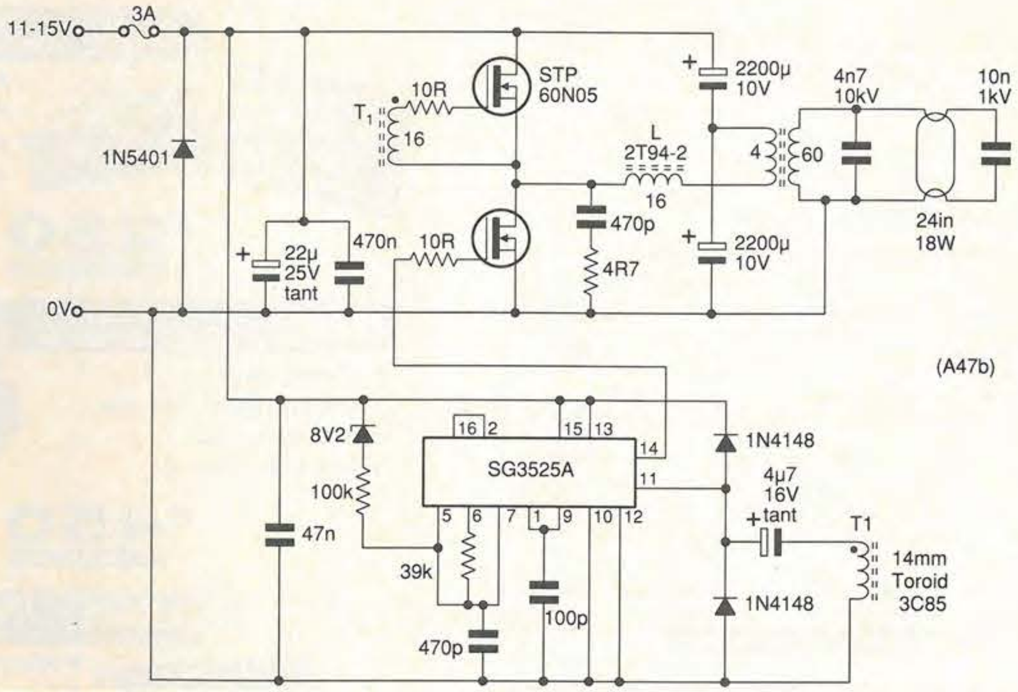


Warning

This circuit must be fed by a 240V safety isolating transformer. Even then it remains potentially lethal so apply and insulate the circuit with utmost care. If you intend using this generator to feed an electric fence, also observe any regulations and guidelines applicable in your country regarding animal welfare and safety in relation to electric fences.

Dc supplies for fluorescent lamps

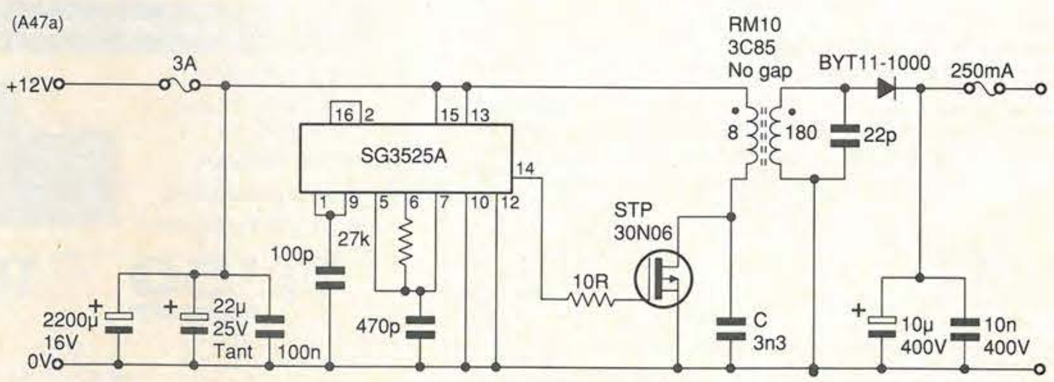
Both these methods of driving fluorescent lamps use the *SC3525* switched-mode power supply control ic, the first varying its frequency instead of the normal duty cycle variation.



(A47b)

12V, 18W inverter. This will accept 11-15V dc input and drives the tube by means of a half bridge, itself fed by the controller at a fixed duty cycle. Starting is by the inductor *L* and the two capacitors across the tube, which form a resonant circuit to produce a high voltage, some of the resulting current going through the tube cathode for preheating. As the tube strikes, its resistance shunts the capacitors and damps the resonant circuit, the reactance of *L* now limiting lamp current and providing some regulation against changes in input voltage. Further regulation is brought about by shifting the controller's oscillator frequency in response to input voltage, which is applied via a 100kΩ resistor and 8.2V zener. Reflected lamp impedance is only a few ohms, so careful layout and capacitors with low equivalent series resistance are needed.

(A47a)



Dc-to-dc converter for compact lamps. Supplying 300V dc for a compact lamp, this converter accepts 12V input and, since the lamps exhibit a degree of supply-voltage regulation, is a simple tracking type. Loss caused by reverse recovery in the output rectifier is avoided here by applying a half-sine reverse voltage. Choose the transformer turns ratio to suit the supply voltage and the capacitor *C* to resonate with the transformer at a frequency to make the half cycle just occupy the fet's off period. If the half cycle is too short, there will be undue stress on the transistor and diode; if too long, it is chopped by the fet's turning on.

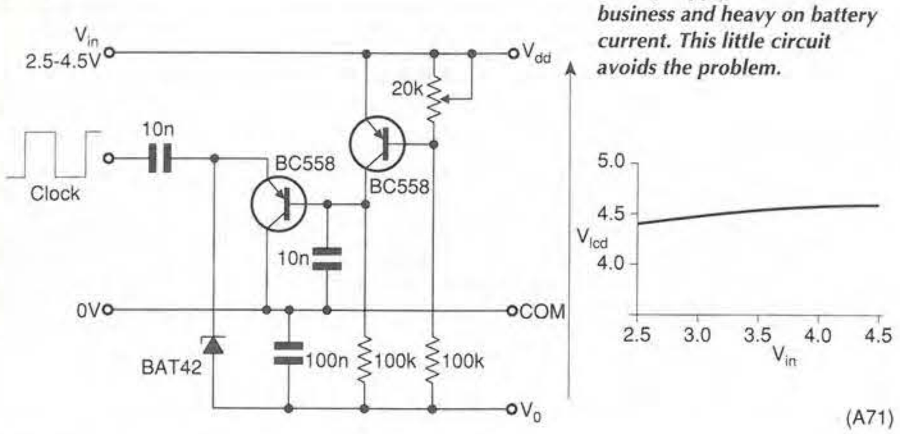
Paul Bennett
Bristol
(A47)

Battery lcd voltage supply

Providing a dot-matrix lcd supply in battery-powered equipment can be wasteful of battery life; it should be adjustable for contrast, it must be stable and it must be around 4.5V below the positive supply. This circuit meets these conditions, line regulation being shown by the graph for a load of 22kΩ.

The clock, whose frequency as shown is 200kHz but is not critical, may already be present in the rest of the circuit, but if not may be generated by an *HC14* or similar gate ic. Temperature coefficient is 0.1V for 7°C. Removing the clock signal powers the circuit down.

David Stephen
Aylesbeare, Devon



Running dot-matrix lcds from a battery supply is an awkward business and heavy on battery current. This little circuit avoids the problem.

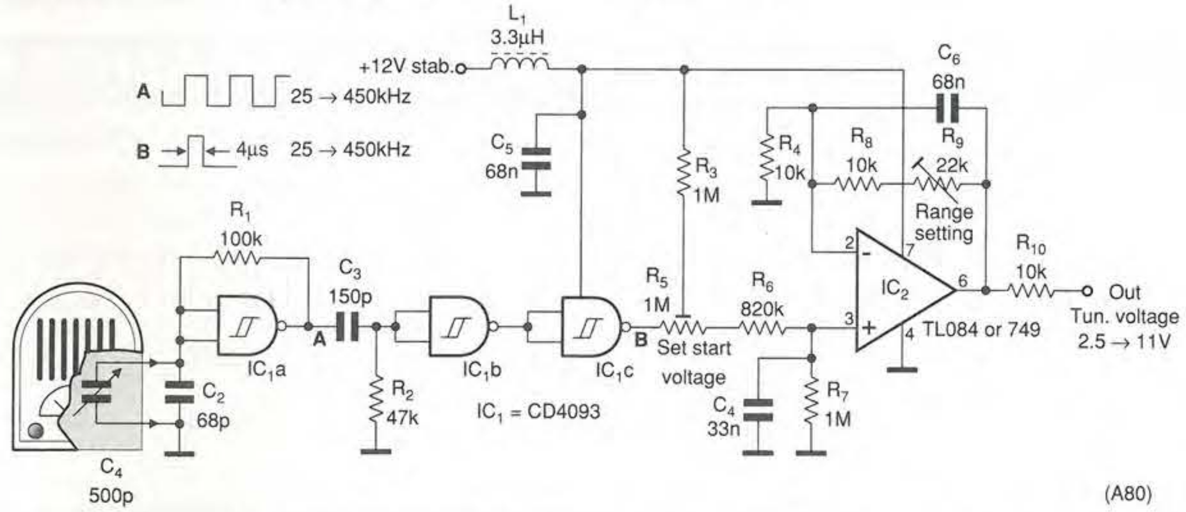
Voltage tuning from a variable capacitor

If you need to incorporate an fm band into an old am receiver, there is the problem of tuning if you use a voltage-tuned fm tuner and the old tuning capacitor is to be retained. This is an easy way of doing it.

An astable flip-flop using IC_{1a} oscillates at 25-150kHz, tuned by the old capacitor and C_2 in parallel. The resultant square wave is differentiated by C_3 and R_2 to give variable-frequency 4μs pulses at the output of IC_{1c}. These go to the op-amp integrator to give a voltage level dependent on the setting of the variable capacitor.

The low end of the range is set by R_5 and the span by R_9.

Vlastimil Novotny
Harrachov
Czech Republic



Using the variable capacitor of an old am receiver to provide a tuning voltage for an fm tuner.

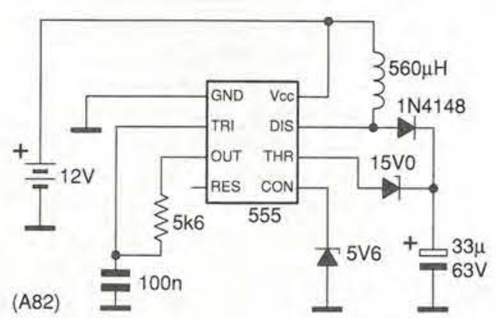
555 switched-mode power supply

In yet another form of existence, this 555 becomes a switched-mode psu, since it has all the necessary components on board.

Having said that, it has been a problem that the discharge transistor seemed to be fully occupied with timing. But driving the timing directly from the output pin frees the transistor to drive the inductor in a standard smps configuration. This makes for an efficient design, since the switching m:s ratio varies and also stops the oscillator completely when there is a low load current; there is therefore a low supply current in that condition.

As shown, the circuit comfortably delivers 20V at 20mA. With a stabilised input supply, there is no need for the 5.6V zener and, if it is omitted, the output will be the 15V from the other zener plus twice the supply voltage divided by three. The zener can be varied to suit the output requirement. ■

Jack Paterson
West Lothian College
Livingston
Scotland



The 555 soldiers on in another guise, this time as a switched-mode supply giving 20V at 20mA.

Wideband isolator

Circulators and isolators are a type of directional coupler with intriguing properties. They are common at microwaves, but they become bulky and expensive at uhf. They would be useful at much lower frequiecies, but they have simply not been available. Ian Hickman's new design covers 0 to 500MHz.

Circulators and isolators are examples of directional couplers, and are common enough components at microwave frequencies. They are three-port devices, the ports being either coaxial- or waveguide-connectors, according to the frequency and particular design.

The clever part is the way signals are routed from one port to the next, always in the same direction. The operation of a circulator – or isolator – depends on the interaction, within a lump of ferrite, of the rf field due to the signal, and a steady dc field provided by a permanent magnet. This is something to do with the precession of electron orbits – or so I gather from those who know more about microwaves. Circulators can be used for a variety of purposes, one of which is the subject of this article.

Figure 1a) outlines a three-port circulator, the arrow indicating the direction of circulation. This means that a signal applied at port A is all delivered to port B, with little coming out of port C. Ideally, no signal will come from port C if the device's 'directivity' is perfect.

What happens next depends on what is connected to port B. If this port is terminated with an ideal resistive load equal to the device's characteristic impedance – usually 50Ω in the case of a circulator with coaxial connectors – then all of the signal is accepted by the termination and none is returned to port B. This means that the 'return loss' in decibels is infinity.

But if the termination on port B differs from (50+j0)Ω, then there is a finite return loss. The reflected, i.e. returned, signal goes back into port B and circulates around in the direction of the arrow, coming out at port C. Thus the magnitude of the signal appearing at port C, relative to the the magnitude of the input applied to port A is a measure of the degree of mismatch at port B.

Because of this characteristic, a circulator with the aid of a source and detector can be used to measure the return loss – and hence the vswr – of any given device under test, as in **Fig. 1b)**. This assumes that the detector presents a good match to port C. If not, it will reflect some of the signal it receives, back into port C – from where it will resurface round the houses at port A.

Given a total mismatch, i.e. a short or open circuit at port B, then all of the power input at port A will come out at port C – but strictly via the clockwise route – bar the usual small insertion loss to be expected of any practical device.

Because it is a totally symmetrical device, the circulator in Fig. 1b) could be rotated by 120° or 240° and still work exactly the same. It doesn't matter which port the source is connected to, provided the device under test and detector are connected to the following two in clockwise order.

An isolator is a related, if less symmetrical, device. Here, any signal in Fig. 1b) reflected back into port C by the detector is simply absorbed. It is not passed around back to port A. As a result, an isolator would actually be a more appropriate device for the vswr measuring set-up of Fig. 1b), although for some applications circulators are preferable.

Microwave circulators with high directivity are narrow band devices. Bandwidths of up to an octave are possible, but only at the expense of much reduced directivity. Circulators and isolators are such useful devices, that it would be great if economical models with good directivity were available at uhf, vhf and even lower frequencies. And even better if one really broadband model were available covering all these frequencies at once.

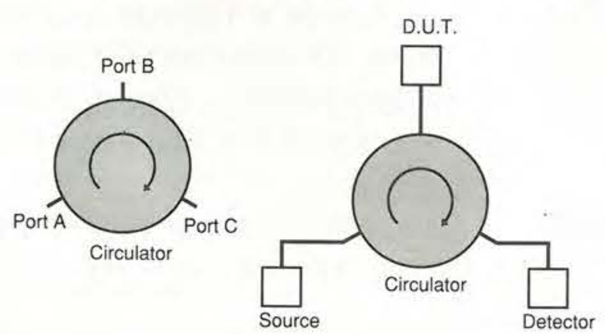


Fig. 1. a) A three port circulator. b) An arrangement using a circulator to measure the return loss of a device under test.

The answer to a long felt need

Though not as well known as it deserves, such an arrangement is in fact possible. It filled me with excitement when I first came across it, in the American controlled-circulation magazine *RF Design*.[1]

This circuit uses three *CLC406* current feedback op-amps – from Comlinear, now part of National Semiconductor – and operates up to well over 100MHz. The upper limit is set by the frequency at which the op-amps begin to flag.

What the article describes is nothing less than an active circuit switchable for use as either a circulator or an isolator, as required. It has three 50Ω BNC ports, and operates from, say, 200MHz, right down to dc, **Fig. 2**.

While at the leading edge of technology when introduced, and still a good op-amp today, the *CLC406* has nonetheless been overtaken, performance-wise, by newer devices. In particular, the *AD8009* from Analog Devices caught my interest, with its unity-gain bandwidth (small signal, non-inverting) of 1GHz.

Of course, if you demand more gain or apply large signals, the performance is a little less – 700MHz at a small signal gain (0.2V pk-pk) of +2, or 440MHz, 320MHz at large signal gains (2V pk-pk) of +2, +10. Still, it seemed a good contender for use in an up-dated version of the circuit described above.

But before going on to describe it, it might be as well to analyse the circuit to show you just how it works.

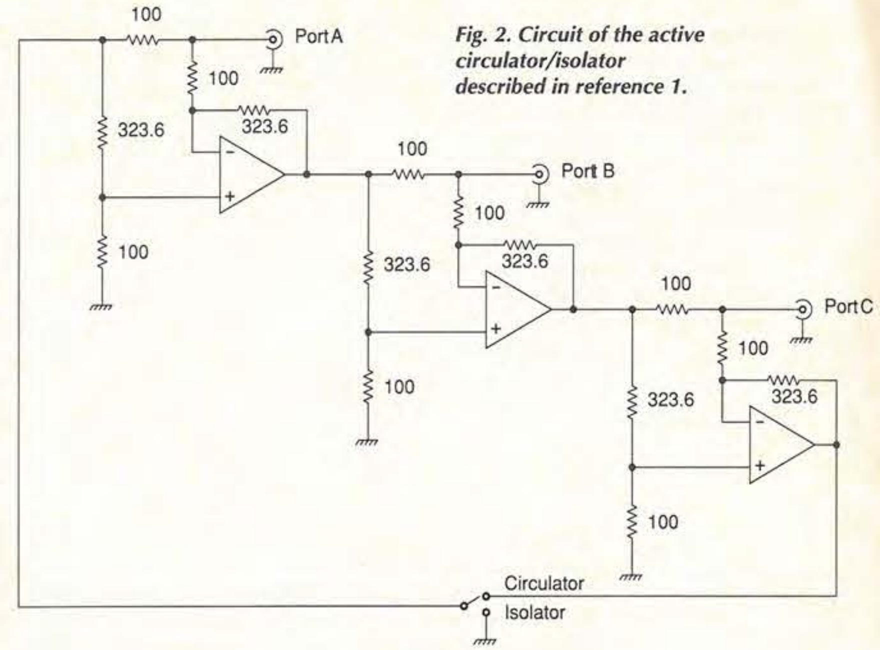


Fig. 2. Circuit of the active circulator/isolator described in reference 1.

How this circulator/isolator works

A feature of this circuit is that it works down to dc. As a result, its operation can be described simply with reference to the partial circuit shown in **Fig. 3**. Here, the voltages may be taken as dc, or as ac in-phase, or antiphase where negative.

Instead of assuming an input voltage and trying to derive the output voltage, or *vice versa*, a useful trick in circuit analysis is to assume a convenient voltage at some internal node, and work forwards and backwards from there. The results then drop out fairly simply – even by mental arithmetic in some cases.

So assume the voltage at the non-inverting input of IC_2 is 100mV. Then the voltage at the output of IC_1 must be 423.5mV. Also, due to the negative feedback, IC_2's output will do whatever is necessary to ensure that its inverting input is also at 100mV.

Figure 3 shows what the output of IC_2 will be, for the cases of a short circuit, or 50Ω, or an open circuit at the port. The short-circuit case is obvious: the resistor at IC_2's inverting input and its feedback resistor form an identical chain to that at the non-inverting input. Thus the output of IC_2 is at +423.6mV, like IC_1, the overall gain is +1, but note that the op-amp is working at a gain is excess of +3.

In the open-circuit case, the net voltage drop across the two 100Ω resistors in series is 323.6mV, so the output of IC_2 must be at 323.6/200×323.6mV negative with respect to the inverting input. Thanks to the careful choice of resistor values, this works out at –423.6mV.

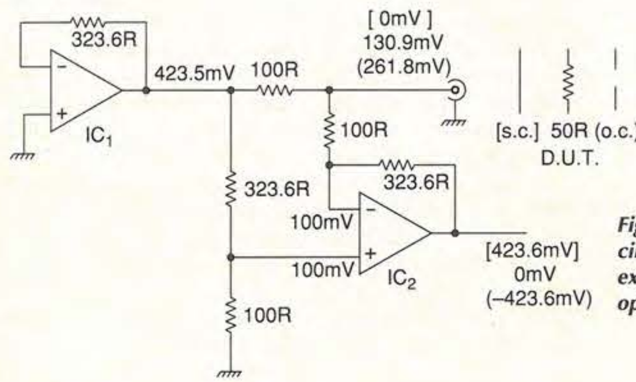


Fig. 3. Partial circuit, explaining circuit operation.

With a 50Ω termination at the port, a line or two of algebra on the back of an envelope may be needed. Let the voltage at the port be *v*. Now equate the current flowing from IC_1 output to the port, to the sum of the currents flowing from there to ground via 50Ω and to the inverting input of IC_2 via 100Ω.

Voltage *v* drops out immediately, defining the current

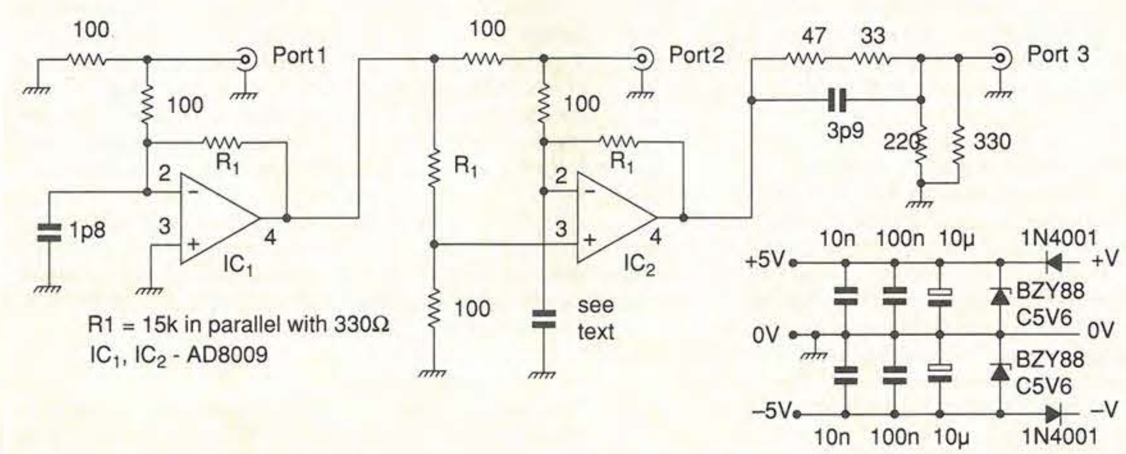
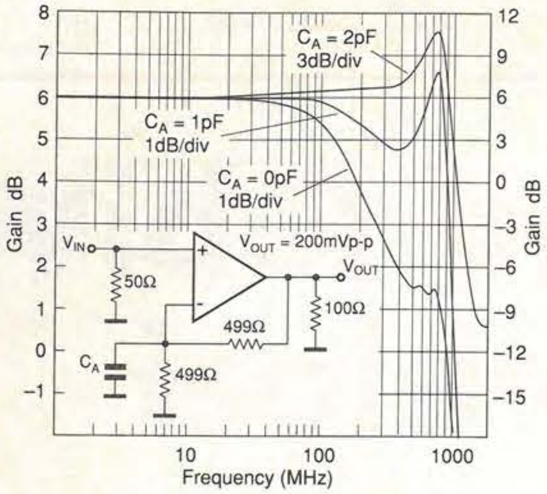
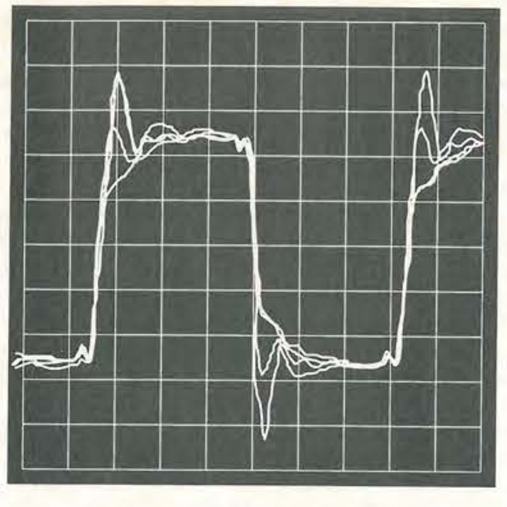


Fig. 4. Circuit diagram of a wideband isolator, usable from 0Hz to 500MHz.

Fig 5a). Bandwidth extension for the AD8009 achieved (for a gain of +2) by adding capacitance from the inverting input to ground.
b) The effect of these three values of capacitance on the pulse response. Horizontal scaling is 1.5ns/div while vertical scaling is 40mV/div.





flowing through the input and feedback resistors of IC_2, and hence the voltage at IC_2's output.

It turns out – again thanks to the ingenious design of the resistive network between each of the op-amps – that the voltage at the output of IC_2 is zero and the corresponding voltage at the device under test port is 130.9mV. Since this is precisely the voltage at a port which produces 423.5mV at the output of the following op-amp, clearly it is the voltage that must be applied to the source input port A – not shown in Fig. 3 – which drives IC_1. Hence the gain from port A to B (or B to C, or C to A) is unity, provided that both the two ports 'see' 50Ω.

Also, if the second port sees an infinite vswr load, the gain from the first to the third port is unity. Effectively, all the power returned from the second port circulates round to the third. At least, this is the case with a circulator.

As Fig. 2 shows, in the case of an isolator, any incident power reflected back into port C is simply absorbed, and does not continue around back to port A.

An updated version

Having obtained some Analog devices *AD8009* wideband current-feedback op-amps, I was keen to see what sort of performance could be achieved with such an up-to-date device. Clearly, it could simply be substituted for the *CLC406* in the circuit of Fig. 2.

But, after careful consideration, it seemed that all the applications I had in mind could be met with an isolator. Now if you are is willing to forego the ability to switch the circuit to operate, when required, as a circulator, then not only are substantial economies in circuit design possible, but also one or two dodges to improve performance at the top end of the frequency range can be incorporated.

So at the end of the day, my circuit finished up as in **Fig. 4**. You can see immediately that as an isolator only, the circuit needs but two op-amps. Also obsoleted are a switch, and a number of resistors, while port C is simply driven by an L pad.

A word about the power supply

But before describing the operation of the rf portion of Fig. 4, a word about the power supply arrangements is called for.

Circuits under development sometimes fail for no apparent reason. This often put down to 'prototype fatigue', meaning some form of unidentified electrical abuse. I have suffered the ravages of this phenomenon as often as most.

The construction of the isolator, using op-amps in small-outline SO8 form, chip resistors and 0805 packaged 10n capacitors, was not a simple task. It involved both dexterity and some eye strain.

I built the circuit using 'fresh air' construction on a scrap of copper-clad FRG used as a ground plane. The thought of having to dive back into the bird's nest to replace an op-amp or two was horrific, so some protection for the supplies was built in.

The series diodes guard against possible connection of the power supplies in reverse polarity, while the zener diodes prevent excessive voltage being applied. The types quoted will not provide indefinite protection from 15V supplies with a 1A current limit, but they will guard against an insidious and often unrealised fault.

At switch-on, some older bench power supplies output a brief spike of maximum voltage equal to the internal raw supply voltage. And after a number of years' use, many power supplies develop a noisy track on the output voltage setting potentiometer. Depending on the particular design, this too can result in a brief spike of maximum output voltage whenever the potentiometer is adjusted. For the sake of a few extra components, it is better to be safe than sorry.

Putting it together

The two op-amps were mounted in between the three BNC sockets placed as close together as possible.

In somewhat cavalier fashion, the ICs were mounted above the ground plane, standing on leads 1, 5 and 8, and also lead 3 in the case of IC_1. These leads had been carefully bent down from the usual horizontal position on a surface mount device, the remaining leads having been bent upwards.

A 10nF 0805-packages chip capacitor was then soldered between the ground plane and each supply lead, leaning in towards the device at an angle of about 60° from the vertical. The leaded 100n capacitors – also four in total, these items of Fig. 4 being duplicated – were then also fitted, to each side of the op-amp to leave space for the chip resistors.

The chip resistors were then fitted, the feedback resistors around IC_1 and IC_2 being mounted on top of the devices, directly between the bent-up pins 2 and 6. As the body length of the 100Ω input resistor to IC_1 was not sufficient to reach the shortened spill of the BNC centre contact at Port A, the gap was bridged by a few millimetres of 3mm wide 0.001in copper tape. The same trick was used elsewhere, where necessary.

If you don't have any copper tape to hand, a little can always be stripped from an odd scrap of copper-clad. The application of heat from a soldering iron bit will enable the copper to be peeled from the board. This is possible with GRP and even easier with SRBP.

Testing the prototype

The finished prototype was fired up and tested, using the equipment briefly described later. Performance up to several hundred megahertz was very encouraging, but it was obviously sensible to try and wring the last ounce of performance from the circuit.

Reproduced from the *AD8009* data sheet, **Fig. 5a)** shows how a useful increase in bandwidth can be achieved by the addition of different small amounts of capacitance to ground from the op-amp's inverting input, at the expense of some peaking at the top end of the frequency range.

Figure **5b)** shows the effect of those same values of capacitance on the pulse response. In Fig. 4, the op-amps are used at a gain in excess or +10dB, so the same degree of bandwidth extension cannot be expected for sensible values of capacitance at the op-amp's inverting input.

After some experimentation, in the case of IC_1 a value of 1.8pF was selected. In the case of IC_2, the value of capacitance was adjusted for best device directivity. This involved terminating port B with a 50Ω termination and tweaking the capacitance to give the greatest attenuation of the residual signal at port C in the 300 to 500MHz region.

As the required value was around 1pF, lower than the minimum capacitance of the smallest trimmers I had in stock, it was realised as two short lengths of 30SWG enamelled copper wire twisted together. The length was trimmed back for optimum directivity as described above, leaving just over 1cm of twisted wire.

The transmission path from port A to B and that from port B to C both showed a smooth roll off above 500MHz, with no sign of peaking.

Isolator performance evaluation

After using the equipment described above to optimise the isolator's performance, some photographs of the screen display were taken for the record. The upper trace of **Fig. 7** shows the output of the tracking generator, connected via two 10dB pads and two coaxial cables connected to the input of the spectrum analyser. These cables are joined by a BNC back-to-back female adapter.

The sweep covers 0 - 500MHz and the vertical deflection factor is 10dB per division. The back-to-back BNC connector was then replaced by the isolator, input to port B, output from port C.

A second exposure on the same shot captured the frequency response of the isolator, Fig. 7, lower trace. It can be seen that the insertion loss of the isolator is negligible up to 300MHz, and only about 3dB at 500MHz. The response from port A to port B is just a little worse, as this path could not use the frequency compensation provided by the 3.9pF capacitor in the output pad at port C.

Figure 8 shows the reverse isolation from port B (as input) to port A (lower trace; with the input, upper trace, for comparison). This can be seen to be mostly 45dB or greater, and better than 40dB right up to 500MHz.

Given an ideal op-amp with infinite gain even at 500MHz, the negative feedback would ensure an effectively zero output impedance. Then, IC_1 would be able to swallow any current injected into its output from port B with none passing via R_1 to port A.

At lower frequencies this is exactly what happens, the lower trace reflecting in part the limitations of the instrumentation. The fixed 2.05GHz oscillator Tr_1 in Fig. 6 is of course running at the analyser's first intermediate frequency. So any leakage from Tr_1 back into the analyser's first local-oscillator output, and from there into the first intermediate

Equipment used for the testing

With such a wideband device, any sensible evaluation of its performance required some form of sweep equipment.

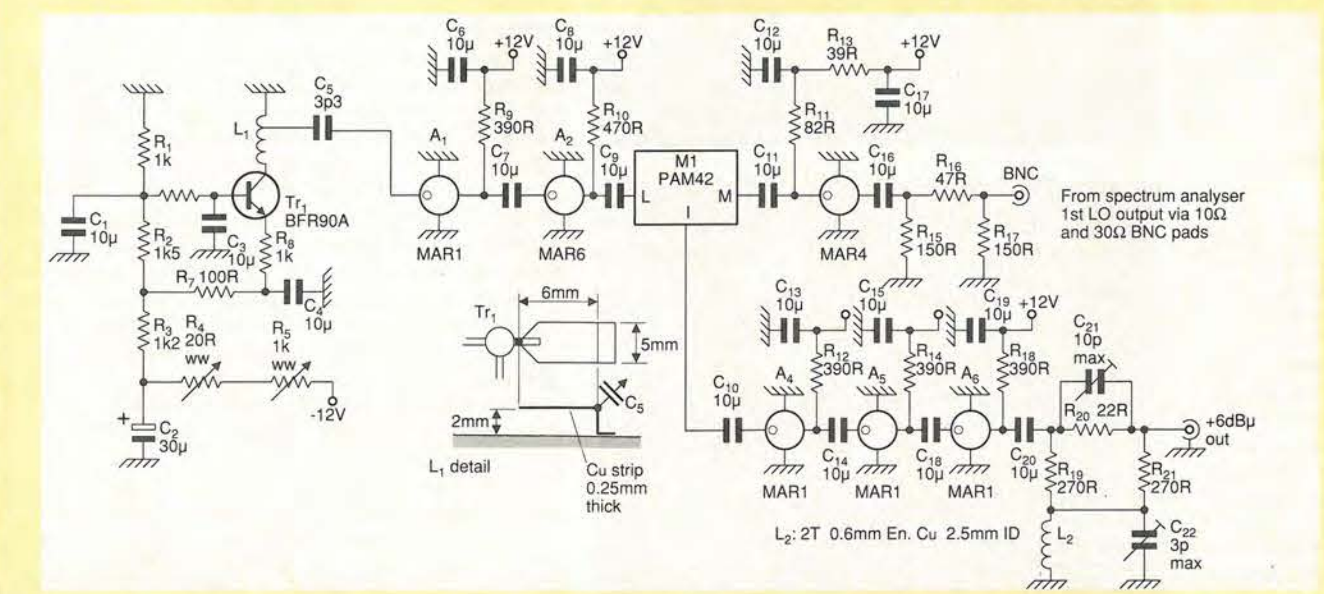
For general rf measurements, I have a Hewlett-Packard 0.1 to 1500MHz spectrum analyser type *8558B*, which is a plug-in unit fitted in a *182T* large screen display mainframe. I bought the mainframe and plug-in as a complete instrument, tested and guaranteed, from one of the dealers in this type of second hand equipment who advertises regularly in this magazine. Being an older instrument, long out of production, it is available at a very modest price, considering its performance.

Unfortunately, this instrument does not include a built-in tracking generator. Those only came in with the introduction of a later generation of spectrum analyser. But it does make a sample of the 2.05 to 3.55GHz first local oscillator available at the front panel.

Some time ago I published a circuit for an add-on for such an instrument.[2] It accepts an attenuated version of the spectrum analyser's first local oscillator output and mixes it with an internally generated continuous wave centred on 2.05GHz.

The output, as the spectrum analyser's first local oscillator sweeps from 2.05 to 3.55GHz, is a tracking output covering the analyser's 0 to 1500MHz input range.

Fig. 6. Circuit of an applique box for an HP 8558B spectrum analyser, providing a 0 - 1500MHz tracking generator output. (Reproduced courtesy* Electronic Product Design*, July 1994, page 17.



frequency stage, is by definition always on tune. Indeed, the purpose of $R_{4,5}$ is precisely to permit tuning of the fixed oscillator – which is not in any way frequency stabilised – to the analyser's first intermediate frequency.

The purpose of the external 13dB pad between the analyser's first local oscillator output and the applique box, and the latter's internal pad $R_{15\text{-}17}$ is to minimise this back-leakage. Despite these precautions, even with the input to the spectrum analyser closed in a 50Ω termination, the residual trace due to leakage is only a few decibels below that shown in **Fig. 8**.

Testing the isolator's directivity

My main use for the isolator is as a means of testing the vswr of various items of rf kit, such as antennas, attenuators, the input and output impedances of amplifiers, etc. To determine just how useful it was in this role, the output at port C was recorded, relative to the input at port A, for various degrees of mismatch at port B, **Fig. 9**.

The top trace is the output level with an open circuit at port B. Comparing it with the upper trace in Fig. 7, it is about 7dB down at 500MHz, this being the sum of the insertion loss from port A to port B, plus the insertion loss from port B to port C – already noted in Fig. 7 as around 3dB.

The three lower traces in Fig. 9 are with a 75Ω termination at port B providing a 14dB return loss, a 50Ω 10dB pad open at the far end providing a 20dB return loss, and three 50Ω 10dB pads terminated in 75Ω. The latter works out as a theoretical 74dB return loss, or close to 50Ω, and the resolution of the system as measured is apparently limited to around 40dB. A return loss of 40dB corresponds to a reflection coefficient of 1%. Now,

$$\rho = \frac{Z_t - Z_o}{Z_t + Z_o}$$

where Z_t is the actual value of the termination and Z_o is the characteristic impedance viz. 50Ω. So ρ=1% corresponds to a Z_t of 51Ω. The dc resistance looking into the string of three 10dB pads plus the 75Ω termination was measured at dc as 50.6Ω.

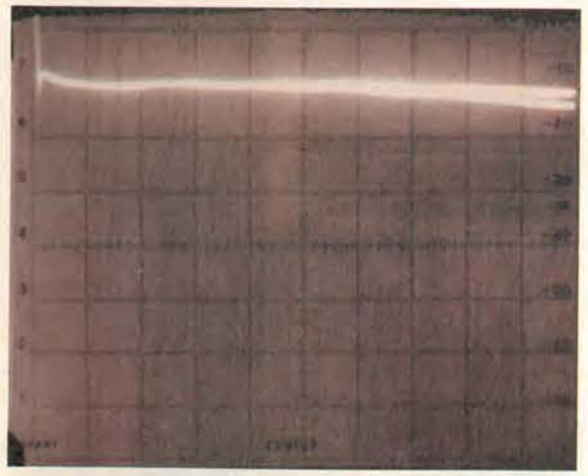


Fig. 7. Upper trace, output of the tracking generator, attenuated by 20dB. Lower trace, as upper trace, but with the signal routed via port B to port C of the isolator. Reference level -2.5dB, 10dB/division, span 0 to 500MHz, intermediate frequency bandwidth 3MHz, video filter medium.

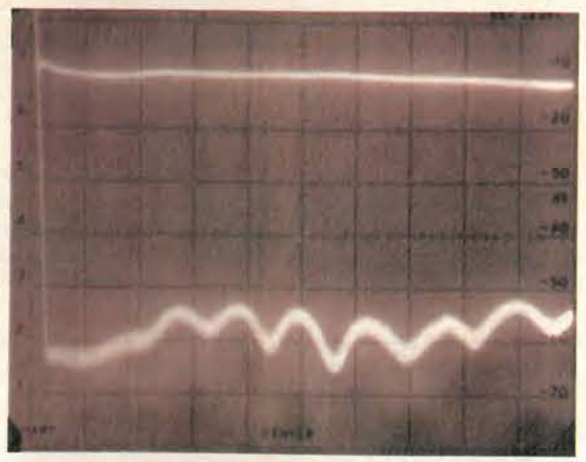


Fig. 8. Upper trace as Fig. 7, for reference. Lower trace, output from port A of the isolator with the input applied to port B. Spectrum analyser settings as for Figure 7 except video filter at max.

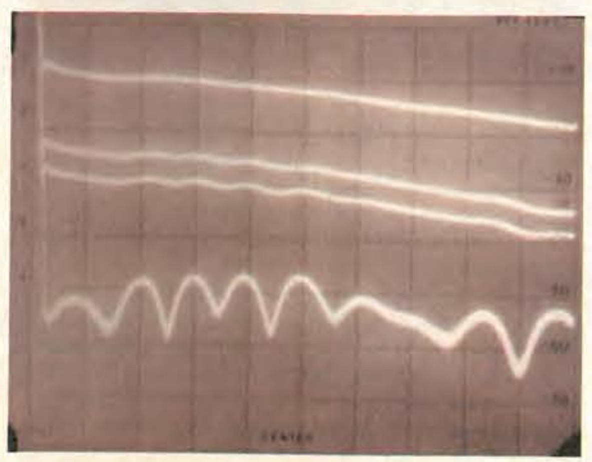


Fig. 9. Traces showing the signal at port C for various degrees of intentional mismatch at port B: with (top to bottom) return loss of 0, 14, 20 and 74dB. Signal applied to port A as in Fig. 7, upper trace. Spectrum analyser settings as for Fig. 8.

Clearly, then, assuming this is still the case at 500MHz, much of the residual signal in the bottom trace in Fig. 9 can be assumed to be due to the error in the characteristic impedance of the pads. These were normal commercial quality, as opposed to measurement laboratory standard.

For the rest, it is down to the limited directivity of the isolator. To maximise this, the chip resistors were all selected to be well within 1%, from the supply of 5% chips to hand. I had originally hoped to be able to select 326.3Ω resistors from the 313.5Ω to 346.5Ω spread of 330Ω 5% resistors. But most were in fact within 1%, hence the need for a parallel 15kΩ to secure the right value.

But the interesting – and indeed vital – point is that the directivity of the system does not depend on the flatness of the frequency response. The fact that the three upper curves in Fig. 9 are so nearly identical and parallel, indicates that the isolator is useful for vswr measurements right up to 500MHz, and perhaps a bit beyond. This is because the directivity depends upon two things.

Firstly, that the balance of the bridge of resistors at the input of IC_2 in Fig. 6 remains constant with frequency. Secondly, that the common mode rejection of the opamp remains high right up to 500MHz. And in view of the excellent results obtained, this certainly seems to be the case.

Using the isolator

The spectrum analyser, together with its hand-made tracking generator was very useful for demonstrating the isolator's performance over the whole band up to 500MHz in one sweep. But the arrangement has its limitations.

Apart from the back-leakage from the 2.05GHz oscillator, already mentioned, there are two other limitations. Firstly, as the 0 to 500MHz sweep proceeds, the frequency of the 2.05GHz oscillator tends to be affected slightly, so that it is necessary to use a wider than usual intermediate-frequency bandwidth in the analyser. Secondly, to maintain a sensibly flat output level, the output is taken from an overdriven string of amplifiers, with resulting high harmonic content. This is normally of no consequence, since the analyser is selective and is by definition, tuned only to the fundamental. But problems can arise with spurious responses due to the presence of the harmonics.

Where a more modest frequency range up to 200MHz suffices, the sweeper described in reference 3 can be used, in conjunction with a broadband detector – perhaps preceded by a broadband amplifier – connected to port C. A successive-detection logarithmic amplifier makes a very convenient detector, and types covering frequencies up to 500MHz are mentioned in reference 4.

For many applications, a swept measurement is not essential, for example when adjusting a transmitting antenna for best vswr at a certain frequency. In this case, any convenient signal generator can be used. At the higher frequencies however, it is best to keep the input to port A to not more than 0dBm.

A receiver can be pressed into service as the detector at port C. Many receivers, for example scanners, include an RSSI facility. In many cases, these make surprisingly accurate logarithmic level meters.

Measuring the level at port C relative to that at port A gives the return loss, and hence the vswr, of the device under test connected to port B. Tuning/adjusting it for maximum return loss will provide a device under test with an optimum vswr. Return loss measurements can be cross-checked at any time by substituting an attenuator(s) and/or 75Ω termination for the device under test, as described earlier.

Could it sing?

Finally, an interesting point about this active circuit. No problem was experienced at any stage with instability. But what about the circulator of Fig. 2? Here, any reflected power at port C circulates back around to port A.

What happens if all three ports are left open circuit? Given that tolerance variations on the resistors could result in a low-frequency gain marginally in excess of unity in each stage, could the circuit 'sing around' and lock up with the op-amp outputs stuck at the rail?

In fact the answer is no, because as Fig. 3 shows, when a port is open circuit, the output of the following op-amp is of the opposite polarity. In this way, the voltage passed on to the next stage is of the opposite polarity to the reflected voltage at the stage's input.

Three inverters in a ring are dc stable, and at frequencies where each contributes 60° phase shift or more, the loop gain is already well below 0dB. Of course, if all three ports are shorted, each stage passes on a voltage – possibly marginally greater – of the same polarity and lock-up is a possibility. But I can't think of any circumstances where one might want to try and use a circulator with all its ports short circuited! ■

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2. March, I, 'Simple tracking generator for spectrum analyser', *Electronic Product Design*, p. 17, July 1994.
3. Hickman, I, 'Sweeping to VHF', *Electronics World*, pp. 823-830, October 1995.
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Adaptive filters explained

An illustration of the usefulness of the application notes on this month's cover CD – the article presented here runs through the basics of choosing designing and applying adaptive digital filters.

Although well-known and widely used, adaptive filtering applications are not easily understood, and their principles are not easily simplified.

Currently, adaptive filtering is applied in such diverse fields as communications, radar, sonar, seismology, and biomedical engineering. These various applications are very different in nature, but one common feature can be noted: an input vector and a desired response are used to compute an estimation error, which is used, in turn, to control the values of a set of adjustable filter coefficients.

The adjustable coefficients may take the form of tap weights, reflection coefficients, or rotation parameters, depending on the filter structure employed.

Despite the diversity and complexity, a simple classification of adaptive filtering does emerge and practical applications can be demonstrated. This article begins by describing four basic classes of adaptive filtering applications and follows with sections that detail various fundamentals, techniques, and algorithms of several selected adaptive applications, **Table 1**.

Classifying adaptive filtering

Applications. Various applications of adaptive filtering differ in the manner in which the desired response is extracted. In this context, we may distinguish four basic classes of adaptive filtering applications, **Figs 1** to **4**.

- Identification
- Inverse modelling
- Prediction
- Interference cancelling

The functions of the four basic classes of adaptive filtering applications follow.

Identification, Fig. 1. The notion of a mathematical model is fundamental to sciences and engineering. In the class of applications dealing with identification, an adaptive filter is used to provide a linear model that represents the best fit to an unknown plant.

The plant and the adaptive filter are driven by the same input. The plant output supplies the desired response for the adaptive filter.

If the plant is dynamic in nature, the model will be time varying.

Inverse modelling, Fig. 2. In this second class of applications, the adaptive filter provides an inverse model repre-

Figs 1-4. The four basic classes of adaptive filtering. In these diagrams, u is input applied to the adaptive filter, y is output of the adaptive filter, d is desired response and e is d–y, which is the estimation error.

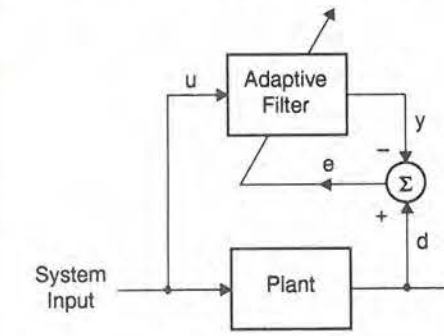

Fig. 1. Identification.

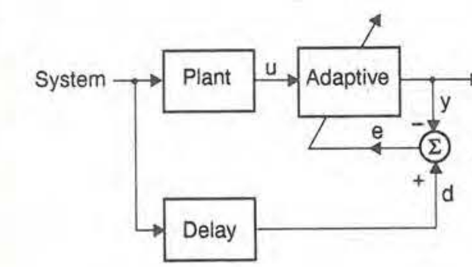

Fig. 2. Inverse modelling.

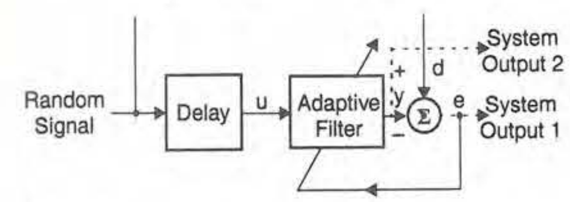

Fig. 3. Prediction.

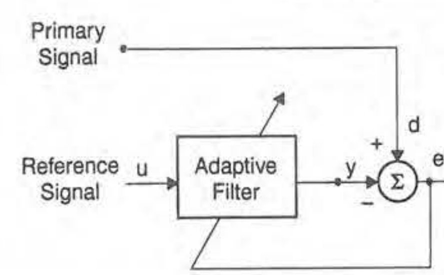

Fig. 4. Interference cancelling.

Table 1. Adaptive filtering applications.

Adaptive filtering class	Application
Identification	System identification
	Layered Earth modelling
Inverse modelling	Predictive deconvolution
	Adaptive equalisation
Prediction	Linear predictive coding
	Adaptive differential pcm
	Auto-regressive spectrum analysis
	Signal detection
Interference cancelling	Adaptive noise cancelling
	Echo cancellation
	Radar polarimetry
	Adaptive beam forming

senting the best fit to an unknown noisy plant. Ideally, the inverse model has a transfer function equal to the reciprocal of the plant's transfer function.

A delayed version of the plant input constitutes the desired response for the adaptive filter. In some applications, the plant input is used without delay as the desired response.

Prediction, Fig. 3. In this example, the adaptive filter provides the best prediction of the present value of a random signal.

The present value of the signal serves the purpose of a desired response for the adaptive filter. Past values of the signal supply the input applied to the adaptive filter.

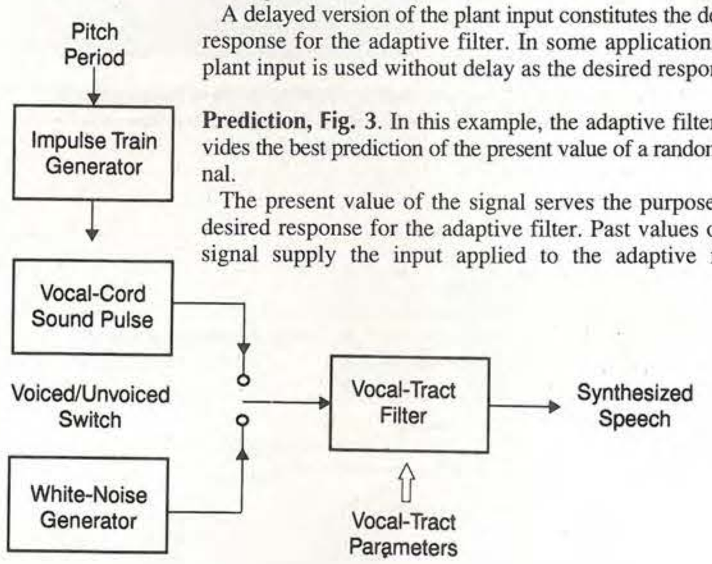

Fig. 5. Simplified model for the speech production process in block form.

Depending on the application of interest, the adaptive filter output or the estimation error may service as the system output. In the first case, the system operates as a predictor; in the latter case, it operates as a prediction error filter.

Interference cancelling, Fig. 4. In this final class of applications, the adaptive filter is used to cancel unknown interference contained in a primary signal, with the cancellation being optimised in some sense.

The primary signal serves as the desired response for the adaptive filter. A reference signal is employed as the input to the adaptive filter. The reference signal is derived from a sensor or set of sensors located in relation to the sensor(s) supplying the primary signal in such a way that the information-bearing signal component is weak or essentially undetectable.

Prediction

The coders used for the digital representation of speech signals fall into two broad classes: source coders and waveform coders. Source coders are model dependent. This means that they use *a priori* knowledge about how the speech signal is generated at the source.

Source coders for speech are generally referred to as vocoders. Vocoders can operate at low coding rates; however, they provide a synthetic quality, with the speech signal having lost substantial naturalness.

Waveform coders, on the other hand, strive for facsimile reproduction of the speech waveform. In principle, these coders are signal independent. They may be designed to provide telephone-toll quality for speech at relatively high coding rates.

In the context of speech, linear predictive coding, or lpc, strives to produce digitised voice data al low bit rates (2.4 to 4.8kbit/s) with two important motivations in mind:

- The use of linear predictive coding permits the transmission of digitised voice over a narrow-band channel.

- The realisation of a low bit rate makes the encryption of voice signals easier and more reliable than would otherwise be the case.

Figure 5 shows a simplified block diagram of the classical model for the speech production process. In this particular example, the sound-generating mechanism is linearly separable from the intelligence-modulating, vocal-tract filter. The precise form of the excitation depends on whether the speech sound is voiced or unvoiced.

Voiced speech sound is generated from quasi-periodic, vocal-cord sound. In the speech model, the impulse-train generator produces a sequence of impulses, which are spaced by a fundamental period equal to the pitch period. In turn, this signal excites a linear filter whose impulse response equals the vocal-cord sound pulse.

An unvoiced speech sound is generated from random sound produced by turbulent air flow. In this case the excitation consists simply of a white noise source. The probability distribution of the noise samples does not appear to be critical.

Figure 6 shows the block diagram of an linear predictive coding vocoder, comprising a transmitter and a receiver. The transmitter first applies a window to the input speech signal. It does this to identify a block of speech samples for processing.

The window used is short enough for the vocal-tract shape to be nearly stationary. As a result, the parameters of the speech-production model may be treated as essentially constant for the duration of the window.

Next, the transmitter analyses the input speech signal in an adaptive manner – block by block – by performing a linear prediction and pitch detection. Finally, it codes the parameters made up of the set of predictor coefficients, the pitch period, the gain parameter, and the voiced-unvoiced parameter, for transmission over the channel.

The receiver performs the inverse operations by first decoding the incoming parameters. In particular, it computes

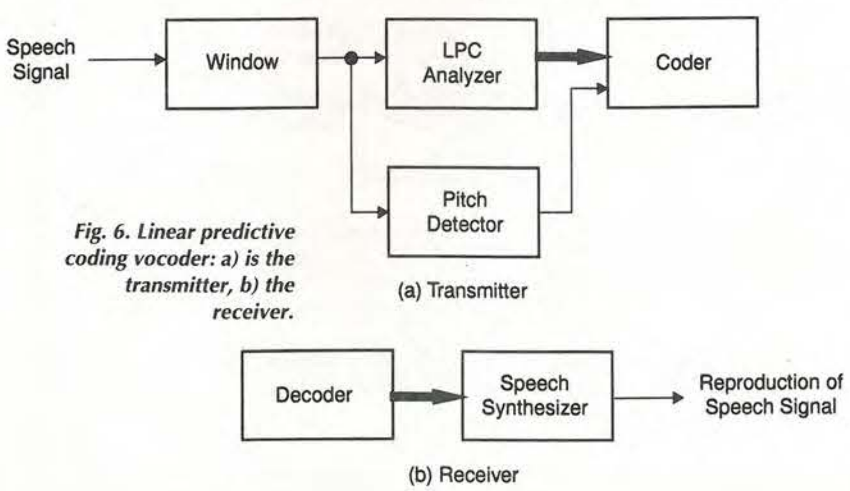

Fig. 6. Linear predictive coding vocoder: a) is the transmitter, b) the receiver.

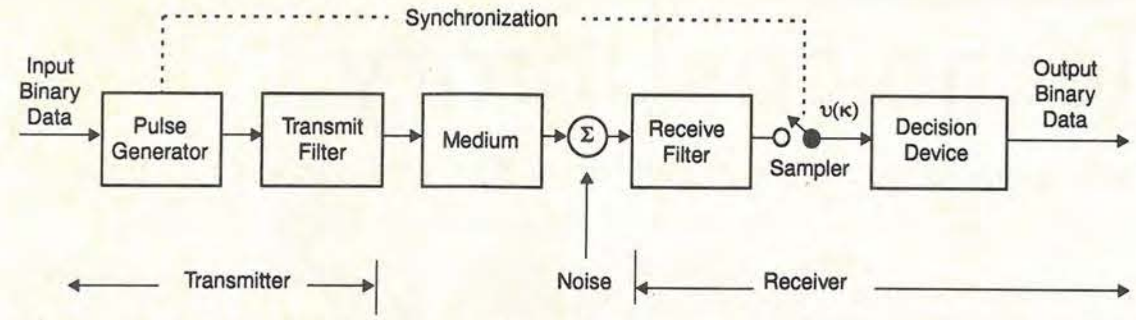

Fig. 7. Baseband data transmission system without equalisation, summarised in block form.

the values of the predictor coefficients, the pitch period, and the gain parameter. It then determines whether the segment of interest represents voiced or unvoiced sound. Finally, the receiver uses these parameters to synthesise the speech signal by utilising the model of Fig. 5.

Adaptive equalisation

During the past three decades, considerable effort has been devoted to the study of data-transmission systems that use the available channel bandwidth efficiently. The objective here is to design the system to accommodate the highest possible rate of data transmission, subject to a specified reliability that is usually measured in terms of the error rate or average probability of symbol error.

The transmission of digital data through a linear communication channel is limited by two factors. One is intersymbol interference, the other additive thermal noise.

Intersymbol interference. Also known as ISI, this is caused by dispersion in the transmit filter, the transmission medium, and the receive filter.

Additive thermal noise. Generated by the receiver at its front end. For bandwidth-limited channels, intersymbol interference seems to be the chief determining factor in the design of high-data-rate transmission systems.

Figure 7 shows the equivalent baseband model of a binary pulse-amplitude modulation, or PAM, system. The signal applied to the input of the transmitter part of the system consists of a binary data sequence, in which the binary symbol consists of 1 or 0. This sequence is applied to a pulse generator, the output of which is filtered first in the transmitter, then by the medium, and finally in the receiver.

Let $u(k)$ denote the sampled output of the receive filter in Fig. 7; the sampling is performed in synchronism with the pulse generator in the transmitter. This output is compared to a threshold by means of a decision device. If the threshold is exceeded, the receiver makes a decision in favour of symbol 1. Otherwise, it decides in favour of symbol 0.

On a physical channel there is always intersymbol interference. To overcome intersymbol interference, control of the time-sampled function is required. In principle, if the characteristics of the transmission medium are known precisely, then it is virtually always possible to design a pair of transmit and receive filters that will make the effect of intersymbol interference arbitrarily small.

For adequately reducing the intersymbol interference, an adaptive equaliser will provide precise control over the time response of the channel.

An adaptive filtering algorithm requires knowledge of the desired response to form the error signal needed for the adaptive process to function. In theory, the transmitted sequence is the desired response for adaptive equalisation.

In practice, however, with the adaptive equaliser located in the receiver, the equaliser is physically separated from the origin of its ideal desired response.

There are two methods in which a replica of the desired response may be generated locally in the receiver. First is the training method, second the decision-directed method.

Training method. In this method, a replica of the desired response is stored in the receiver. Naturally, the generator of this stored reference must be electronically synchronised with the known transmitted sequence.

Decision-directed method. Under normal operating conditions, a good replica of the transmitted sequence is being

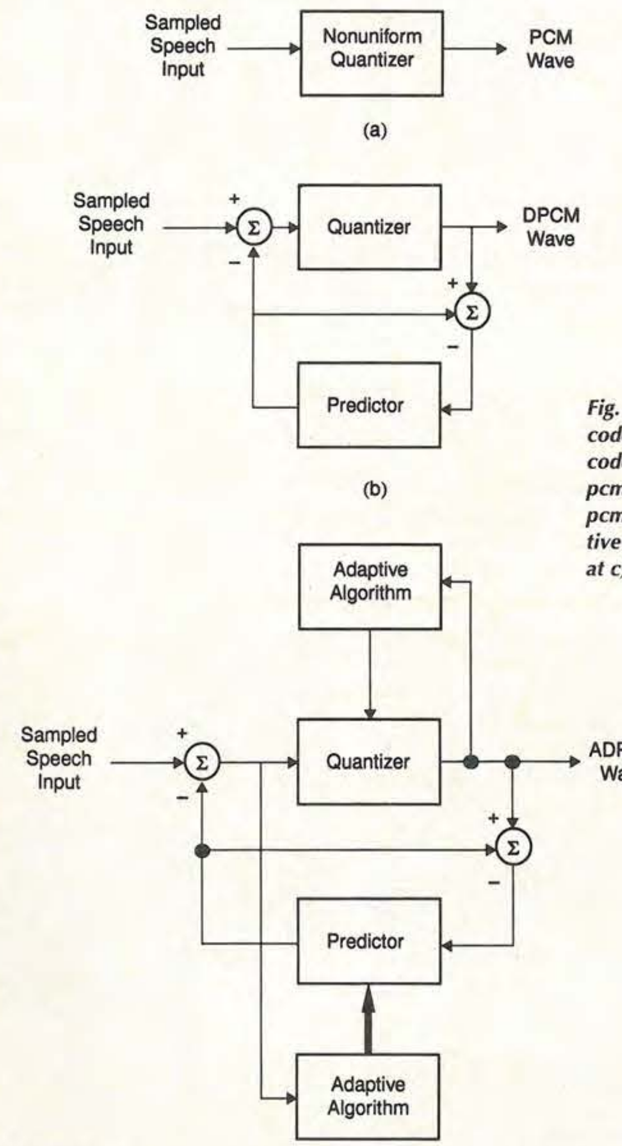

Fig. 8. Waveform coders, with pulse-code modulation, or pcm, at a), differential pcm at b) and adaptive differential pcm at c).

Continued on page 244

Zilog Technical Library

Information representing one company's complete set of data and applications book, brought to you by leading UK Zilog distributor and design house, Gothic Crellon.

The CD ROM given away free with this issue* is the complete Zilog Technical Library CD-ROM from Gothic Crellon. It contains not only data on the company's microcontrollers, digital signal processors and communications controllers, but also a host of application notes relating to all three product areas. These notes includes a caller ID system complete with software listings, a discussion of how to use a microcontroller as an I^2C bus master and a high-performance PC COM port design. Development tools for the various product areas are also listed, together with a number of demonstrations.

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System requirements

Windows platforms:
- 386, 486 or Pentium-based personal computer
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- 8MB Application RAM, 4MB hard disk space and CD-ROM drive

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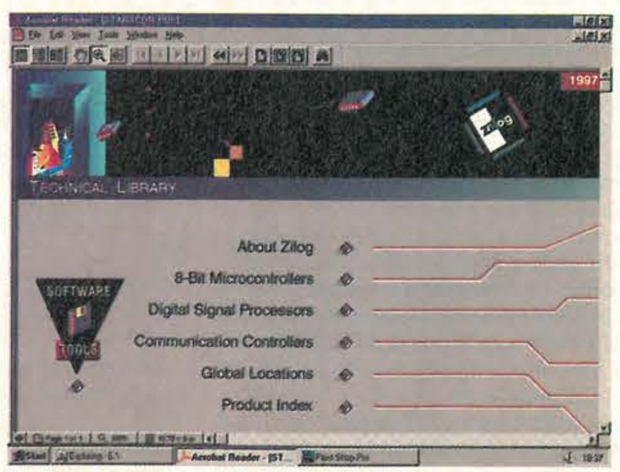
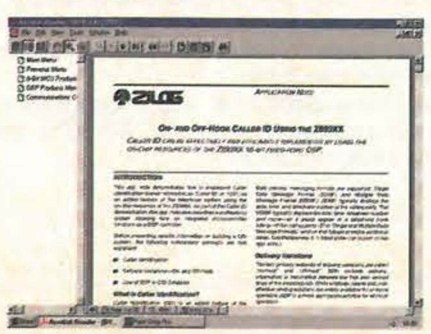


Being based on the widely used Adobe Acrobat file reader, Zilog's data and applications suite has been developed to run smoothely on a number of platforms, including Unix and the MAC.



Stereo audio monitor

When making audio recordings, it is useful to have an at-a-glance overview of the phase and amplitude relationships between the two stereo channels – particularly if there is a chance that the two channels will be combined to produce mono. Robert Kesler's Dotscope provides that information, together with a direct mono compatibility indication.

Visual stereo monitors are used in sound recording and broadcast studios to monitor the phase and amplitude relationship of the two audio channels of a stereo signal.

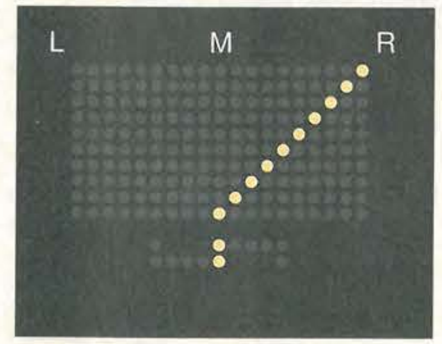
These features are not revealed by simply listening in stereo. If the stereo channels were opposite in phase to each other, the program could still sound acceptable in stereo. But if the same two channels were combined and listened to in mono, the signals would cancel each other out.

A simple visual stereo program monitor can be set up using an oscilloscope in X-Y mode. The sum of the left and the right channels is fed to the Y vertical inout and their difference feeds the X input. In this way, a mono signal appears as a vertical line while a signal on the right or left-hand channel only appears as a line tilted right or left by 45°. Two anti-phase inputs produce a horizontal line.

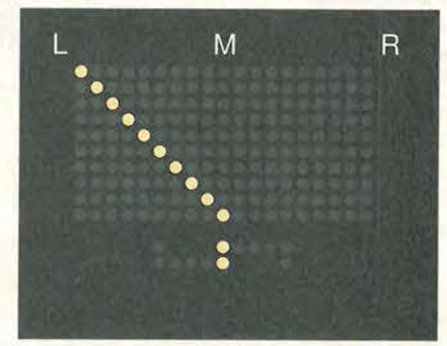
Note that the effect produced by feeding the X and Y chhannels with the sum and difference signals could be reproduced by tilting the oscilloscope by 45°, and feeding the right and the left channels to the X and Y inputs, as in **Fig. 1**. Using sum and difference signals simply provides a more elegant solution that tilting the display.

Of course a usable instrument for this purpose should have a gain control to allow it to cover the dynamics of the programs. It also needs an automatic illumination suppressor, to avoid burning out the phosphor layer in the centre of the screen when there is no input signal.

A pure sinusoidal input of slightly different phase on both channels produces a narrow ellipse. This



Only the right channel is present



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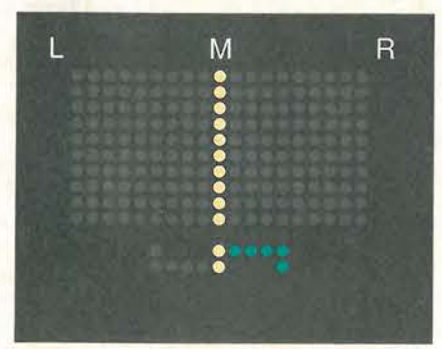
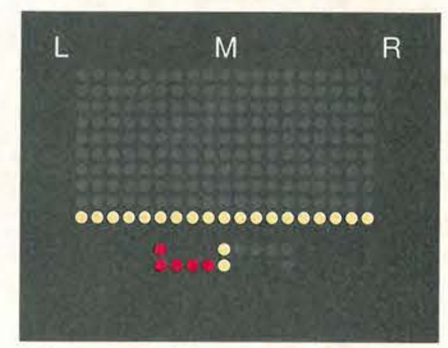


Left and the right channels identical in intensity and phase



Left and right channels are identical in intensity, but of opposite phase

Examples of Dotscope displays with various input signals.

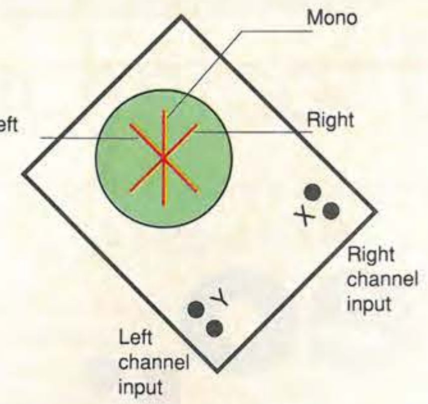


Fig. 1. Tilting an oscilloscope like this has the same effect as leaving it upright and feeding the and Y channels with left+right and left–right signals.

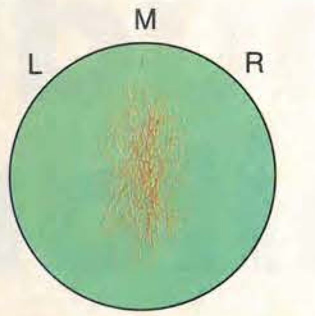


Fig. 2. A good stereo signal looks like a loosely wound clew of thread. A mostly anti-phase signal on the other hand – i.e. one that is not stereo compatible – looks similar except that it is wide and flat.

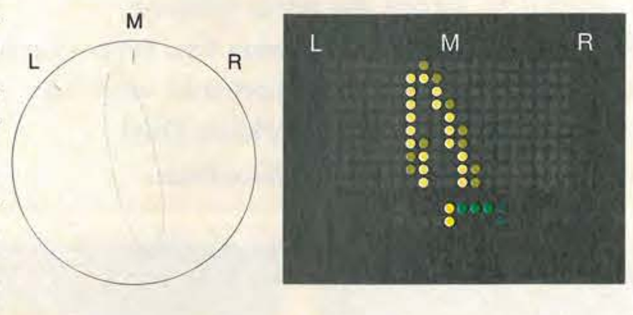
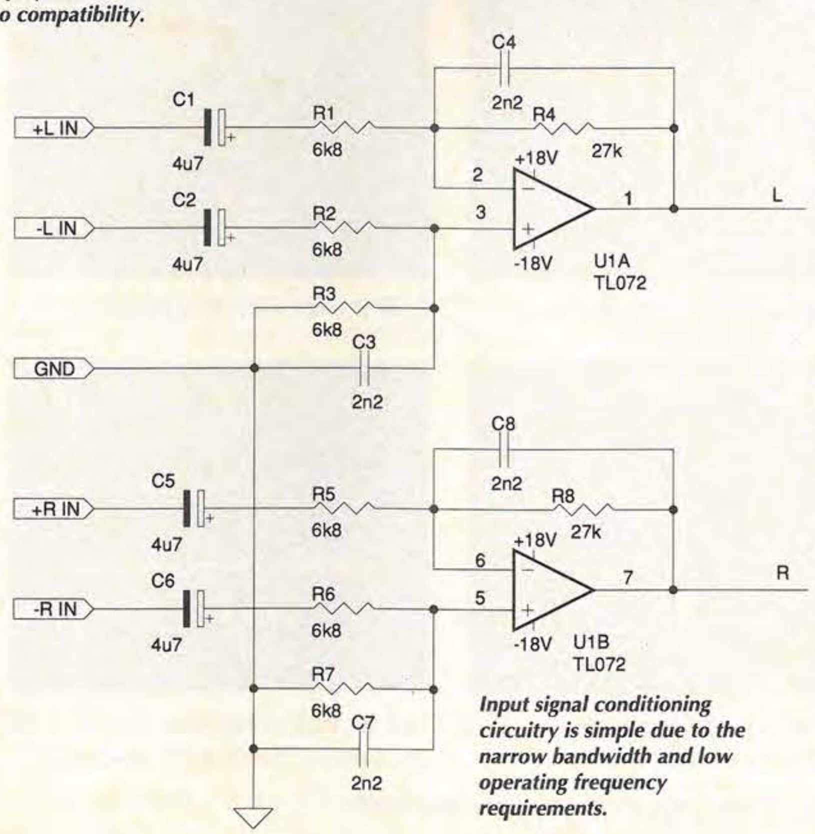


Fig. 3. Representation of a clean, slightly out of phase signal as it would appear on a conventional oscilloscope, left, and on the Dotscope, right. Note the stereo/mono bar graph at the bottom of the dotscope display indication mono compatibility.



Input signal conditioning circuitry is simple due to the narrow bandwidth and low operating frequency requirements.

ellipse becomes circular if the phase difference is 90°. A complex signal from more than one sound source is represented by an irregular clew, **Fig. 2.**

For a good stereo signal, the squiggles of the clew are roughly contained within an ellipse. This elipse has its longer axis pointing in the direction of the signal source. A flat shape, pointing left-to-right means that the two channels are out of phase. This is what the sound engineer wants to avoid most.

The Dotscope alternative

In the monitor described here, a led-matrix display replaces the crt. The matrix was chosen because it was cheaper, but it later turned out to offer other advantages.

Surprisingly, the low resolution makes interpreting the display easier. The chunky picture shows the overall quality of the signal, without the clutter of unnecessary detail. For a sound engineer, speed of informatiion assimilation is important since there are many other things to take into acount like the performers and peak-meters.

Due to the fact that the displayed picture is basically symmetrical, the lower part of the led matrix has been omitted. Doing this halves the number of needed without sacrificing resolution, simplifying the design and reducing costs.

A second indicator is incorporated into the design. This indicator monitors the average phase difference between the channels and displays it on a multi-colour bi-directional led bargraph (see panel).

If the stereo is mono compatible, the green half of the bar graph shows the degree of compatibility. The left-hand half of the indicator is red to draw the operator's attention if the signal is out of phase.

System elements

Four *LM3914* bar graph drivers are used in the instrument. With a little overstatement, this IC could be called a decoded analogue-to-digital converter. Basically it is made up of ten comparators, driving ten open-collector outputs. The ten comparators are tied to ten, twenty, or whatever, percent of the reference voltage connected between the R_{HI} on pin 6 and R_{LO} on pin 4.

There is an in-built programmable reference generator. Its output, REF_{OUT} on pin 7, is not used here, but the same pin is used to program the output current limiter.

Pin 9, the mode input, is used to choose between the bar mode and dot mode. In this design, all four *LM3914*s are used in dot mode. As a result, the mode pins must be left floating.

Finally, note that the first led output has a considerable leakage current in dot mode – up to 0.45mA – so these outputs are equipped with bleeding resistors.

Input amplifiers. Simple amplifiers with electronic symmetry are adequate for such monitoring. Frequency bandwidth of the instrument is set relatively narrow; this monitor concentrates on the middle frequencies, which are the most relevant for the stereo perception.

Summing amplifiers. On the output of IC_{2A} there is a signal proportional to the difference between the right and the left inputs, while the IC_{2B} output delivers the difference between the left and the right inputs. This is actually, the same signal but in opposite phase. Op-amp IC_{3A} produces the sum of the same signals.

Automatic gain control. The automatic gain control used is simple, yet it ensures that the picture fills the whole display

Continued over page

Bidirectional bar graph features low current drain

Sometimes it is necessary to display the results of a measurement showing the value in two opposite directions, in the same way that a centre-zero meter does.

Two of circuits shown here are designed for this purpose. They not only provide bidirectionality, but also a ten-fold reduction in current consumption. When all leds are switched on, the conventional 3914 bar graph draws ten times the current needed by one led since the diodes are parallel connected. These new configurations have the diodes series connectedso current demand is reduced to that needed by one led. The trade off, of course, is that the supply voltage needs to be raised high enough to be able to turn on all the series leds.

The basic circuit is shown in **Fig. A**. By leaving the mode pin floating, the *LM3914* is configured in dot-mode, that is, only one of the open collector outputs is active, at any time. However, in contrast with the usual circuit, not all the leds' anodes are connected to the supply line. The leds are connected in series, in one or two groups between output associated with the centre, origin and the output or outputs associated with the end or ends of the bar graph. The intermediate taps are connected to the other outputs. In this way, all the leds between the active output and the supply line light, creating a bar above or below the point of origin.

The centre point of the bar-graph, the origin, need not be in the actual centre. It may in fact be associated with any output pin, depending on what you need. The bar-graph can be made symmetrical, asymmetrical, or simply a low current but otherwise conventional one-way bar graph. For the latter, connect the supply to the led1 output, and all the cathodes pointing to the same output.

If, on the other hand, you connect the supply next to the led-10 output, the bar will shrink instead of grow with increasing input voltage.

For the circuitry for reference voltage and input voltage scaling and shifting, refer to the manufacturer's data sheets and application notes.

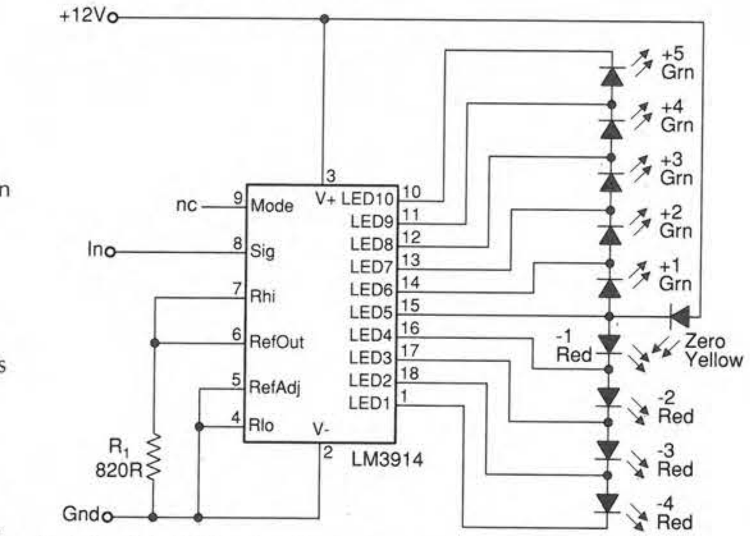


Fig. B. Modified bidirectional bar graph. The zero led is always lit. For a symmetrical display omit led +5 and short the connections.

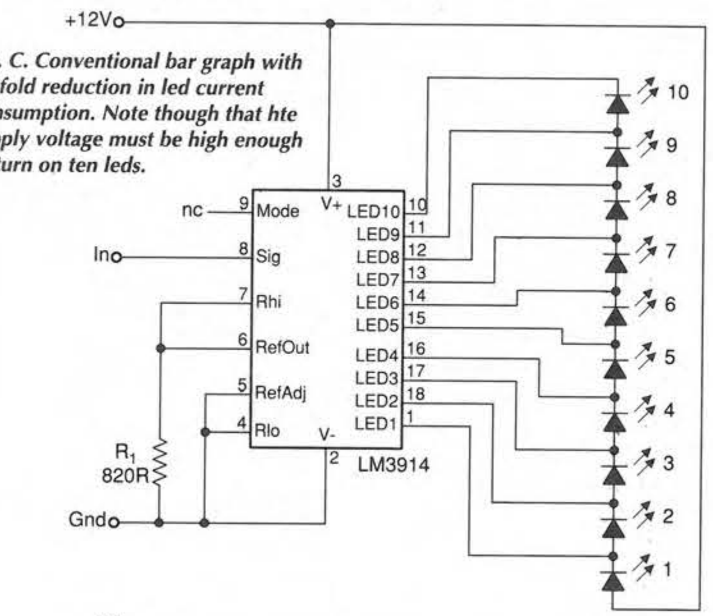


Fig. C. Conventional bar graph with tenfold reduction in led current consumption. Note though that hte supply voltage must be high enough to turn on ten leds.

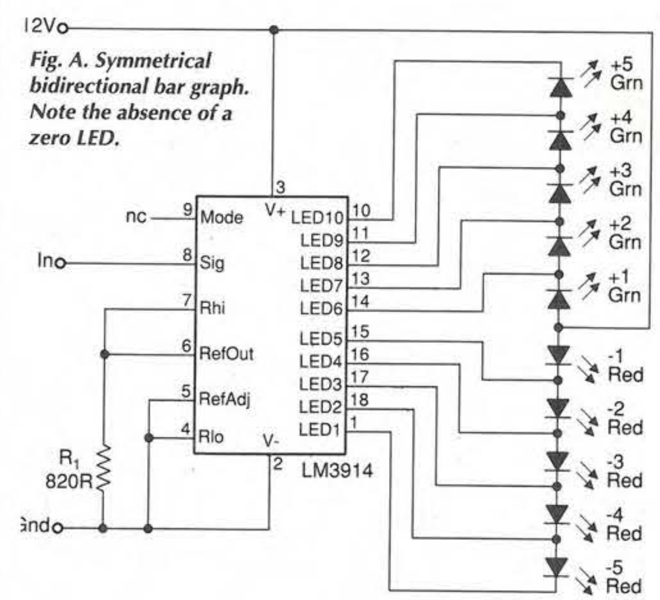


Fig. A. Symmetrical bidirectional bar graph. Note the absence of a zero LED.

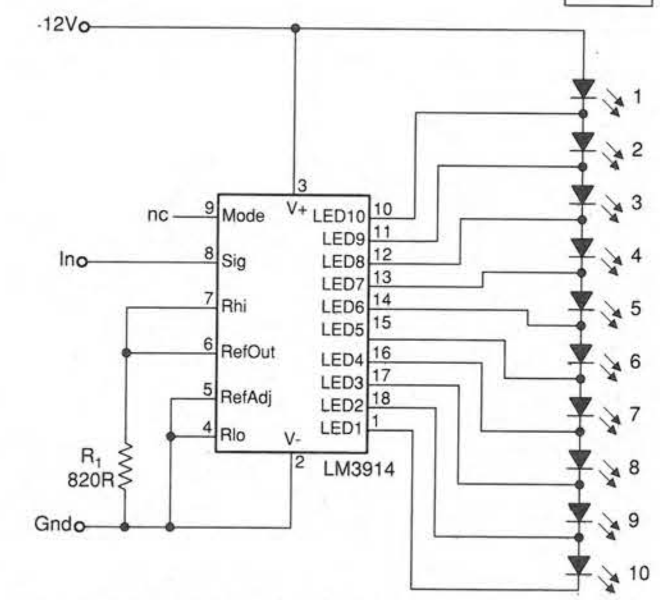
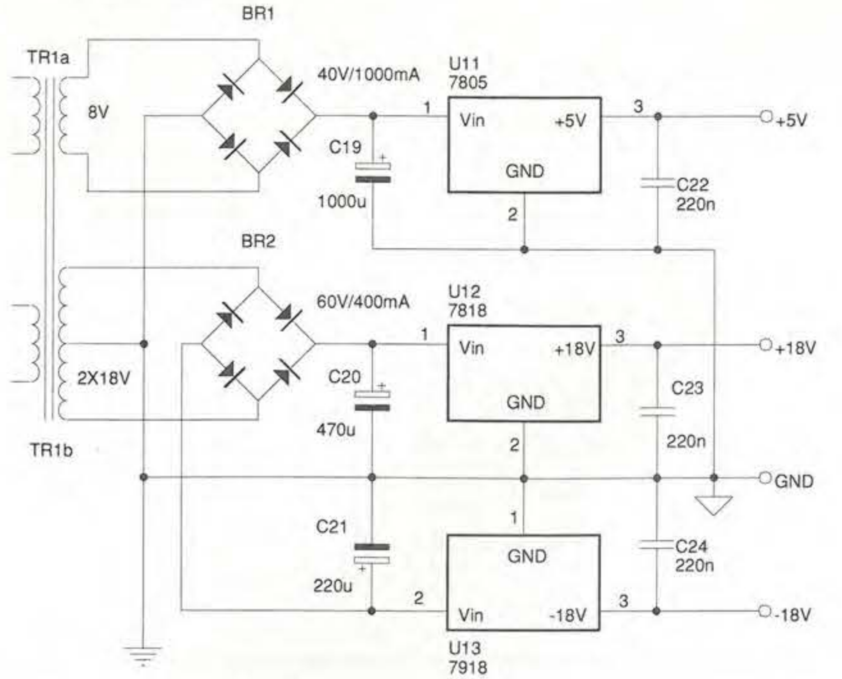
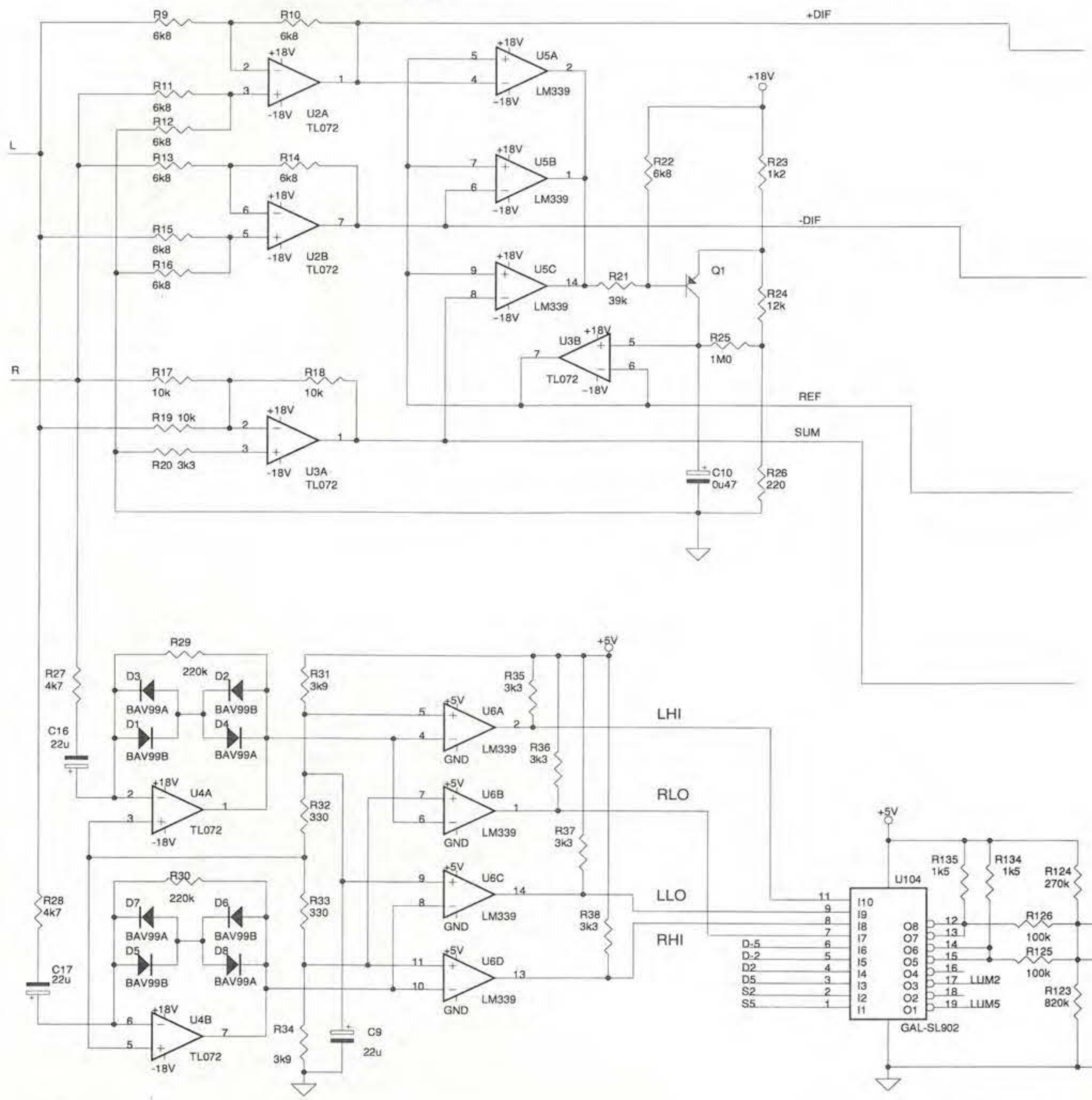


Fig. D. Shrinking bar graph with tenfold reduction in led current. The length of the bar reduces with increasing input voltage. Again, the supply voltage must be high enough to turn on all leds.



Power supply needs no comment except that newcomers to the three-terminal regulators shold note that if the device is more than a couple of inches away from the transformer, extra ceramic 0.1mF input capacitors should be added near to the input pins.

At the top of this diagram are the sum and difference stages and their associated comparators. Note the reference input feeding the comparator non-inverting inputs. Below is the illumination control and stereo-mono compatibility logic array, together with its input conditioning.



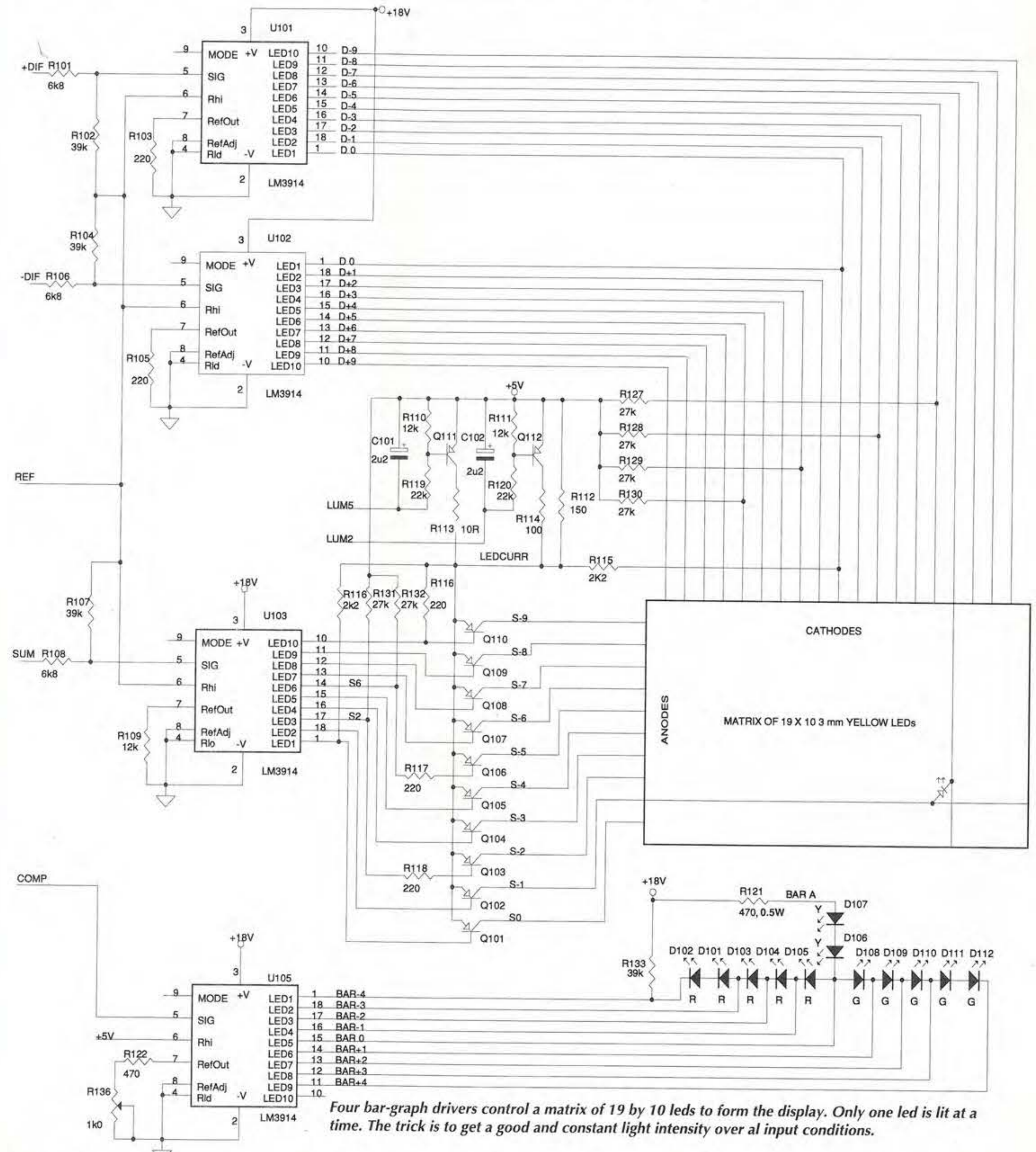
over an input range of about 32dB. It drives the reference voltage of all three dot-matrix-driver analogue-to-digital converters, so that the reference voltage promptly follows any increase of the highest of input of the same converters. It also assures that recovery follows slowly.

This technique is very efficient and easier to implement then synchronously controlling two or three voltage-controlled amplifiers.

Limiter amplifiers. For each channel, an amplifier and two comparators provide a logic level output for inputs above the threshold and no output for all inputs below it.

The two outputs are used for the average phase measurement, representing stereo/mono compatibility. The threshold is introduced to avoid erratic display when one or both inputs is too low, or missing.

Dot-matrix driver a-to-d converters. Two *LM3914* drive the columns; one is fed with the difference of the left and the right channel, the other with the same signal of opposite phase.



Four bar-graph drivers control a matrix of 19 by 10 leds to form the display. Only one led is lit at a time. The trick is to get a good and constant light intensity over al input conditions.

A part of the reference voltage is added to both inputs. This results in the the first led output of both ICs being driven when the input signal is zero. These two outputs are connected in parallel drive the central, zero column.

The other outputs are driven one at a time, as the input voltage changes. Note that only one column is driven at any one time.

The vertical a-to-d converter is configured similarly, but it drives the rows via inverting transistors. As the picture is symmetrical around the central point, only half of the picture is displayed, i.e. the components that would be needed to convert and display the rows below the zero line are omitted.

Similarly, as noted for the columns, only one row is driven at any one time. But I should emphasize here that no row is driven during the negative half periods. In other words, for almost half of the time no led is lit at all.

Dot-matrix display. The dot matrix display is composed of ten lines by nineteen columns of yellow leds. They should be narrow angle types, and of same intensity class.

All the anodes within a row and the cathodes within a column are tied together. The columns are driven directly by the two *LM3914s*, while the rows via inverting transistors.

A single led at the crossing of the driven row and of the driven column is lit at any one time. The illumination control circuit determines the current.

Illumination control. As noted above, only one led is lit at any one time and almost half of the time no led is lit at all. This would cause an apparent loss in light intensity. To compensate, the leds must be driven with higher currents than allowed in continuos operation. But under very low, or no input condition, when only a few leds share the current, the current must be reduced to a safe level.

The number of active columns and rows is monitored by the GAL inputs 1 to 6. The GAL in turn controls the current source made up of Tr_{111} and Tr_{112}. This current source determines the current of the led-matrix display, **List 1**.

Stereo/mono compatibility bar graph. The remaining part of the GAL converts the phase difference of the two channels into a voltage, which drives the bi-directional multicolour bar graph. It displays a right-bound green bar to indicate mono-compatible stereo. Two yellow center leds are always lit. When one or both channels missing, below the threshold, or totally different, these are the only leds lit on the graph. Finally, the bar graph displays a left-bound, red bar for anti-phase channels.

Implementing the design

Resistors throughout should be 2% tolerance or better. I suggest a sandwich construction: one pcb with the leds, and the other one carrying the rest of the components, except the power supply.

In my design, the 3mm leds are placed at 0.15-inch pitch, so the display board is about 85 by 80 mm. Using surface-mount resistors, diodes and transistors, the main pcb need not be bigger than the display board. The two boards are linked via a 40-pin connector.

I placed a sheet of smoked acrylic glass in front of the sandwich and sprayed the pcb black before soldering the leds in place. The whole construction without the power supply is only about 30mm deep, so it may be mounted alongside the VU or peak meters or in a small stand-alone box, including the power supply. ■

List 1. These are the equations needed to program the GAL.

```
>
>PINS   /* in sequential order: pin 1 to pin 20 */
>
>S5,  S2,  D5,  D2, D-2, D-5,   RLO,  RHI,   LLO, GND,
>LHI, Q2, Q2A,  Q1, Q1A, HELP1, LUM2, HELP2, LUM5, VCC
>
>EQUATIONS
>
>Q1 =          /RLO * /RHI *  LLO *  LHI
>      +        RLO * RHI  * /LLO * /LHI
>
>Q1A = Q1             /* parallel output for more drive */
>
>Q2 =          /RLO * /RHI * /LLO * /LHI
>      +        RLO *  RHI * LLO  *  LHI
>
>Q2A = Q2             /* parallel output for more drive */
>
>/* Open collector output simulation using two (not quite) buried nodes:
>HELP1 and HELP2  */
>
>HELP1 = D5 * S5 * D-5
>HELP2 = D2 * S2 * D5  * S5 * D-2 * D-5
>
>/* The following equations use the product-term enabled tri-stating
ability
>of the 16V8 GAL */
>/* The output values of LUM2 and LUM5 are constants (GND), the output
>driver being enabled  */
>/* only if the term in parentheses is true  */
>
>IF (/HELP1) LUM5 = GND
>/* OPEN-COLLECTOR simulation using product term HELP1 */
>IF (/HELP2) LUM2 = GND
>/* OPEN-COLLECTOR simulation using product term HELP2 */
>
```

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HP3763A	Error Detector	£800.00
HP3764A	Digital Transmission Analyser	£3,200.00
HP3770B	Telephone Line Analyzer	£200.00
HP3780A	Pattern Generator/Error Detector	£1,100.00
HP3781A	Pattern Generator	£500.00
HP3782A	Error Detector	£500.00
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Marconi 2828A	Digital Simulator	£200.00
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Marconi OA2805	PCM Regenerator Test Set	£250.00
Marconi TF2019C	Noise Generator - Many Filters Available	£250.00
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Meguro	MK612A VTR Jitter Meter	£75.00
Phoenix	5500A Telecommunications Analyzer	£1,500.00
Racal 202, 205	State Logic Analyzer	£150.00
Schlumberger	7700 140 m Bit/s Filtermeter/Generator	£125.00
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Siemens	D2155 Level Meter/W3155 Tracking Oscillator	£150.00
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Tek 834	Programmable Date Comms Tester	£150.00
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W&G	SPM15 Level Meter	£800.00
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Racal 9302	RF Millivolt Meter	£350.00
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Farnell	2085 AF Power Meter	£40.00
Farnell	DM131 Digital Multimeter	£30.00
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Marconi	6587 Levelling Amp	£25.00
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Philips	PM2504 Electronic VA Meter	£75.00
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Philips	PM5108 Function Generator 1Hz-1MHz	£95.00
Philips	PM5142 0-01Hz LF Synthesizer	£60.00
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Trio	CS/1566A 20MHz Oscilloscope	£60.00
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W&J	SM/1622 Signal Monitor	£30.00
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W&J	TFC103 Telephone Frequency	£30.00
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W&J	SM/6108 Signal Monitor	£30.00
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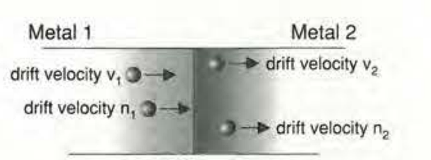
CIRCLE NO. 130 ON REPLY CARD

Using Peltier devices

Peltier-effect devices represent a rival solid-state method for cooling small items. Richard Lines discusses how they work and explains how to apply them. His design is illustrated by a design example for cooling a ccd.



Due to Peltier effect, electrons crossing the dissimilar-metal junction energy change in energy level, producing or removing heat depending on the direction of current flow. With dissimilar metals, the effect is negligibly small, but when applied to a special semiconductor junction, Peltier effect becomes a useful means of extracting heat.

Thermoelectric coolers are convenient solid-state devices capable of pumping heat up a thermal gradient. They are very useful for cooling small items like charge-coupled device sensors, photodiodes and semiconductor lasers.

The principle of operation is closely related to the Seebeck effect, where a temperature difference between two junctions generates an emf. This effect was described in relation the thermocouple in my previous article.

The Peltier effect involves passing a current through a junction which produces a heating or cooling effect at the junction depending on the direction of the current.

The basic principle is outlined in **Fig. 1**. This illustrates electrons flowing from left to right across a metal junction. The solid state properties of the metal dictate the average energy of the conduction electrons – E_1 for metal 1 and E_2 for metal 2. Thus the electrons in metal 1 will bring energy up to the junction at a rate of,

$$E_{in} = n_1 E_1 v_1$$

and similarly energy will be carried away from the junction,

$$E_{out} = n_2 E_2 v_2$$

and the difference between these two terms is the energy taken from or given to the atoms in the lattice. Since the current must remain the same over the junction you can say that,

$$I = n_1 \times v_1 \times q = n_2 \times v_2 \times q$$

where q is the electronic charge, and I is the current.

Net heat flow from the junction is,

$$E_{out} - E_{in} = \frac{1}{q}(E_2 - E_1)$$

So the heat removed by a Peltier device is directly proportional to the current and the difference in energy carried by the conduction electrons.

The Peltier effect is due to the bulk properties of the two metals; in effect the heat capacity of the conduction electrons changes on crossing the junction. With normal metals, say copper/iron, the Peltier effect is very small and difficult to measure. In practice the normal ohmic heating effect is much larger. However, in semiconductors the effect is exaggerated and can be exploited as a useful method for cooling

A thermoelectric cooler consists of an array of p-n junctions wired in series between two thermally conductive ceramic plates. The p-n junctions are so wired that heat is absorbed on one plate, transferred through the semiconductor junctions, and leaves the device via the other plate; the device is a heat pump.

Heat is absorbed at the reverse biased junctions and emitted at the forward biased ones. A semiconductor material with a low bandgap is required – usually bismuth telluride – so that enough minority carriers can be generated at the operating temperature.

As the device is reversible it can also be used for heating, allowing an item to be maintained at constant temperature for ambient variations above or below the set point. For the sake of a consistent nomenclature, I will refer to the thermoelectric cooler surface connected to the controlled item as the cold plate and the surface connected to the heat sink as the hot plate, regardless of whether the thermoelectric cooler is cooling or heating the item.

The two main thermal parameters of interest for any thermoelectric cooler are the maximum temperature that can be maintained across the device and the amount of heat that can be pumped. The maximum temperature differential is quoted for no external heat load on the cold plate i.e. nothing being cooled.

Under these conditions all the electrical power input to the thermoelectric cooler is being used to overcome its internal reverse heat leak and heat input by radiation and convection through the area of the cold plate. This is the least efficient situation since virtually by definition the thermoelectric cooler is doing no useful work.

The maximum temperature differential figure ΔT_{max} is usually quoted for a given hot plate temperature in both a vacuum and dry nitrogen; running in a vacuum eliminates heat input by convection and allows slightly colder temperatures to be achieved. For the single-stage thermoelectric cooler as illustrated in **Fig. 2** the numbers would be typically 60 and 70°C for use in nitrogen and *in vacuo* respectively.

For applications needing a larger temperature drop, there are multistage devices available looking like small pyramids; the hot plate for the top layer is the cold plate for the layer underneath. A law of diminishing returns sets in; a three-stage device might have a maximum differential of 100° C but a six stage device may only manage 130° C.

How much heat?

The second parameter is the heat that can be input to the cold plate to reduce the temperature differential to zero, known as Q_{max}. Under these conditions all the electrical power is being used simply to prevent the cold plate getting warmer than the hot plate.

Since the cold plate temperature is the same as the surroundings there is no heat input due to convection or radiation and the thermoelectric cooler will be working efficiently. The maximum heat input scales with the area of the cold plate and is typically 2-3W/cm^2. There are tiny devices consisting of a single p-n junction which will pull only a few hundred milliwatts, up to larger 6cm^2 units rated at 100W.

Specifying a thermoelectric cooler

In order to select a suitable thermoelectric cooler it is necessary to determine how much heat is to be transferred over the required temperature difference. Two things really need to be known from the start. One is the cold plate temperature, which is decided by the application. The second is the method for getting the heat away from the hot plate, since this will define the hot plate temperature and thus the temperature drop across the thermoelectric cooler.

The second point usually boils down to the choice between a heat sink and a water loop. The water loop is more effective since the hot plate needs to be only a fraction of a degree warmer than the water even for quite modest flow rates to enable the heat to be disposed of.

A heat sink has to run warmer than the surrounding air to function, so the thermoelectric cooler has to work harder to achieve the same cold plate temperature. However the heat sink is more practicable in portable applications.

An example

As an example, consider a charge-coupled device image sensor to be cooled to –20°C (253K) from room temperature with a heat sink, **Fig. 3**. For room temperature, assume 20°C.

The heat sink is to run 10°C warmer than the surrounding air, so the temperature drop across the thermoelectric cooler will be 50°C. The ccd has case dimensions 20 by 15 by 5mm and exactly fits on the cold plate. It has 20 pins, all of which are connected to circuitry at room temperature by copper wires 0.25mm diameter and 100mm long.

Both ccd and thermoelectric cooler assembly are contained in dry nitrogen. Further, the ccd runs from a 10V supply with a current consumption of 5mA.

You are now in a position to estimate the total heat input to the cold plate. There are several terms involved. First is the active heat generated by the ccd due to its power consumption. In this case P_{active} is simply the voltage×current product, or 10V by 5mA=50mW.

As an aside, the temperature difference between the actual silicon slice and the cold plate is assumed to be zero. This means that that the thermal resistance, junction to case, is assumed very low. Usually, this is true, but there may be applications where the device dissipation changes in use, causing uncontrollable temperature changes in the silicon.

Remember that the sensor usually has to go outside the chip package! There are some photodiodes and ccds available which include a thermistor sensor bonded internally to the silicon chip to get around this problem.

There are now the passive components to be considered. These are the radiation, convection and conduction terms. You will need to know the exposed surface area of the ccd.

$$\begin{aligned} area &= 20\times15 = 300\text{mm}^2 \text{ (front face)} \\ &+ 20\times5\times2 = 200\text{mm}^2 \text{ (sides)} \\ &+ 15\times5\times2 = 150\text{mm}^2 \text{ (ends)} \end{aligned}$$

Total exposed area, A is 650mm^2.

Radiation. The exposed surface area is treated as a black body which will absorb radiation emanating from the surroundings at room temperature. This term is calculated using the Stefan/Boltzmann radiation law;

$$P_{rad} = A\sigma\left(T_{hot}^4 - T_{cold}^4\right)$$

where A is the area already calculated, T_{hot} is the background temperature assumed as 293K and T_{cold} is the ccd temperature at 253K. Symbol ρ is Stefan's constant, which is,

5.67×10–8W/m^2/K^4.

Absorbed radiation is found to be 0.12W. This is a worst case result since the surface will not behave perfectly as a black body, but it is obviously better to overestimate the heat absorbed.

Convection. Unless the unit is operated in a vacuum there will be circulating air currents inside the assembly. The convection component is given by,

$$P_{conv} = Ah(T_{hot} - T_{cold})$$

where A is the exposed area, h is the convection coefficient is 21.7W/m^2/°C. Temperatures T_{hot} and T_{cold} can be either °C or K. The convection term is found to be 560mW.

Convection is often the largest contribution. It can be removed by operation in a vacuum; this also solves the problem of frosting up as the assembly is cooled past the dew point.

Conduction. The connecting wires to the ccd and temperature sensor form a heat leak which must be accounted for. Any retaining structure used to hold the ccd in place will also need to be considered.

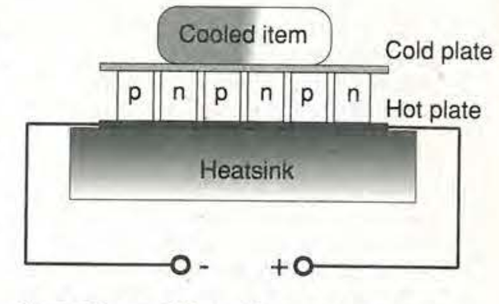


Fig. 2. When used for cooling, a Peltier device produces heat that usually needs to be removed via a heat sink.

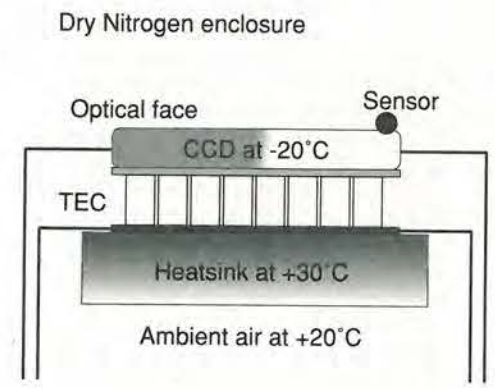


Fig. 3. Using a Peltier-effect device to cool a ccd.

Allowing three sensor wires, there will be 23 copper wires going to the ccd assembly. The equation describing heat flow for conduction is,

$$P_{cond} = \frac{ka}{l}\left(T_{hot} - T_{cold}\right)$$

where k is the thermal conductivity of copper is 386W/m/°C, a is the area and l is the wire length, at 10cm. In this case, area a is 23 times the cross sectional area of a 0.25mm diameter wire.

Putting in the numbers shows that the conduction heat component will be 175mW. Note that using wire of twice the diameter will increase this heat leak by a factor of four.

Thus the total heat entering the cold plate will be $P_{active}+P_{rad}+P_{conv}+P_{cond}$ giving a grand total of 0.905W. This number should be treated as an approximation as all the terms have considerable errors. The active load could be in error by a factor of two either way since IC power consumptions are not that well defined as a rule. For the radiation term the emissivity has been assumed to be unity; the convection depends to some extent on the shape, angle and state of the surfaces.

Selecting a cooler from the range

Knowing the heat input, you are now in a position to consult the manufacturers' data sheets and choose a suitable thermoelectric cooler.

The information is normally presented graphically by a set of curves. These relate the current flowing through the thermoelectric cooler and the heat input with the temperature differential produced.

Figure 4 shows a typical set of curves. It so happens that the shape is very much the same for all single-stage thermoelectric coolers so the one set can be used for many devices simply by normalising the axes to suit.

The *y* axis is the temperature drop produced typically normalised to 60°C, so the 50°C needed in our application will be represented as a horizontal line at 0.83.

The *x* axis is the thermoelectric cooler current normalised to I_{max}. The relevance of I_{max} is that above this current the cooling effect actually falls off. Peltier cooling increases linearly with current but unwanted ohmic heating increases as the square of the current; at I_{max}, the ohmic heating becomes greater than the Peltier cooling.

In Fig. 4, the curves of the temperature/current characteristic for heat loads are normalised to Q_{max}. The outermost curve is with no heat load so at I_{max} the temperature drop will be ΔT_{max}. This is the top right point on the graph.

The curves are then shown in steps of 0.2 for Q/Q_{max}. A finer spacing would have been better but Lotus 1-2-3 only plots six curves at a time. The situation with the heat input $Q=Q_{max}$ is represented at the bottom right corner where at I_{max} the cooling effect has been reduced to zero by the heat load.

For copyright reasons, these curves are not prepared from any one manufacturer's data. Rather they are plotted from a simple formula developed by looking at the performance of several single stage thermoelectric coolers from various manufacturers;

$$\frac{\Delta T}{\Delta T_{max}} = 2\frac{I}{I_{max}} - \left(\frac{I}{I_{max}}\right)^2 - \frac{Q}{Q_{max}}$$

As far as I know, this equation is not a recognised one, but it does have some basis in reality. The first two terms on the right hand side represent Peltier cooling and ohmic heating respectively. When differentiated with respect to current, these reduce to zero at I_{max}.

The heat loading term goes inversely as the temperature drop as would be expected from the conduction and convection terms. Errors due to other terms are ignored.

The optimum thermoelectric cooler performance is obtained on the line marked 'optimum'; this line has a 1:1 slope for normalised temperature *versus* current.

For the example used here, T/T_{max} is 0.83 so I/I_{max} becomes 0.83. The ratio Q/Q_{max} is found by interpolating between the curves and is seen to be 0.14. The required value of Q_{max} is now available as 0.905W heat input divided by 0.14, which is 6.4W.

Now you have enough information to refer to the manufacturer's literature and select a thermoelectric cooler having a ΔT_{max} of 60°C or more and a Q_{max} of 6.4W, or slightly greater.

For example, the Marlow Industries *MI 1061* has the following characteristics,

ΔT_{max}	64°C in dry nitrogen
Q_{max}	6.4W
I_{max}	5.3A at approximately 1.9V
Cold plate	13×15mm²

When these numbers are normalised and plotted on the curves, the following are derived,

$\Delta T/\Delta T_{max}$	0.78
Q/Q_{max}	0.14
I/I_{max}	0.72

This is a point just above the optimum line giving the 50°C drop for a thermoelectric cooler current of 3.8A. As a matter of interest the Marlow data sheet predicts a current of 3.7A to the accuracy to which the charts can be read.

I should point out that this thermoelectric cooler has a cold plate smaller than the ccd quoted; this will mean there is some exposed area at the back of the ccd which will behave as an extra heat leak. It may be necessary to go back and recalculate the heat input after the thermoelectric cooler has been chosen.

If there is any doubt as to whether a thermoelectric cooler will deliver the expected performance then it is essential to consult the manufacturer's data. This is especially true of the multistage units.

If you are working to a tight budget that allows you to buy only one thermoelectric cooler, it is worth getting one that can pump more heat into than is really needed in case your heat input estimate on the cold plate is too optimistic.

The *MI 1023* has Q_{max} of 9.2W, and would allow for a margin of safety.

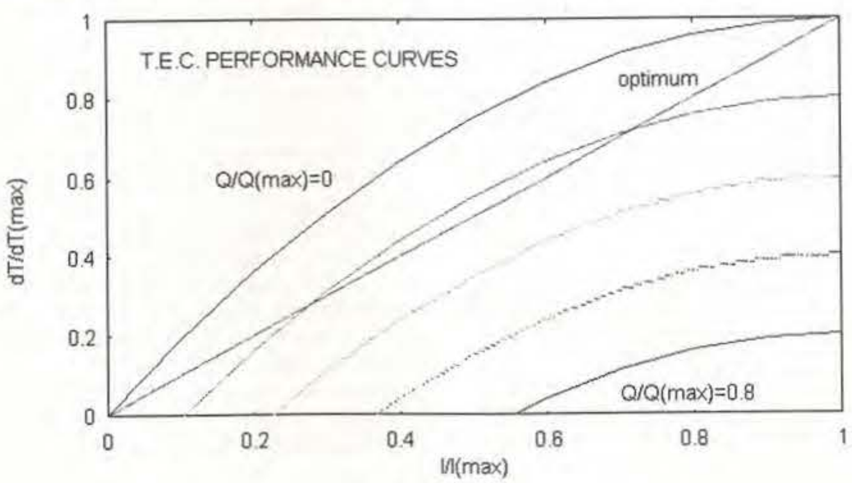


Fig. 4. Typical normalised Peltier device performance curves. Note that at I_{max}, ohmic heating becomes greater than Peltier cooling.

Choosing a heat sink

Possibly the most common reason for disappointing performance with a thermoelectric cooler is inadequate provision for removing heat from the hot plate.

Manufacturers usually quote the thermoelectric cooler parameters with a hot plate temperature of 25 to 30° C and anomalies can be expected if this temperature is wildly different.

A more insidious effect is the possibility of thermal runaway if the heat sink is too small. The thermoelectric cooler becomes less efficient as the temperature differential increases; this leads to a warming of the heat sink making matters worse.

There comes a point where the hot plate warms up at a faster rate than the cold plate is cooled; increasing the thermoelectric cooler current now causes a warming of the cold plate. This results in a phase reversal in the control loop and the system locks up at full current. In this state the only option is to switch off and wait for things to cool down.

This effect is separate from the effect of running the thermoelectric cooler above I_{max} and can occur long before this current if the heat sink specification is seriously deficient.

Heat emerging from the hot plate is the sum of the dc power in and the heat entering the cold plate. To a first approximation, the thermoelectric cooler behaves as a resistor given by the volts at I_{max} which for the *MI 1061* will be 0.36Ω.

This value is not perfectly constant for all conditions; the resistance appears somewhat higher for low values of Q/Q_{max}. But the difference can be ignored if an oversized heatsink is fitted. Thus the expected heat output in the example is,

$$0.905\text{W}+3.8\text{A}^2\times0.36\Omega=6.1\text{W}$$

Knowing this number, the remainder of the problem is solved using normal heat sink design procedures. From the original problem details, the air temperature is 20°C and the heat sink 30°C so the maximum thermal resistance that can be accepted is 30–20°C, i.e. 10°C, divided by 6.1W. A heat sink capable of dissipating 1.5°C/W or more would be a good choice.

What about water?

As this is going to be quite a large heat sink, it may be worth considering a water loop. If the loop is using tap water at 15°C and you allow 2°C due to the thermal resistance from the hot plate to the water, the thermoelectric cooler now has to produce a ΔT of 37°C.

Going back to the curves, you will find that the thermoelectric cooler current falls to 3A and the total heat to be disposed of is 4.2W. Remembering the heat capacity of water is 4.2 joules/cc/°C this is a flow rate of only 1cc per second if the water is assumed to heat up by 1°C in crossing the hot plate.

Electronic considerations

Thermoelectric coolers of the same outside dimensions and thermal properties can be made from many small p-n junctions in series, or just a few large ones. Devices in the latter category are cheaper to make and, in my experience, slightly more robust.

However, the resulting resistance can be very low – less than an ohm as in this example. This can make designing the driving electronics less straightforward – especially if a power-efficient design is required. Simple linear circuits tend to dissipate a lot of power. Switch-mode designs naturally give much better results. Remember that with very low resistance units the effects of long cable runs can seriously increase the overall power consumption.

The *MI 1013* happens to be especially convenient, with its specification of 8.5V at an I_{max} of 1A, but is quite expensive at around £60 in one-off quantities. It is also a bit small for this example. It is a single stage device with a Q_{max} of 4.8W and ΔT_{max} of 61° C.

If the thermoelectric cooler is used in a situation, where it can be required to heat or cool, then bear in mind that the devices are actually much more effective at heating than cooling.

This has implications for the gain settings in the control electronics. The servo gain can be expected to be much higher with the device heating – so beware of servo oscillations.

Mounting a Peltier device

Thermoelectric coolers are brittle, fragile and expensive. The heat sink should be milled flat and the thermoelectric cooler held down with the minimum of compression.

Even though they are reversible, devices are always mounted with the wires on the hot plate to prevent a significant heat leak. They should never be subjected to tension or shear forces, or exposed to temperatures much above 100°C since the p-n junctions are soldered together with special low melting point solder.

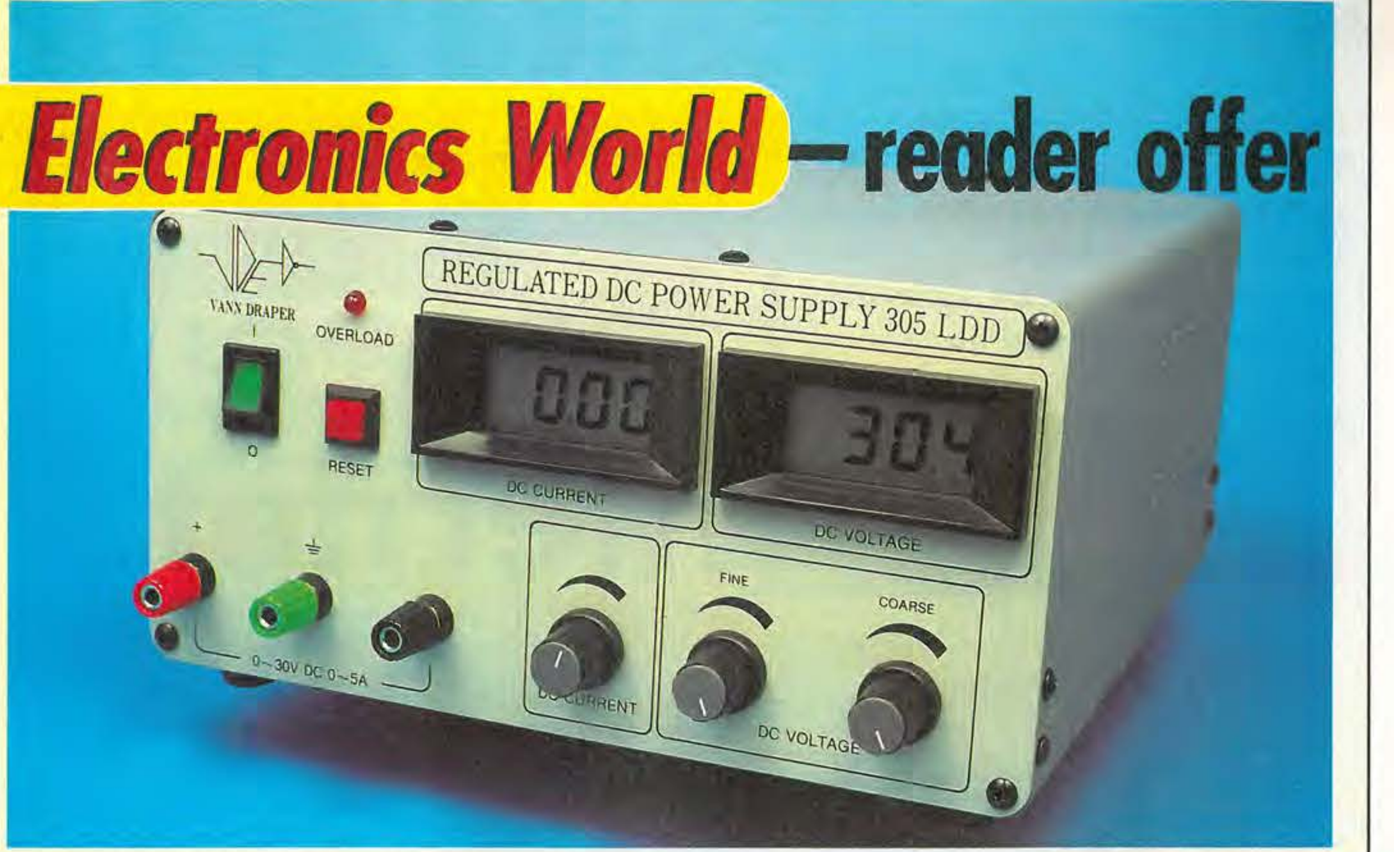
If heat is removed using a water loop then some consideration should be given to protecting the thermoelectric cooler in the event of a leak or pump failure. Farnell and RS stock some useful bimetallic cutouts which are suitable if size is not a problem.

Some devices have both plates metallised with copper enabling the unit to be soldered down. This enables a very good thermal contact to be made and has the extra advantage that no supporting structure is required thus minimising heat leaks to the cold plate. Needless to say use of the manufacturer's special solder is mandatory. Heat conductive glue (RS 850-984) is an option.

Understandably, you may be reluctant to solder together an expensive thermoelectric cooler and ccd until the system is proven. It is usually possible to devise some means of compressing the ccd – or whatever device – to thermoelectric cooler and its heat sink. Inevitably though, the device used to compress the components adds to the heat input to the cold face.

If the system can be made to work in this state then results are always slightly better when the supports are removed and the components glued or soldered together. ■

Electronics World – reader offer



30V, 5A bench power supply with dual displays – 25% discount

For a limited period, Vann Draper is offering over 25% discount on the 305 LDD – a bench power supply featuring digital display of both voltage and current. Normally, the 305 retails at £186.82 including VAT but it is available to *Electronics World* readers filling in the adjacent coupon at the 25% discount price of £139 – fully inclusive of VAT *and* delivery.

Infinitely variable between 0 and 30V – with coarse and fine controls – and adjustable between 0 and 5A, the 305 LDD has a ripple figure of typically 10mV. Its load regulation is also excellent, at typically ±0.2%.

Accuracy of the supply's dual 3.5-digit liquid crystal displays is 0.1 decimal digit. The output can handle a continuous short-circuit, overloading at 5.5A ±0.5A. When the overload circuit is activated, it causes both audible and visual alarms, resettable via a push-button on the front panel.

Dimensions of the 305 LDD are 310 by 260 by 120mm and its weight is 5.5kg. Housed in a light-grey steel enclosure, the unit is built to comply with UL, CSA and TUV safety standards.

Please address orders and all correspondence relating to this offer to Vann Draper Electronics at Unit 5, Premier Works, Canal Street, South Wigston, Leicester LE I 8 2PL, telephone on 0116 2771400, or fax 0116 2773945.

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New processors are on the horizon. Cyril Bateman tells you where to search if you want to know whether they will be worth the wait, and he relays news on hardware bugs, Microsoft problems and circuit CAD software – in particular filter design tools.

Hands-on Internet

Intel has released a fix for both *Classic* and *MMX* processors to overcome the 'FOOF' bug, reported last month on CNET. This bug allows malicious code to stall or freeze up the computer, which cannot then respond to a CTRL-ALT-DELETE, or any other command. The only remedy is to switch off then reboot, which can result in lost disk clusters. Both *Pentium* and *MMX Overdrive* processors are also affected, but not the *i486, Pentium Pro* or *Pentium II*.

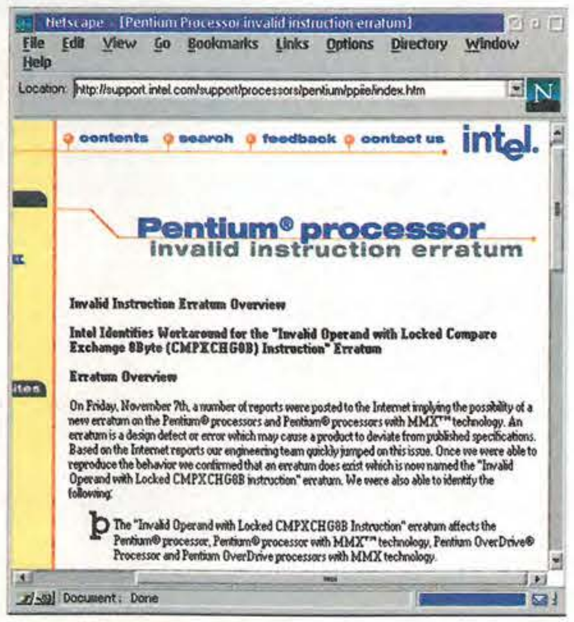

Fig. 1. The Intel bug now has a work-around. Read this overview and the technical description pages for details.

Since code that prompts this bug does not exist in commercial software, Intel's report claims that users of commercial software cannot be affected.[1] The company has now renamed this bug 'Invalid operation with Locked CMP X CHG 8B instruction'. Full details of the problem, together with a software work-around to restore normal error handling of this instruction, are on its overview and technical description web pages, **Fig. 1**.

According to News.Com, the German magazine C't reports a further freeze-up bug affecting certain Cyrix 6x86 processors.[2] This problem, which has been acknowledged by Cyrix, results in a freeze up if a series of legal instructions are issued in an illegal sequence. Since compilers should not allow such illegal sequences, this is considered to be a minor problem, and should not affect end users.

New processors: but are they better?

As Intel's plans for *Pentium II* processor speed increases are unveiled, criticism of minimal performance gains from the present 'slot 1' versions emerge. An article on News.Com entitled "Pentium II: Dynamo or dud", presents critiques for both present and planned processor releases.[3]

By autumn 1998, processors on the current slot-1 motherboards using the 'LX' chipset are expected to run at 450MHz.[4] A version of the chipset without internal cache is planned for lower-cost systems.

The new slot-2 motherboard system will be aimed at workstation applications. Based on the BX chipset and running a 100MHz system bus, its processors, with perhaps 1Mbyte of cache, will run initially at 400MHz.

Fig. 2. More and even faster Intel processors will soon arrive. Read these forecasts before committing yourself to a new system.

Both 'LX' and 'BX' chipsets support the new advanced graphics port, or AGP, providing vastly improved graphics capability, **Fig. 2**.

The combined effects of increased performance and demand for sub $1000 desktop machines has resulted in Toshiba's decision to phase out its Infinia range and quit the consumer desktop machine market. The company will now concentrate on its Equium business desktop range. The Infinia inventory will be sold off, so watch out for bargains.

Microsoft under pressure

External pressures on Microsoft have increased. Reported by News.Com[5], Ralph Nader has targeted Microsoft in an attempt to curb what he says are Microsoft's abusive business practices. Meanwhile IBM, Lotus, Oracle and Sun have united to produce an alternative to the 'Windows' interface, **Fig. 3**.

Fig. 3. National consumer groups join in the Microsoft bashing affray. Ralph Nader, the famed consumer campaigner from the seventies, takes on Microsoft.

In an eight-page report on News.Com,[6] Nader is quoted as accusing Microsoft of being "uniquely ruthless". Nader rose to fame by attacking the poor safety record of American cars in the seventies. His involvement may well trigger an escalation of evidence needed to support claims against Microsoft's supposed undue marketing pressures and Internet Explorer licensing.

The IBM, Lotus, Oracle and Sun consortium[7] plans to produce a standard network computing desktop interface to be called 'Webtop'. It promises to let developers create applications to run on any Sun Java-based network computing device, including network computers, networked pcs and palmtops.

The specification will be licensed to developers and Oracle has stated that it will give it away for free. No delivery date is announced and details are scant as yet.

Circuit discoveries

I mentioned last month that a wealth of component data and circuit application notes can be found on Internet. While many of these articles can be downloaded as PDF files, Sirius[8] now supplies these on cd-rom, which many

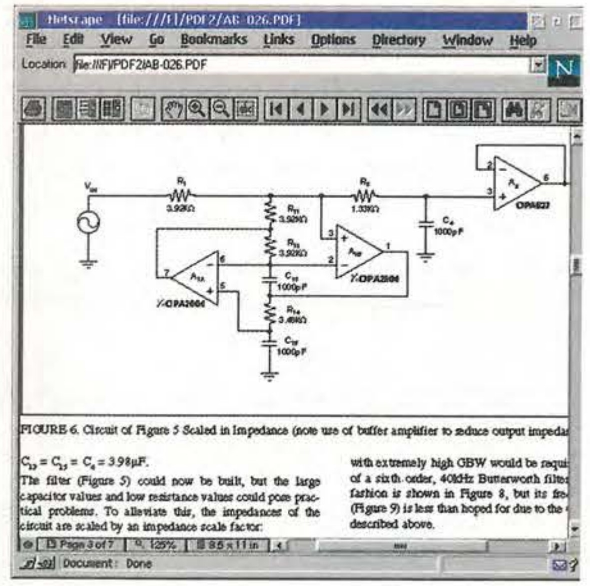

Fig. 4. Basic third-order building block for a superb linear phase, anti-alias audio filter. Two blocks cascaded provide sharp cut-off, yet are flat within 0.2dB to 20kHz.

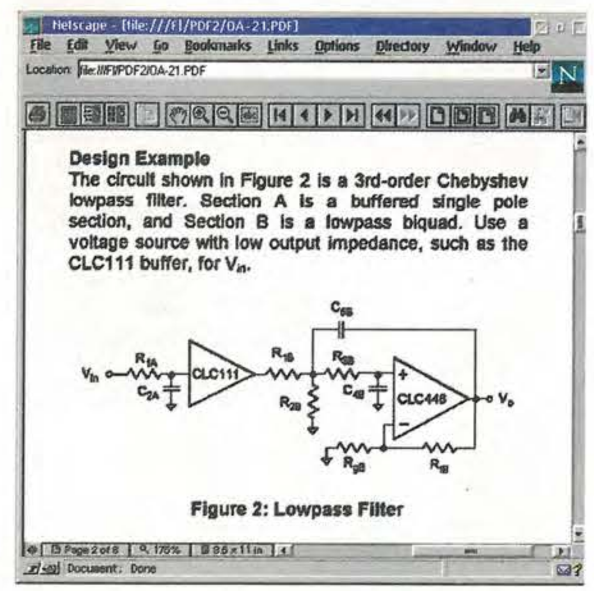

Fig. 5. Application design support for Sallen-and-Key filter design. Optimise your filters against component sensitivity and op-amp band-width.

Where to surf

1	Intel Corporation	http://support.intel.com/support/processors/pentium
2	Cyrix chip hit by bug, too	http://www.news.com/News/Item/0,4,16347,00.html
3	Pentium II: Dynamo or dud?	http://www.news.com/News/Item/0,25,16543,00.html
4	Intel to Rev Pentium II Speeds	http://www.techweb.com/wire/news/1997/11/1114pentium.html
5	Microsoft under the gun	http://www.news.com/News/Item/0,4,15416,00.html
6	Nader calls MS "uniquely ruthless"	http://www.news.com/News/Item/0,4,16378,00.html
7	Bashing Windows	http://www.wcmh.com/lantimes/97/97nov/711b001b.html
8	Sirius CD-ROM Collection	http://www.tds-net.com/main/us/sirius.htm
9	Icesoft I.C. Data library	http://www.semi.com.tw
10	National Semiconductors	http://www.national.com
11	Burr-Brown Corporation	http://www.burr-brown,com
12	GangGang Australian Shareware	http://www.ganggang.com.au
13	QikDrawCad	http://www.qikdraw.sa.com.au
14	MetaCrawler Search Engine	http://www.metacrawler.com

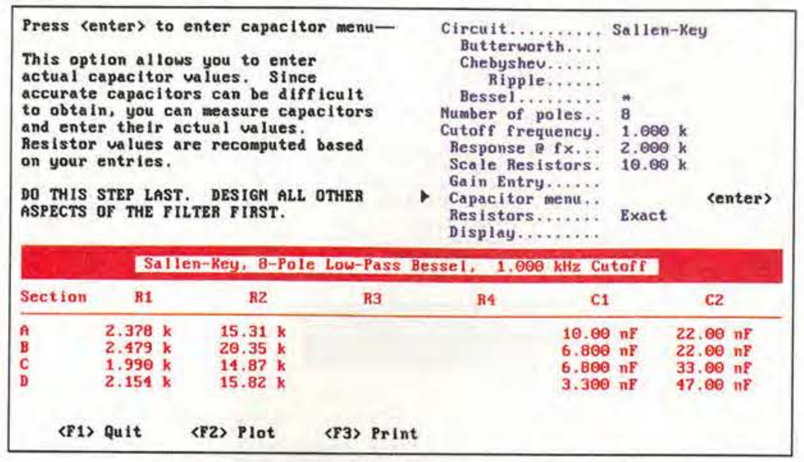

Fig. 6. Sallen-and-Key or multiple feedback filter design software facilitates 'what-if?' explorations. Simply respond to the prompts in the upper screen, immediately find component values below.

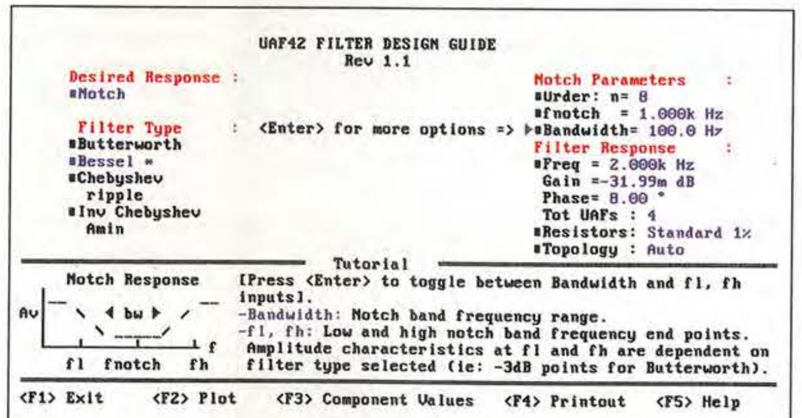

Fig. 7. Dedicated for use with the 'UAF42' package; simply respond to the prompts. Push <F2> for an instant performance plot, <F3> to see required parts list.

Fig. 8. This well organised shareware repository specialises in Australian software. It provides links to many 2D specialist drawing packages.

users prefer. Sirius is part of TDS-NET, which I reported as a vast source of on-line datasheets and application notes in August 1997.

This cd-rom collection now hosts data on some 200000 integrated circuits and discrete devices. It also includes more than 235000 pages of information from 38 manufacturers world-wide. Four cds each containing 600Mbyte of technical information make up the set. This collection is updated each two months.

Another extremely large data source, Icesoft can be found at the semi.com.tw Web site, based in Taiwan.[9] This site provides links directly to the original maker's site so that you can download data or application notes. It requires you to pre-register on-line. You are then supplied with your password by e-mail. Mine arrived almost instantly by the way.

Since this site permits data searching by description, part number, classification or function, you need not know the manufacturers name, or the device's part number. Using this resource, I was able to identify and download six application notes needed for this column in just over a quarter of an hour.

Designing active filters

Designing active filters can require repetitive iterations of relatively awkward simultaneous equations. Traditional Spice-based simulators are of little help for switching filter designs and can be slow when designing continuous mode filters. Fortunately, application notes and dedicated software for designing such filters can be quickly downloaded.

One good overview of filter design, called "A basic introduction to filters – active, passive and switched-capacitor", can be found in application note 779 from National Semiconductor.[10] This 22 pages of advice can be downloaded as AN-779.PDF.

While switched capacitor filters can be extremely useful, a major draw back may be the level of noise breakthrough at the switching frequency. This can be reduced by adding passive or active time continuous filters, or it can be avoided completely by performing filtering using only time-continuous techniques.

Unfortunately, designing active time continuous filters requires analysis then choice of the appropriate circuit configuration and filter function. One particularly elegant technique, especially for use as an anti-aliasing filter, uses immittance conversion. This technique results in devices called generalised immittance converters or GICs.

The GIC method offers superior noise gain characteristics, and can be designed to ensure minimal amplitude and phase deviations over the desired frequency band. This method is frequently used as an anti-alias filter in high quality compact disc audio players.

For a good description with a design method for this technique, download application note AB-026A from Burr Brown.[11] This details a worked example of a third order linear phase filter, easily cascaded, to provide an excellent sixth-order linear-phase response, **Fig. 4.**

Sallen and Key filters

Perhaps your needs are more modest and can be satisfied using the traditional Sallen and Key configuration. Two National Semiconductor application notes detail design techniques to overcome two sources of performance degradation found using Sallen-Key designs.[10] These sources are sensitivity to component parasitic elements and the finite gain bandwidth of the op-amps used.

Application note OA-27, "Low-Sensitivity Lowpass Filter Design", outlines techniques for minimising the filter's sensitivity to component tolerances – especially with time and temperature changes. The companion application note OA-21, "Component Pre-Distortion for Sallen-Key Filters", deals with component changes needed to compensate for the op amps finite bandwidth. Both voltage and current feedback amplifier systems are supported, **Fig. 5.**

Simulation and design software

I mentioned earlier that continuous time filter design could be time consuming. Can this process be simplified using software?

One easy solution for both Sallen and Key and the quite similar multiple-feedback filters may be found using a piece of software called *FilterPro*. This can be downloaded from Burr-Brown.[11] Supported by application note AB-034B, this dos-based program allows very quick and easy "what-if?" variations to be evaluated, and simulated results can be plotted on screen or via a printer, **Fig. 6.**

Burr-Brown produces an integrated circuit, comprising four op amps together with the precision on-chip capacitors and matched resistors needed to build state-variable filters. This device is called the *UAF42*.[11] The company's *Filter42* software, supported by application note AB-025C, automates the design task.

The *UAF42* provides three filter sub-circuits together with a fourth uncommitted precision op amp. Using this dos software, combinations of these sub-circuits can be used to quickly design low-pass, high-pass, band-pass and notch filters. Using "what-if?" variations, filter configurations, can be quickly simulated and the results viewed on screen, **Fig. 7.**

Both filter software packages are contained within the downloaded file FILTER.EXE and accommodate Butterworth, Bessel and Chebyshev filter responses.

Towards a better searching

Many sites collate data for circuit simulation packages. But computer-aided drafting software, now almost mandatory for all designers, is less well serviced.

One Australian site offers a large number of shareware packages originated by Australian writers, so is doubly different, making the Ganggang[12] site one to visit. Their download link to QikDrawCad[13] is worthy of evaluation, **Fig. 8.**

One search engine not previously visited is MetaCrawler.[14] This tool is highly commended in a CNET search engine evaluation. Perhaps I should explain that metacrawler simultaneously passes your request to several major search engines, collates these results, discards redundant or broken links, before presenting your results. Advanced searching allows you to control the number and presentation of search results it presents. ■

Adaptive filters explained

Continued from page 222

produced at the output of the decision device in the receiver. Accordingly, if this output was the correct transmitted sequence, it may be used as the desired response for the purpose of adaptive equalisation. Such a method of learning is said to be 'decision directed' because the receiver attempts to learn by employing its own decisions.

A final comment pertaining to performance evaluation: A popular experimental technique for assessing the performance of a data transmission system involves the use of an eye pattern. This pattern is obtained by applying the received wave to the vertical deflection plates of an oscilloscope, and a saw-tooth wave at the transmitted symbol rate to the horizontal deflection plates.

The resulting display is called an eye pattern because of its resemblance to the human eye for binary data. Thus, in a system using adaptive equalisation, the equaliser attempts to correct for intersymbol interference in the system and thereby open the eye pattern as far as possible.

Adaptive differential pcm

In pulse-code modulation, or pcm, which is the standard technique for waveform coding, three basic operations are performed on the speech signal. These are sampling, quantisation and coding.

The operations of sampling and quantisation are designed to preserve the shape of the speech signal. As for coding, it is merely a method of translating the discrete sequence of sample values into a more appropriate form of signal representation.

The rationale for sampling follows from a basic property of all speech signals – they are band limited. This means that a speech signal can be sampled in time at a finite rate in accordance with the sampling theorem. For example, commercial telephone networks designed to transmit speech signals occupy a bandwidth from 200 to 3200Hz.

To satisfy the sampling theorem, a conservative sampling rate of 8kHz is commonly used in practice.

In pcm, as used in telephony, the speech signal is sampled at the rate of 8kHz, nonlinearly quantised, and the coded into eight-bit words, as in **Fig. 8a)**. The result is a good signal-to-quantisation noise ratio over a wide dynamic range of input signal levels. This method requires a bit rate of 64kbit/s.

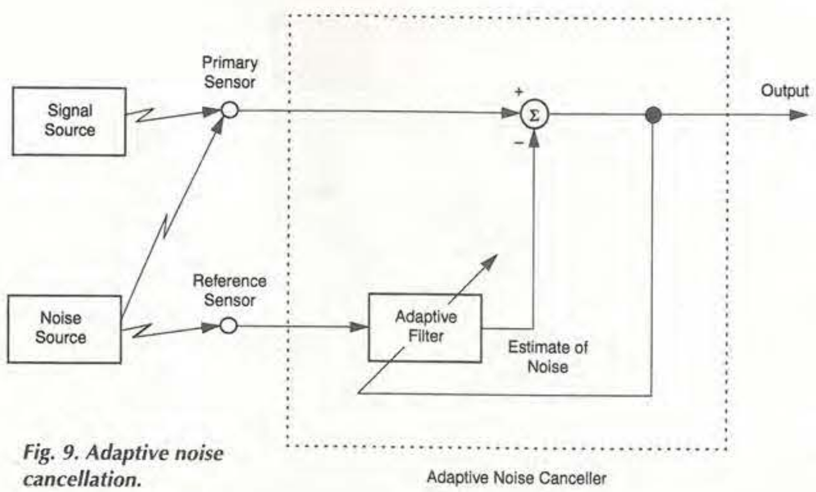

Fig. 9. Adaptive noise cancellation.

Differential pulse-code modulation. Abbreviated to dpcm, this is another example of waveform coding. It involves the use of a predictor as shown in Fig. **8b)**.

The predictor is designed to exploit the correlation that exists between adjacent samples of the speech signal in order to realise a reduction in the number of bits required for the transmission of each sample of the speech signal. It does this while maintaining a prescribed quality of performance. This is achieved by quantising and then coding the prediction error that results from the subtraction of the predictor output from the input.

If the prediction is optimised, the variance of the prediction error will be significantly smaller than that of the input signal, so a quantiser with a given number of levels can be adjusted to produce a quantising error with a smaller variance than would be possible if the input signal were quantised directly as in a standard pcm system.

Likewise, for a quantising error of prescribed variance, dpcm requires a smaller number of quantising levels than pcm. Differential pcm uses a fixed quantiser and a fixed predictor. A further reduction in the transmission rate can be achieved by using an adaptive quantiser together with an adaptive predictor of sufficiently high order.

Adaptive differential pcm, or adpcm, can digitise speech with toll quality (eight-bit pcm) at 32kbit/s, Fig. **8c)**.

Adaptive noise cancelling

As the name implies, adaptive noise cancelling relies on the use of noise cancelling by subtracting noise from a received signal, an operation controlled in an adaptive manner for the purpose of improved signal-to-noise ratio.

Ordinarily, it is inadvisable to subtract noise from a received signal because such an operation could produce disastrous results by causing an increase in the average power of the output noise. However, when proper provisions are made, and filtering and subtraction are controlled by an adaptive process, it is possible to achieve a superior system performance compared to direct filtering of the received signal.

Basically, an adaptive noise canceller is a dual-input, closed-loop adaptive control system as illustrated in **Figs 9** and 1. The two inputs of the system are derived from a pair of sensors – a primary sensor and a reference sensor.

The primary sensor receives an information-bearing signal $s(n)$ corrupted by additive noise $v_0(n)$. The signal and the noise are not correlated with each other. The reference sensor receives a noise $v_1(n)$ that is not correlated with the signal $s(n)$ but correlated with the noise $v_0(n)$ in the primary sensor output in an unknown way:

$$E[s(n),v_1(n-k)]=0, \text{ for all } k \text{ and}$$
$$E[v_0(n)v_1(n-k)]=p(k)$$

where, as before, the signals are real valued and $p(k)$ is an unknown cross-correlation for lag k.

The reference signal $v_1(n)$ is processed by an adaptive filter to produce the output signal $y(n)$. Filter output is subtracted from the primary signal $d(n)$, serving as the desired response for the adaptive filter. The error signal is defined by:

$$e(n)=d(n)-y(n)$$

The error signal is used, in turn, to adjust the tap weights of the adaptive filter, and the control loop around the operations of filtering and subtraction is thereby closed.

Note that the information bearing signal $s(n)$ is indeed part of the error signal $e(n)$. Now, the adaptive filter attempts to minimise the mean-square value (average power) of the error signal $e(n)$. The information bearing signal $s(n)$ is essentially unaffected by the adaptive noise canceller.

Hence minimising the mean-square value of the error signal $e(n)$ is equivalent to minimising the mean-square value of the output noise $v_0(n)-y(n)$. With the signal $s(n)$ remaining essentially constant, it follows that the minimisation of the mean-square value of the error signal is indeed the same as the maximisation of the output signal to noise ratio of the system.

The effective use of adaptive noise cancelling therefore requires that the reference sensor be placed in the noise field of the primary sensor with two specific objectives in mind.

One objective is that the information-bearing signal component of the primary sensor output is undetectable in the reference sensor output. The other is that the reference sensor output is highly correlated with the noise component of the primary sensor output. Moreover, the adaptation of the adjustable filter coefficients must be near optimum.

Noise-cancelling applications

Consider the two useful applications of the adaptive noise cancelling operation that follow.

Cancelling 60Hz interference in ecg. In electrocardiography, commonly used to monitor heart patients, an electrical discharge radiates energy through a human tissue. The resulting output is received by an electrode.

The electrode is usually positioned such that the received energy is maximised. Typically, however, the electrical discharge involves very low potentials. Hence extra must be exercised in minimising signal degradation due to external interference.

By far the strongest form of interference is that of a 60Hz periodic waveform picked up by the receiving electrode from nearby electrical equipment. **Figure 10** shows a block diagram of the adaptive noise canceller used to reduce the harmonics.

Reducing acoustic noise in speech. At a noisy site, such as the cockpit of a military aircraft, voice communication is effected by the presence of acoustic noise. This is particularly serious when linear predictive coding is used for the digital representation of voice signals at low-bit rates.

The noise corrupted speech is used as the primary signal. To provide the reference signal, a reference microphone is placed in a location where there is sufficient isolation from the source of speech.

Echo cancellation

Almost all conversations are conducted in the presence of echoes. An echo may not be distinct, depending on the time delay involved. If the delay between the speech and the echo is short, the echo is not noticeable but perceived as a form of spectral distortion or reverberation. If, on the other hand, the delay exceeds a few tens of milliseconds, the echo is distinctly noticeable.

To see how echoes occur, consider a long-distance telephone circuit depicted in **Fig. 11**. Every telephone is connected to a central office by a two-wire line called the 'customer loop.' The two-wire line serves the need for communications in either direction. However, for circuits longer than 35 miles, a separate path is necessary for each direction of transmission.

Accordingly, there must be provision for connecting the two-wire circuit to the four-wire circuit. This connection is accomplished by means of a hybrid transformer, commonly referred to as a hybrid.

Basically, a hybrid is a bridge circuit with three ports. If the bridge is not perfectly balanced, the 'in' port becomes coupled to the 'out' port, giving rise to an echo, **Fig. 12**.

Basically, the principle of echo cancellation is to synthesise a replica of the echo and subtract it from the returned signal, **Fig. 13**, for only one direction of transmission.

The adaptive canceller is placed in the four-wire path near

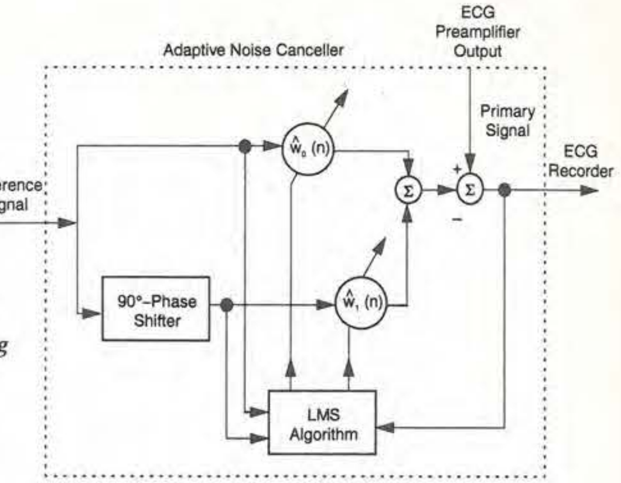

Fig. 10. Adaptive noise canceller for suppressing 60Hz interference in electrocardiography, after Widrow, et al, 1975.

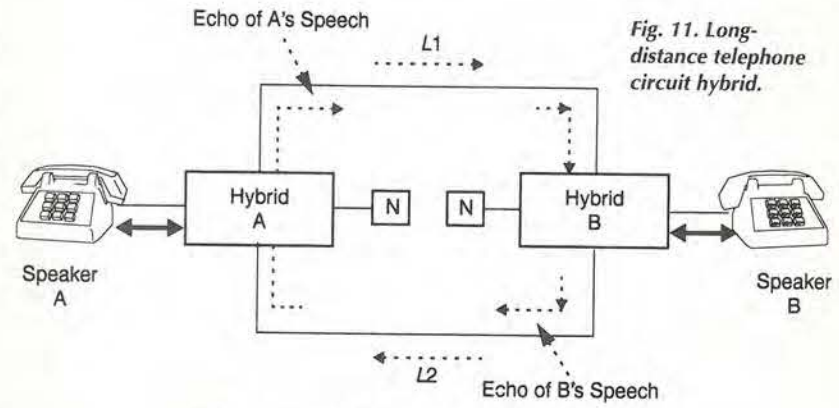

Fig. 11. Long-distance telephone circuit hybrid.

The boxes marked *N* are balancing impedances.

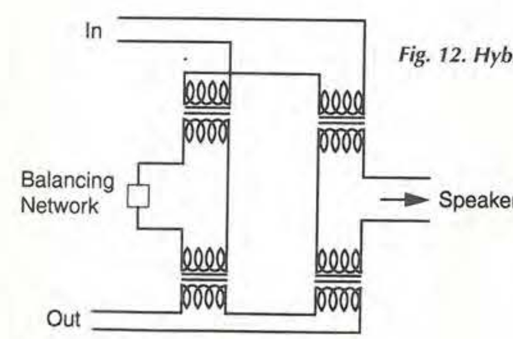

Fig. 12. Hybrid circuit.

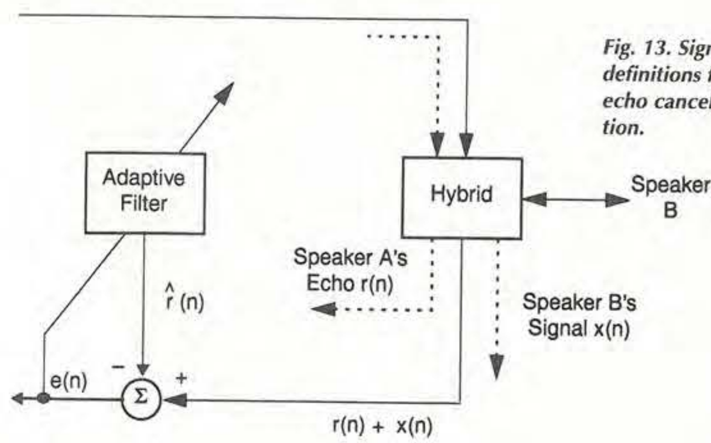

Fig. 13. Signal definitions for echo cancellation.

the origin of the echo. The synthetic echo is generated by passing the speech signal from speaker A through an adaptive filter that ideally matches the transfer function of the echo path.

Passing through the hybrid, the reference signal results in the echo signal. This echo, together with a near-end talker signal x, constitutes the desired response for the adaptive canceller. Synthetic echo is subtracted from the desired response to yield the canceller error signal.

In any event, the error signal is used to control the adjustments made in the coefficiencies of the adaptive filter. For the adaptive echo cancellation circuit to operate satisfactorily, the impulse response of the adaptive filter should have a length greater than the longest echo path that needs to be accommodated.

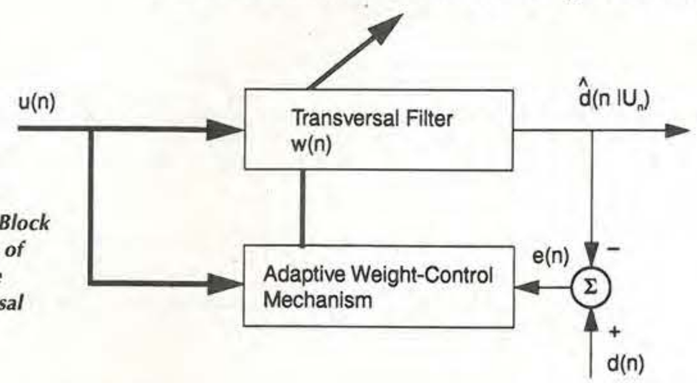

Fig. 14. Block diagram of adaptive transversal filter.

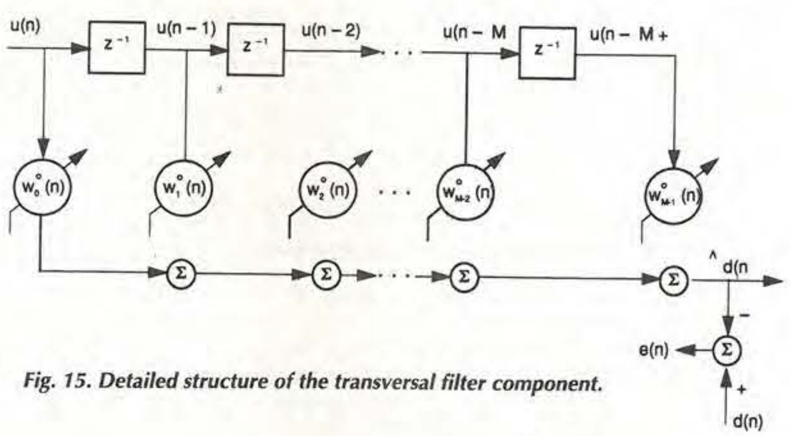

Fig. 15. Detailed structure of the transversal filter component.

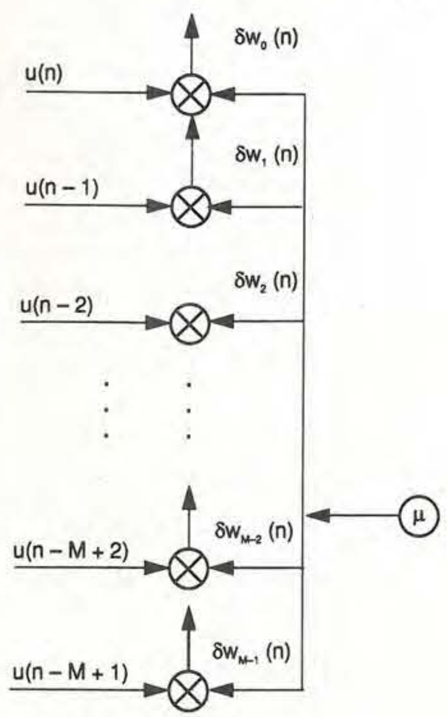

Fig. 16. Details of the adaptive weight-control mechanism.

More on adaptive filters

- Haykin, S. 'Adaptive Filter Theory,' Prentice-Hall International Inc, 1991, pp. 17-20, 31-40, 49-55, 303.
- Honig, M. and Messerschmitt, D. 'Adaptive Filters: Structures, Algorithms, and Applications,' Boston: Kluwer Academic Publishers, 1984.

Least-mean square algorithm

The well-known least-mean-square, or lms, algorithm is an important member of the family of stochastic gradient-based algorithms.

A significant feature of this algorithm is its simplicity. It does not require measurements of the pertinent correlation functions, nor does it require matrix inversion. Indeed, it is the simplicity of the lms algorithm that has made it the standard against which other adaptive filtering algorithms are benchmarked.

The operation of the lms algorithm is descriptive of a feedback control system. Basically, it consists of a combination of two basic processes. One is an adaptive process, which involves the automatic adjustment of a set of tap weights. The other is a filtering process, which involves: (a) forming the inner product of a set of tap inputs and the corresponding set of tap weights emerging from the adaptive process to produce an estimate of a desired response, and (b) generating an estimation error by comparing this estimate with the actual value of the desired response. In turn, the estimation error is used to actuate the adaptive process, thereby closing the feedback loop.

Correspondingly, it is possible to identify two basic components in the structural constitution of the lms algorithm, **Fig. 14**.

First you have a transversal filter, around which the lms algorithm is built. This component is responsible for performing the filtering process. Second, there is a mechanism for performing the adaptive control process on the tap weights of the transversal filter.

Details of the transversal filter component are presented in **Fig. 15**. The tap inputs from the elements of the M-by-1 tap input vector $u(n)$, where M–1 is the number of delay elements.

Figure 16 presents details of the adaptive weight-control mechanism. Specifically, a scaled version of the inner product of the estimation error and the tap input is computed. The result obtained defines the correction applied to the tap weight. The scaling factor used in this computation is called the adaptation constant or step size parameter.

The tap-weight vector computed by the lms algorithm executes a random motion around the minimum point of the error performance surface. This random motion gives rise to two forms of convergence behaviour for the lms algorithm – convergence in the mean, and convergence in the mean square.

It is important to realise, though, that the 'mis-adjustments' are under the designer's control. In particular, the feedback loop acting around the tap weights behaves like a low-pass filter, with an average time constant that is inversely proportional to the step size parameter u.

Hence, by assigning a small value to u, the adaptive process is made to progress slowly, and the effects of gradient noise on the tap weights are largely filtered out. This, in turn, has the effect of reducing the mis-adjustments. ■

Designers interested in Zilog's dsp, comms and miocrocontroller solutions, contact Gothic Crellon at 3 The Business Centre, Molly Millars Lane, Wokingham, Berkshire RG41 2EY, tel. 0118 978 8878, fax 0118 977 6095.

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ACTIVE

A-to-d and d-to-a converters

12-bit, 5MHz sampling a-to-d. *ADS803* from Burr-Brown is a 12-bit converter offering a 69dB s:n ratio and 82dB to beyond the Nyquist frequency, sampling at 5MHz. It has an internal reference and may be programmed for 2V pk-pk input for best spurious-free dynamic range or 5V pk-pk for lowest referred noise of 0.09 lsb rms, or any range between. An over-range flag for high input can be used to reduce the front-end gain to compensate. Digital error correction reduces differential linearity error to a typical ±0.25 lsb. Burr-Brown International. Tel., 01923 233837; fax, 01923 233979. ***Enquiry no 501***

Delta-sigma a-to-d. Crystal Semiconductor's *CS5529* 16-bit, delta-sigma analogue-to-digital converter allows a reduction in size over types of converter using other techniques. This is a low-power device using a single 5V, or ±2.5V, rail, includes a digital filter and self and system calibration, the converter being controllable from a 3V system. A 6-bit latch output allows control of switches and other devices and there is an SPI and Microwire three-wire interface for programming bipolar/unipolar working, calibration and output word rate. Sequoia Technology Ltd. Tel., 0118 9258000; fax, 0118 9258020. ***Enquiry no 502***

Digital signal processors

Digital potentiometers. Replacing four mechanical slider potentiometers, Xicor's *X9408* chip contains four non-volatile digital potentiometers, each having 64 settings, the *X9418* version being a dual, 64-setting type. Noise figure for both is –140dB/Hz, standby current 1μA. Sixteen eight-bit registers that hold the wiper positions in eeprom may also be used to store other data. Settings are carried out by means of a 400kHz, two-wire interface and the devices may be programmed to return to previous settings at switch-on or to go to a preset starting point. End-to-end resistance is 10kΩ, each resistor in the array having a value of 158.5Ω. Xicor Ltd. Tel., 01993 700544; fax, 01993 700533. ***Enquiry no 503***

Memory chips

Industrial serial eeprom. Microchip's *24AAXX* family of 1.8V I²C serial eeproms operate in the –40°C to 85°C temperature range and are available in densities of 128b to 16Kb. Standby and active currents are 500nA and 500μA, write speed 1-10ms, maximum clock frequency 400kHz and a claimed one million erase/write cycles. There is hardware write protection and, barring accidents, the ability to retain data for 200 years. Arizona Microchip Technology Ltd. Tel., 0118 92155858; fax, 0118 9215835. ***Enquiry no 504***

Microprocessors and controllers

Otp controllers. Two microcontrollers in the Temic *C51* range work at 40MHz at 5V or 16MHz at 3V. *TSC87C51/52* are compatible with existing *C51s*, these two offering 4K and 8K of eprom, being compatible with mask rom products. Temic has a factory programming service and the devices are provided with programming tools compatible with the Intel Quick-pulse algorithm. IEC Micromark Electronics Ltd. Tel., 01628 76176; fax, 01628 783799. ***Enquiry no 505***

32-bit risc starter. NEC has a starter kit to introduce the V850 32-bit risc controllers. The *EB-V853* is said to contain everything needed to allow people to start application development and testing the performance of the *V853*, which NEC says is one of the most flexible in the range. There is a compact cpu board with sram to hold the user's program, rom with a debugger monitor, a flash rom self-writing circuit, an RS232 port, connections to all cpu signals, a display and an 8-bit input port with dil switch to simulate input signals. The *V853* has an 8-bit d-to-a converter accessible from the cpu board, five 16-bit counter/timers, an interrupt controller and a two-channel pulse-width modulator. Software provided includes a C compiler with C source debugger and demo program. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290. ***Enquiry no 506***

C161 plus Newly introduced in the Siemens *C161* range of 8-bit microcontrollers is the *C161R1*, which comes with a two-channel, multiplexed I²C bus and SPI interfaces, a usart, an 8-bit, 4-channel analogue-to-digital converter, timers and a real-time clock, which allows the device to be 'woken up' from standby. There are also 3Kbyte of ram. At 16MHz, execution time for a command is around 125ns. Siemens plc. Tel., 0990 550500; fax, 01344 396721. ***Enquiry no 507***

Motors and drivers

Variable-speed control ics. GEC Plessey offers two integrated circuits for variable-speed control of motors in 'white goods' and in general-purpose inverters. *SA828* is a three-phase pwm generator for low-cost, efficient ac induction motor drives, the *SA838* being a single-phase version, which is also used in uninterruptible supplies. The *828* switches at 24kHz to give ultrasonic power switch operation and two standard waveforms, sine and sine+third harmonic, are available, the latter being a method of increasing motor power. No intervention by a microprocessor is needed unless frequency or waveform need to be changed. An evaluation board is provided. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095. ***Enquiry no 508***

PASSIVE

Passive components

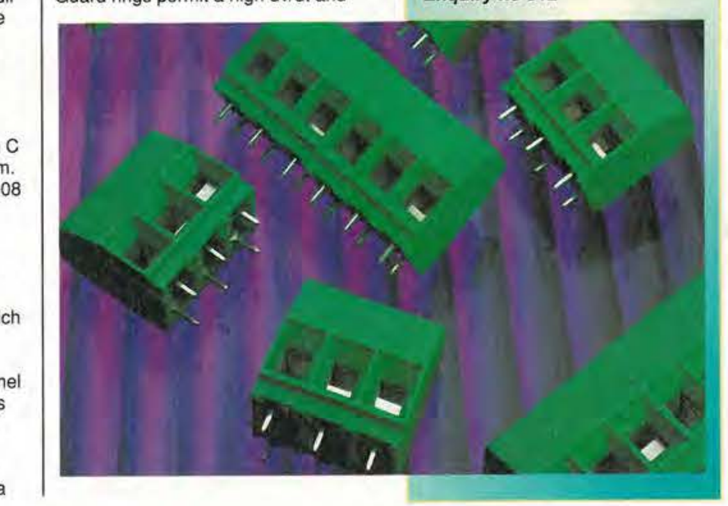
Miniature Schottkys. Zetex *ZHCS500* is an SOT-23 0.5A (6.75A pulsed) Schottky diode having a forward voltage of 550mV. Reverse recovery takes 10ns when switched from 500mA to –500mA and power dissipation is 330mW at 25°C ambient. Zetex plc. Tel., 0161-622 4444; fax, 0161-622 4469. ***Enquiry no 509***

Power Schottkys. Ixys has a new range of power Schottky diodes rated from 10A to 320A at 100V in TO-220, SOT-227B and TO-247 packages. Guard rings permit a high d*v*/d*t* and junction temperature may be 175°C with guaranteed avalanche ratings. Applications lie in low-voltage rectification in smps and as free-wheel diodes in low-voltage converters. GD Rectifiers Ltd. Tel., 01444 243452; fax, 01444 879722. ***Enquiry no 510***

Snap-in electrolytics. Nover's *LS* snap-in electrolytics combine a ripple rating of up to 4.8A and a rated life of 2000h at 85°C in a 22-35mm diameter snap-in case. Values are in the 47μF-22000μF range at ±20% tolerance and at voltages between 16V and 400V; leakage current is under 0.01*CV*. Anglia. Tel., 01945 474747; fax, 01945 474849. ***Enquiry no 511***

PCMCIA transformers. Transformers by Pulse, said to be the smallest available, are designed for Type 2 PCMCIA cards and are mounted in the middle of a card by pick and place equipment. Each transformer in the range will carry out transmit and receive functions and matches

Terminal blocks. High-current terminal blocks in the Phoenix Contact *MKDSP10* range carry 57A at 690V. Pitch is 0.4in and each position has two pins to increase mechanical stability and enhance the current capacity. Two and three pole versions are available and interlock to allow the construction of larger blocks. They accommodate solid wire of 0.5mm to 1.6sq.mm. and multi-stranded wire to 10sq.mm, versions also being able to take 2.3mm test probes. Onboard Electronics Ltd. Tel., 01256 818222; fax, 01256 840610. ***Enquiry no 512***



transceiver chips made by the leading companies. Silicon Concepts Ltd. Tel., 01428 751617; fax, 01428 751603.
Enquiry no 513

Audio products

Stereo, 115dB a-to-d. For professional use, AKM's *AK5392* 128-times oversampling analogue-to-digital converter uses the company's dual-bit technique to retain the low distortion of single-bit devices but with a wider dynamic range. *AK5392* resets itself when power is applied and phase detection of the clock ensures correct synchronisation, which is useful when several devices run together in slave mode. Sampling rate is 54kHz and sinad 100dB (s:n 115dB). Dynamic range is 115dB, and stopband attenuation 110dB. Asahi Kasei Microsystems Ltd. Tel., 01923 226988; fax, 01923 226933.
Enquiry no 514

Navigation systems

GPS development. GEC Plessey has a development system for the hardware and software of a 12-channel Global Positioning System receiver. *GPS Architect* consists of a GPS receiver board using the GEC Plessey *GP2000* chipset, the *GP2010* rf section, a *DW9255* saw filter, the *GP2021* 12-channel correlator and an *ARM60-B* 32-bit risc processor. There is enough rom and ram to avoid memory shortage during development and three serial ports are provided for connecting to a pc to assist with software downloading, differential GPS correction data input and display. An active antenna is provided, as are power supply and cables. The C source code may be embedded in a receiver. Even having acquired 12 channels, half the processing power is unused to allow the equipment to be combined with comms systems. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.
Enquiry no 515

Connectors and cabling

Parallel interface cables. GTK has a range of high-speed, bidirectional parallel interface cables for connection to IEEE 1284 pcs and peripherals. They use 25W D-type, 35W Centronics and 36W mini Centronics connectors and are made in lengths from 1.5m to 10m. Temperature rating of the double-shielded cables is –20°C to 75°C and they withstand 500V rms for a minute. Current rating is 1A. GTK (UK) Ltd. Tel., 01344 304123; fax, 01344 301414.
Enquiry no 516

Fine-pitch connectors. From Flint comes a range of 0.5mm pitch connectors in 130 combinations of ways, heights and configurations. *JAE WR* connectors are available in 30-120 circuits and may be stacked to give parallel board-to-board spacing from 4mm to 9mm. They are made in right-angled and vertical styles and a 'floating' type takes up misalignments of up to 0.25mm in X and Y planes. Current handling is 0.5A (0.3A for the floating type), minimum insulation resistance 100MΩ and dielectric voltage up to 500V ac. Contact resistance is 50mΩ. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.
Enquiry no 517

D-type subminiatures. *Subminiature D* connectors by ITT Cannon are available with combinations of pin types, being modular in form. Options include variations in current capacity up to 40A, 50Ω or 75Ω coaxial, high voltages and optical fibre, all with choice of crimp or solder connection and straight or right-angle pins. There are various materials and finishes for the shells and the connectors meet NASA, ESA and medical equipment requirements for outgassing and residual magnetism. PEI-Genesis UK. Tel., 01797 322003; fax, 01797 321589.
Enquiry no 518

Parallel-to-SCSI connector. *TransIT* is a parallel-port-to-SCSI interface for connecting portable SCSI equipment such as removable hard disks, cd-roms and scanners to a notebook or pc parallel port. Data transfer is at the rate of 900KB/s with a burst rate of 1.5MB/s. Power is by SCSI Term Power and up to seven SCSI devices may be daisy-chained. Drivers for dos, Windows and OS/2 are included. Shuttle Technology Ltd. Tel., 0118 9770441; fax, 0118 9771709.
Enquiry no 519

Displays

Wide-angle lcd. NEC's *NL6448AC33-24* is a 10.4in flat-panel lcd that has a viewing angle of 80° from normal in any direction, that is 160° overall; in this region, there is no preferred viewing angle. It is a 640 by 480 pixel screen and a built-in digital interface to give 262 000 colours. The display is nearly three time as bright as a typical pc panel, luminance being 190cd/m^2 and contrast ratio 150:1. Sunrise Electronics Ltd. Tel., 01908 263999; fax, 01908 263003.
Enquiry no 520

Bench multimeter. Thurlby Thandar's *Model 1604* benchtop multimeter is an auto-ranging instrument with a large, bright display. Scale length is 40 000 and it offers true rms measurement in the audio band, a basic accuracy to within 0.08% and resolutions of 10μV, 10mΩ and 0.1μA. Functions include relative measurement, max/min storage and a function to allow readings to be held on display each time a new test point is probed. Frequency up to 40kHz may be measured to within 0.1Hz and there is an isolated RS-232 interface to a pc; data logging software is available. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.
Enquiry no 524

Display driver kits. Flat-panel display driver kits from Anders allow users to run any tft or stn graphics display in minutes, using a Goldstar panel or any other. Kits include everything needed to drive the display: driver, interconnection, controller board and bios and any cables needed, the kit being configured for the specified panel. A family of controller cards in the kit are for ISA, VESA and PCI buses, which covers every current type of panel. Both low-voltage differential signalling and Panelink cards are available to allow a high-resolution display to be driven remotely. Anders Electronics plc. Tel., 0171 3887171; fax, 0171 3872951.
Enquiry no 521

100-led bargraphs. New *led bargraph* displays by Lumex contain 101 or 103 led chips in a 106mm-long package, providing a fast display comparable in accuracy with slow analogue meters. The chips are mounted on 1mm centres and overall height is 5.99mm. Colours are red, green and yellow in combination and lenses and drivers can be specified. Lumex Opto/Components Inc. Tel., 001 847 359-2970; fax, 001 847 359-8904.
Enquiry no 522

Thin tft monitor. GTT produces the *GTM-121* thin-film transistor liquid-crystal monitor, which is a 12.1in active-matrix type having a resolution of 800 by 600 and giving 262144 colours; the range of sizes is now complete at 9.4in to 15.1in. A full range of controls are incorporated and a non-volatile memory stores size and set-up functions. As an option, a resistive touch facility may be added. Craft Data Ltd. Tel., 01494 778235; fax, 01494 773645.
Enquiry no 523

Hardware

Thin fan. Sanyo Denki's *San Ace 140L* long-lived cooling fan has been slimmed down from its earlier 51mm and is now 38mm thick, being 140mm square. It will, it is said, live for 100 000h at 60°C and is made for use on 12V, 24V and 48V supplies in 1900rev/min or 2600rev/min versions. Maximum air flow from the faster versions is 4.5m^3/min. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.
Enquiry no 525

Racks. Vero's *Network Support Rack* is a floor-standing enclosure designed to hold several computers in tower or desk-top cases, monitors, keyboards and an uninterruptible power supply; it also contains a fire protection system. Cable management, power distribution, thermal management and filtering are all catered for. The rack can hold a 500kg load, comes in heights of 37U and 42U and in widths of 600mm and depths of 600 or 800, being compatible with 19in and ETSI racking in combination. High-quality locks are used and, optionally, powered locks operated by keypad. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.
Enquiry no 526

Test and measurement

Digital pressure measurement. Digital manometers by Yokogawa, the *MT120/110*, use a silicon-resonant sensing technique to achieve an accuracy within ±0.02%. In addition to the facilities found in the *MT110*, the *MT120* also has a digital multimeter to measure 1-5V dc and 4-20mA, a 24V dc power supply for transmitters and sensors and percentage error display for automated field calibration. Both measure gauge pressure to 3000kPa, absolute and differential pressure down to 1kPa in gases and liquids. A 1024-point memory is incorporated for data and calibration storage. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.
Enquiry no 527

Torque/angle measurement. From Schatz GmbH comes the *Blue Box*, a portable torque and torque/angle measuring instrument designed to work with Autocode transducers, which automatically transmit their vital statistics to the instrument, including range, model, serial number, calibration details and date of calibration. Measurement is in the 1Nm-10000Nm range and all the above information is supplied to the *Blue Box* with no intervention on the part of the user. Until, that is, an inappropriate transducer is connected, when a warning is given. Results are printed or passed by way of an RS232 interface to a pc. The instrument may also be operated as a slave from the pc. Scientific Electro Systems Ltd. Tel., 01702 335174; fax, 01702 431105.
Enquiry no 529

Optical encoder output. Control Transducers has a range of small, add-on pcbs to complement its family of optical encoders, conferring the facility to use a standard encoder with a range of output features for which a special board often must be made. Output features available from eight types of board include a line driver, quadrature decoding, clock and direction of motion converter, the ability to work from 7-32V rails and several others. The boards are very small and are mounted in layers to give a compact assembly. Control Transducers. Tel., 01234 217704; fax, 01234 217083.
Enquiry no 528

Literature

PCI/ISA cards. Signal acquisition and signal generator cards for PCI and ISA bus are available from LeCroy for use with pcs and are described in a new short catalogue. There are 14 cards with up to 16Mbyte of acquisition memory, transferring data to pc dram at 100Mbyte/s, 8-bit cards sampling at 500Msample/s with 2Mbyte of memory and 12-bit cards sampling at up to 100Msample/s. LeCroy Ltd. Tel., 01189 344882; fax, 01189 348900.
Enquiry no 530

Materials

Death to solder balls. *Carapace EMP 110 LGXM* is liquid photoimageable soldermask that eliminates solder balls forming during the reflow process, by which time it is too late to do much about it. *Cara...* combines a light colour mask with a matt finish to make a solder mask to resist ball adherence, thereby increasing the chances of obtaining high yields in ball-grid array device mounting. The roughness of the surface and the lower absorption of thermal energy in the light colour are responsible for the mask's performance. Electra Polymers and Chemicals Ltd. Tel., 01732 811118; fax, 01732 811119.
Enquiry no 531

Cleaner. *EnSolv* spray cleaner is a non-flammable, fast-drying type with low surface tension for good penetration. It was developed for use as an aircraft cleaner and is very suitable for use with electrical components and as either immersion or spot cleaning for populated boards. It is an economical, 'green', low toxicity replacement for 1.1.1., trichlorethane and HCFC141b. Croftshaw Solvents Ltd. Tel., 0181 508 5564; fax, 0181 508 5559.
Enquiry no 532

Printers and controllers

Thermal printer. Offering an improvement in speed, noise (55dB) and reliability over types using impact printing, Panasonic's *EPL1901S2* thermal printer is meant for use in hand-held terminals, point-of-sale and other ticket-printing applications; it is 82.6 by 41 by 15.8mm in size and weighs 60g. If needed to replace an impact printer, there is a mounting plate for the purpose. Printing speed is 6.8lines/s and resolution 12dots/mm and paper widths up to 58mm. Panasonic UK Ltd. Tel., 01344 853157; fax, 01344 853081.
Enquiry no 533

Production equipment

Bare board tester. Polar Instruments has the *CITS500* controlled-impedance test system for use in pcb production. It will verify both single-ended trace impedance and the differential impedance of balanced traces, simply and with none of the setting up needed with conventional time-domain reflectometry. The system is controlled by Windows-based software to measure the reflection of fast pulses, in which the user selects the appropriate test file, positions the probes on being prompted and clicks the mouse or footswitch, a series of tests progressing automatically to give a graphical view of the characteristic impedance against a pass profile. Results are also present in opto-isolated rear-panel signals for use by a factory's data-logging system and may be printed. Many accessories are supplied in the package or are optional. Polar Instruments Ltd. Tel., 01481 53081; fax, 01481 52476.
Enquiry no 534

Board inspection. Electronic assemblies are simpler to inspect using the *Alpha 3Di* twin-camera system. The two cameras give vertical and angled views of the subject, the angled one providing dual magnification, controlled by a foot switch, and 360° rotation. Variable halogen illumination is applied via an optical-fibre cable and the X-Y table has an anti-static mat and wrist strap connection is provided. Video appears on the rear of the 250mm PAL monitor, with output for video printer or built-in computer for storage of screen images in .BMP files. Alpha Metals. Tel., 0181 6656666; fax, 0181 6654734.
Enquiry no 535

125mA charge pump. Maxim's *MAX1680/1* are high-frequency charge pump voltage converters supplying up to 125mA when doubling or inverting inputs of 2-5.5V and meant to supply analogue measurement and amplifier circuits. Both need only two ceramic capacitors, total board area needed being 0.06sq.in. The *1681* operating frequency is selectable at 500kHz and 1MHz, while the *1680* choice is 125kHz or 250kHz. Output resistance of both types is 3.5Ω. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.
Enquiry no 536

Power supplies

Pos/neg 3.5A regulator. *MSK 5200* 3.5A voltage regulators are a series of 350mV dropout, fixed-voltage devices that combine positive and negative outputs in one package. Output combinations are based on 3.3V, 5V, 5.2V, 10V and 12V positive and 5V, 5.2V, 10V, 12V and 15V negative, all outputs being internally trimmed to within ±1%. Internal short-circuit and thermal protection are provided and an electrically isolated case is used. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.
Enquiry no 537

600W module. Intended for use in racks, file servers and base station transmitters, Coutant Lambda's *PD600* single-output ac/dc power supply module includes EN55022 level B filtering, power factor correction, monitoring and signal-generation as standard. To form a complete power source, only a pcb, heat sink and output connector are needed, the whole being able to fit into a 3U rack or 1U tray. In the event of another power supply failing, the PD600 will operate as the main supply. Regulation and stabilisation are both 0.5%, ripple 200mV maximum, inrush current 25A and typical input current 3-4.2A depending on input voltage. Coutant Lambda Ltd. Tel., 01271 865656; fax, 01271 864894.
Enquiry no 538

A.c. supplies. *Chroma 6400* a.c. power supplies are programmable in frequency from 45Hz to 500Hz, the range consisting of five models for 375VA to 3000VA output power at voltages from 0-150V to 0-300V. Distortion is under 0.3% and the pfc provides a power factor of 0.98. Programming is by way of GPIB, RS232C and analogue interfaces and there is protection against the natural hazards of power supplies, including a fan. Glassman Europe Ltd. Tel., 01256 883007; fax, 01256 883017.
Enquiry no 539

1.25V adjustable regulator. Semtech's *SC431* is an adjustable shunt regulator and is a direct

replacement for the other *431*s, having an operating current range of 100μA-150mA and adjustable output voltage of 1.24-20V and 2.5-37V by means of two external resistors. Output impedance is 0.25Ω and the very sharp turn-on characteristic is like that of a zener diode. Voltage tolerances of the device is 0.5%, 1% and 2% and there is a choice of four packages. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.
Enquiry no 540

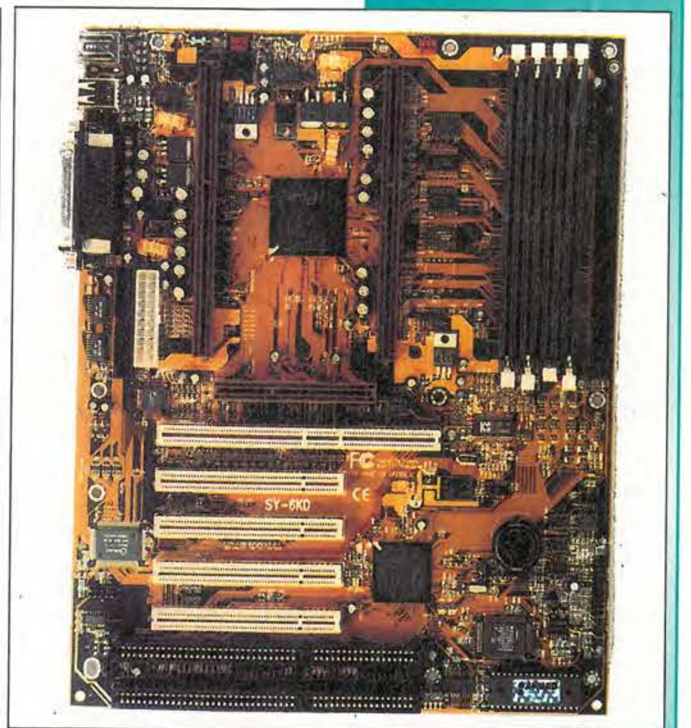
Protection devices

Battery protection. *PolySwitch LR4* resettable fuses by Raychem provide protection against current overload in battery packs, being a third the size, Three times faster acting and having half the resistance of earlier *PolySwitch* devices, and are of a new design to satisfy the requirements of oems using AAA cells. Hold current ratings are up to 7.3A in the *LR4* family and maximum operating voltage 15-20V dc, with a maximum interrupt current of 100A. Also introduced are *TAC* devices for AAA NiCd cells, having a new cap design for use on batteries with or without buttons. There are three devices: the 1.7A *TAC170-09*, the 1A *100-09* and the 2.1A *210*. Raychem Ltd. Tel., 01973 572692; fax, 01973 572209.
Enquiry no 541

Solderable voltage suppressors. Harris's new *ML* and *MLE* leadless, multi-layer transient suppressors are made using a termination process that enables the devices to meet solderability needs of conventional surface-mounting processes and testing. The new process allows nickel plating of only the end terminations while retaining the 50% to 70% fillet height needed for gold testing. They replace larger zener diodes and the leadless construction eliminates the normal lead parasitics to allow the suppression of very fast transients, complying with IEC 1000-4-2, MIL-STD-883C and others. Resistance when on is 1-10Ω. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.
Enquiry no 542

Mass storage systems

Rewritable optical disk drive. Panasonic has launched a portable version of its PD drive, the 650Mbyte rewritable optical disk and cd-rom drive. *LF1500EPB* is meant for removable secondary data storage and cd drive for the notebook market; since it uses the standard 25-way parallel printer port, it is also suited to shared use for multiple office pcs. It will be compatible with Panasonic's DVD-ram drives shortly to be introduced, which will read and write to existing PD disks. Software is supplied. Controls on the unit allow its use as an audio player. Panasonic Industrial (Europe) Ltd. Tel., 01344 853827; fax, 01344 853313.
Enquiry no 541



Switches and relays

Miniature relay. *G5V-2*, a miniature signal relay by Omron has double-throw, double-pole contacts handling 1μA-2A and up to 125V ac or dc. Two models are available: the *-2* low-sensitivity version handling up to 60W with a 500mW coil rating for operating voltages from 4.5V dc to 24V dc or 580mW at 48V dc. The *G5V-2H1* type switches 24W with a coil rated at 150mW from 4.5V to 12V dc, 200mW at 24V and 300mW at 48V. Packages are sealed and measure 11.5mm high by 20.5mm by 10.1mm. Onboard Electronics Ltd. Tel., 01256 818222; fax, 01256 840610.
Enquiry no 543

Rf relays. Teledyne's *RF3XX* series bypass relays, which operate in the band 0-3GHz, are provided with an internal bypass link to avoid the need for a bypass link on the board. An external link is unavoidably longer, involves the use of differing materials in the path and several direction changes for the signal – a bad idea at 3GHz. The relays provide the bypass in normally open and normally closed versions. Repeatability is ±0.1dB. Package is TO-5, with a ground pin.
Enquiry no 545

Television components

Digital satellite interface. A *network interface module* is announced by GEC Plessey, forming a reference design for the front end of a set-top box for digital satellite television reception. It complies with the dvd standard, accepting 950-2150MHz L-band in and producing MPEG2 transport stream out, no alignment being necessary. In five ics, the L-band input is converted to a 485MHz if, which is I/Q processed down-converted to give I and Q channels at 40MHz bandwidth. Each channel is then passed to an a-to-d converter and to a qspk/fec decoder to produce the MPEG data. Microphony is reduced by the use of printed inductors. Visual Basic evaluation software is provided and generic C coding for production. An evaluation kit is available. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.
Enquiry no 546

Computer board-level products

Dual Pentium motherboard. Soyo has a motherboard for dual Pentium systems up to 333MHz which uses the latest high-speed graphics interface, the Accelerated Graphics Port, and the intelligent i/o bus. The *SY-6KD* is based on the 82440LX chipset and is in the standard ATX form and, in addition to the AGP, has five PCI and two ISA expansion buses. There are four 168-pin dimm sockets for up to 1Gbyte of dram. An AMI PCVI bios supports multiple boot from IDE, SCSI, cd-rom or floppy disk and an optional Lan-Desk Client Manager is also available for networking. Soyo UK Ltd. Tel., 0181 4819720; fax, 0181 4819725.
Enquiry no 544

Transducers and sensors

Fibre sensors. Matsushita's *UZF* range of fibre sensors is increased by three specialist types with long-range sensing amplifiers and a large array of heads. They use the *UZFRE21* diffuse reflective head with a 0.8mm diameter sleeve on an M3 screw thread, giving a maximum range of 13mm and the facility to detect a standard target of 0.01mm diameter. The *UZFRH7* has a stainless steel armoured sleeve and optical-fibre cable operating at temperatures between –60°C and 350°C, sensing at a range of 88mm. Lastly, the *UZFRL4* is a fixed-focus type to detect any object, regardless of colour or surface condition, at a range of 4.5-8mm. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.
Enquiry no 547

Light-dependent resistors. Invac cadmium sulphide ldrs come in miniature open-frame and enclosed versions and have a response matched to that of the human eye at 560nm to make them suitable for lighting control. *VAC54* is of the miniature type with epoxy coating and operating from –30°C to 70°C at up to 150V dc. Minimum dark resistance is 20MΩ, 105kΩ at 10lux; rise and decay times are 20ms and 30ms. Anglia. Tel., 01945 474747; fax, 01945 474849.
Enquiry no 548

COMPUTER

Data acquisition

PCIbus cards. United Electronics *PowerDAQ* data acquisition cards are said to be the first to fully utilise the PCI bus, thereby removing bottlenecks inherent in the ISA bus. The cards are based on a 24-bit Motorola *DSP56301* running at 66MHz and having an on-chip PCI interface; the dsp processor connects via a high-speed internal data bus to the system logic, which is in a fpga. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.
Enquiry no 549

Programming hardware

Production programmer. Stag has a new version of the *P803* programmer for use in production, this one using two, eight-socket plug-in modules to increase its flexibility. It will now gang-program between eight and sixteen 3V or 5V devices simultaneously. For stand-alone or remote working, the P803 has an embedded processor for rapid programming; sixteen *28F010* flash chips can be processed in six seconds. Ram is 4Mbyte as supplied and may be expanded to 16Mbyte using standard simms. The device library can be updated and is in non-volatile memory. There are visual and audible alarms to indicate completion; software for operation and data comms is available for all types of computer. Stag Programmers Ltd. Tel., 01707 332148; fax, 01707 371503.
Enquiry no 550

Software

Neural networks. *Neuropredictor* by SignalBox is particularly good at pattern recognition and prediction and is an advance in that it cost less than earlier examples of this type of software and uses rather less expensive hardware than of yore, running under Windows on an ordinary pc. This network is based on radial basis function architecture, in which locally tuned, overlapping receptive fields allow training of one part of the net without having another part regress. In this way, learning is easier, ability to find patterns and trends, even in the presence of noisy data, is enhanced and training is faster. So much faster, in fact, that the software can be run on an unmodified pc to give answers to 'what if?' input. SignalBox Ltd. Tel., 01709 898989; fax, 01709 897787.
Enquiry no 551

Improved *LabWindows*. National Instruments has improved the virtual instrument development software *LabWindows/CVI* in version 5.0. In the new version, an Instrument Driver Wizard generates VXI plug&play instrument drivers automatically, these drivers reducing test times by over 50% and keeping track of the virtual instrument settings to reduce traffic on the bus. A Channel Wizard interactively defines input transducers, handling conversion and scaling. Among other enhancements is the multithreading facility, which enables data to be acquired or other continuous operations to be performed separately, apart from other tasks the processor is carrying out; full priority is afforded the acquisition board in this way and no data is lost. National Instruments UK. Tel., 01635 572400; fax, 01635 524395.
Enquiry no 552

SPEAKERS' CORNER

What are the best shapes and materials for woofer enclosures? John Watkinson relays his view of the best options.

As the wavelength of sound at low frequencies is long compared to the dimensions of a typical loudspeaker drive unit, an unbaffled driver is ineffective as air simply flows from one side of the diaphragm to the other as it moves. The solution is to enclose one side of the diaphragm so that this cancellation cannot occur.

A sealed enclosure is the only one which allows an accurate phase response. The sealed enclosure is sometimes referred to as an infinite baffle, but this is quite incorrect. **Fig. 1a)** shows that radiation from a driver in a true infinite baffle is into an acoustic half-space, whereas from a sealed box **1b)** it is omnidirectional into a whole sphere. The acoustic impedance experienced by the driver, and hence its output, is quite different.

The sealed enclosure acts as an air-spring in parallel with the stiffness of the driver's own suspension. This raises the fundamental resonance of the system, raising the lowest reproducible frequency.

Resonant frequency can be reduced by increasing the diaphragm mass, but this results in an inefficient unit. Another approach is to make the stiffness of the drive unit very low so that the air-spring dominates. This results in the so-called acoustic suspension loudspeaker.

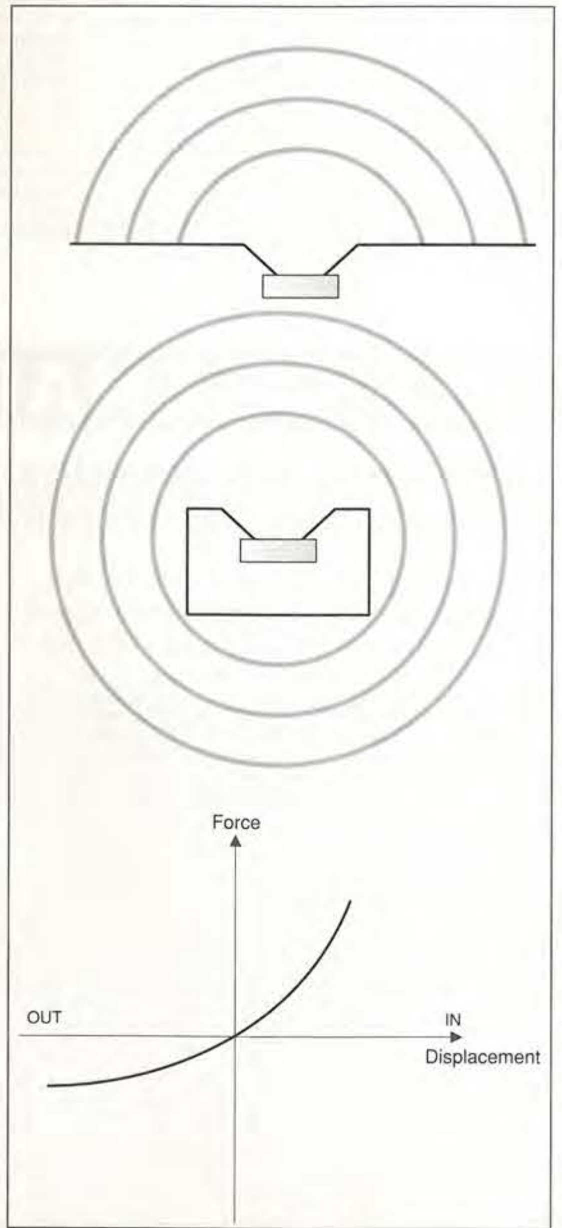


Fig. 1. Section a), top, depicts radiation into half space in a true infinite baffle. In b), radiation from a sealed enclosure is into the whole sphere. Air spring in an enclosure, c), is non-linear because box volume is not proportional to displacement.

Air-spring drawbacks

Acoustic suspension was claimed to be more linear than the spider of the drive unit, but this is not so.

Figure **1c)** shows that an air spring is non-linear because the pressure increase is greater for a given inward movement than the pressure drop for the same outward movement. This is great for truck suspensions, but causes distortion in loudspeakers.

The resonant frequency can be lowered by addition of a critical amount of a material such as wool inside the enclosure. The wool must be teased out so that it fills the entire volume.

Wool's specific heat is much greater than that of the air and so the temperature of the air cannot change with pressure. The stiffness of the air-spring is reduced and the fundamental resonance goes down.

If the resonance is still too high, then it can be artificially lowered using signal processing techniques which are readily incorporated into an active speaker.

The surface area of an enclosure is an order of magnitude greater than that of the diaphragm and so the enclosure can radiate very effectively if it is not correctly designed. The goal is to prevent vibration of the enclosure walls and this can be done by stiffness, mass or damping or a combination of these.

A loudspeaker can easily be tested for enclosure colouration. You simply rap the panel with your knuckles in various places. On a well designed unit the only evidence of the blow should be a painful hand. Unfortunately too many loudspeakers emit a lot of sound due to structural resonances and flexing.

Many are not stiff enough

Many traditional loudspeakers have really quite poor structural stiffness. The flat wooden panel is the weakest method of resisting pressure known. **Figure 2a)** shows that flat panels bend readily under internal pressure because a large deflection causes only a small change of length.

A spherical or cylindrical shape cannot flex because wall deflection can only occur with a serious change in wall length. Aerosols and airliners also adopt this solution.

At one time enthusiasts used concrete drain pipes as

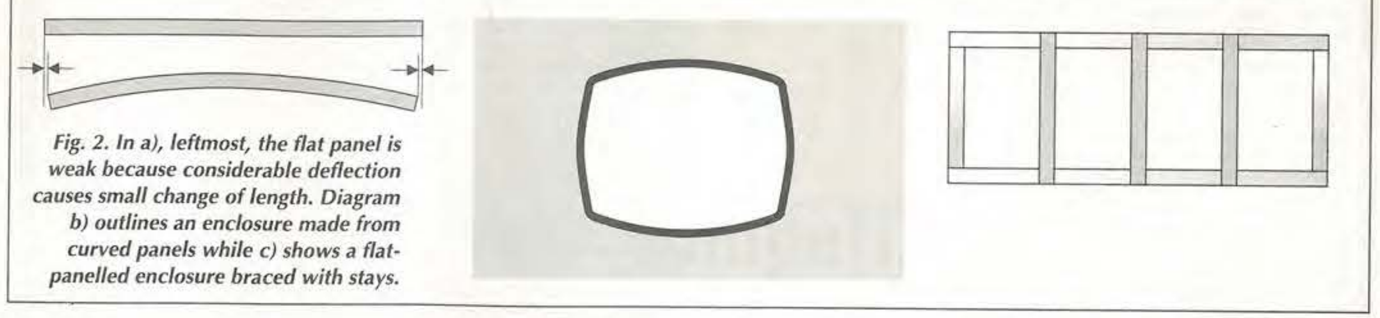


Fig. 2. In a), leftmost, the flat panel is weak because considerable deflection causes small change of length. Diagram b) outlines an enclosure made from curved panels while c) shows a flat-panelled enclosure braced with stays.

enclosures with considerable success – although the result was hardly portable. Single or compound curvature construction is always stiffer than flat panels, as you will notice from the shapes of car bodies. Useful loudspeaker enclosures can be made by joining spherical or curved sections as in Fig. **2b)**.

Flat panels are often used for economy or ease of home construction. Better results can be obtained by suitable bracing. Figure **2c)** shows the use of stays, a technique used to brace the flat areas of locomotive boilers in the firebox area. In loudspeaker, stays can be 10mm dowel.

Domestic builders can easily incorporate dowel stays as they can be glued into through-drilled holes and cut off flush. The dowels have a secondary purpose of preventing the acoustic wadding from sagging.

The size v performance conflict

The conflict between size and performance can be reduced by using suitable materials and design.

If the internal volume of a loudspeaker can be increased, the fundamental resonance will go down in frequency. This can be done without making the outside dimension larger, but by making the walls thinner.

Thin-wall woofer design is still relatively new, but allows huge leaps in performance with respect to size. Curved shapes get their stiffness through converting flexing into length changes and these can only be resisted in a thin wall by using a material with low modulus of elasticity.

Metal is a natural choice here, allowing a structure both stiffer and lighter than wood, and with a wider range of shapes available. Pressed steel has many advantages as it is strong and cheap, although the tooling costs put it out of the reach of the home builder.

Where moderate quantities are required, glass-fibre has much to recommend it because it can easily be fabricated with compound curvature and an excellent finish. ■

TRAC Design Competition Winner

One of six winners of the TRAC design competition, Mike Button, has devised a new solution to ssb modulation and demodulation that is only practicable using the TRAC concept.

A new demodulator for single sideband

The arrival of the TRAC devices on the electronics market has led to the possibility of solutions which have hitherto been only possible by either complex pcb with many op amps and passive components or by expensive digital signal processors.

As a radio amateur I have been intrigued by the possibility of obtaining single sideband modulation or demodulation without recourse to expensive filters. The 'third method' of obtaining the sum and difference frequency components of two signals at low intermediate frequencies is now a low cost possibility.

Behind ssb

If two sinusoidal signals of different frequencies are multiplied together it can be shown that the result of the multiplication is given by the following,

$$A\sin xB\sin y = AB\frac{\cos(x-y)+\cos(x+y)}{2} \qquad (1)$$

$$A\cos xB\cos y = AB\frac{\cos(x-y)-\cos(x+y)}{2} \qquad (2)$$

$$A\sin xB\cos y = AB\frac{\sin(x-y)+\sin(x+y)}{2} \qquad (3)$$

These equations show that the result of the multiplication comprises of two new frequency components, one the sum and the other the difference of the two frequencies.

Each of these three equations represents the mathematics for an amplitude-modulated signal when a perfect balanced modulator is used to mix a low audio frequency with a high frequency. In practice, no mixer can be made perfect. Consequently, there will always be a component of the high frequency in the equation.

Assuming the higher frequency to be y and K to be the leakage factor of the mixer then an amplitude-modulated signal will comprise of the following components,

$$\mathrm{K}\sin yAB\frac{\sin(x-y)+\sin(x+y)}{2}$$

An amplitude-modulated signal comprises upper and lower sidebands plus some component of the modulating frequency. Because each of the sidebands contains all the intelligence of the modulating frequency it was soon realised by both professionals and the amateur radio fraternity that the bandwidth of the modulated signal could be reduced to less than half if only one sideband was transmitted. Reducing the bandwidth meant that allotted radio frequency bands could be used more efficiently – i.e. more channels in a given radio band.

Currently, most ssb radio transmitters and receivers use one of two methods to to remove the unwanted components produced by the mixers. A good example of a direct-conversion receiver is presented in the February issue of *Electronics*

More winners

These are the remaining five winners of the TRAC design competition. Each will receive a TRAC development kit worth £600.

400Hz three-phase exciter. Ben Sullivan's exciter is primarily for aircraft equipment testing. Its three-phase oscillator can lock to an external signal and has an out-of-lock indicator.

Amplitude modulator. Designed by Franck Bigrat, this modulator is implemented from the mathematical formula for an AM signal.

Digitally controlled audio preamplifier. Andrew Wilkes' amplifier uses one TRAC IC to produce a four input-source stereo preamp with tape deck support and a 4-bit, i.e. 16 log step, volume control.

Power meter. Charles Bacon's design is a meter for measuring power dissipation in a transistor. It makes use of TRAC's log, addition and antilog abilities to calculate power dissipated in real time, by multiplying observed current with voltage.

Two tone oscillator. This entry is a circuit is for linearity testing of wireless transmitters, for example at hf. Designed by Ian March, it uses TRAC's log function as limiter.

Four TRAC devices are used in the design, each having 20 functional blocks, giving a total of 80 functional blocks. The number of function blocks used is 40. The number of unused blocks is 5, giving 35 unusable blocks. These are pin functions of TRAC device 1.

Pin	Connection	Description	Pin function
1	INPUT	Sinusoidal input to be demodulated (Sig)	*Asinx*
2	INPUT	Sin wave input local oscillator (LO)	*Bsiny*
3	Link to 15	Signal	*Asinx*
4	NIP	Local oscillator	*Bsiny*
5	AUX		
6	AUX		
7	DIF	Signal differentiated to give 90° phase shift	d/dx(*Asinx*)=*KAcosx*
8	DIF	LO differentiated to give 90° phase shift	d/dy(*Bsiny*)= *LBcosy*
9	AUX		
10	AUX		
11	AMP	Gain adjust to give phase shifted signal the same amplitude as unshifted signal	*Acosx*
12	AMP	Gain adjust to give phase shifted signal the same amplitude as unshifted signal	*Bcosy*
13, 14	-		
15	Link from 3	Sig	
16	Link to 68	Sig inverted	*–Asinx*
17-24			

Pin functions of TRAC device 2.

Pin	Connection	Description	Function on pin
23	Link from 15	Signal	*Asinx*
24	INPUT	Input dc bias	*E*
25	ADD	Addition of bias to signal (inverted)	–(*E+Asinx*)
26	NIP	DC bias	*E*
27	LOG	Positive log of signal plus bias	{+log}(*E+Asinx*)
28	LOG	Inverted log of dc bias	{–log}(*E*)
29	ADD	Log of signal plus bias divided by dc bias	{–log}(1+*A/Esinx*)
30	NEG	Positive log of dc bias	{+log}(*E*)
31	Link from 4	Local oscillator	*Bsiny*
32	Link from 26	DC bias	*E*
33	ADD	Adds dc bias to local oscillator (inverted)	–(*E+Bsiny*)
34		-	
35	LOG	Positive log of local oscillator plus bias	{+log}(*E+Bsiny*)
36	Link from 28	Inverted log of dc bias	{–log}(*E*)
37	ADD	Log of LO plus bias divided by dc bias	{–log}(1+*B/Esiny*)
38	Link from 29	Log of signal plus bias divided by dc bias	{+log}(1+*A/Esinx*)
39	ADD	Sum and negate log outputs, ie multiply	{+log}((1+*A/Esinx*).(1+*B/Esiny*)) ={–log}(1+*A/Esinx*+*B/Esiny*+ *AB/E/Esinx.siny*)
40	Link from 30	Positive log of dc bias	{+log}(*E*)
41	ADD	Multiply result by *E* (negated)	{–log}((*E+Asinx+Bsiny*+ *AB/Esinxsiny*)
42	Link from 32	DC bias	*E*
43	ANT	Antilog of multiplication	*E+Asinx+Bsiny*+ *AB/Esinx.siny*
44	NEG	Negated dc bias	*–E*

World. This method relies on the inherent low-pass filtering of an audio amplifier.

Another method, preferred because of its greater sensitivity and frequency selection is the use of intermediate frequency, or frequencies, employing high gain amplifiers with very close tolerance filters to remove the unwanted components.

The 'Third Method' of modulating and demodulating ssb signals has been known since the beginning of the century, but up to now the other two methods have been cheaper or easier to implement.

You can see that by adding or subtracting the result of equations 1 and 2 above, then either the frequency sum component or the frequency difference component can be isolated. In this way, an audio signal can be retrieved from a ssb signal by 'mixing' the ssb signal with a frequency equal to the original modulating frequency. Alternatively, an ssb signal can be produced from an audio frequency mixed with the required radio frequency.

TRAC calculations

Constraints. The *TRAC* device provides an inversion for both the log and antilog functions – i.e. a negative output is obtained for a positive input, and that both functions give a '0V' output for a '0V' input. This caused me to revise my understanding of elementary mathematics, which led me to believe that the log of zero was minus infinity and the antilog of zero was one.

Furthermore, if a number is divided by itself, as in $\log x - \log x = 0$ the resulting antilog should be 1. This infers that the zero level input to the log and antilog functions represents a unity-level signal. After delving further into the misty past of my O-Level mathematics, I realised that if you scaled the inputs and outputs to meet the available logarithmic range, any multiply or divide calculation is possible.

Once I understood that a '0V' input is the low end of the logarithmic dynamic range and that a '1.4V' was the high end then all was clear. Investigations using the simulator gave a dynamic range for the log/antilog function of 87dB.

The fact that the log/antilog functions were inverting also produced wrong results, until I realised that if a multiplication was required to a negative log output then the multiplicand had to be subtracted, rather than added. Conversely division needs an addition.

Also, the log and antilog functions needed the input to be wholly negative or wholly positive to obtain a correct multiplication or division.

Defining the inputs. Let the two input signals be,

$$A.\sin x \qquad (4)$$
$$B.\sin y \qquad (5)$$

The input to the log function must be wholly negative or positive and a dc bias voltage is necessary to lift the ac inputs above 0V. Let E be the dc bias voltage.

Functions. To obtain the necessary cosine function, as required by equation 2, the two ac input signals need to be differentiated. i.e. $\cos x = d/dx(\sin x)$.

$$A\cos x \qquad (6)$$
$$B\cos y \qquad (7)$$

There are now available the necessary functions to calculate the sum and difference components of the two inputs. To ensure that the inputs to the log functions never go below the zero level, a dc bias must be added to the inputs prior to performing a log function. Adding bias voltage E to equations 6 and 7 gives,

$$E + A\sin x \qquad (8)$$
$$E + B\sin y \qquad (9)$$

where E must be greater than the larger of A or B.

To keep the signals within the dynamic range of the TRAC, these signals have to be divided by E prior to multiplication.

$$1+\frac{A}{E}\sin x \qquad (10)$$

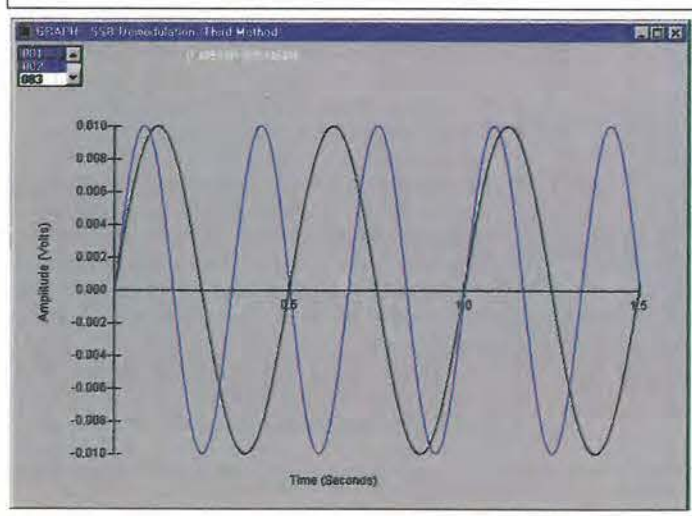
$$1+\frac{B}{E}\sin y \qquad (11)$$

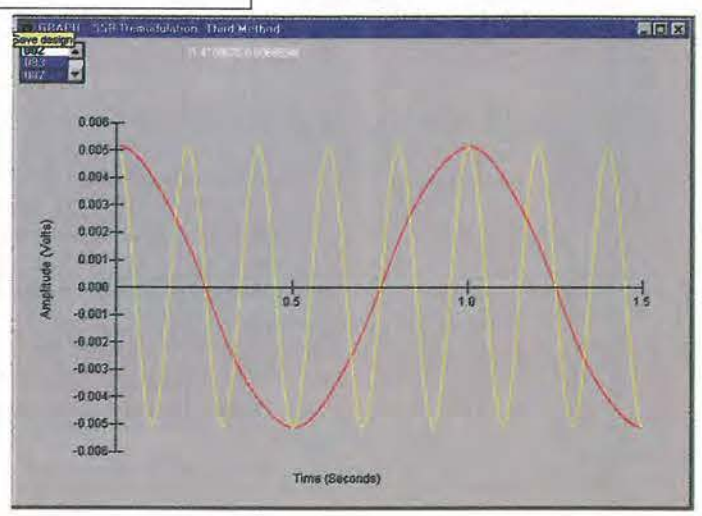
Multiplying equations 10 and 11 gives,

$$1+\frac{A}{E}\sin x+\frac{B}{E}\sin y+AB\frac{\sin x\sin y}{E^2} \qquad (12)$$

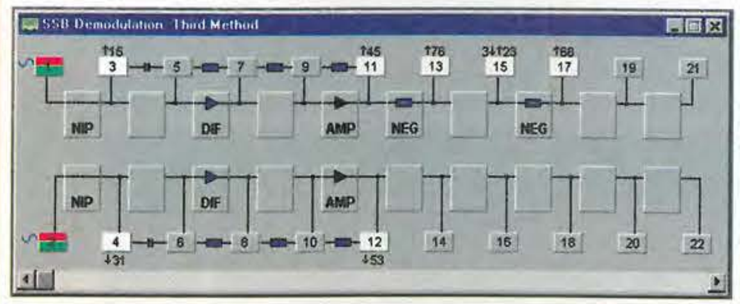
Pin functions of TRAC device 3.

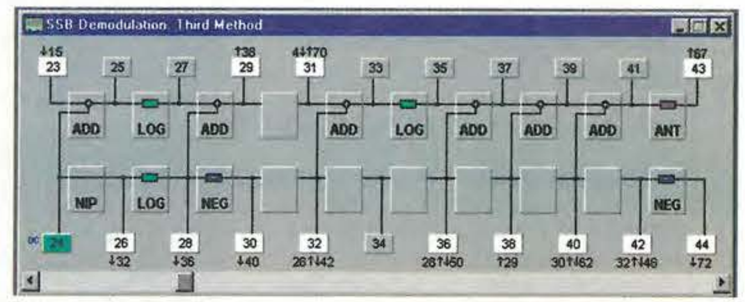
Pin	Connection	Description	Pin function
45	Link from 11	Signal	*Acosx*
46	Link from 42	DC bias	*E*
47	ADD	Addition of bias to signal (inverted)	–(*E+Acosx*)
48			
49	LOG	Positive log of signal plus bias	{+log}(*E+Acosx*)
50	Link from36	Inverted log of dc bias	{–log}(*E*)
51	ADD	Log of signal plus bias divided by dc bias	{–log}(1+*A/Ecosx*)
52			
53	Link from 4	Local oscillator	*Bcosy*
54	Link from 46	DC bias	*E*
55	ADD	Add dc bias to local oscillator (inverted)	–(*E+Bcosy*)
56			
57	LOG	Positive log of local oscillator plus bias	{+log}(*E+Bcosy*)
58	Link from 50	Inverted log of dc bias	{–log}(*E*)
59	ADD	Log of LO plus bias divided by dc bias	{+log}(1+*B/Ecosy*)
60	Link from 51	Log of signal plus bias divided by dc bias	{+log}(1+*A/Ecosx*)
61	ADD	Sum and negate log outputs, ie multiply	{–log}((1+*A/Ecosx*).(1+B/E.cos y)) = {–log}(1+*A/Ecosx*+ *B/E.cosy+AB/E/E.cosx.cosy*)
62	Link from 40	Positive log of dc bias	{+log}(*E*)
63	ADD	Multiply result by *E* (negated)	{+log}((*E+Acosx+Bcosy*+ *A.B/Ecosx.cosy*)
64			
65	ANT	Antilog of multiplication	–(*E+Acosx+Bcosy*+ *A.B/Ecosx.cosy*)
66			

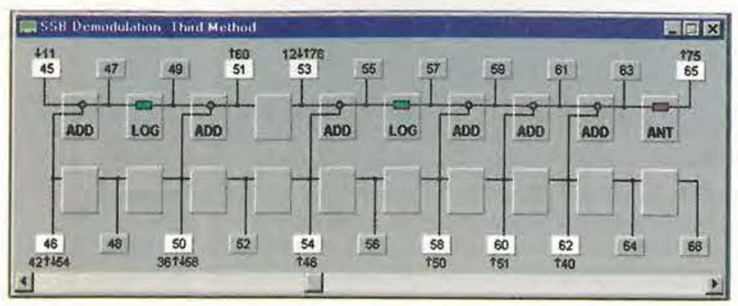


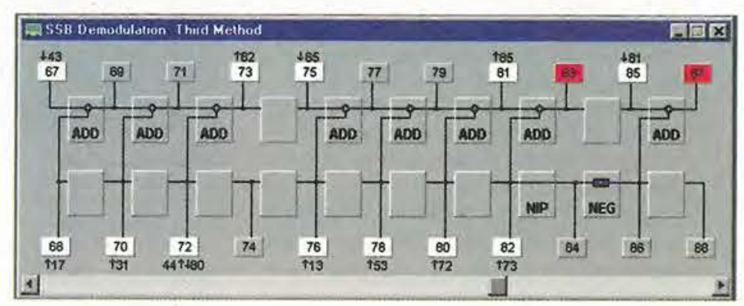


Displays of the ssb modulator/demodulator produced by the TRAC simulator, inputs on the left, outputs on the right. Pin 1 is the higher frequency curve on the left and pin 2 is the lower. On the right, pin 83 is the lower frequency and pin 87 the higher.









Each of the four TRAC devices needed to implement the ssb modulator and demodulator. Numbering on the pins indicates which device is which.

More on TRAC

These comments have been added by David Winch of Fast Analog Solutions to help you get to grips with Mike's design more easily.

Using TRAC's LOG and ANT functions to multiply. In theory, you can multiply by adding logs and then taking the antilog. This is also true with TRAC, *with provisos.*

The transfer function of the LOG cell is,

$$V_{out}=0.07474375(\log_{10}(V_{in})+10)$$

Here the logarithm base is shown as 10, but any other base is applicable, if you adjust the constants.

For the ANT cell, the transfer function is,

$$V_{out}=10^{(V_{in}/0.07474375)-10}$$

Again, any log base is applicable. The ADD cell is self explanatory. The operational limit of the LOG cell is $|V_{in}|<1.4V$ while that of the ANT cell is $0.1V<|V_{in}|<0.8V$. Equally importantly, the operational limit of the ADD cell is $|V_{out}|<1.4V$.

For inputs from 0.1V to 1.0V, most usable logs are in the range 0.6V to 0.75V, so adding two together would take the signal outside the operational limit of the ANT function. This can be corrected by 'dividing by one', i.e. subtracting the log of 1.0V.

However, if the two V_{in} signals exceed about 0.25V, their logs will exceed 0.7V and so their sum will be outside the operational limit of the ADD function. To remove this possibility, the 'dividing by one' must be done before the ADD.

When multiplying more than two inputs together, this 'dividing by one' must be performed after all but the final input.

Using the LOG and ANT functions to multiply, divide and raise to powers. Some people have told us they have experienced difficulties using TRAC to multiply and divide by adding and subtracting logarithms. Let me try to make things clearer.

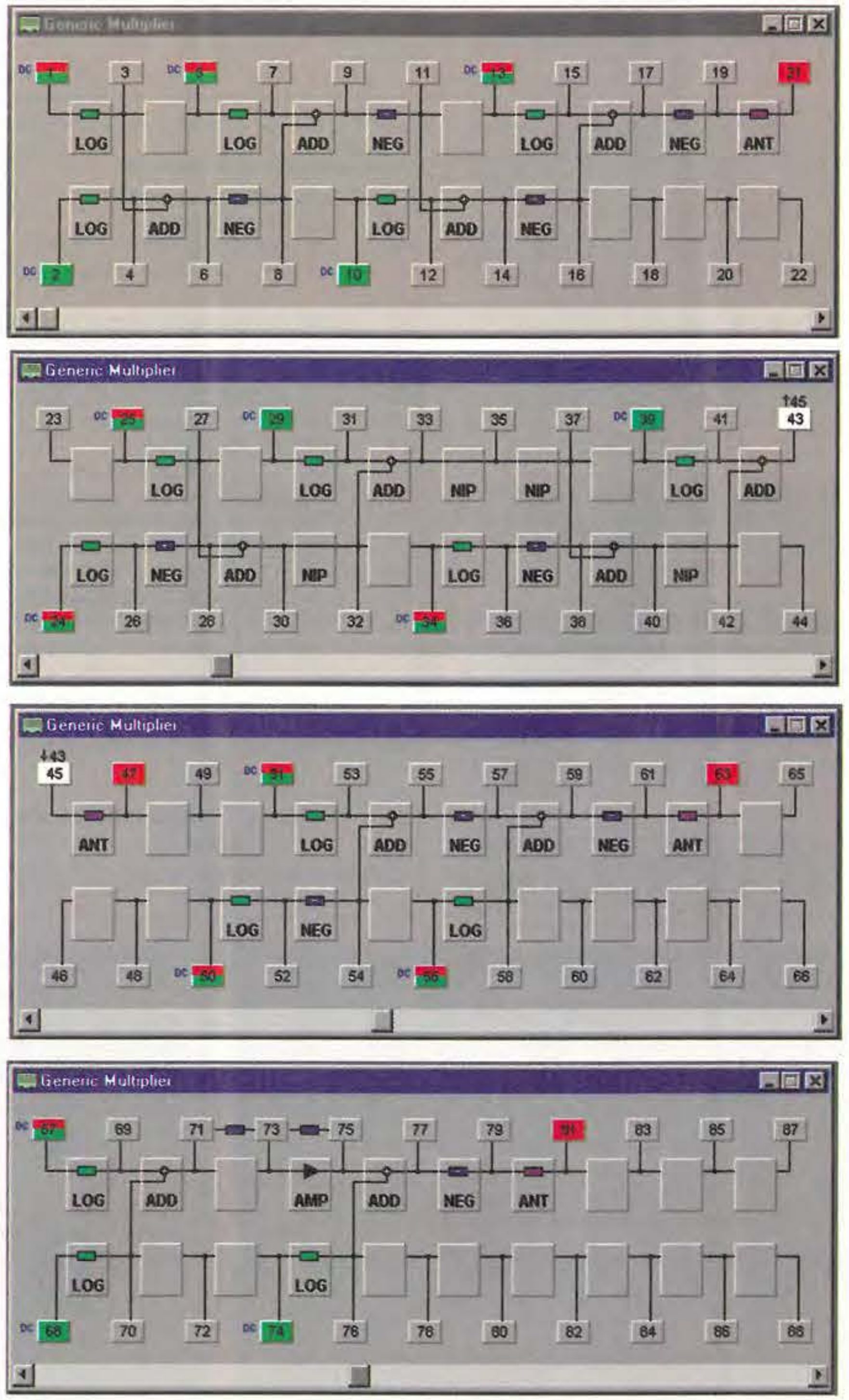
If you have used logs to base ten, or natural logs, to multiply numbers you may not have realised how much of a fortunate coincidence it is that the log of 1 to any base is zero. If this doesn't make sense, let me put it another way. What would you expect the answer to be if you multiplied 0.5 volts by 0.4 volts? Did you say 0.2 volts? Now, what should the answer be if you multiply 500mV by 400mV? Should it be 200 000mV or µV?

Now do you start to see what happens when we multiply quantities rather than numbers? It depends on what you decide 'one' is!

Deciding what 'one' is means working *relative* to a fixed point. When multiplying *numbers* we work relative to 1, so using their logs we work relative to 0. So effectively we do nothing, or more probably we don't even realise we are doing nothing.

So what's different with TRAC? Well nothing actually. Using the TRAC LOG cell, the log of 1V is approximately 750mV, and the log function has a gain of approximately 75mV per decade, or approximately 23mV per octave.

The log function is not to any particular base. It just obeys the rule that multiplying the input by a constant changes the output by a constant amount, no matter what the original input. The TRAC antilog cell follows the inverse of the same curve.



Studying these implementations should help you grasp the idea of TRAC's log function not being to any particular base – if you haven't already of course. From the top, they are a generic multiplier, a generic divider, an alternate 'multiply-and-divide' and 'raise-to-a-positive-power' circuits respectively. Note that the generic divider overflows into the third TRAC chip by one element.

So the log of 0.5V is approximately 727mV and the log of 0.4V is approximately 720mV. The sum of these is 1447mV but the antilog of 1447mV would never get to the 2000 megavolts indicated by the mathematics. We have not worked relative to what 'one' is. If 1V is 'one' then we need to calculate,

{(log(0.5V)–log(1V)}+
{log(0.4V)–log(1V)}+log(1V)

Putting in quantities we get,

{727mV–750mV}+
{720mV–750mV}+750mV

Or in other words,

–23mV–30mV+750mV=697mV

and the antilog of 697mV is 0.2V. And if you think about it, this is what you do with numbers. Only that you work relative to log of 'one' is zero.

How does this apply to practical TRAC designs? Well, if you are multiplying you would need to subtract the log of 1V from each multiplicand and then add it back in once at the end. Practically you would just not subtract from the last multiplicand.

When dividing, you would need to subtract from every term and put it back in once at the end as well, but in this case you would practically add the log of 1V to every divisor but not the numerator at the beginning.

Raising to a power be it positive of negative, greater or less than unity, the same applies. Subtract the log of 1V, do the processing and add the log back in once again.

A further practical problem is raised by the ADD cell of the TRAC. Its output saturates at approximately 1.4 volts. So you need to be careful exactly where in the design you subtract the log of 1V.

In multiplication, subtraction needs to happen before the log of the next multiplicand is added; in division, addition needs to happen after the log of the next divisor has been subtracted; and the same general principles apply to raising to a power.

If you think back to when you used a slide-rule, you'll remember that the easiest calculations were the ones with alternate multiplication and division. The same is true using TRAC. If you alternately subtract and add logs, and if there is one more multiplicand than divisor, the need to 'divide by one' is removed and the ADD function never saturates.

The simple TRAC designs shown here should clarify the situation even further.

Multiplying by E gives,

$$E+A\sin x+B\sin y+AB\frac{\sin x\sin y}{E} \tag{13}$$

Subtracting equations 4 and 5 from above gives,

$$E+AB\frac{\sin x\sin y}{E} \tag{14}$$

Subtracting E and expanding gives,

$$\frac{AB}{2E}\left(\cos(x-y)+\cos(x+y)\right) \tag{15}$$

Performing the same functions on equations 6 and 7 gives,

$$\frac{AB}{2E}\left(\cos(x-y)-\cos(x+y)\right) \tag{16}$$

Adding or subtracting equations 15 and 16 gives,

$$\frac{AB}{E}\cos(x-y) \tag{17}$$

Thus the sum or difference frequency components are thus obtained.

In summary

Note that the differentiation function is used to obtain a $\pi/4$ phase shift (sin to cos function) and the amplitude of this function's output is frequency sensitive. The signal to be demodulated must, therefore, not vary in frequency by a significant amount or the amplitude of the cosine function will vary causing distortion on the output.

At the final 470kHz intermediate frequency of most receivers, this should not be a problem as the audio signal in the range 300Hz to 3.4kHz gives only a 0.7% deviation.

Provision of an ssb modulator using this method is not possible because the frequency deviation – and hence the amplitude deviation – of the $\pi/4$ phase-shifted output of an audio signal is too large.

Another look at the theory shows that if an audio signal $A\sin x$ is mixed with another constant frequency of $B\sin y$ and $B\cos y$ the resulting functions will be,

$$A\sin xB\sin y=AB\frac{\cos(x-y)+\cos(x+y)}{2}$$

$$A\sin xB\cos y=AB\frac{\sin(x-y)+\sin(x+y)}{2}$$

By filtering out the sum or difference component of the two equations, the sine and cosine functions of the audio frequency are available. Provided the frequency offset provided by the frequency y is taken into account then the functions thus obtained can be used as inputs to the *TRAC* functions given by this design.

Investigation of the data sheets show that the available bandwidth for low level signals is 4MHz. It should, therefore, be possible to perform a direct conversion on the 160m amateur band (1.8MHz) and possibly on 80 metres (3.5MHz). ■

Pin functions of TRAC device 4.

Pin	Connection	Description	Function
67	Link from 43	Result of 'sin' multiplication	*E+Asinx+Bsiny +AB/Esinx.siny)*
68	Link from 17	Sig inverted	*–Asinx*
69	ADD	Subtract Sig from result	*–(E+Bsiny+AB/Esinx.siny*
70	Link from 31	Local osc	*Bsiny*
71	ADD	Subtract local osc from result	*–(E+AB/Esinx.siny)*
72	Link from 44	Negative dc bias	*–E*
73	ADD	Subtract dc bias from result. Sum of sum and difference frequencies obtained.	*AB/Esinx.siny =AB/E(cos(x–y)+cos(x+y))*
74			
75	Link from 65	Result of cosine multiplication	*E+Acosx+Bcosy+ AB/Ecosx.cosy*
76	Link from 13	Cosine of Sig inverted	*–Acosx*
77	ADD	Subtract cosine of Sig from result	*–(E+Bcosy+AB/Ecosx.cosy)*
78	Link from 53	Cosine of local oscillator	*Bcosy*
79	ADD	Subtract cosine of local oscillator from result	*–(E+AB/Ecosx.cosy)*
80	Link from 72	Negative dc bias	*–E*
81	ADD	Subtract dc bias from result The difference of the sum & difference frequencies obtained.	*AB/Ecosx.cosy =AB/E(cos(x–y)–cos(x+y))*
82	Link from 73	The sum of the sum & difference frequencies.	*AB/E(cos(x–y)+cos(x+y))*
83	ADD	Add the sum and difference to obtain the difference component of the signal and LO – **the required result**.	*AB/Ecos(x–y)*
84	NIP		
85	Link from 81	Difference of sum and difference frequencies	*AB/E(cos(x–y)–cos(x+y))*
86	NEG	Invert	*–AB/E(cos(x–y)–cos(x+y))*
87	ADD	Subtract the sum and difference to obtain the sum component of the signal and LO – **the required result.**	*AB/Ecos(x+y)*
88			

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The multi-national company that designs the technology behind some of the biggest names in mobile communications is undergoing phenomenal growth and is looking for the most highly motivated, practical and methodical PCB layout engineer to help complete their new Systems Team. If you are from a background which has given you a broad, yet detailed knowledge of communications technology, maybe as a test engineer, or as an apprentice and are looking to broaden your horizons and stretch your skills, then you must **send your CV to Max Pritchard now.** Experience of PADS would be useful, but is not essential.

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M3, M4 **To £37,000**

Join the successful and expanding development arm of possibly the best known mobile communications company in Europe. They have over 75,000 employees world-wide and have activities in more than 100 countries. This site specialises in the development of cutting edge mobile communications products, which include Digital Cellular (GSM), Personal Communications (PCM), Personal Communications Systems (PCS), RLL and Satellite Communications. Due to continued investment in R&D they currently require software engineers at all levels up to Principal Engineer. Communicaitons experience is not essential, but is highly desirable. Essential skills are C software development in a real-time environment and an understanding of structured methods. Desirable skills include the following:

* Mobile Communications
* Case Tools
* GSM / UNITS / DECT / CT2
* DSP

In return for this expertise you will receive excellent renumeration, ongoing training to maintain your skills at the cutting edge of the industry and the benefits of working in a development environment with a high degree of involvement, autonomy and support. **Contact Toby Walker.**

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M4 Corridor **To £40,000 + Benefits**

This widely acknowleged and recognised telecommunications specialist supports the global telecommunications industry with innovative products and services. A Handset Specialist is required. With an experience of over 5 years you will have a thorough knowledge of GSM / CDMA network design and have experience with either GSM / CDMA or handset manufacture in the area of GSM / CDMA handset testing for feature set functionality and performance testing. Although based in Bristol the position will require you to spend time in the field. You will therefore by required to be able to start quickly and work independently almost immediately. In return you will achieve an exciting career opportunity whilst receiving an excellent salary plus pension, annual bonus and share option, medical insurance and where appropriate relocation. **Call Nick Dixon.**

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Berks **c£30,000**

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M3 / M4 Corridor **To £40,000**

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Highly successful manufacturer of electromechanical defence equipment require a production engineer. HNC / Degree qualified in Electronic Engineering, with an understanding of the latest PCB manufacturing techniques. You will be able to lead by example undertaking assembly/test work to solve problems, provide on line support to production and review and improve on the electrical and electronic aspects of the manufacturing process. **Contact Alan Jupp.**

RF TEST TECHNICIAN

Buckinghamshire **To £25,000**

This company is a well respected RF products and systems manufacturer. They require an RF Test Technician to optimise and test the company's Mast Head Amplifiers. This will mean ensuring the receiving and transmitting frequencies conform to required customer specification (1 to 2 Ghz). You will need to be aware of the individual and team production targets as defined in the production plan and to consistency meet those targets. Part of your work will be to carry out the practical diagnosis following specific diagnostic practice in a logicial and competent manner to ensure swift resolution. To do this you will need to recognise the importance of consistent quality in all aspects of the job. You will work as a member of the production team ensuring sound communication within the department and flag up problems which might affect the team's overall performance. If you have the ability, experience (network analysers) and drive for this demanding role, **call Nick Dixon.**

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Southampton **To £35,000**

This company is a fab-less semiconductor company and respected design consultancy. Due to expanding growth and opportunities within the semiconductor division, they are looking for an engineer with good experience of CMOS full custom IC design. You will have more than 3 years' experience in a variety of IC design, mature customer contact whilst maintaining good self-motivation. BiCMOS skills would be useful, but not essential. You can grow with this company whilst benefiting for an excellent salary package, relocation, healthcare and a pleasant science park location in a flexible small company culture. Be part of an exciting future, **call Nick Dixon today.**

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Middlesex **To £30,000**

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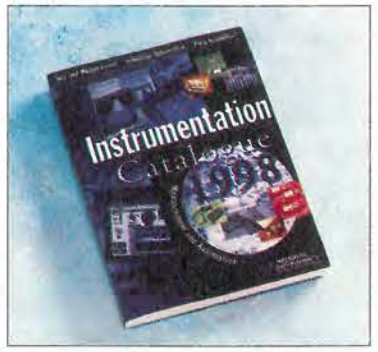

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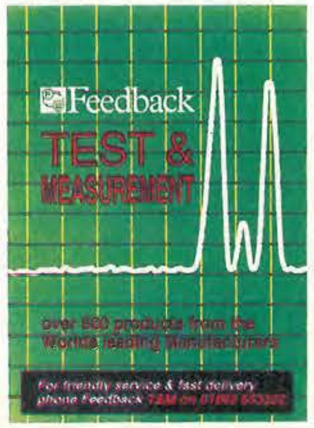

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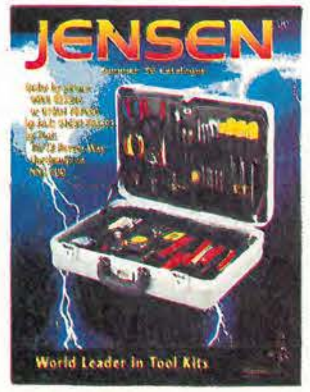

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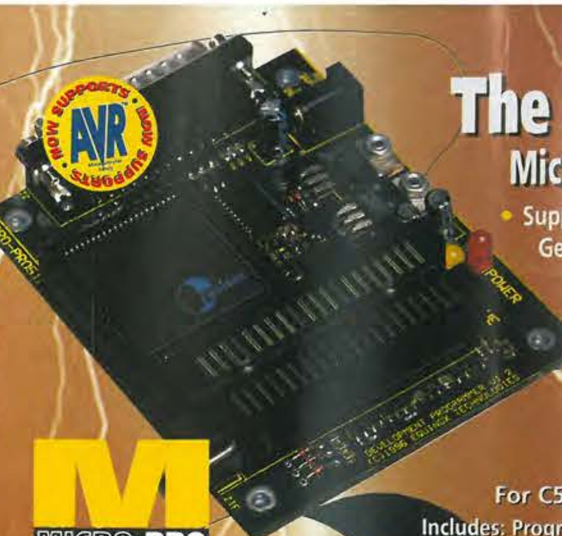

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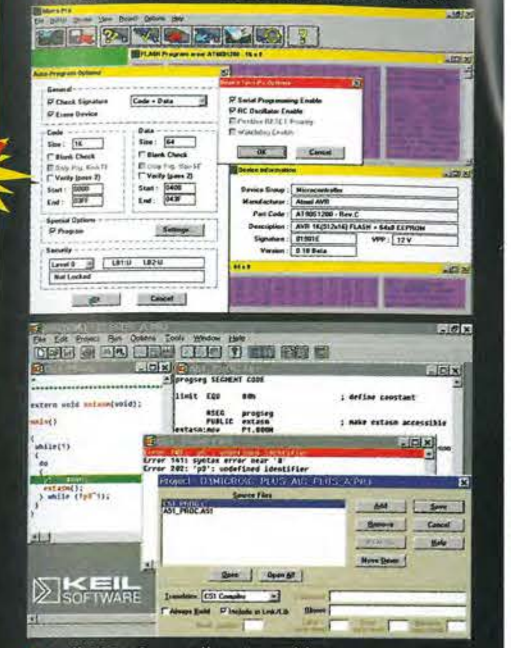

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