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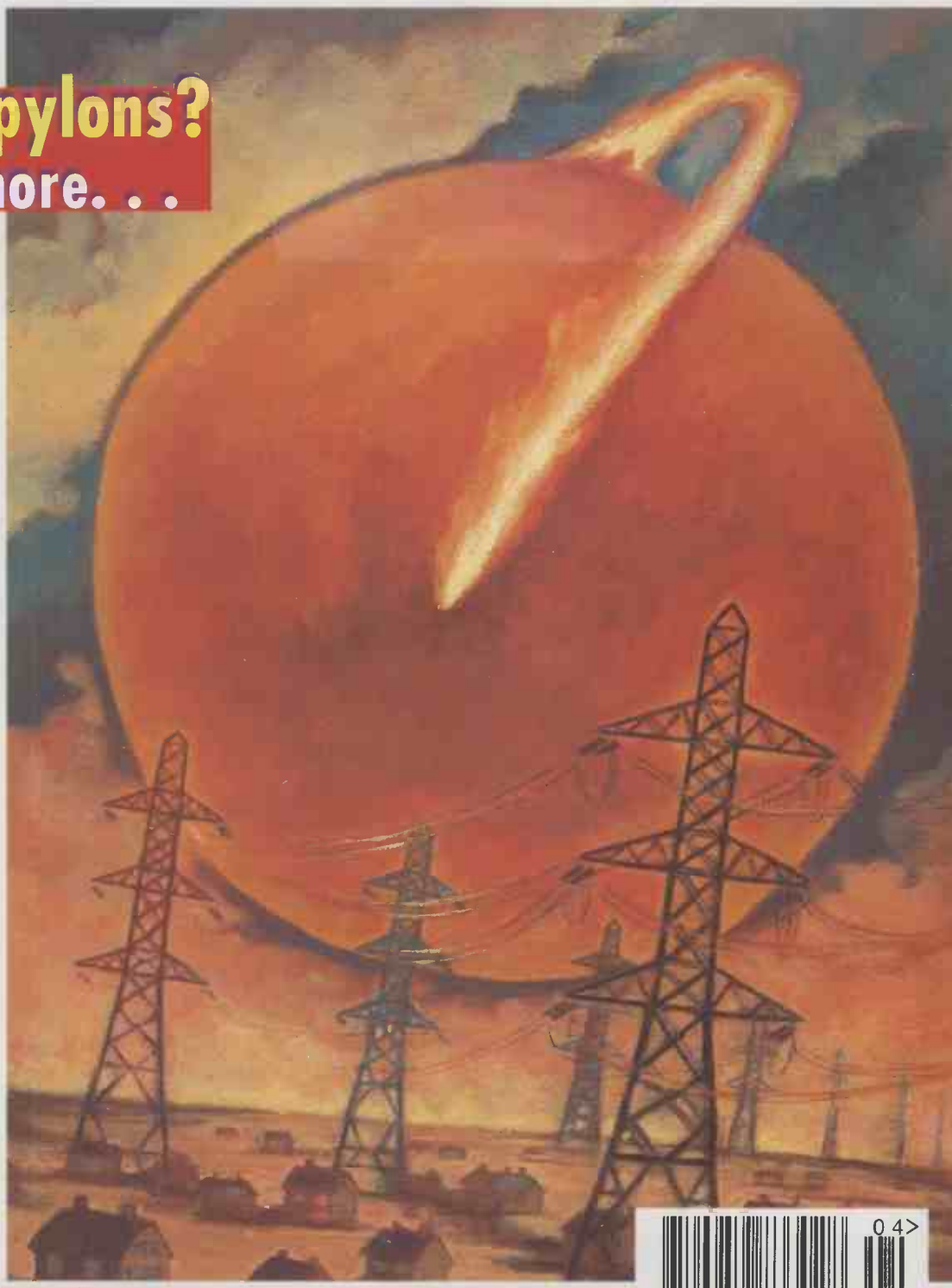

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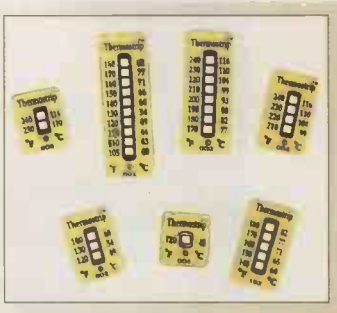

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REED BUSINESS PUBLISHING

In the interest of the customer

It used to follow that what was good for the customer was good for profits. The customer must always come first. It is therefore surprising that the consumer software industry has a such bad record in this respect.

Of course, in any technology driven industry, such as computer software, customers have to be led somewhat. They have to be persuaded to give up their favourite DOS programme, for example, and move to something 'bigger and better', and in the long run it often really is a good move, in terms of speed and flexibility.

But that's not the real problem. What we have to contend with, quite frankly, is poor quality, over-selling and indifferent after sales service. The customer comes a poor second.

There's also little regard for the customer's equipment. Sloppy development results in applications requiring larger than necessary amounts of computer resources. Do programmes really need to be that big? Do we really have to have loads of ram? The software is delivered on a huge pile of disks, or increasingly on CD-rom. But the time is surely not far away when the setup instructions will read 'Place CD-rom #1 into drive and press enter'.

Maybe part of the problem is the impressive software development tools available today such as Visual Basic, Delphi and Visual C++. They enable suprisingly fast development of new products, but this tends to bring about a false sense of confidence. Prototypes can be up and running in hours and lots of features can be bolted on. But the more features and facilities that a programme has the more meticulous the testing has to be. Development tools can have bugs as well! Inadequate test methodology often results in uncertain interaction between applications. There is surely an analytical way of predicting how applications inter-react.

There have been cases recently when clearly product had been released before it was ready. Every industry is subject to commercial pressures, none more so than software. But shipment of immature product can cause misery. For example, a recently marketed operating system did not contain all the device drivers it needed for Soundblaster and some CD-rom drives. The 'Plug and Play' feature became a nightmare. One punter, I heard of, tried to load the software from CD-rom. Half way though it stopped because it didn't recognise the CD-ROM drive. It left him in a total state of limbo that took days to sort out. I somehow don't think that he was alone.

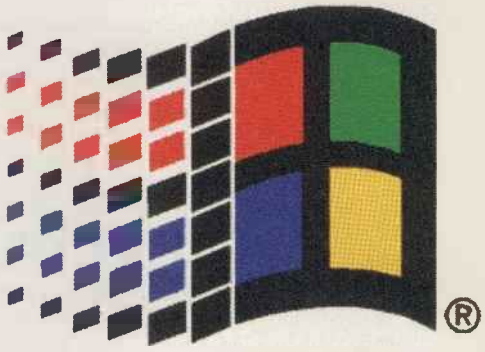
Some products are hyped to a dangerously high level, raising customer expectations, only to have them dashed later. Of course, the software world is highly competitive and fast moving. Millions of dollars can be made overnight with the right break. Recent examples are Netscape and the UK company who wrote some software that would bar child access to dubious parts of the Internet.

I've heard it said that the marketing costs for any software package start at around half a million pounds. It's hard to do it for less, which makes it high risk. But looking at it from the user point of view, we need to know whether the programme really will run on a 386 with 4 Meg of ram, for example, and what applications will it not work with? We should not have to rely on the software press to tell us these things.

Now a gripe about customer support. How often have you heard from a support line "We know about the bug, there are no real workarounds, but it will be corrected in the next version". And how long do you have to wait for an answer? Furthermore, companies who used to have free call facilities on 0800 are now migrating to the more lucrative 0898 lines at the customer's expense.

Coupled with this, companies only usually give 'Limited Warranties' with their software packages. These warranties cover the cost of the floppy disks and maybe the original cost of the software but little else. There is little or no liability if it doesn't work to your satisfaction. It would be interesting if this situation could be tested in court to see if customer's 'statutory rights' were being upheld – it would probably uncover a can of worms! Maybe the answer is some sort of code of practice whereby customers could obtain bug-fixes free of charge by mail or download for at least a year after purchase.

Any improvements in quality and customer service will inevitably cost money and companies will try pass it on to their customers in some way. But I think that it is a price worth paying. Software represents a large investment



for individuals and companies alike, and we are becoming more and more dependent on it.

Quite soon software will be available that will take decisions for us, called 'software agents'. What if they don't work properly! I think that there is still a 'start-up' and 'get rich quick' mentality in the software business. After all it's one of those few industries that even today can be started in the garage or spare bedroom. The focus is firmly on developing product as quickly as possible and getting it out of the door before anyone else does the same. Support does not really feature much.

However, the software industry has come a long way, and the lead needs to be taken by the large companies to improve customer service and set an example. Maybe survival will depend on it one day. Quality and customer service issues are not as glamourous as the technology, but they need attention – ***now***. ■

Peter Marlow

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Mosfets enhance video compression

Mosfets used directly as calculators could simplify video compression systems following work at the Defence Research Agency, DRA, in Malvern.

DRA has used a twin floating gate mosfet circuit as a vector quantiser to calculate the Euclidean distance between two points. The floating gate device, fabricated using standard foundry processes, exploits a characteristic that is comparable to the Euclidean distance metric.

Gillian Marshall, a member of the research team said: "With standard analogue systems feeding a digital signal processor (DSP), there is a large bottleneck at the analogue to digital converter. The new vector quantiser does all the calculations in analogue, only converting the final compressed data to digital for transmission."

The benefits claimed for the approach include a computation rate 20 times that of typical digital signal processors and a power consumption that is less than one-tenth.

Applications that could benefit from the approach include video conferencing where large amounts of analogue information is transmitted down telephone lines, and cost sensitive systems where a fast A/D converter is too costly.

The scheme exploits the fact that current through the fet is proportional to the square of the difference between the gate voltage and the threshold voltage. In turn, the Euclidean distance squared equals the square of the difference between an input point and a reference point. Hence, if the input point, represented as a voltage, is applied to the gate and the reference is the threshold, the distance measure is proportional to the device's current.

Various parts of the VQ have been constructed by the research team, and have worked well. However a full scale system will have to wait for further funding.

Chaos keeps communications secure

Chaos theory promises the ultimate in secure communications, enabling systems to emit signals indistinguishable from background noise.

Researchers at the University of Birmingham's school of electronic and electrical engineering have developed a communications system that chaotically encodes a digital data stream. At the same time, it hides the signal within a noise-like structure. This is desirable especially for military applications where the 'enemy' would not even know communications are taking place.

Dr Jim Edwards, leading the research said: "Encoded signals may look like noise, but are in fact deterministic if both the structure of the encoder and the initial conditions are known. Being short of one or both of these makes prediction difficult."

He further pointed out that traditional 'secure' systems are not in fact because enough information is available for signal reconstruction.

The chaos system offers enhanced security since the initial conditions must be known exactly. Any slight difference and the system quickly diverges. This is comparable with the chaos theory example that says weather cannot be predicted without knowing all the starting conditions which may include a butterfly's wings beating in Australia.

The claimed bit error rate (BER) of the current system is 1 in 10,000 at a signal-to-noise ratio of 10dB. The University is working on a system where an acceptable BER is obtained for negative signal-to-noise; in other words, the noise has more power than the signal. This would give truly undetectable communications.

Synchronising the transmitter and receiver, critical with chaotic systems, is not a problem according to Edwards: "Because the system is digital, it tends to self-synchronise."

Richard Ball
Electronics Weekly

Plasma displays for wall mount tvs. The first plasma displays suitable for use in tvs will be mass-produced by Fujitsu from October at an initial $5000 price tag.

The displays are the world's only 42in plasma panels available commercially. Although the company has had 21in displays available for two years, they are considerably more expensive than crts and are not used by tv makers.

At 42in, however, the screens are bigger than crts and, naturally much thinner. Fujitsu's panel is only 75mm thick, allowing a tv to be hung on the wall. The company is currently supplying panel samples to tv manufacturers, including Thomson, Nokia, Philips and Bang and Oluffsen in Europe.

Unlike thin-film transistor alternatives, plasma displays have a wide viewing angle and are therefore useful for public information displays as well as tvs.

3D graphics add-on for pcs

VideoLogic will be selling this summer a £300 add-on card that brings 3D picture realism to pcs. At the heart of the boards will be a 3D graphics processor which the UK company has developed in partnership with NEC.

The two companies have adopted an approach to 3D rendering which reduces the high speed synchronous dynamic ram buffer memory requirement, and removes a fundamental memory bandwidth bottleneck.

"The consequence of no longer requiring z-buffer memory can result in a $30 to $60 saving in s-d-ram cost," said Trevor Wing, VideoLogic's group marketing director.

In the PowerVR 3D rendering architecture, VideoLogic has removed the need for storing picture depth information in a z-buffer. Instead, it implements in real-time the necessary hidden surface calculations. According to Wing this is possible because the design uses an array of 32 processor elements which can operate on each pixel independently.

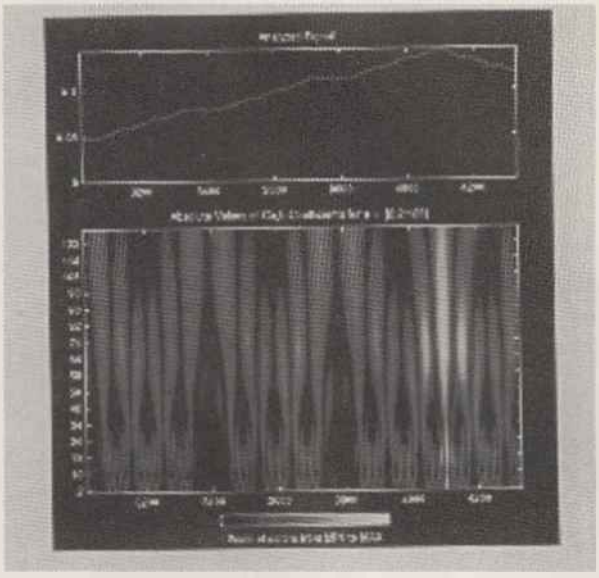
The NEC chips will hit the market at the same time as another UK-developed 3D graphics processor, the Permedia from 3DLabs.

NEC has integrated the complete hidden surface and polygon texturing functions into two devices which require just 2Mbyte of synchronous d-ram buffer memory. NEC will start sampling the first chips this quarter and a single chip version for the pc market will be available by the summer.

In contrast, 3DLabs' Permedia chip incorporates a 16-bit z-buffer, and also has an overall s-d-ram requirement of 2Mbyte. It is targeted at $300 pc add-on card designs.

The first pcs to incorporate the NEC chip set will be launched next year, according to Wing, who added: "Two of the big three pc suppliers are already evaluating the 3D chips." VideoLogic has existing video partnerships with IBM and Compaq Computer.

Richard Wilson
Electronics Weekly



MathWorks has announced a toolbox claimed to be the first product to make wavelet analysis a practical engineering tool.

In these photographs, wavelets have been used to compress and decompress fingerprint data with little degradation, and fractal signals are decomposed with various scales (stretches) of wavelet.

Wavelet transforms are an up-and-coming technique for data compression and analysis. They transform signals into a sum of small, overlapping waves and are claimed to be more effective for analysing non-continuous waveforms than traditional Fourier methods.

In addition to supporting advanced applications, the Wavelet Toolbox is said to offer engineers unfamiliar with wavelets an easy way to try out the transforms on their own problems.

Slow but less volatile growth for semiconductors

The semiconductor market will grow more slowly, but with reduced volatility. So says Sergio Vicari, European application specific product manager of Texas Instruments. For Vicari the main source of growth will be computer sales: "The increasing electronic content of products, as well as emerging markets will also contribute."

He predicts that computer sales will rise from just under 100 million units in 1996 to some 200 million by 2000. The increasing semiconductor content is extrapolated from the trends over the past two decades. His figures show growth of 4.8 per cent per year in the eighties rising to 6.3 percent per year in the nineties.

What of the total semiconductor market? Vicari said: "It depends on the industry growth. Fifteen per cent per year will mean the market is $275bn by 2000, 20 percent will take it to $350bn."

Digital signal processing chips is one of TI's core businesses and these come under Vicari's wing. "The driving force for dsp sales comes from wireless communication products and hard disc drives, but high efficiency motor controllers are likely to become a large sector." Electric motors currently consume 50 percent of the world's electricity. dsp based controllers will double their efficiency, Green pressures and lifetime costing are making them popular in new installations.

"The DSP market grew 68 percent to $1.6bn in 1995 and I predict this will increase to $9.4bn by 2000. During this time the unit cost of a DSP will drop from $12.2 to $8.8."

However, he admits the difficulty in predicting the DSP market: "The 1994 prediction for 2000 was between $6bn and $7bn."

Experts to pick design language

The protracted process of defining analogue circuit modelling extensions to the VHDL digital system design language reaches a crucial point this week. A panel of experts will be presented with two competing proposals for a language specification.

The IEEE 1076.1 language design committee is to ask independent experts and users to choose between the Jade language, championed by Mentor Graphics' subsidiary Anacad, and the Opal alternative, supported by Analogy, Cadence and Compass Design.

The experts will make a choice by the end of March with a full Language Reference Manual, LRM, to follow by July. An IEEE ballot on the LRM could then be completed in the following six months.

Andy Patterson, Analogy's European technical director, said most arguments appeared to be supporting Opal and he was hopeful a firm choice would be made on schedule. "The committee is being spurred by the analogue Verilog efforts with Verilog-A having been published this month," he said.

Pentium Pro flaw

In the same way that the famous Pentium flaw was first brought to public attention by an academic – Professor Nicely – another professor from San Francisco State University has pointed out a flaw associated with the Pentium's successor – Pentium Pro.

Intel conceded last week that it had not responded properly to the professor but claims that the 'few complaints' it has received result from incorrect use.

The reported problems arise when the Orion chip-set is used with the Pentium Pro microprocessor in server applications.

According to Intel they only arise when certain add-on cards – which are not recommended for use by Intel – are used in the application.

It was claimed that Pentium Pro servers made using the Orion chip-sets were resulting in systems that operated at half the speed of previous generation Pentiums.

Intel concedes that the add-on cards can cause problems with the Pentium Pro/Orion combination resulting in sluggish performance but says there is nothing wrong with Orion and that it is not being re-engineered to speed up performance.

However, the company intends, later this year, to launch a new chip-set for use in Pentium Pro-based servers.

Non-slewing

Giovanni's article 'Non-slewing power amplifier' in the March issue contained a couple of minor inaccuracies. In Fig. 1, there should be no 200Ω resistor in the right-hand CSA circuit. In Fig. 4, the unmarked resistor is 3.3kΩ. Apologies.

Campaign for anti-theft chips

In response to the fastest growing area of crime in the UK, the magazine *Computer Weekly* has begun an anti-chip theft campaign bringing together the police, chip makers, insurance firms and computer manufacturers and buyers.

The idea of the scheme is threefold: to show computer owners how to secure their equipment, to lobby chip and module makers to mark their products, and promote anti-theft techniques.

The valuable parts of a computer are the simms and, to a lesser extent, the cpu. The police have already produced advice to computer owners to assist them in securing their property.

The real breakthrough will come with simms that are tagged or become unusable away from their host.

Metropolitan police commissioner Sir Paul Condon said: "I truly believe that if consumer goods can be designed and manufactured so that they are useless to anyone other than the owner, then we could bring about a complete reversal of the figures."

Marking, tagging or putting intelligence onto the simm pcb would seem to be a waste of time as mobile phone thieves already 're-chip' their swag. This involves removing the identification prom from the phone and replacing it with one holding another identity. There is therefore no reason to believe that simm thieves could not transfer chips to new pcbs.

The need is for memory chip makers to incorporate some form of security device into the chips, but this seems unlikely until the voice of the user becomes impossible to ignore.

Steve Bush, *Electronics Weekly*

1800MHz access for cellular carriers

Cellular operators Vodafone and Cellnet have succeeded in gaining access to radio frequencies in the 1800MHz band – a move seen as crucial in their battle with newer operators Orange and Mercury One-2-One.

"This is important to us and we intend to use any spectrum for new products and areas (of coverage)," said a spokesman for Vodafone.

As well as reserving two 10MHz blocks in the 1800MHz band for possible allocation to Orange and Mercury at the end of 1997, the government intends to make two further 11.5MHz blocks available to Vodafone and Mercury. This will be first access to the relatively under-populated 1800MHz band for Vodafone and Cellnet which depend on the increasingly congested spectrum below 900MHz for their analogue and digital GSM services.

"The government wants to set out a strategy for a fair allocation of spectrum on the basis of need between all four mobile phone operators," said science and technology minister Ian Taylor.

Vodafone and Cellnet intend to move all users from their older and cheaper analogue networks to digital services by the year 2005. This will lead to greater congestion in the digital 900MHz bands, as four out of five UK mobile phone users are connected to analogue networks.

As well as seeking proposals for new use of the 1800MHz band the government will also make additional frequencies in the 900MHz band available to the two operators.

Managing the move from analogue to digital is the biggest challenge for Vodafone and Cellnet who, like all operators, are facing falling profitability, according to market researcher CIT.

RW, *Electronics Weekly*

In Brief

Interactive traffic information

Japan is to launch the world's first on-line interactive traffic information service that uses telephone lines in April this year. Dubbed Advanced Traffic Information Service (ATIS), the system will supply information to pcs and in-car units via land lines and cellular links.

Power pc off the desktop at IBM

In a review of the future of PowerPC, IBM is reported to have decided to de-prioritise the microprocessor as a cpu for desktop personal computers.

Instead, IBM is said to be concentrating on Intel's x86 for desktop pcs and is focusing its PowerPC effort on workstation and server applications and as an embedded microcontroller.

EMC testing backlog

EMC test houses are heavily oversubscribed – many up to six months in advance, now that the EMC directive is in force.

ERA Technology's civil test facility in Leatherhead is currently booked until August and SGS in Durham is full until July and both are working three shifts per day. Test slots are booked on a first come/first serve basis.

Any company committed to using test houses to CE mark their products, and expend their time slot with incomplete tests or a failed product, may find themselves out in the cold.

Windows for hand-helds

Several major computer and telecommunications companies are planning to introduce hand-held computer devices based on a secret operating system under development at Microsoft.

The operating system, code-named Pegasus, is Microsoft's third attempt to develop a small operating system based on Windows for use in handheld computers and smart telecommunications devices.

Microsoft is expected to unveil Pegasus by the middle of this year. ■

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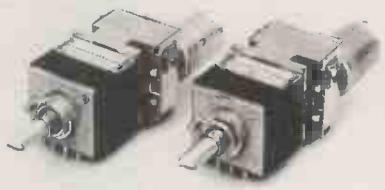

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RESEARCH NOTES

Jonathan Campbull

Signalling a rethink of array receiver design

Innovative signal processing techniques developed by researchers at the University of Southern California are set to turbo-charge the performance of conventional array signal-receivers. The work carried out by Jerry Mendel and former USC doctoral student Mithat Dogan could well affect everything from how the military locates far-away submarines, to how we track objects in space, to how we design more efficient home antennas receiving signals from direct satellite broadcasts.

The invention works by combining 'higher-order statistics' with correlations of readings from adjacent detectors in an array. By returning to the fundamentals of physics and signal processing, the researchers have shown that, using the known geometry of an actual array, it is possible to compute correlations not only between pairs of physically present detectors, but also between a physically present detector and a non-existent, virtual detector. Or even between pairs of virtual detectors. This gives a small array of detectors a much larger scope.

Arrays are an attractive alternative to large dishes for picking up very faint signals. The method is to use a multitude of detectors, each one served by a much smaller radiation collector, and spread these detectors out in an array, either in a line, or in two or three dimensions.

Arrays can cover an area or volume far larger than any possible dish, and though the signal picked up by each detector is faint, engineers can construct the network so that the faint traces received by each individual site reinforce one another, creating an instrument that can perform like a single massive dish.

Mendel and Dogan's software, called a virtual cross-correlation computer, works only if the distance to the signal-producing target is large in relation to the size of the array of detectors. The detectors in the array must be tuned to a relatively narrow bandwidth too – listening to only a limited range of sound pitches, light colours or radio frequencies.

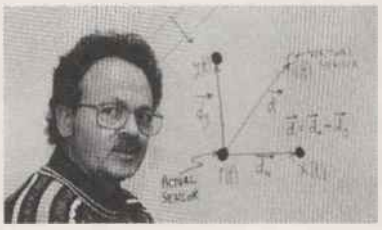
Finally, the signal being received must be of a specific kind – namely, non-Gaussian. The Mendel-Dogan invention functions to suppress Gaussian signals and preserve non-Gaussian ones.

If these three conditions are met – and they frequently are in real-world sonar, radar and other array detectors – major improvements in performance are possible, according to Mendel. More targets can be located than before, closely spaced targets can be resolved, and interfering noise can be suppressed.

As well as improving the performance of existing array detectors, the virtual cross- correlation computer concept can be used to design new, more efficient antenna arrays. For example, a 20-by-20 planar array, which would normally require 400 elements, can be implemented with a 10-by-10 array using only 100 elements.

"But the technique also has an aesthetic appeal," says Mendel. "It uses the hidden, internal structure of a signal that is unknown, to, in effect, decode itself. It uses the characteristics of the array used to detect this signal to bootstrap the array's efficiency. Even if it ultimately proves to have no uses at all, I find the technique highly satisfying to contemplate."

Better reception from distant sources looks possible with the new approach to array design.



Jerry Mendel at USC has analysed exactly how a receiver array works.

Making photons interact is first step to quantum computer

Physicists at Caltech, Pasadena, have taken a step closer to quantum computing with testing of an optical gate whose output depends on the polarisation state of two photon inputs.

Photons normally do not interact. But the team led by professor of physics H Jeff Kimble at Caltech, has found that they can be made to strongly influence each other when brought together with an atom inside an optical cavity.

To be useful in computing, any legitimate logic gate must display an essential feature called conditional dynamics, where the output must depend upon both inputs. In an optical quantum logic gate, the output state of each photon must depend on the input state of both photons.

Kimble's group has showed strong conditional dynamics for an atom in an optical cavity formed by two highly reflective mirrors, one of which allows partial transmission of light. The scientists sent pairs of photons through the cavity, and investigated the states of the photons when they re-emerged, showing that the output state of each photon depended on the polarisation of both input photons. *Cont'd over...*

In effect, the cavity functioned as a rudimentary logic gate at the single photon level. Changing the photons' polarisation is analogous to flipping the bits in conventional computers.

This is the first demonstration of conditional dynamics at the single-quantum level, and while many complex problems remain to be solved before even primitive networks of quantum logic gates could be built, the result is being seen as a significant first step in quantum computing. Even if it doesn't lead to a practical route to quantum computing, the researchers say optical quantum logic gate will definitely have a role in specialised applications in optical communication.

Not a remote possibility

How we laugh as we remember those days when we used to have to pull ourselves up from our chairs and drag the 3m or across the room to press the channel changer on the tv with a finger. Now we just reach for the remote control and... hang on, I know it's here somewhere.

Unfortunately, as increasing numbers of household devices and even light switches become remote controllable, keeping track of them all is becoming more and more difficult. Universal remote controllers are a great idea – if you have small fingers and a photographic memory for densely packed keyboards. But a researcher in the Department of Electronic Engineering, The Chinese University of Hong Kong, has proposed a solution that could be easy to implement and simple to use.

In the system proposed by C S Choy, a temporary link is established between the remote control and the target appliance. So any further key presses are only recognised by that one device.

Typically, audio-visual system offer many functions sometimes calling for tens of keys on a remote control. Choy proposes dividing these into types, according to the nature of control, such as on/off, +/- volume, and numeric.

The resulting smart universal remove control could use an optimum number of programmable keys to keep its bulk and complexity down. Through a learning process, each key could send out different commands according to the appliance being controlled.

So far Choy has built a remote controller, based on a Motorola 68701 with 2K eprom, 192 bytes built in ram and three i/o ports, which he has used to control light switches.

But the concept could form the basis for a universal controller that is much simpler than anything currently around.

C S Choy is in the Department of Electronic Engineering, The Chinese University of Hong Kong, Hong Kong.

Soft touch brings robot breakthrough

Much work has gone into designing robot grippers that are sensitive to force so that, for example, the robot can pick up an egg and hold it firmly without breaking it. Now two US researchers have found an answer that was easily to hand all the time – robot fingertips.

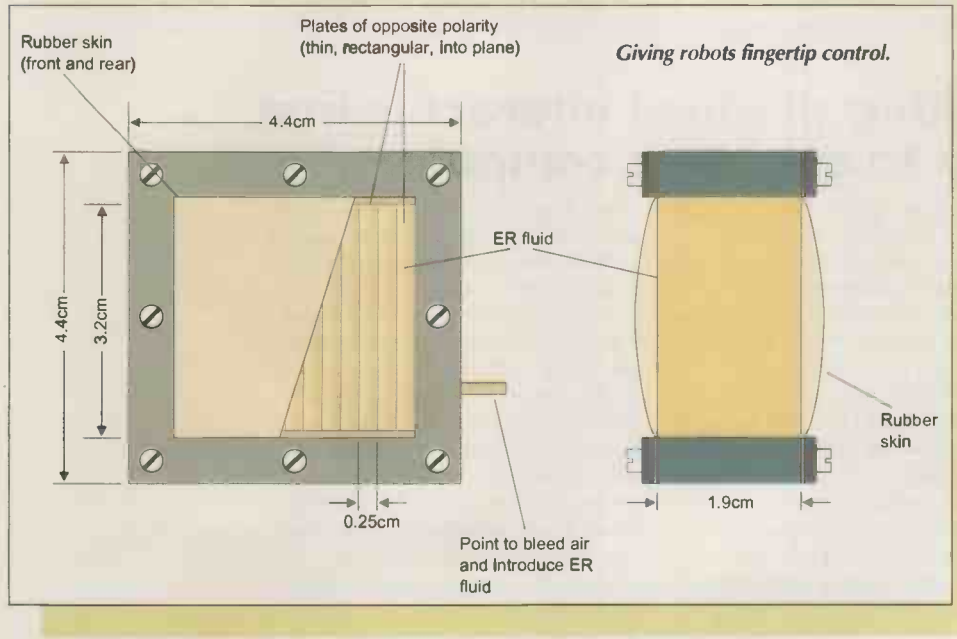
The fingertips are actually an electrorheological fluid of particles of polymers suspended in a dielectric fluid. In the presence of a strong electric field, their behaviour changes from that of a viscous, approximately Newtonian fluid to that of a plastic, with a finite shear strength as well as a viscous coefficient.

Prasad Akella and Mark Cutkosky had observed that the ability of human hands to make contact smoothly is partly due to fingertips that deform and dissipate energy.

Taking this as their starting point, the two workers have now produced their latest prototype fingertip that seems to reproduce that effect in a robot ('Contact transition control with semiactive soft fingertips', *IEEE Transactions on Robotics and Automation*, Vol 11, No 6, pp. 859-867).

The soft fingertip consists of a non-conducting rubber skin containing the fluid, with the electric potential applied across a series of plates oriented perpendicular to the skin surface. As the skin is pressed, er fluid is forced to flow between the plates with a resistance that varies with the applied voltage. A second membrane at the back side of the plates provides a restoring force that returns the system to a standard equilibrium configuration when unloaded.

Building a fingertip whose stiffness and damping properties can be directionally controlled still remains a challenge. Even so, the researchers report that the current generation of fingertips can provide compliance and damping that are very similar to human fingertips.



Giving robots fingertip control.

More information from P N Akella who is now at the Manufacturing Center, General Motors Corporation, 30300 Mound Road, Warren MI 48090, USA or email at akella@gmr.com. The research was carried out in the Department of Mechanical Engineering and the Center for Design Research, Stamford University, Stamford CA 94305, USA.

Magnetism motivates microactuator research

Researchers into microelectro-mechanical systems (Mems) at the Berkeley Sensor & Actuator Center (Bsac) have developed a powerful microactuator that uses magnetism as the actuating force and can be batch-manufactured in relatively simple processes.

Mems specialists Jack Judy, Richard Muller and Hans Zappe at Bsac report that their microactuator has so far demonstrated forces and displacements far larger than those generated by most electrostatic microactuators. In addition the microactuator can be be fabricated using conventional electroplating, lithography, materials and equipment.

Novel features of the technology are that actuation can be controlled by a remote magnet – a hand-held permanent magnet was used in some of the experiments – and that structures can be actuated in three dimensions: ie movement is not restricted to the plane of the wafer.

The microactuator itself is essentially a polysilicon cantilever beam, or flexure, onto which a magnet is formed at the free end. That magnet interacts with an external magnetic field, bending the flexure.

Fabrication is straight-forward in that the magnetic layer of NiFe layer is simply electroplated onto the silicon at the end of a process which is already in use to produce chips of polysilicon resonant structures.

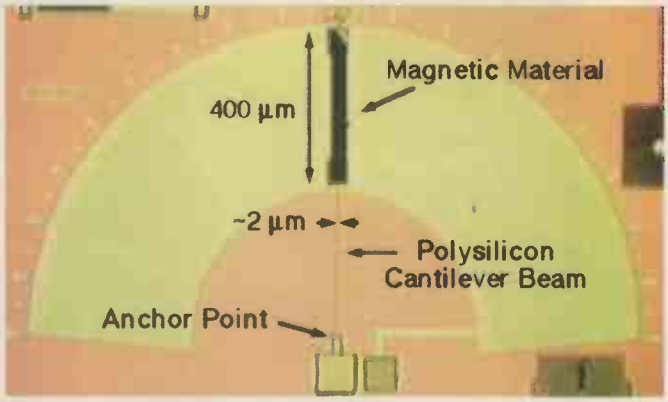
Using an external magnet to provide the actuating force means surface-to-surface interactions such as those found in linear and rotary variable-capacitance, and variable reluctance structures, are not required – so fabrication is easier.

The external magnet can also be used to activate many devices simultaneously – though that also means that control of independent microactuators will require miniaturised sources of magnetic fields, perhaps even on-chip sources.

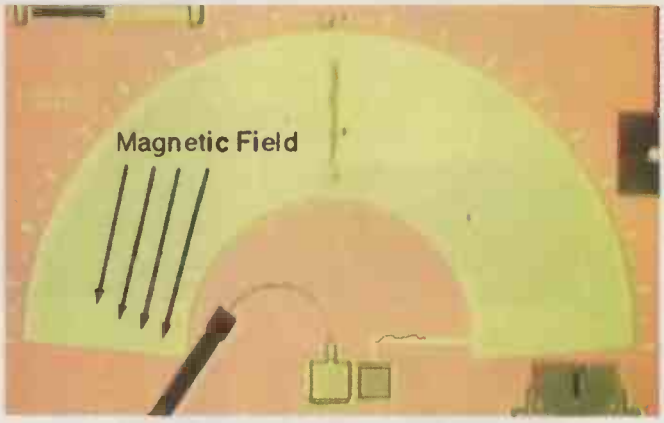
So far the tip of an 800μm-long cantilever has been deflected over a distance of 1.2mm and rotated through an angle greater than 180° under an imposed torque of 0.185nNm ('Magnetic microactuation of polysilicon structures', *Journal of Microelectromechanical Systems*, Vol 4, No 4, pp. 162-169)

The team is hopeful that similarly fabricated magnetically-actuated microstructures might be applied to micromanipulators, microgrippers, magnetometers or microphotonic systems.

Jack W. Judy can be contacted at 497 Cory Hall 2041 Francisco, Apt. #5 Berkeley, CA 94720-1770, USA or j.judy@ieee.org



Before application of the external magnetic field...



...while after the external magnet is applied the beam is deflected. Deflection is not restricted to the plane of the wafer.

Planners get ready for road rage

Road rage seems to be the most extreme example of an ever-increasing aggression on the highways. So how are road planners reconciling their computer models of happy 'model' drivers giving way at junctions with a cheery wave, to the reality of the bumper to bumper stand-offs which increasingly are the norm.

At MIT in the US, they might have an answer. Because MIT engineers have developed a state-of-the-art traffic simulator that actually mimics the behaviour different drivers, aggressive, careless, timid or fast- and how they affect traffic flow.

The traffic simulator, which runs on a workstation, is called Mitsim for short (microscopic traffic simulator) and treats traffic as a set of individual vehicles, or particles, allowing each vehicle to move according to its own characteristics. The more common macroscopic simulator treats traffic like a fluid, assigning one set of characteristics to the entire stream of cars. Mitsim is more lifelike because it allows for differences in vehicles' movements as dictated by drivers' personalities.

As each vehicle enters the simulated road system, it grabs a packet of vehicle characteristics that determines how it will act in certain circumstances. Not only does each vehicle have a size, type, occupancy level and destination, it also has driver characteristics. These include desired speed, propensity to yield to other vehicles, lane-changing behaviour and route decisions. There's even a driver impatience factor that makes each driver's choices more realistic. ■

Professor Denis Henshaw recently proposed that radon gas could be concentrated by high electromagnetic fields from overhead electricity supply lines. Radon causes lung cancer by ingestion of short-range alpha particles, whereas the cancer usually linked to pylons is leukemia, implying particles penetrating much deeper into the body. And if radon is highly significant, shouldn't there be a higher incidence of Leukemia in the West Country, where radon is more prevalent? There isn't. To test for radon, Prof. Henshaw tracked alpha particles. But could other factors have affected his results? Anthony Hopwood presents his case.

POWER LINES particles and CANCER

"The first suggestion that power lines might cause disease was made in 1976..."

In my original feature in the November 1992 issue of *Electronics World*, I proposed that the electric and magnetic fields around power lines intensified natural background radiation in their vicinity.

My observations were based on many hours of measurement of the background ionising radiation close to overhead power lines over the period 1990-91 which happened to be close to the peak of solar cycle 22 sunspot maximum.

This was serendipitous because the effect is most marked at solar maximum, and the rather crude portable Geiger counter I used would not have detected any effect as the sun went off the boil in its approach to the present minimum of its 11 year cycle.

Figure 1 shows the large change in solar activity expressed by the number of sunspots. This can also be measured as the solar flux at 10.7cm wavelength. At maximum, the solar flux approaches 300, whereas it is now around 70 – a change of over 400%.

The first suggestion that power line routing might cause disease in populations close by was made in 1976 from studies in the Denver area – which happens to be a mile above sea level and therefore has less protection from the atmospheric layer. Since then, the debate has continued, but the necessary scientific proof of a credible disease mechanism has been absent.

There have been numerous theories to explain the increasing epidemiological evidence. Most have concentrated on internal cel-

lular effects observed in the presence of alternating magnetic fields, and have involved free radicals, melatonin or chemical changes in the living cell. Other theories have suggested that the electrodynamic fields have damaged cell function by precipitating pollutants from the atmosphere. Some have suggested that electric and magnetic fields per se are damaging, and that a new disease mechanism is implicated.

There is no argument that electric power lines and distribution systems create strong electromagnetic and electrostatic fields in their vicinity. Overhead power lines are a highly visible source of this radiated energy. Some 'supergrid' lines carry up to 800A per phase at 440kV and spread an electrodynamic footprint over 100 meters either side of the centre line.

There is also no argument that charged secondary atomic particles are influenced by ambient electric and magnetic fields. It was the alteration in the numbers of charged particles detected on my continuous cosmic ray monitor by the passage of electrically charged clouds that first gave me the idea of investigating whether the more intense electrodynamic fields round overhead power lines affected natural background radiation nearby.

"Why had no one noticed this (power line) effect before?"

The results were surprising. In simple terms, a horizontal geiger tube with an L/D ratio of about 13 and a low energy cut off at about 60keV showed a background rate increase of up to three times either side of the line, compared with the rate outside the electrodynamic footprint.

Why had no-one noticed this effect before? There are two main reasons.

Firstly, only a third of the charged particle flux detected by the tube came from the sun, whose high energy particles were fissioned by collision with the atmosphere to give the 'cosmic drizzle' of lower energy charged particles reaching the ground. These low energy but still biologically damaging particles have been ignored by cosmic ray physicists because they were only interested in ultra high energy particles which could not be replicated easily on earth.

Secondly, environmental researchers were only concerned with picking up radioactive particles from pollution sources like Chernobyl, and deliberately set their Geiger tubes vertically upwards to minimise the natural background rate variation due to the sun, ground and changes in atmospheric pressure.

The key to my observations was to use a long thin Geiger tube aligned to the geomagnetic field as a coincidence detector to improve the detection statistics for down-coming solar particles against the background radiation from the ground.

"For the first time, I could see the rate change as the sun rose..."

Textbooks suggest that typical rate variation in the UK due to solar emanations is about 3 per cent – a figure confirmed when I first set up the Geiger tube on a 7 metre pole with an east-west axis in June 1989. By October 1989, the rate variation stayed maddeningly around 3 per cent while the sun fulminated at solar maximum.

Turning the tube to a geomagnetic NS axis made a magical difference. For the first time I could see the rate change as the sun rose, and track active areas across the solar disc by the 14 day rate change they produced.

Interaction between charged particles and the geomagnetic field was also apparent during magnetic storms. During the great auroral display of 8/9 November 1991, an individual auroral ray from the geomagnetic zenith passed over my detector and increased the count by about 20 per cent for the few minutes it was focussed on my sensor.

This may have been the first time that an auroral beam of particles has been detected on the ground.

All this – plus the continuous recording of atmospheric electric field alongside particle rate – led me to try and find whether power lines could alter the solar particle rate in their vicinity.

After publication of my results and conclusions in *EW+WW*, a debate started. The Swedish Radiation Protection Institute went out into the Scandinavian winter and found there was a background radiation anomaly near power lines.

Nearer home, the NRPB was more sceptical. It did carry out joint field tests with me with inconclusive results. Some of the tests were flawed because they did not include anticoincidence counting on the multiple tube arrays. The real problem was that between my field work in 1990-91 and their tests in 1993, the solar flux had dropped by 70 percent so the effect was difficult to detect with relatively unsophisticated sensors.

Since then, I have been working to improve my detector, as have the Swedes. Although it is early days, we now have two different types of sensor to plot any radiation anomalies near power lines.

My own instrument uses two closely matched independent Geiger tubes driving separate counters, as well as a coincident pulse monitor. Earlier work had suggested that the change in particle rate near a power line was most marked at the low energy end of the spectrum – below 100keV.

At sunspot minimum, the mix of particles entering the atmosphere still varies with solar activity. The most sensitive ground level indicator of solar particle flux is the geomagnetic field. This is easily measured. Conditions are logged from 'quiet' to 'storm' on a K index published monthly for every three-hour period. Another index of incoming solar plasma is the ionosphere. Its condition can be monitored by recording changes in high-frequency radio propagation from day to day. These two indices, plus the rates from fixed particle counters produce a clear signal when extra particles are entering the atmosphere to suggest when field measurements are best made.

The twin Geiger tube detector now used has two matched tubes with an L/D ratio of about 12:1. One tube has a plastic protective case, and the other has one of copper to give a differential screening effect of about 4:1 at the low energy end of the spectrum.

Under 'quiet' conditions, the two tubes count within 2 per cent over several hours away from a power line. The rate variation between tubes stays within 5 per cent close to the 11kV line crossing my garden under geomagnetically quiet conditions. When there is a geomagnetic disturbance, the balance changes, with a differential rate of at least 10 per cent in favour of the lightly screened tube.

The instrument has only been under test since the beginning of January, and with a quiet sun, there have been no major magnetic storms so the 10 per cent count differential is

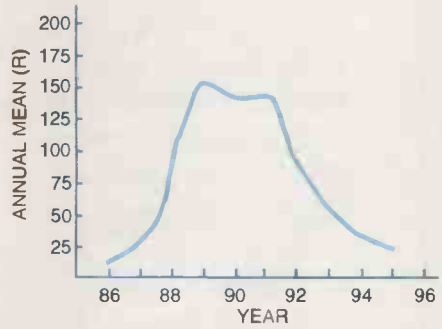


Fig. 1. Large changes in solar activity over an eleven year cycle have been linked to a delayed eleven year cycle for breast cancer.

a reasonable result which can only improve as the new solar cycle gets into gear.

I mentioned earlier some of the other research in this field. Recently a paper was published by Professor Denis Henshaw suggesting that domestic wiring was able to concentrate alpha particle emitters like radon gas in its vicinity. The particles were detected using a sensitised plastic which is pitted by alpha particles, the standard method for detecting radon emanations.

"...it is likely that they did not all come from radon"

Prof Henshaw proposed that the source of his particles was radon gas, which is certainly present in most homes. Given that the tracks were etched by alpha particles close to electric leads, it is equally possible that they did not come exclusively from radon. I suspect that his observations complete the penetrating particle fission chain which starts in the upper atmosphere and which I measured outdoors above 50keV – the low energy cut off for my

"Concorde... routinely reduces height if a solar flare occurs..."

Geiger tube charged particle detectors.

If this proves to be the case, there is a complete chain of potential ionising radiation cell damage from the sun to the wall socket.

So what other evidence is there that the sun can produce sufficient radiation to harm susceptible individuals? The atmosphere is a very effective screen which protects life on earth from the damaging emissions of the sun.

Solar background radiation exposure is already monitored for airline crews. Concorde, which flies higher than other commercial jets, has a solar particle monitor onboard and routinely reduces height if a solar flare occurs or particle rates exceed set limits.

I mentioned that charged solar particles are concentrated at high geomagnetic latitudes, and can be seen as aurorae when the magnetosphere intercepts solar plasma ejected during flares and coronal mass ejections. The geomagnetic intensification effect implies that so-called radiation cancers should be more common in industrial nations at high geomagnetic latitudes.

Cancer statistics from the IARC seem to confirm this, Fig 2.

Further evidence implicating the sun comes from a Russian paper by T.P. Ryabyh and N.B. Bodrova in 1993 outlining a delayed solar cycle for breast cancer in women. Much earlier was the first paper linking power lines and cancer published in Denver USA in 1976. Its significance is that the 'Mile High City' is between 5-6000 feet, where solar background radiation is at least four times that at sea level.

"What is needed now is properly funded research..."

I am also sure it is no accident that the best statistics to date for a link between power lines and cancer come from Scandinavia which is highly electrified and at a high geomagnetic latitude.

What is now needed is properly funded research into the symbiotic reaction between electric power and background radiation using the best radiation metrology. In my opinion, the intensification of natural background radiation by the electric and magnetic fields associated with electrical installations provides the missing link between human cell damage and eventual disease in some people living and working under the aegis of the pylon.

The evidence is mounting, and won't go away. ■

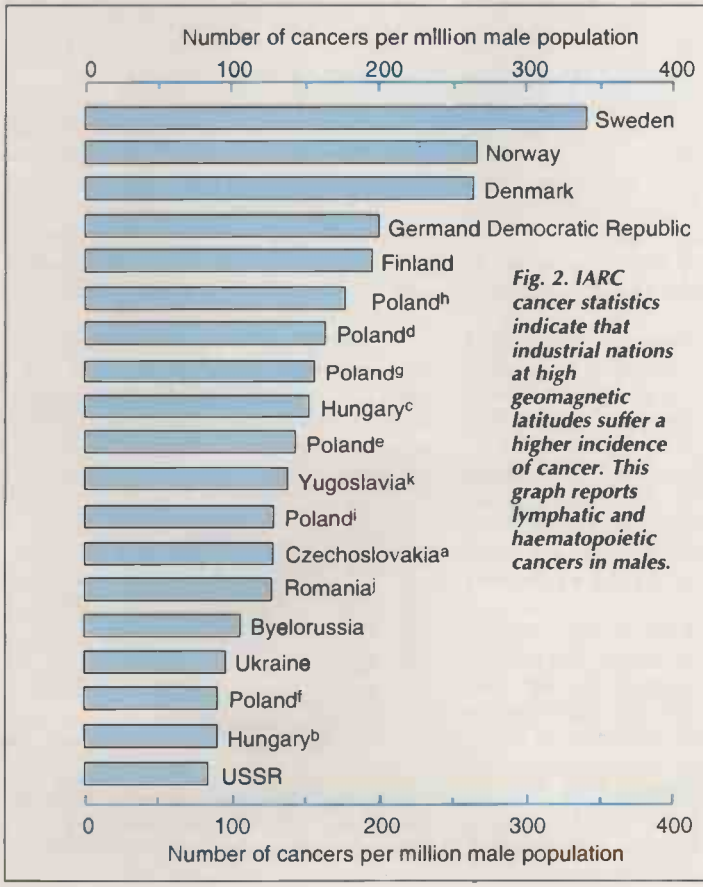
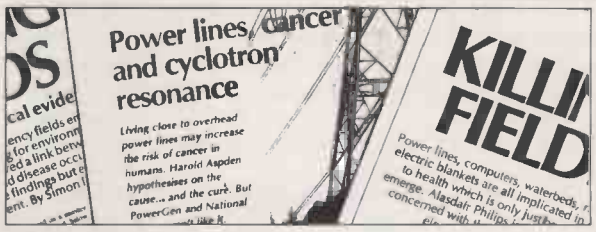


Fig. 2. IARC cancer statistics indicate that industrial nations at high geomagnetic latitudes suffer a higher incidence of cancer. This graph reports lymphatic and haematopoietic cancers in males.



More information

Photocopies of earlier articles on non-ionising radiation published in *Electronics World* are available from SoftCopy for £7.50 fully inclusive. Totalling 25 pages, these A4 copies comprise five articles from the Killing Fields series covering: introduction, biophysics, epidemiology, microwaves, politics and causes. Also included is Anthony Hopwood's 1992 article – 'Radiation focused by power lines'. Send postal order or cheque payable to SoftCopy to 1 Vineries Close, Cheltenham, Gloucester GL53 0NU.

GRAPHIC DISPLAYS

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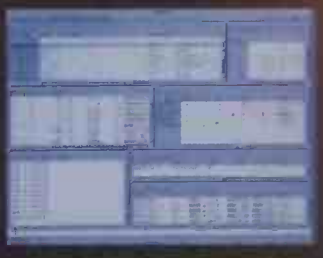

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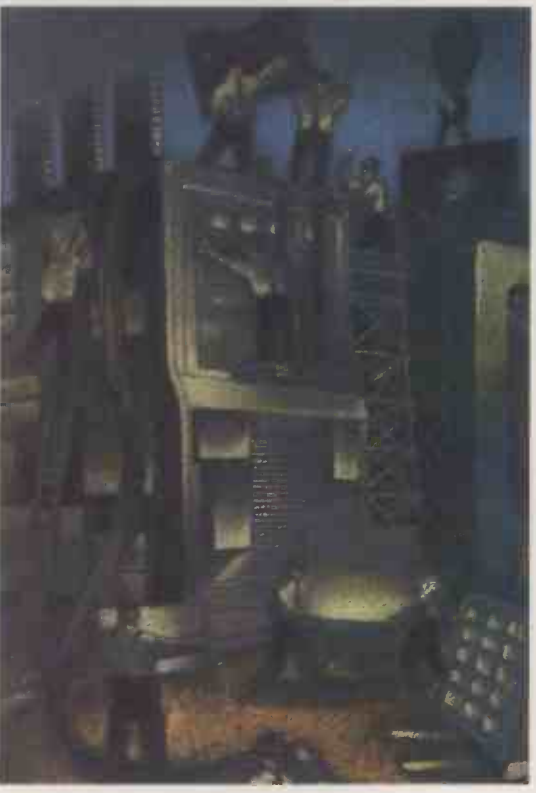
22bit analogue-in for PCs



Simon Bateson and Andrew Woodward run through the design stages needed for achieving very high resolution analogue-to-digital conversion via a PC's LPT port – and at a relatively low cost.

A large number of analogue-to-digital converter designs have been published in electronics journals, either as free-standing units or incorporated into other test and measurement equipment. These have mostly been of 8-bit resolution, based on devices such as the *ZN425*, which can achieve sampling rates suitable for audio.

For even moderate measurement quality, 12-bit converters are needed, such as the *AD1674* which can achieve 10µs conversion at around £35. Higher resolution and higher speeds are generally very expensive; the 'audio optimised' 16 and 18bit converters have good linearity but less good dc characteristics.

Manufacturers provide cards to fit inside pcs for analogue input and output, but it is very difficult, hence expensive, to obtain optimal performance in the electrically noisy environment of a computer.

There are many applications where high resolution is required but speed is uncritical. Here, the 'voltmeter' a-to-d converters are appropriate and ICs like the *7106* and *7135* have provided excellent performance for many years.

Recently, the development of low speed, high resolution converters has moved forward and some devices offer extremely high sensitivity, resolution and self-calibration facilities. Analog Devices' *AD7710AN* is a sigma-delta analogue to digital converter with an on-chip programmable gain amplifier. Given a suitable environment, this device can achieve 22bit resolution – the equivalent of 0.25ppm.

In addition, the *7710* can provide total rejection of superimposed periodic interference and better than 16 bit non-linearity at over 15 samples per second. This sampling rate is adequate for many process and experimental uses, making the converter relevant for mechanical, thermal and chemical sensors, panel meter applications and research.

Resolutions of the a-to-d converter versus sampling rates, at gains of 1-4.

Rate (Hz)	Resolution (bits)
10	21.5
25	20
50	19.5
100	18.5
250	15.5
500	13
1000	11

Noise rejection

It is well known that integrating converters such as the voltage-to-frequency, delta-sigma and dual-slope types have the ability to reject periodic noise. They do this because the output is proportional to the average, integrated, input voltage over the measurement period.

If the measurement period is a multiple of the local supply waveform period, the converter rejects this frequency and its harmonics. For this reason, the dual-slope integrating converters used in ordinary digital panel meters all run at a similar speed, giving about three readings per second and rejecting both 50Hz and 60Hz interference.

The clock frequency must be an exact multiple of the line frequency or cancellation will be incomplete and errors will appear as before. A point that is often overlooked in the implementation of integrating converters is that they rely *totally* on the short-term stability of the clock oscillator.

Crystal and *LC* oscillators fulfil all practical requirements, but c-mos inverter and other *RC* oscillators must be carefully designed for low short-term drift and phase noise.

Noise pickup in the form of 50/60Hz interference is very common in high-impedance sensors. Among these are ion-selective electrodes, clinical electrodes and piezo transducers as well as in low-level industrial sensors such as strain gauges and katharometers.

Noise is induced capacitively in high-impedance transducers and magnetically in low impedance circuits. Applying comput-

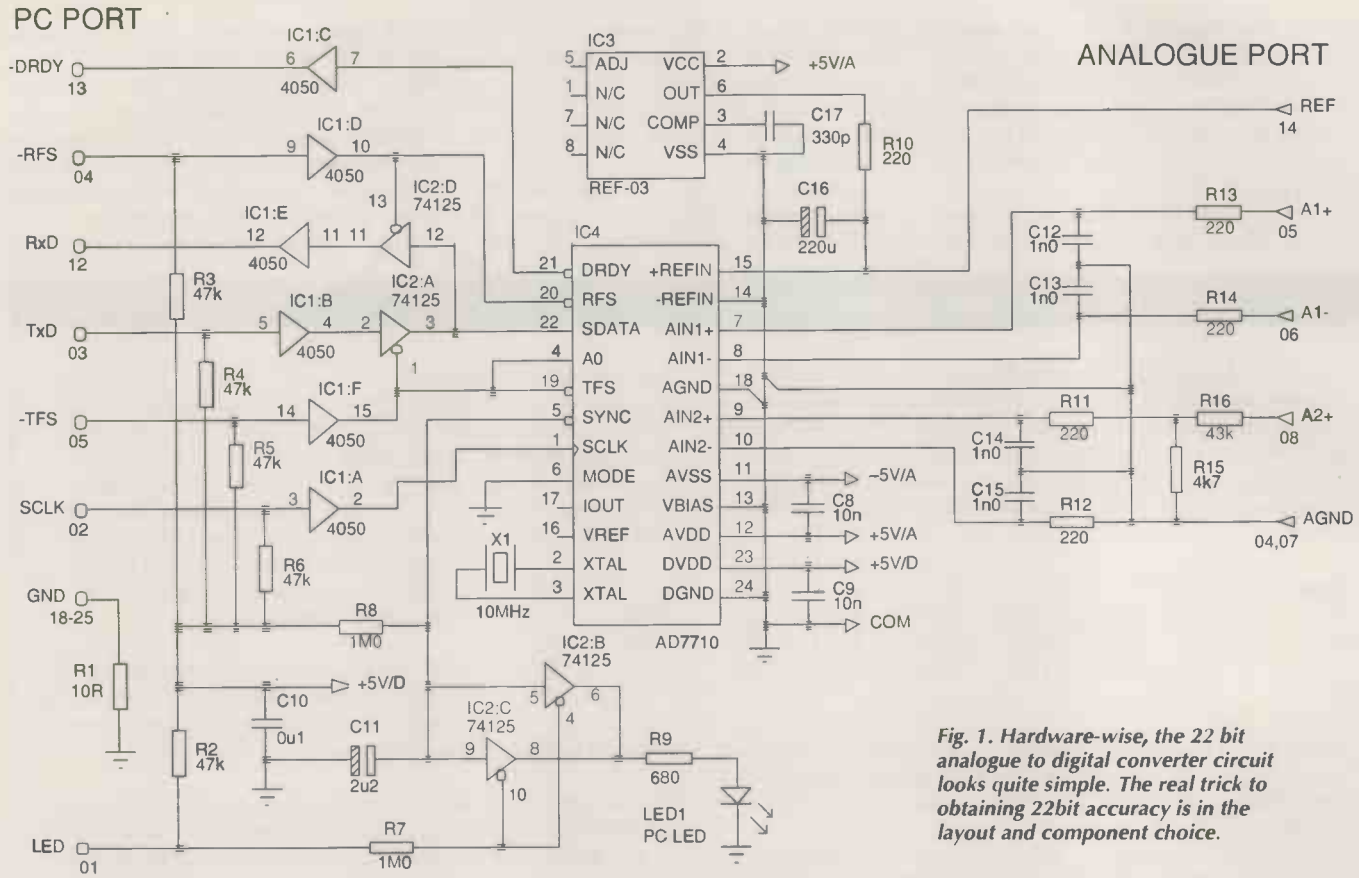


Fig. 1. Hardware-wise, the 22 bit analogue to digital converter circuit looks quite simple. The real trick to obtaining 22bit accuracy is in the layout and component choice.

erised data-collection systems in industrial plant or laboratory environment implies the interconnection of numerous mains-powered devices. Errors in input layout and grounding procedures can cause further problems and the resulting earth loop interference can be difficult to eliminate.

The successive approximation converter – the most common type found in pc cards – is not inherently differential or able to reject cyclic noise. Although pickup can be reduced by standard techniques such as balanced transmission and filtering, once the signal contains cyclic noise, the only really effective converter is one with inherent ac rejection.

This article details the design of a practical implementation of the *7710* converter which connects conveniently to a standard Centronic printer port. The necessary control software for a pc is also detailed. For readers interested in constructing the unit, pcbs are available, as is a fully featured Windows controller detailed below.

Measurement principles

This design, **Fig. 1**, uses an external *REF-03* reference for the maximum stability and minimal noise. The converter is a 'sigma-delta' or '1-bit' converter. It comprises a differential amplifier, an integrator and a comparator, **Fig. 2**.

The system is a negative-feedback loop which tries to keep the net integrator charge at zero. It does this by balancing charge injected by the input voltage with charge removed by alternately applied positive and negative reference voltages.

When the analogue input voltage is zero, the only charge source is via the switched reference voltages. Assuming ideal components, the resulting duty cycle of the modulator switch will be 50%. Changes in input voltage cause linearly proportional variations in duty cycle. In the *AD7710AN*, an on-chip digital filter derives a rolling average of the modulator duty cycle.

An on-chip microcontroller allows software control of sampling frequency. The more clock periods available for the filter to calculate an average from, the closer to the true input the result will be. Consequently the converter gives its lowest noise and best resolution at low conversion speeds.

It is important to realise that, due to this averaging effect, a sudden change in input will not be reflected in an instantaneous output change. At a sampling speed of 12 readings per second, the effective bandwidth is about 3Hz. However, the inherent noise of normal signal sources means that faster measurements would be meaningless at the voltage levels measurable with this converter – the individual readings would differ significantly due to noise and would need averaging anyway.

An additional facility of the converter is a programmable gain amplifier, pga, providing seven software programmable gains from 1 to 128. It is not really an amplifier, but uses multi-sampling to achieve the same effect. Consequently it is extremely accurate.

Converter resolution and noise

Resolution of the converter is calculated by finding the standard deviation of a number of readings. For signals with a mean value of

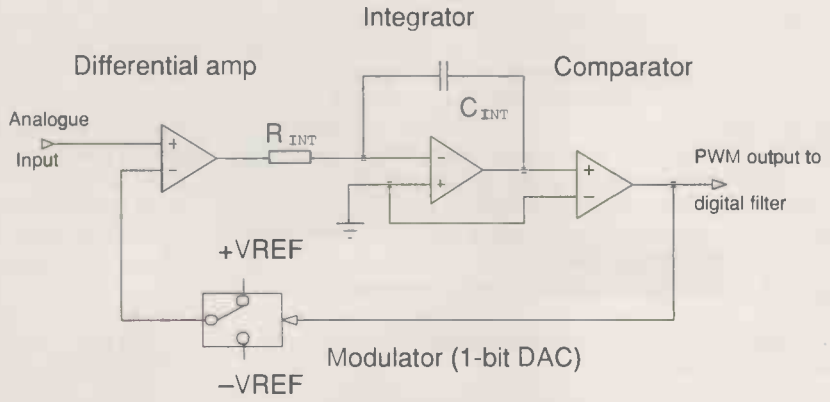


Fig. 2. Sigma-delta a-to-d converter principles. Theoretically, the '1bit' output can produce any desired resolution.

zero this value equals the rms 'noise' amplitude. The available dynamic range is then defined as the ratio of full scale deflection to rms noise. This was found in the production prototype to be around 132dB or 22 bits at a unity gain, worsening to 105dB or about 17.5 bits at a gain of 128.

A converter capable of a practical resolution of around 22 bits must be built up with considerable attention to sources of digitally induced noise. At a pga gain of 128 and with a typical reference voltage of 2.5V, one bit corresponds to just over 2nV and it is possible to reliably measure to 0.2μV without external pre-amplification.

Circuit-board layout – in particular the power supply earthing sequence – is critical, which is why the pcb is being made available.

The digital filter has a $(\sin x/x)^3$ response, **Fig. 3**, and rejects noise frequencies lying within the notches. For the greatest possible rejection, it is possible to retune the filter periodically, under microprocessor control, to track the mains frequency. However, for most requirements, the fixed notch frequencies suggested below are more than adequate.

When correctly tuned, the converter will

Converter configuration and programming

For a full explanation of the facilities of the AD7710 family readers are referred to the Analog Devices data sheets. The chip incorporates a microcontroller which programs the digital filter and operates various mode switches in the converter. It is programmed with a 24 bit control word which must be sent completely, msb first, and which I have split into three bytes:

Byte	MSB	-	-	-	-	-	-	LSB
1	MD2	MD1	MD0	G2	G1	G0	CH	PD
2	WL	RO	BO	B/U	FS11	FS10	FS9	FS8
3	FS7	FS6	FS5	FS4	FS3	FS2	FS1	FS0

MD2, 1 and 0 set the calibration mode. Normally, on power-up and after calibration these read 000 and the device is in normal operating mode. The other modes of interest to us are as follows:

Bit pattern 001 instigates self calibration. The input selected by CH is shorted to analogue ground internally, a conversion run and the result stored as a zero offset coefficient. The input is then connected to V_{ref} internally, converted and stored as a full scale coefficient. Calibration is then complete and the microcontroller uses these values when translating converter values for transmission. Due to thermal effects and contact potentials, there is a residual offset of a few microvolts between internal and external 'shorted inputs'. This is not important when the converter is used with a pc application program since extra coefficients can be saved by the program to remove the offset error. However, if perfect raw data is needed or if external signal conditioning is performed with drift-prone equipment, the 'system calibration' options are preferable. These are initiated by sending control words with the following mode bits:

Bit pattern 010 causes a system offset calibration. This would be sent after the external system input had been zeroed, for instance by grounding with a reed relay. The conversion result is stored as a zero offset coefficient and the converter returns to normal operation.

Bit pattern 011 causes a system full scale calibration. This would be sent after the external system input had been set to full scale, for instance by connecting to a reference voltage with a reed relay. The conversion result is stored as a full scale coefficient and the calibration is complete. Clearly any sort of signal source or conditioning can be included in this loop. In a transmission photometer, for example, zero could be a closed shutter and full scale, direct lamp illumination. By repeating system calibration every minute or so, long term drift of any component is eliminated. External system switching is facilitated in this board design by the output Darlingtons which can operate relays as needed.

G2, 1 and 0 set the PGA gain in a binary sequence, from 1 to 128 – the default at power-on.

CH is the channel select, 0 = channel 1 which is the default at power-on. It should be appreciated that although the converter has two physical inputs it is not practical to use both continuously at full speed, nor is it sensible to use an analogue input multiplexer. The converter core needs to settle after a step change in input, unlike a successive approximation type, and this takes four measurement periods. If many inputs must be measured it is best to use several converters concurrently and read their outputs in turn.

PD sets the power-down mode which stops conversion to reduce power consumption but retains calibration coefficients. It defaults to normal operation (0) at power-on.

WL controls the output data word length and defaults to 16 bits (the most significant, of course) at power on. When set to 1, all 24 bits are transmitted though the last few bits are normally noise.

R0 switches a 20μA current source on pin 17 and is not used in this design.

BO switches a 100nA current source to A1+ input which would typically be used to detect whether a low resistance sensor such as a thermocouple had burnt out and become open circuit.

B/U sets bipolar or unipolar mode, defaulting to 0 (bipolar). It does not alter the converter analogue section at all, just the output coding which is binary in unipolar and offset binary in bipolar. With a +2.5V reference, the unipolar differential input range is 0 to +2.5V; in bipolar it is –2.5 to +2.5V.

FS11-FS0 is a 12-bit value which must lie between 19 and 2000 and which defaults at power-on to 325. The 10MHz master clock is divided by this value and then by 512 to define the converter update frequency. Suitable decimal values for FS are 1562 (12.5 readings / second) and 1302 (15 readings per second). The filter notches occur at multiples of the update rate so 1562 will reject 12.5Hz, 25Hz, 37.5Hz, 50Hz etc., while 1302 is appropriate for 60Hz rejection. The ac response of the converter depends on the first notch frequency, such that the -3dB frequency is 0.262 × the first notch frequency. Hence with a 12.5Hz update rate, the useable bandwidth is dc to 3.3Hz. Note that noise increases with update rate and pga gain, reducing the dynamic range of the converter. The best combination of sensitivity and dynamic range occurs for a pga gain of 4 and an update rate of 12.5Hz. In combination with instructions to set bipolar mode, perform a self-calibration and select channel 1 as input, the resulting control word is 0010 1000 1000 0110 0001 1010 or, in hex, 28 86 1A.

completely reject mains frequency interference to greater than 150dB. However it cannot accept noise peaks far outside its common mode range without suffering modulator overload and consequent non-linear intermodulation. If high amplitude spikes do appear on the signal, some simple analogue filtering will also be needed.

Interfacing to a PC

Data communication with the *AD7710* is via a serial input/output pin, but several extra lines are needed to control data flow. For this reason, and since speed is not important, we found it most convenient to use a pc printer port with manually programmed serial communication. Analog Devices recommends that all digital lines to and from the converter are buffered. This was found to be essential, both to reduce noise-inducing transient currents from the converter and to prevent latchup if the data lines go high before the converter is powered.

High voltage *4050/4049* buffers are required. If ordinary c-mos gates or buffers are used, current passes through the input protection diodes to the supply rail which then powers up the digital side of the *7710* and sends it into scr latch-up. You will notice that spare gates in the *74HC125* are used to drive a front panel led – not functional, just something to flash. Latch-up-inducing input current here is simply limited by a large resistor.

Naturally, the pc is not the only possible host; the prototype was used with an *8052*-based single-board microcontroller. The programming instructions shown in the Basic listing should make application to other systems quite easy. **Table 1** is a list of pin functions as used by the pc and by the converter in this design.

Converter input impedance

The converter's programmable gain amplifier is a useful inclusion. However, it is important to understand that the input current taken increases at high gains as multiple sample are taken by the integrating capacitor. Hence the input impedance decreases and this can induce loading errors.

In many applications, the loading error will be constant and can be calibrated out of existence. However, when the source impedance varies with output as is the case with some deflection bridge circuits, the variation in loading error will induce non-linearity.

Integrated circuits are incapable of being produced to high levels of absolute accuracy, so the exact input impedance cannot be quoted. It is about 720kΩ at unity gain, 360kΩ at a gain of two and reaches a minimum of 90kΩ at gains of eight and above.

Where high input impedance is important, for instance, pH electrodes, ionisation detectors and electrometers we recommend the use of an external buffer amplifier such as the *AD549* 'electrometer buffer amplifier'. This exhibits an extreme input impedance of $10^{15}\Omega$ and which can be incorporated into the self-calibration loop as discussed below in order to eliminate drift.

Table 1: Connections between the Centronics port and 7710 converter.

D25 connector	Centronics function	PC 8255 register line	Converter function	PC port Inverts!
1	/Strobe	C0	Red LED	*
2	DB0	D0	SCLK	
3	DB1	D1	TxDATA	
4	DB2	D2	/RFS	
5	DB3	D3	/TFS	
6	DB4	D4	Ext1	
7	DB5	D5	Ext2	
8	DB6	D6	Ext3	
9	DB7	D7	Ext4	
10	/ACK in	S6	ExtIn	
11	BUSY in	S7	grounded	*
12	Paper End in	S5	RxDATA	
13	On Line in	S4	/DRDY	
14	/LF/CR	C1	not used	
15	ERROR in	S3	link to 16	
16	/INITIALISE	C2	link to 15	
17	/SELECT	C3	not used	*
18-25	GROUND		GROUND	

Notes: For LPT1, the 8255 port addresses are 888 (data) 889 (status) 890 (control). For LPT2, the addresses are 632 (data) 633 (status) and 634 (control). The link between C2 and S3 can be toggled and checked to verify hardware connection of the converter.

Analogue input connections

The IC has two inputs, either of which will operate over a wide range of voltages. For instance, the output of a strain-gauge bridge connected between 0 and 5V can be measured on input 1. The 2.5V common-mode voltage is ignored and the pga gain can be set to 128 for microvolt resolution.

The common-mode range extends from +5 to –5V. The pcb design makes the fully differential input 1 available directly and without protection on a 15-way D 'multi-function' connector. Input 2 is fed via an attenuator from the D connector and also from separate input terminals or a front panel BNC connector.

Because of its grounded attenuator, input 2 is not differential. The attenuator division ratio is not exact due to the relatively low converter input impedance. This is overcome, of course, by the self-calibration facility. The full circuit diagram of the converter is shown in Fig. 2 which also clarifies the multiple power supply regulation.

Implementing self-calibration

Self-calibration is a facility which can be added to any microprocessor-controlled equipment, but which is generally reserved for high-accuracy systems. The commands for self-calibration are explained in the panel discussing the set-up and control word for the *7710*. These commands result in a linear converter response.

Naturally, self-calibration does not imply traceable calibration or comparison with anything except the system's own reference. Thus, for instance, although the voltage reference used in this system has a very small guaranteed drift with temperature and time, it has a relatively wide initial voltage tolerance.

A typical ratiometric panel meter IC would inherently deliver a zero reading at zero input and a full scale reading when the input is equal to the reference – equal to minus one count, to be pedantic. This reference is typically 1V or 100mV, derived through a preset potentiometer from a bandgap voltage reference IC, the preset being adjusted to calibrate the meter.

A normal preset would not be sufficiently stable for this design. You can make a more stable system by connecting the *REF-03* directly to the *7710* to make an 'approximately 0-2.5V' converter.

Data fed from the converter to the supervising pc are simply 24 bit numbers. The process of 'absolute calibration' is to apply zero volts and an accurate near-full-scale

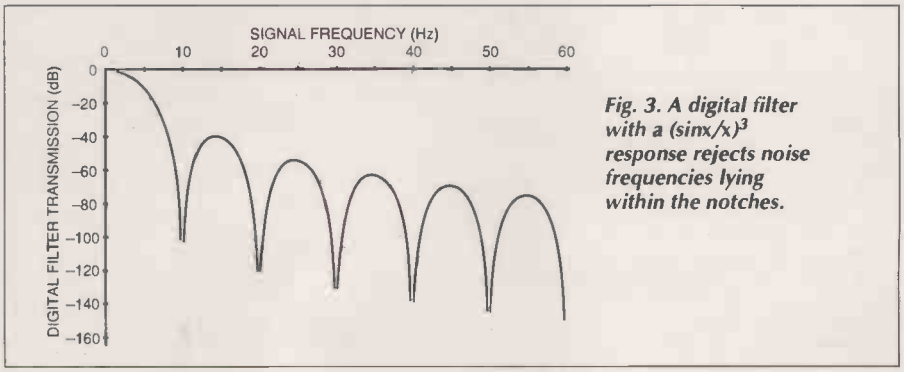


Fig. 3. A digital filter with a $(\sin x/x)^3$ response rejects noise frequencies lying within the notches.

voltage from a high-quality calibrator. Next, note the corresponding numerical values coming into the pc and insert appropriate conversion factors in the pc program to display correct absolute values.

These conversion factors can be held in an initialisation file. If the converter is incorporated in a larger system, overall system calibration can be done the same way and an initialisation file held for any set-up. This facility is fully utilised in the Windows application program to present a virtual instrument with relevant units – a strain gauge load cell output displayed in kg, for instance.

Non-linear calibrations, incorporating corrections for the well-known non-linearities of thermocouples, for instance, can be dealt with in a couple of ways. If the polynomial coefficients are known they can be included in the user's program. Alternatively, the system can be calibrated at several fixed points and the polynomial coefficients calculated by least squares curve fitting.

Digital input/output facilities

As there are several spare lines available on the Centronics port and some space on the pcb it was thought well worthwhile to add a few buffered digital inputs and outputs. There is little to say about these except that the *MPSA14* can carry 300mA and hold off 30V which makes it capable of switching relays. Don't forget to add a recirculation diode across the coil.

A single protected digital input is included to allow external triggering. Its cost is negligible and it has been found very useful for automation experiments.

The 12V regulated supply is also available to power external signal conditioning. It should not be misused as a robust bench power supply since if the *AD7710* analogue

This basic Basic routine allows communication between the 22bit a-to-d converter and a PC via the printer port.

```
DECLARE SUB screenmeter (volts!)
DECLARE SUB flash ()
DECLARE SUB digitalout (code)
DECLARE SUB digitalin (level$)
DECLARE SUB convertersetup ()
DECLARE SUB setbit (port!, bit!)
DECLARE SUB clearbit (port!, bit!)
DECLARE SUB getword (adval)
DECLARE SUB waitconverter ()
COMMON SHARED dataport, statusport, controlport, outword, s$
REM a program to test the Centronics converter functions
REM written on August 2 1995
REM printer port addresses for lpt1
dataport = 888
statusport = 889
REM only used for front panel red LED
controlport = 890:
REM initial setup - _RFS, _TFS high, rest low
outword = 12:
REM ******** Start of Program *********
OUT dataport, outword
waitconverter
convertersetup
screenmeter (0)
FOR i = 1 TO 10: flash: NEXT
DO UNTIL INKEY$ = ""
   getword adval
   REM omit for unipolar
   adval = adval - 2 ^ 23:
   REM use 2^24 for unipolar
   volts = adval * 2.5 / (2 ^ 23):
   LOCATE 16, 27
   PRINT "converter INPUT IS:  ";
   PRINT USING "#.######"; volts;
   PRINT " VOLTS"
   flash
LOOP
SUB convertersetup
md$ = "001":    REM mode; 001 = int zero, self-calibration
pg$ = "000":    REM PGA gain, set to 1x here
ch$ = "1":      REM channel selection, set to channel 2
pd$ = "0":      REM power-down, turned off
wl$ = "1":      REM word length, set to 24 bits
ro$ = "0":      REM RTD excitation current, turned off
bo$ = "0":      REM burn-out detection current, turned off
bu$ = "0":      REM bipolar/unipolar, set to bipolar
f1$ = "0110":   REM first 4 bits of filter '6'
f2$ = "0001":   REM middle 4 bits of filter '1'
f3$ = "1010":   REM last 4 bits of filter 'A'
s$ = md$ + pg$ + ch$ + pd$ + wl$ +
   ro$ + bo$ + bu$ + f1$ + f2$ + f3$
REM filter is set here to 61A = 1562 decimal,
REM ie 12.5Hz sampling rate
REM set up dataport with _RFS, _TFS high
OUT dataport, outword:
REM clears dataport bit 3, ie takes _TFS low
clearbit dataport, 3:
REM now clock out control word (24 bits)
REM by toggling SCLK line
   FOR i = 1 TO 24
   IF MID$(s$, i, 1) = "1"
                  THEN setbit dataport, 1 ELSE clearbit
   dataport, 1
   setbit dataport, 0: REM SCLK line
   clearbit dataport, 0
   NEXT i
setbit dataport, 3:      REM return _TFS high
END SUB
SUB clearbit (port, bit)
valbit = 2 ^ bit
outword = outword AND NOT valbit
OUT port, outword
END SUB
SUB digitalin (level$)
S6 = INP(statusport) AND 64: REM comes in on line S6
IF S6 = 64 THEN level$ = "low" ELSE level$ = "high"
END SUB
SUB digitalout (code)
REM this sends out a code from 0 to 15 on the Darlingtons
outword = outword AND 15:  REM ensure top 4 bits are off
outword = outword OR (code * 16): REM place top 4 bits out
OUT dataport, outword
REM this is a bit simple because it's not set bit by bit
REM so there will be an 'off' glitch every time.
END SUB
SUB flash
OUT controlport, 1: FOR i = 1 TO 10000: NEXT
OUT controlport, 0: FOR i = 1 TO 10000: NEXT
END SUB
SUB getword (adval)
waitconverter
adval = 0
clearbit dataport, 2:    REM take _RFS low
FOR i = 0 TO 23
   setbit dataport, 0:    REM take SCLK high
   IF INP(statusport) AND 32
      THEN adval = adval + 2 ^ (23 - i)
   REM read statusport bit 5,
   REMadd its value to adval, MSB first
   REM take SCLK low again
   clearbit dataport, 0:
NEXT
setbit dataport, 2: REM return _RFS high
END SUB
SUB screenmeter (volts)
SCREEN 12
LINE (0, 0)-(639, 479), , B
LINE (100, 100)-(539, 379), 5, BF
LINE (170, 220)-(470, 270), 0, BF
LOCATE 11, 27
PRINT " AD7710AN BASIC TEST PROGRAM "
END SUB
SUB setbit (port, bit)
valbit = (2 ^ bit)
outword = outword OR valbit
OUT port, outword
END SUB
SUB waitconverter
notready:
   IF INP(statusport) AND 16 THEN GOTO notready
END SUB
```

supply dips below the digital side for an instant it changes from a data converter into a thyristor and gets very hot.

Lines to the connector include current-limiting 22Ω resistors to provide some protection. Only a few tens of milliamps are available and decoupling capacitors will be needed on the external circuitry.

Test program written in Basic

A listing is given for a minimal test program. This routine operates the converter by somewhat agricultural data transmission methods but it serves to illustrate the important points.

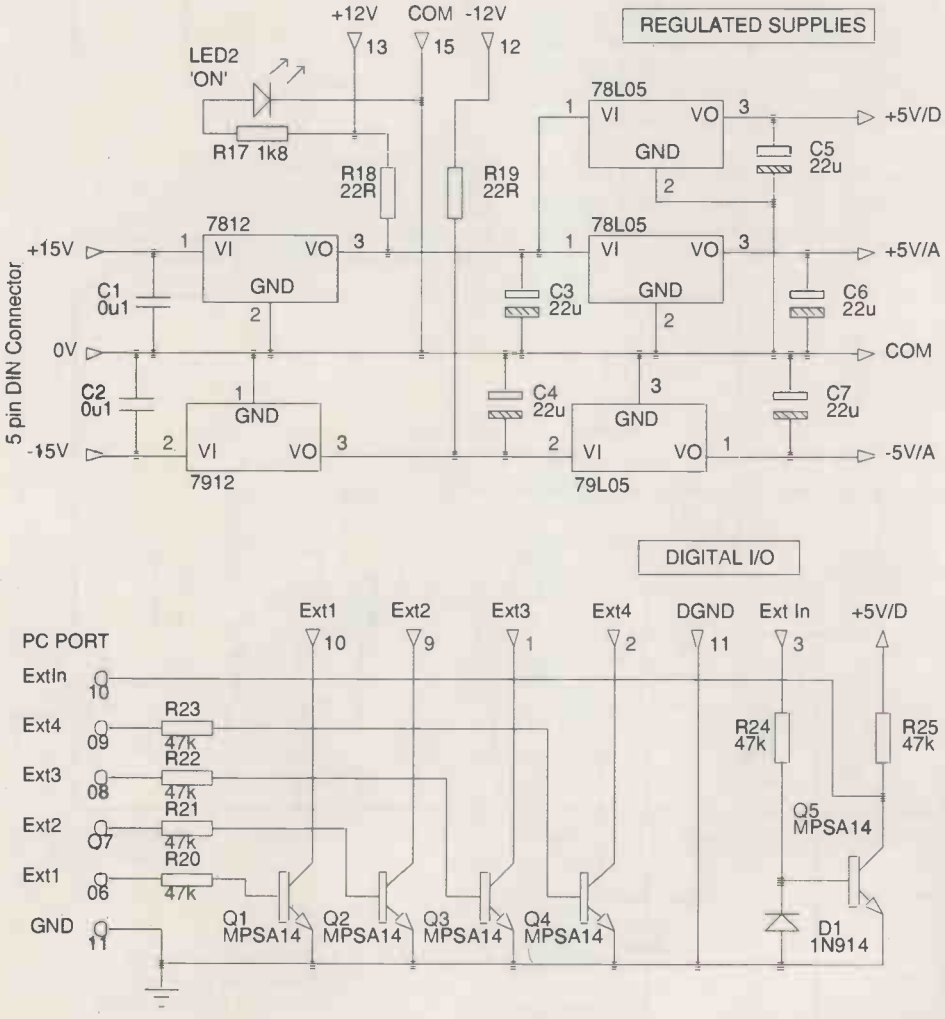
Initially, the parallel port is set up for normal action and no communication. The program waits for the converter to indicate readiness by taking /DRDY low. A 24 bit set-up word is sent to the converter by taking the transmit frame synchronisation signal /TFS low, setting TxDATA high or low and toggling serial clock line SCLK for each bit. After all 24 bits have gone /TFS is returned high.

Converter values are read in a similar fashion, by taking the receive frame synchronisation signal /RFS low and clocking data into the computer by toggling the SCLK line.

A possible cause of confusion when working with the parallel printer port is that the pc hardware inverts some of the lines. Because of this, a bit set in the output register may not come out of the socket high. Port lines chosen for this design are mostly non-inverting. **Table 1** provides information on the port lines.

Summary

Ultra-high resolution, high accuracy analogue measurement used to be the preserve of very expensive and exotic equipment, supplied by companies like Fluke, Hewlett Packard and Solartron. While the extremes of quality measurement must stay with such companies, this hardware/software approach should provide performance in excess of most conceivable professional and amateur requirements – at a relatively low cost. ■



Note: pins marked ○—
are on rear panel PC port connector
pins marked ▷—
are on front 'Multifunction Connector'

'Ext In' is very high impedance and needs external pull-up or pull-down resistor

Regulator and digital i/o for the 22bit a-to-d converter. Note that the regulators are fed from a separate dc power supply.

Benefiting from a new high-side switching device – namely a Treeswitch – this economical battery charger allows fast charging of NiCd and NiMH and reduces 'memory effect' in NiCd cells.

Fast charger for NiCd and NiMH

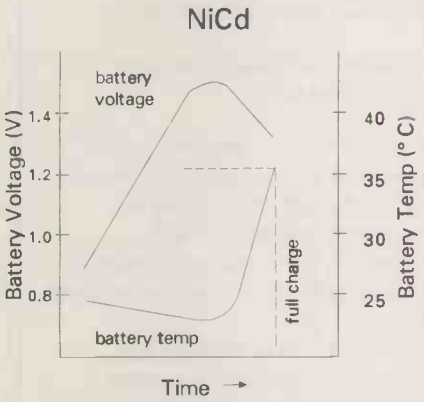


Fig. 1. Charging characteristics of NiCd cells indicate that battery voltage starts to fall significantly just before full charge.

Generally, nickel-cadmium cells, or NiCd cells, are trickle charged at 0.1C for about 14h., **Fig. 1**. With better understanding of battery chemistry the trend is shifting towards rapid charging at higher rates – 1C and greater for example – especially in the professional market.

The new generation of 'smart charger' employs an ASIC, often in combination with a microcontroller to optimise battery management. The methods used to detect end of charge are dv/dt inflexion, temperature and time.

We found that the monitoring of temperature to detect end of charge is as effective as the dv/dt method, which can be a problem for NiMH as the inflexion point is not well

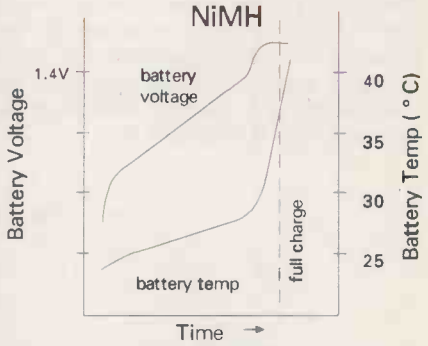


Fig. 2. With NiMH cells, charging characteristics show that the voltage fall at full charge is much less significant than with NiCd alternatives to temperature change is a more useful indicator of cell charge status.

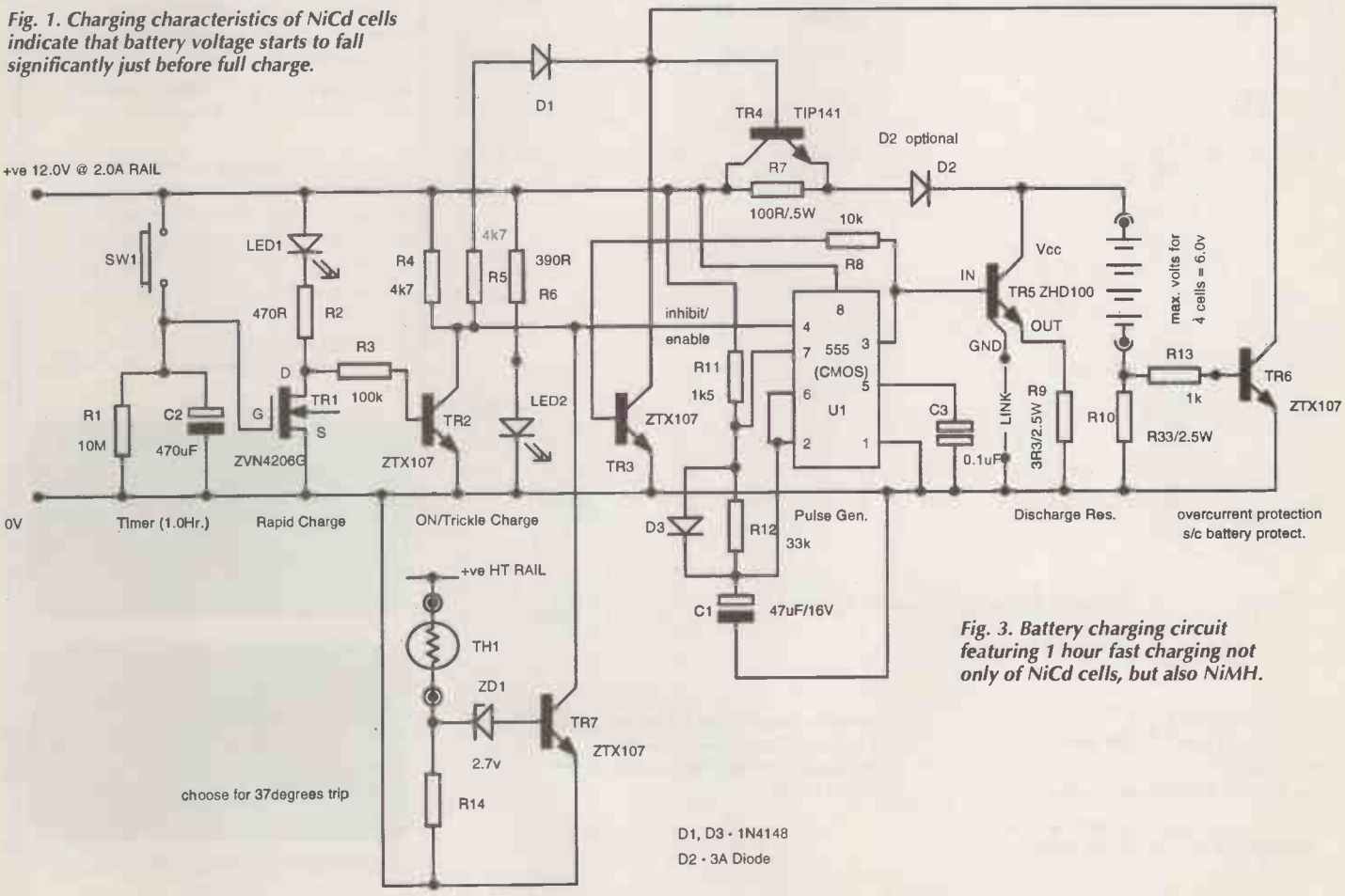


Fig. 3. Battery charging circuit featuring 1 hour fast charging not only of NiCd cells, but also NiMH.

defined, **Fig. 2**. However, as the 38°C end of charge temperature for nickel-metal-hydride cells, NiMH, is greater than the 35°C, of NiCd cells, using the temperature-only method results in a slight undercharge for NiMH batteries.

Benefits of charge/discharge cycles

Figure 3 shows a circuit capable of charging four 'AA' size cells in series within 1hr – less if the cells are not fully discharged. When rapid charging, the circuit supplies a 3s charge, then 10ms discharge current pulse. This repeated discharging during charging reduces or removes the memory effect of in NiCd cells cells.

At switch on, the charger defaults to trickle charging at 70mA, so it can be used as a simple conventional charger. When SW_1 is momentarily closed a 1.2h timer is enabled and the charger goes into rapid charge mode, charging at a 1C rate of about 1A.

Temperature of the battery rises as it nears full charge. When it reaches 35°C, the unit reverts to trickle charge and stays indefinitely in this maintenance mode – until SW_1 is closed again.

The prototype unit was set so that the timer was greater than that required to charge NiMH from zero depth of discharge. This ensured the cells would be charged to maximum, whatever the initial state of the cells.

The unit has been in use for some time and has successfully recharged both NiCd and NiMH cells – some of which would not hold charge using conventional trickle chargers. In fact, we observed that only cells showing signs of physical leakage damage could not be recharged – others, even very old ones can be charged with varying degrees of success.

A unique feature of this circuit is the incorporation of a new device called a Treeswitch. Designated the *ZHD100*, this discrete semiconductor comprises a bipolar power device with a mosfet input (see panel). This topology enables the discharge circuit to be implemented easily.

For safety reasons, the unit will not allow rapid charging if there are any short circuit cells in the stack; it defaults to trickle charging.

A 12V, 1A power supply is suitable for driving the circuit shown. To charge larger capacity cells – C and D sizes for example – a psu with the same current capacity of the cell to be charged is recommended for 1h fast charging. ■

Treeswitch – a bipolar high-side driver with mosfet input characteristics

'Treeswitch' is a term describing a new monolithic semiconductor structure combining the benefits of mosfet and bipolar transistor technology. Invented at Zetex, the device is a high-side switch featuring high input impedance and bipolar transistor power switching characteristics.

Unlike most previous bipolar/mosfet combinations, the Treeswitch is a four terminal device. Initial products in the range, namely the *ZHD100* and *ZDHD100*, are single and dual high side switches operating from supply voltages to 80V with continuous currents to 250mA.

To achieve the integration in a rugged and cost-effective manner the designers developed a new technology platform – the structural integration of Zetex's matrix bipolar transistor technology and a mos input stage. The result is a patented bi-mos structure offering the combined advantages of a ground referenced standard logic level mos input with a V_{cc} referenced low output voltage drop.

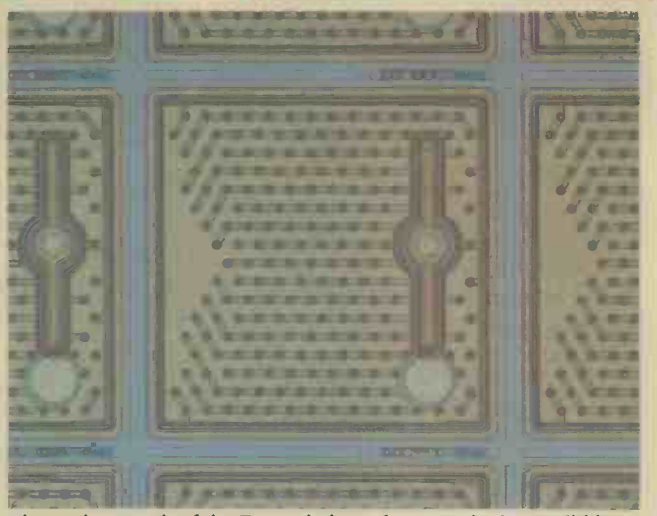
This seemingly simple integration task was complicated by conflicting process requirements and the presence of unwanted parasitic interactions, requiring extensive simulation and verification testing to optimise the device performance.

The advantage of separating the collector and source is that the low saturation voltage of the transistor can be fully exploited. This is in contrast to an igbt structure where the voltage drop is at least a volt.

Pulling the gate positive with respect to the source attracts electrons into the p-type material below it which forms an n-channel between the base and source. This allows base current to flow and turns the transistor on.

The transistor has Zetex's matrix architecture, which results in unusually low saturation voltage. The fet sits in the middle of it and is small in comparison, limiting the amount of current it can pass. This makes the fet approximate a constant-current device, removing the need for a drain-base current limiting resistor.

Housed in SOT223 or Zetex's SM8 packages, the first devices to be released are designed with ruggedness in mind, being able to switch over two amps depending on the duty cycle and drive supplied.



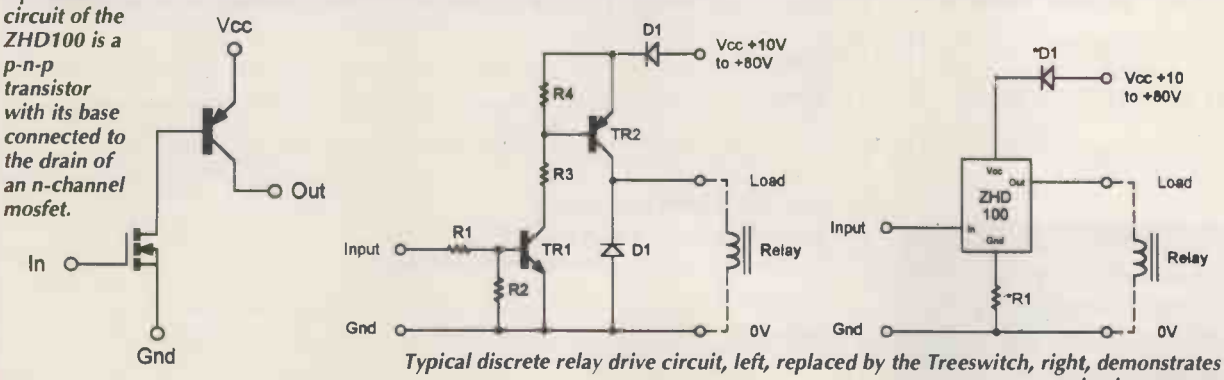
Photomicrograph of the Treeswitch – a four-terminal monolithic device combining the benefits of mosfet input impedance with bipolar switching characteristics.

At high voltages the device dissipation is predominantly from the 10mA base current. If the full output current of the device is not required, this can be reduced by adding a series resistor between the source and ground.

Although there are other high-side switches on the market, co-inventor of the Treeswitch David Casey said: "The matrix architecture results in a very small chip compared with its competitors. The small chip leads to a low device cost."

Structure of the Treeswitch is shown below left, followed by an example of how the device saves components in a typical relay driving application..

Equivalent circuit of the ZHD100 is a p-n-p transistor with its base connected to the drain of an n-channel mosfet.



Typical discrete relay drive circuit, left, replaced by the Treeswitch, right, demonstrates the saving in components.

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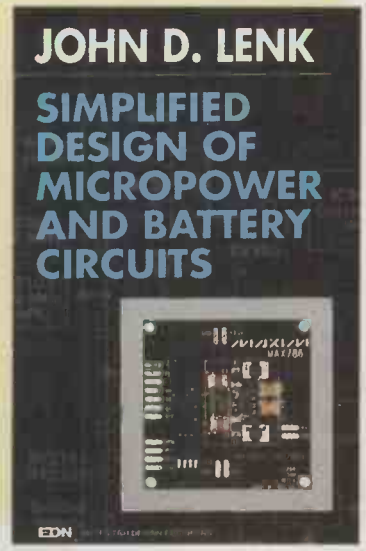

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Designing valve RIAA preamps

Delving further into valve preamplifier design, Morgan Jones shows how to produce a no-compromise balanced design combining the benefits of valves and transistors.

An RIAA preamplifier, to last month's philosophy, needs three individual stages. A cascode or a μ-follower are both possibilities for the input stage, but initially, it is advisable to use a common cathode triode for simplicity. The second stage can be the same, but the third will need to be a cathode follower for reasons that will become apparent later. You can now draw a circuit diagram for the complete RIAA stage, **Fig. 1**.

The 75μs hf loss is formed by the combination of R_4, R_5, and C_3, whereas the 3180μs, 318μs pairing is formed by R_8, R_9, and C_5. The calculation of these components is simple, but you must remember to account for hidden components. Eaxmples of these are the output impedance of the valve, and Miller input capacitance of the next stage in parallel with strays.

Calculation of 75μs component values

The entire pre-amplifier is based on the *E88CC* dual triode, and for the dc conditions chosen for our common cathode triode input stage, r_a equals 6kΩ. This is in parallel with the 100kΩ anode load resistor, so Z_{out} is 5.66kΩ.

To calculate the capacitor needed for the 75μs time constant, you need to find the total Thévenin resistance that the capacitor sees in parallel, as shown in **Fig. 2**.

For the moment, you can ignore C_1. It will be accounted for later. Capacitor C_3 sees the grid-leak resistor R_5 in parallel with the series combination of the output impedance of the preceding valve and R_4. As is usual, you will make the grid-leak as large as is allowed, so R_5 equals 1MΩ.

You are now free to choose the value of R_4. Impedance Z_{out} needs to be a small proportion of R_4, otherwise variations in r_a will upset the accuracy of the equalisation. Too large a value of R_4 will form an unnecessarily lossy potential divider in combination with R_5. At high frequencies, capacitor C_3 is a short circuit, and so the additional ac load on the input valve will be R_4. A good value for R_4 is 200kΩ, and it has the bonus of being available both in 0.1% E96 series, and 1% E24 series. Very few E24 values are common to the E96 series. In combination with R_5, this gives an acceptable loss of 1.6dB, while not being an unduly onerous load for the input stage.

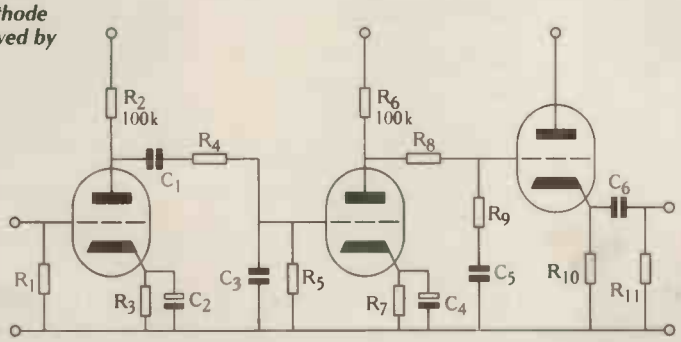


Fig. 1. Basic RIAA preamplifier stage incorporates two common cathode stages followed by a cathode follower.

The capacitor now sees 200kΩ and 5.66kΩ in parallel with 1MΩ, giving a total resistance of 170.58kΩ. Dividing this value into 75μs gives the required capacitance value of 440pF, but you must subtract the stray capacitance of the next stage.

Gain of the second stage is 29, and C_{ag} is 1.4pF, so the Miller capacitance will be 30 times 1.4pF which is 42pF. In addition to this, the cathode, the heaters, and the screen are at earth potential, and will be in parallel with this capacitance. $C_{g\text{-}k+h+s}$ is 3.3pF, and you ought to allow a few pF for external strays. A total input capacitance of 50pF would be reasonable.

Total capacitance required is 440pF minus 50pF, or

390pF, so a 390pF 1% capacitor is acceptable.

Earlier, the effect of coupling capacitor C_1 was ignored, but this must have some effect on the Thévenin impedance seen by the 390pF capacitor. You could use such a large value that its reactance was negligible compared to the 200kΩ series resistor, but a more elegant method is to move its position slightly, **Fig. 3**.

The capacitor now only has to be negligible compared to 1MΩ. The 75µs delay corresponds to a –3dB point of approximately 2kHz, so it is at this frequency that the values of other components are critical. At 2kHz, a 100nF capacitor has a reactance of approximately 800Ω, which is less than 0.1% of 1MΩ. If you had not moved the capacitor, you would have needed a value of 470nF simply to avoid compromising RIAA accuracy.

Interaction problems

The second stage is direct coupled to the cathode follower, so you do not need to worry about interaction between a coupling capacitor and the 3180µs, 318µs pairing. This is fortunate, since 3180µs corresponds to 50Hz, which is close to our 1Hz cut-off. These time constants are sufficiently close that they would interact significantly.

The other reason for using a cathode follower is its low input capacitance. Any stray capacitance across the 3180µs, 318µs pairing will cause an additional high frequency roll-off. In the 75µs network, you were able to incorporate the value of stray capacitance into your calculations, but in this instance this is not possible, and it is therefore essential that stray capacitance is so small that it can be ignored. The full equation for the input capacitance of a cathode follower is,

$$C_{input} = C_{ag} + (1 - A)C_{gk}$$

For a cathode follower, A_v approximates to $\mu/(\mu+1)$; for an *E88CC*, μ is approximately 32, resulting in a gain, an A_v of 0.97. Capacitance C_{ag} is 1.4pF, and C_{gk} is 3.3pF. The C_{gk} term is negligible at 0.1pF, and so the input capacitance is virtually independent of gain at 8pF – including an allowance for strays.

The equations that govern the 3180µs, 318µs pairing are delightfully simple, CR is 318×10^{-6}, and the upper resistor is $9R$. Loss at 1kHz for this network is 19.05dB, **Fig. 4**.

You should now check whether the 8pF stray shunt capacitance is sufficiently small not to cause a problem. To do this, you need to employ a slightly circular argument.

First assume that it will not cause any interaction. If this is true, then the frequency at which the cut-off occurs will be so high that C in the network is a short circuit. If it is a short circuit, you can replace it with a short circuit, and calculate the new Thévenin output impedance of the network.

Since the ratio of the resistors is 9:1, the potential divider must have a loss of 10:1, and the output impedance is therefore one tenth of the upper resistor. If you assume that the upper resistor will again be 200kΩ while neglecting Z_{out} of the previous stage, the Thévenin resistance that the 8pF stray capacitance sees at high frequencies is 20kΩ, this gives an hf cut-off of 1MHz.

As a rough rule of thumb, once the ratio of two interactive time constants is ≥100:1, the response error caused by interaction is inversely proportional to that ratio. A ratio of 100:1 causes an error of approximately 0.1dB.

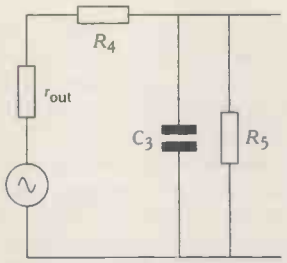


Fig. 2. Determining RIAA's 75µs time constant involves finding the total Thévenin resistance that the capacitor sees in parallel.

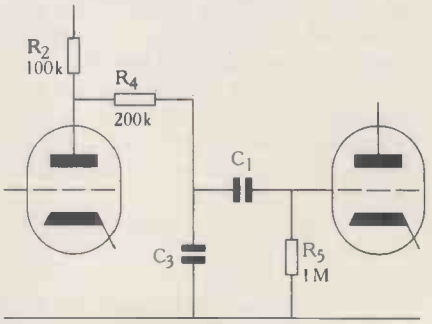


Fig. 3. Moving the coupling capacitor rightwards in the network reduces interaction.

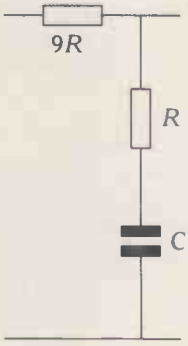


Fig. 4. These values produce the RIAA 3180 and 318µs time constants.

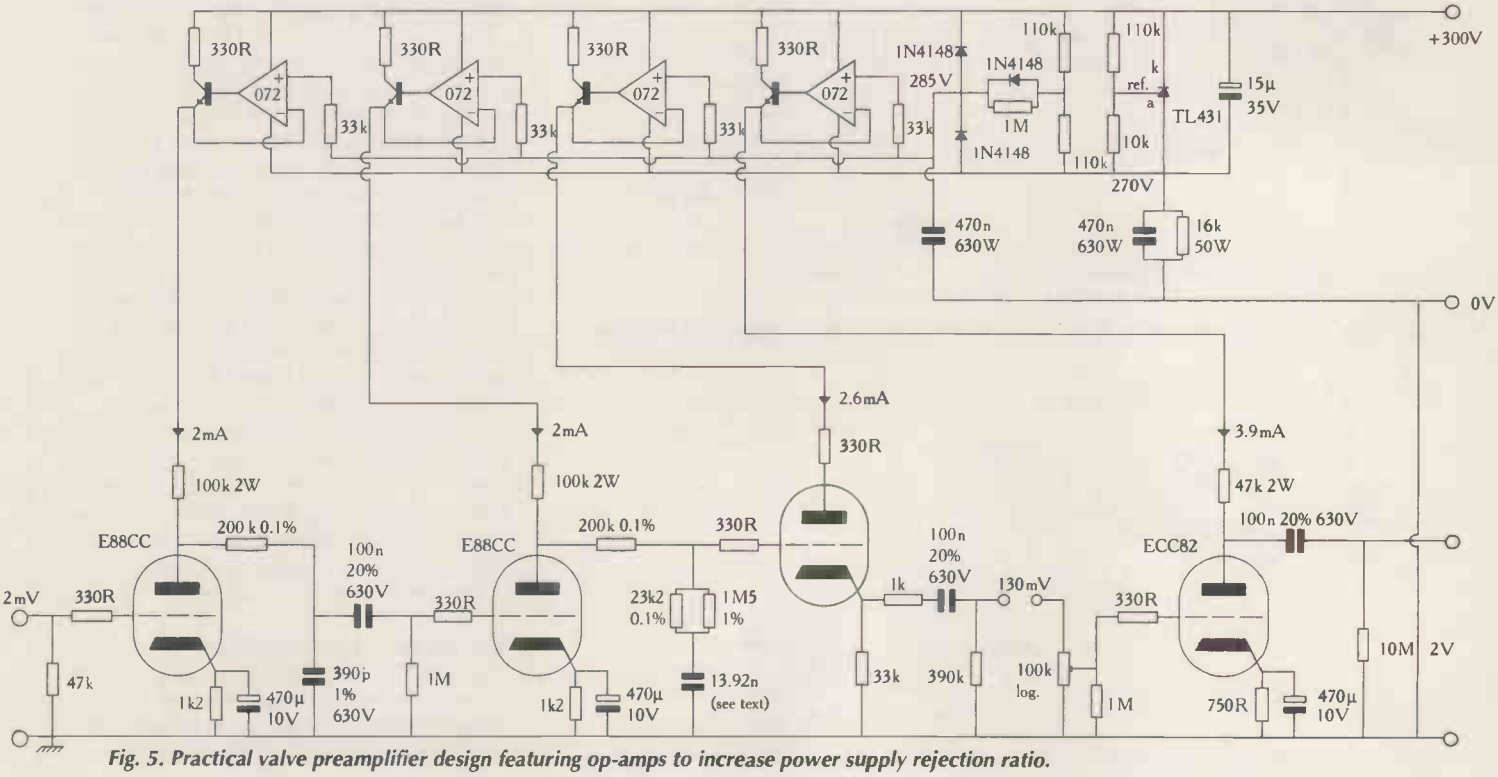


Fig. 5. Practical valve preamplifier design featuring op-amps to increase power supply rejection ratio.

In this example, the ratio of 1MHz to the nearest time constant of 318µs (500.5Hz) is 2000:1. You can now safely ignore interaction and go on to accurately calculate the values for the 3180µs, 318µs pairing.

If the network were driven from a source of negligible resistance such as an op-amp, ideal values for the resistors would be 180kΩ and 20kΩ, since these are both members of the E24 series. The capacitor could then be 16nF with only 0.6% error. Unfortunately, the source has appreciable output resistance, so you will again choose 200kΩ as the upper resistor and accept whatever values this generates for the lower two components.

Since the second stage is identical to the first, output resistance is 5.66kΩ, making a total upper resistance of 205.66kΩ. The lower resistor will therefore be 22.85kΩ, and the capacitor 13.92nF.

A resistance of 22.85kΩ can be made from a 23.2kΩ, 0.1% resistor in parallel with a 1.5MΩ, 1%. A capacitance of 13.92nF can be inconveniently made from a pair of 6.8nF in parallel with 330pF. You can now draw a full diagram of the preamplifier stage with component values, **Fig. 5**.

Equalisation networks for RIAA invariably generate awkward component values, requiring much manoeuvring to nudge them accurately onto the E24 series.

Power supply rejection ratio

Although individual stages have been designed and interconnected to form an audio system, each stage requires power. Supplies are always derived from a common source.

No practical source has zero output resistance, although ac mains is a good approximation The issue of a common power supply with non-zero output resistance is crucial. It implies that as a given audio stage draws a varying supply current in sympathy with the audio signal, a voltage will be developed across the source resistance of the supply.

Although attenuated by individual stage rejection ratio, this voltage is now an input to all other stages. If power supply rejection ratio, psrr, is low while the signal gain between stages is high as in an RIAA stage, the loop gain via the power supply may be greater than unity. This results in oscillation.

Traditional power supplies used a shunt capacitor to define their source impedance, resulting in increased source impedance at low frequencies since,

$$Z_{source} = \frac{1}{2\pi fC}$$

Therefore instability would be more likely at low frequencies, although the non-zero effective series resistance of the normally electrolytic supply capacitors could provoke high-frequency instability if not bypassed.

Modern designs use regulators giving excellent Z_{source} down to dc. However, because the error amplifier must have a response falling with frequency in order to maintain its own stability, Z_{source} is inductive and rises with frequency, and hf instability is a possibility.

Summarising, any practical common supply will always have non-zero output resistance. System stability is only maintainable if individual stages have sufficient psrr to that common supply. It is useful to define two new terms:

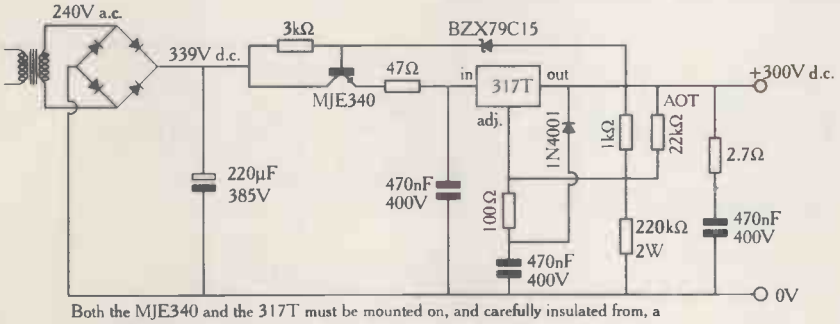
- Intrinsic psrr: the psrr due to the topology of an individual stage.
- Common supply psrr: intrinsic psrr plus any added psrr – by whatever means – to the common supply point.

Any common cathode stage possesses intrinsic psrr by virtue of the potential divider formed by r_a and R_L, but an *E88CC* operated such that r_a is 6kΩ, and R_L is 100kΩ only results in an intrinsic psrr, referred to the output, of 24dB. Using the same valve as a µ-follower could improve this to 50dB, a differential pair might improve the 24dB figure to 64dB depending on valve matching. Used as a cascode, the valve's 24dB figure would be degraded to zero.

Any given stage may have its common supply rejection ratio increased by an arbitrary amount using individual filtering or regulation. Apart from expense, it does not matter whether the common supply rejection is made up mostly from intrinsic psrr, or added psrr via filters or regulators.

Extreme methods might even include individual mains transformers and supplies for each stage. This increases common supply rejection ratio to the ac mains, the common supply point. Use of a dedicated spur from the electricity supply company cable head would be a means of reducing Z_{source}.

A more elegant and considerably cheaper method of improving common supply rejection ratio is to add the high intrinsic psrr of an op-amp to stage intrinsic psrr by supplying each stage via a voltage follower op-amp. This was illustrated in the previous diagram. In order to obtain a low Z_{source}, a regulator is used at the common supply point, **Fig. 6**.



Both the MJE340 and the 317T must be mounted on, and carefully insulated from, a substantial heatsink such as a piece of 3mm thick aluminium angle extrusion.

Fig. 6. Practical 300V regulated ht supply incorporating a floating 317 adjustable regulator.

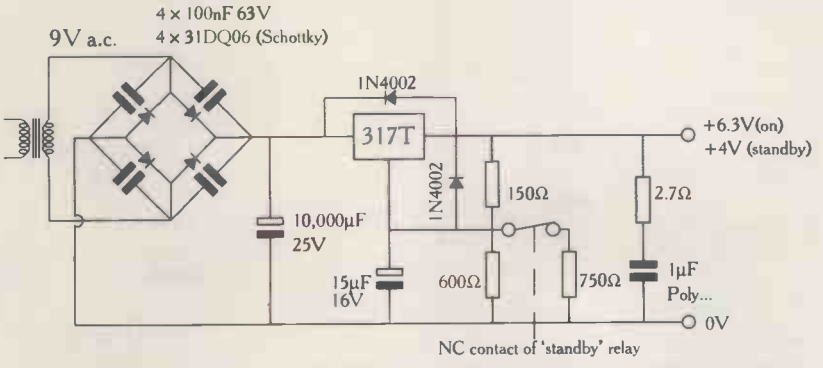


Fig. 7. Using a rectifier and dc regulator for the 6.3V heater supply eliminates hum problems associated with traditional ac heater drives.

Practicalities and performance

For optimum performance, valve pre-amplifiers should have a 'standby' mode, whereby the heaters are supplied with approximately 63% of operating heater voltage. This ensures a minimum of gas molecules within the vacuum. These molecules become ionised when ht is applied, accelerating them to the cathode, resulting in stripping of the cathode emissive surface. As a result, they should be kept to a minimum. At switch-on, ht is applied, and the heaters are restored to full voltage, **Fig. 7**.

A dual colour led was fitted as a power indicator with its green led lit by the permanently applied heater supply, and the red led in series with the lower leg of the ht sink resistor for the op-amps. Switching the pre-amplifier on therefore results in an orange glow similar to the colour of a valve heater, but a pure red glow would indicate heater supply failure.

The preamplifier was designed to be as simple as possible while retaining quality. It works well. Paired with a Garrard *301* on a solid plinth, using an Ortofon *Quattro* moving coil cartridge in a unipivot arm designed and built by me, the complete LP system was comparable to a £2,000 cd-based system.

The balanced preamplifier

Although logic dictated optimum system topology for the RIAA stage, individual stage design is flexible. Audio stage complexity can usefully be traded against power supply com-

plexity for a given common supply psrr requirement. In this respect the differential pair is most useful and has the added bonus of reducing the number of coupling and decoupling capacitors required. This naturally leads to...

Balanced working and cables

Balanced working is commonly used in broadcast and recording studios to protect audio signals from external electromagnetic interference. It is particularly useful for low-level signals such as microphones.

A balanced source is simply one where each terminal of the source has balanced impedances to ground. Frequently, the only path to earth from the terminals is via stray capacitances, and the source is then floating. Connecting cables for balanced systems therefore have two signal wires or legs, and an overall screen to maintain this balance. The input stage of the following amplifier also has its stray impedances carefully balanced to ground and will either be a differential pair or a transformer.

When you immerse the connecting cable in an electromagnetic field, an identical noise current is induced into both wires. The series resistance of the cable is the same on each leg, and the shunt capacitances and resistances to ground are also equal, so the noise current develops a voltage of identical amplitude and phase on both legs at the amplifier input. This common mode signal is then rejected by the differential pair or transformer, whereas the wanted audio signal is differential mode and is amplified.

Typically, a moving coil cartridge produces approximately 200µV at 1kHz and 5cm/s, but before RIAA equalisation, the level at 50Hz is approximately 15dB lower at 36µV. Achieving the goal of inaudible hum on a signal at this level is not trivial. The cartridge is a balanced device, so why unbalance it?

You should immediately rewire the output cable of the pick-up arm to maintain this balance by discarding any coaxial cable. The connecting cable from arm base to preamplifier should be replaced by a twisted pair, with overall screen, for each channel.

A cable construction I use has twisted pair covered with a braid electrostatic screen. Both cables are then threaded down one overall braid screen. Braiding also hold the cables together and further aids screening, while a nylon braid is fitted over the top to prevent handling noise.

The braid should not have voids, so most antenna cables are unsuitable. Broadcast quality video cable or multicore umbilical cable, are both ideal sources of non-voided braid. Once the plastic outer sheath has been removed, the braid will easily concertina off the inner conductors.

A professional quality metal bodied 5-pin DIN or XLR plug is ideal for connecting this cable to the preamplifier, although the cable entry will usually need to be enlarged. Ideally, the screen should be connected to mains earth at the pick-up arm end, but this is not quite so critical in a balanced system.

Incidentally, within the arm tube, most pick-up arms twist all four thin, non-screened wires from the cartridge together, because this makes the wire easier to handle. Crosstalk between channels would be improved by twisting channels individually as they pass down the arm tube, but retaining the four wire twist required for low friction as the wires pass through the bearings to the output cable.

This form of rewiring is especially beneficial for moving coil cartridges and will help hum rejection even if the preamplifier is unbalanced.

Basic preamplifier compromises

If you really want to achieve a significant improvement on the basic preamplifier, you will need to look closely at the fundamental design and reconsider some of the compromises that were initially made.

- Intrinsic psrr was not maximised.

- Individual anode currents were set quite low in order to minimise total current consumption, so that the preamplifier could be powered from an associated power amplifier. This meant that g_m for each stage was low, and noise was not minimised.

- Metal film resistors were used in the anode load resulting in excess noise, although most of this was shunted by r_a. To eliminate excess

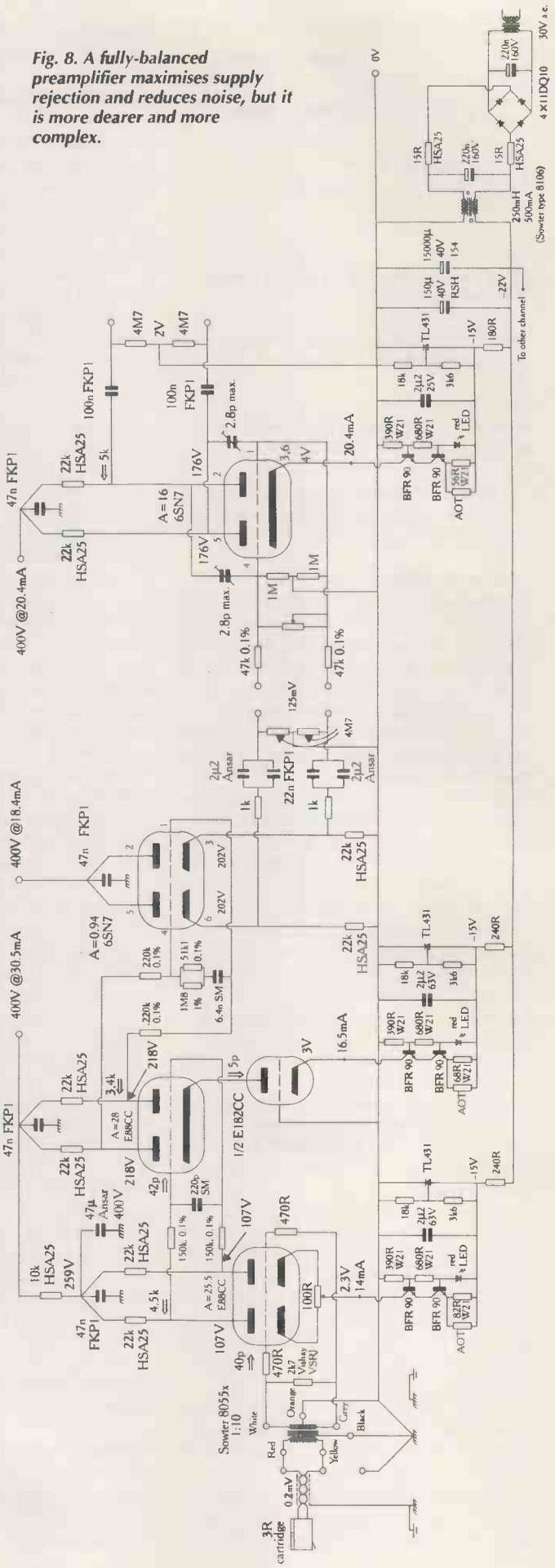


Fig. 8. A fully-balanced preamplifier maximises supply rejection and reduces noise, but it is more dearer and more complex.

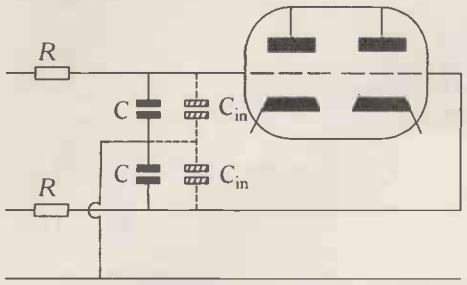


Fig. 9. Implementing the 75µs time constant in balanced mode.

noise, wirewound components should be used for any resistors with significant dc across them.

● Individual stages were kept simple, but linearity was therefore not optimum.

The balanced preamplifier seeks to address all of the above points but does not place such a high priority on simplicity or cost, **Fig. 8**.

The input stage

In order to reap the full benefits of balanced working, a moving coil step-up transformer for 3Ω cartridges was especially designed for this preamplifier by Sowter Transformers of Ipswich. Correctly terminated, the first batch of type *8055x* had a frequency response that was flat ±0.1dB from 12Hz to 100kHz, while the high-frequency phase response was pure delay ±1° to 50kHz.

The *8055x* transformer also has an electrostatic screen between primary and secondary and its stray capacitances to ground have been balanced. This results in excellent rejection of common mode noise on the connecting wires from cartridge to preamplifier.

The first stage has a semiconductor constant current sink to enhance common mode rejection and the grid-leak arrangement is a little unusual. If you were to assume zero winding resistance for the input transformer, then a grid-leak connected to one valve would also serve as the grid-leak for the other, so only one resistor is required.

Since winding resistance is not zero, you move the single grid-leak to the centre tap of the transformer, which is a point of zero ac and dc potential. This assumes perfect transformer balance.

Any noise current passing through this resistor will develop a voltage that is applied equally to both inputs of the differential pair and will be rejected. If the resistor is large, then a larger noise voltage will be generated, and the input stage may no longer be able to reject it. This problem is solved by reducing the grid-leak resistor to 0Ω and connecting the transformer centre tap directly to ground.

Because the circuit is dc coupled, it has become necessary to include a dc balance control which should be set to equalise the anode voltages of the second stage.

The resistors marked AOT, for adjust on test, in the cascode constant current sources will only need to be set once for correct anode voltages, since they correct for individual variations in V_{be} and led voltage

Second stage and 75µs time constant

In order to direct couple the first stage to the second, the cathode of the second stage must be at an elevated voltage. It seems foolish not to use a constant current sink in this position.

Initially, an *E88CC* triode constant current source was considered, but the calculated r_a of 200kΩ was thought to be insufficient, so an *ECF80* pentode was substituted. This increased r_a to 10MΩ. However, the *ECF80* was then being operated very close to its maximum rating. As mentioned before pentodes are noisy. The final design therefore uses half of an *E182CC,* configured as a hybrid triple cascode.

Because the second stage valve is directly coupled to the first, the second stage does not have grid-leak resistors. You therefore avoid the 1.6dB excess loss suffered in the basic preamplifier's 75µs network.

The 75µs time constant is achieved in a balanced fashion, with the shunt capacitor mounted directly onto the valve base with leads as short as possible in order to reduce stray capacitance. Similarly, the bodies of the series resistors are as close as possible to the valve pins. This means that they also perform the function of grid stopper resistors. The best way to understand the equalisation is to redraw the circuit as two unbalanced networks, **Fig. 9**.

The values for *R* and *C* are calculated exactly as before. However, observe that you could break the centre tap of our added capacitors away from ground, which would leave two capacitors in series. These can be replaced with a single capacitor of half the value. A noisy ground is now less able to inject noise into the audio signal.

An additional advantage is that there is now no dc across the capacitor, so a lower voltage rating may be used if desired. For this design it was convenient to use a series resistor of 150kΩ, thus needing a 220pF capacitor between the grids to set 75µs.

Pairing 3180 and 318µs

Since this pairing is achieved in a balanced fashion, the value of the capacitor is halved, and it has virtually no dc across it, which makes it much easier to find close tolerance components.

Because it was desirable to use a balanced 3180/318µs pairing, twin cathode followers were required, resulting in a balanced output from the RIAA stage. Most power amplifiers are push-pull and therefore include a phase-splitter. In the light of this, why not keep the signal balanced all the way into the power amplifier, and discard the problematical phase splitter?

Volume control and output stage

The volume control now has to be balanced, using matched fixed series resistors and a variable shunt to form a potential divider. This has the disadvantage of a high output resistance when set for a sensible input resistance and will cause hf loss if ignored.

The output stage uses a *6SN7* configured to give A_v of 16. Capacitance C_{ag} is 4pF for the *6SN7,* so the input capacitance C_{in} equals 68pF, not including any allowance for strays. The series resistors have been set to 49.9kΩ, giving a 47kHz hf cut-off, which is too low. To meet the 0.1dB loss at 20kHz criterion, you would need a C_{in} of less than 19pF. Alternatively, you would need to reduce the series resistors to 13.5kΩ, increasing the loading on the disc stage. One solution is to partially neutralise the C_{ag} capacitance by adding capacitance from each grid to the opposite anode using small trimmer capacitors.

Note that neutralisation is positive feedback and if it is not applied with care, the stage will turn into an oscillator. For the range 1 to 3.5pF, PTFE trimmer capacitors are readily available. If one of these trimmers is set with its vanes two thirds meshed, a capacitance of approximately 2.4pF results. This is sufficient to reduce input capacitance to an acceptable value.

Ideally, a square wave should be applied between ground and one input of the volume control. The other input should be grounded and the second capacitor adjusted until the output waveshapes are matched as viewed on an oscilloscope. Layout is crucal here.

An alternative to neutralisation would be to revert to using an *ECC82*, which has an intrinsically lower C_{ag} and a slightly lower gain, thus reducing C_{in}. Whichever course is taken, the volume control must be as close as possible to the valve in order to minimise external stray capacitances. Unscreened wires must be used.

Constant current sinking

Although a 'ring of two' circuit could have been used as a sink for the first stage, each transistor would then have been operated at a very low voltage. But operating transistors at a low voltage is undesirable. It makes the circuit more susceptible to rf overload, due to the depletion region within the transistor being narrowed. This increases output capacitance. These factors demand the use of a subsidiary negative supply. A superior cascode constant current sink using rf transistors can then be used, making a virtue out of a necessity.

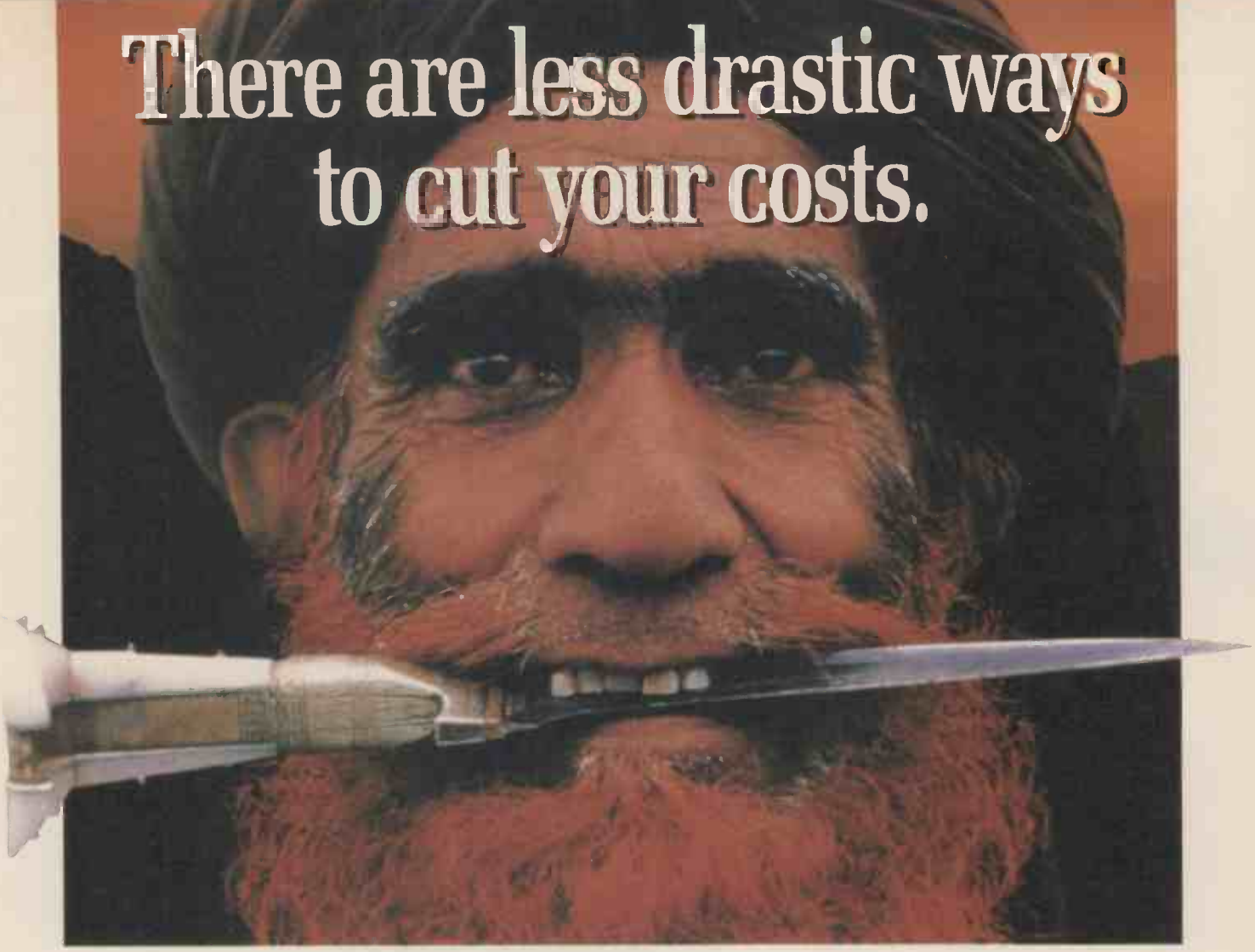
Noise on the subsidiary supply must be minimised, so a choke input supply was chosen. Potentially, the reactance of the choke and the 10,000µF smoothing capacitor form a resonant circuit. This resonance is critically damped by adding the 5.6Ω series resistor to the choke and transformer resistances.

The minimum current requirement of the choke is neatly solved by the use of a *TL431* shunt regulator for each stage. This ensures that a constant current is drawn – even when the ht is switched off. ■

Further reading

Wright, Allen. 'The tube pre-amp cookbook' 1994

Morrison, J. C. 'Siren song: A phono preamplifier for hedonists.' Sound Practices, 1993, Number 3, P3 - 9; Number 4, P6.



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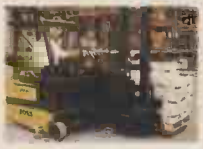

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ISSCC the highlights

Roy Rubenstein reports on the world's top electronics innovation event – the International Solid State Circuits Conference.

If there is one event in the world's electronics calender worth attending it is the International Solid State Circuits Conference – ISSCC – held in San Francisco. It is hard to imagine where else one could gain such a comprehensive overview of the latest analogue and digital circuit techniques and devices.

'Systems on a chip' was this year's conference theme. The opening session reviewed circuit design in the areas of multimedia, electronic imaging and TVs.

The keynote speech, given by NEC's vice president for semiconductors, Dr Hajime Sasaki, addressed multimedia. That much-touted phrase, multimedia, embraces all the emerging applications that manipulate text, graphics and video once encapsulated as ones and zeros.

Personal computers form the present, most common embodiment of multimedia. Sasaki's belief is that multimedia will come to predominate in home and work environments.

His presentation outlined the technology road map of the likely device that will be processing multimedia in the year 2010. His 'multimedia complex' device integrates and extends, common components found in present day PCs, namely the microprocessor, memory, three dimensional graphics accelerator and moving-image (such as video) processing circuitry.

While such a device may appear an obvious development, what is perhaps less so is the technical challenges its accomplishment presents.

All-time top ten circuits

One of the traditions of the ISSCC is the evening session where a panel tackle such weighty issues as 'Is Electronic Imaging at a Watershed?' and 'What is the Best Memory Type for Graphics?'.

This year, by far the best attended session – and certainly the most entertaining – was one that set out to name the ten most significant analogue circuits and circuit techniques. The criteria used included the need to have influenced other circuits and still be relevant today.

The panelists, which included Minoru Nagata, director of Hitachi's Central Research Laboratory and Bob Pease, the analogue guru at National Semiconductor, each selected three. The audience also contributed suggestions and the overall list were then voted on.

The resulting analogue top ten is:

1. Bandgap reference/regulator
2. Differential pair
3. Translinear circuits
4. Current mirror/source
5. Switch capacitor circuits
6. Pole splitting compensation
7. Cascode
8. Negative feedback amplifier
9. The power cord!
10. Integrator

0.07µm geometries by the year 2010

First, Sasaki projected present trends for device parameters such as integration densities, processing performance and power consumption, to gauge the likely system-device in the year 2010.

By then CMOS feature size will be 0.07µm, allowing hundreds of millions of transistors to be integrated on a single integrated circuit. The most advanced process technologies used today have 0.35µm feature sizes, achieving transistor densities up to ten million.

The intricacies involved in designing a 500 million transistor device is expected to be hundreds of times more complicated than that of present day microprocessors.

Looking next at processing performance, Sasaki observed that microprocessors have achieved an astonishing thousandfold improvement since 1980.

During that time, microprocessors have evolved instructions which when executed perform more than a single operation. Hence the emergence of microprocessor measures such as the millions of operations per second, or Mops, in addition to the traditional instructions per second metric, or Mips.

MIPS – slower growth

Sasaki believes that the astonishing Mips progress achieved to date will not continue since the instruction level parallelism that can be extracted from typical software code is rapidly being approached. He expects that in the next 15

years, an improvement of only a factor of 20 can be expected.

However, he sees no reason why the number of operations executed cannot progress at the staggering pace seen to date. Such progress will be achieved as multimedia function blocks are coupled to the main processing unit.

Extrapolating the processing trends, the multimedia complex can be expected to achieve 100 billion instructions/s and 1000 billion operations/s. To better gauge such a figure, Texas Instruments' most powerful multimedia processor, the *TMS320C80*, can attain a peak performance of 4 billion operations/s.

In turn, to sustain such processing rates the memory will need to supply the processing unit with tens of thousands of megabytes per second. Such transfer rates will not be possible between adjacent ICs, observed Sasaki, rather the memory will have to be integrated on-chip.

Yet a further challenge to be met is having the complex consume only 1W, necessary if it is to be used in portable battery-powered equipment.

Even if progress in low power techniques is maintained until 2010, a further order of magnitude reduction has to be found if the stringent 1W target is to met.

Interestingly, the solutions Sasaki outlined to attain such a multimedia complex, including integrating ample on-chip store and evolving present low power circuit techniques, were already in evidence in present papers at this year's ISSCC. Meeting the target specification will not be easy. As Sasaki puts it: "Developing the multimedia complex is a challenging target. We have so many things to do."

Variable voltage threshold techniques

CMOS has always been seen as a low power process technology. The success of VLSI, with the integration of millions of transistors on a device, has made CMOS hotter under its ceramic collar than it would like to be.

The most common approach to tackle device power consumption is by reducing its operating voltage. A recent example is the 433MHz Alpha processor from Digital which operates its processor core at 2V even though the device and its I/O is supplied with 3.3V. And it still consumes 23W.

With a reduced supply voltage comes a corresponding reduction in the voltage threshold, V_{th}. For CMOS, V_{th} is the voltage at which the device changes state.

Reducing V_{th} of a transistor increases its speed. However, the downside is the exponential increase in leakage current, and hence standby power consumption.

At ISSCC, a number of papers highlighted approaches that vary V_{th}. All use a reduced V_{th} when high performance is required and a high V_{th} in standby mode, when reducing leakage current is a primary concern.

One ISSCC example is a processor developed by Nippon Telegraph and Telephone (NTT) for mobile phones. The device is normally in one of two modes: strenuously active when digital encoding and decoding speech or, more commonly, in a sedate state awaiting a call.

The processor features a DSP core and an embedded processor. The DSP core is supplied with 1.1V and is implemented in a low threshold voltage CMOS (V_{th} = 0.25V), whereas the embedded processor is implemented using a higher threshold one.

In the wait mode the DSP is inactive; a high voltage MOSFET isolates it from the supply rail, drastically reducing its leakage current. Here the embedded processor takes over.

Implemented using a higher threshold logic, the embedded processor has a corresponding lower standby current. Moreover, having less to do, it operates at a lower frequency, further saving power.

According to NTT, simply reducing the voltage from 3.3V to 1V reduces the device's energy consumption by one third. Energy consumed being the appropriate measure for the handset. However, employing a multi-threshold logic scheme, energy consumption is reduced to one tenth overall.

Cellular neural network

The world may have gone digital but for applications where high accuracy is not a requirement, an analogue approach can win hands down in terms of speed and power consumption. Moreover if implemented in standard CMOS technology, any requirement to integrate digital circuitry becomes straightforward.

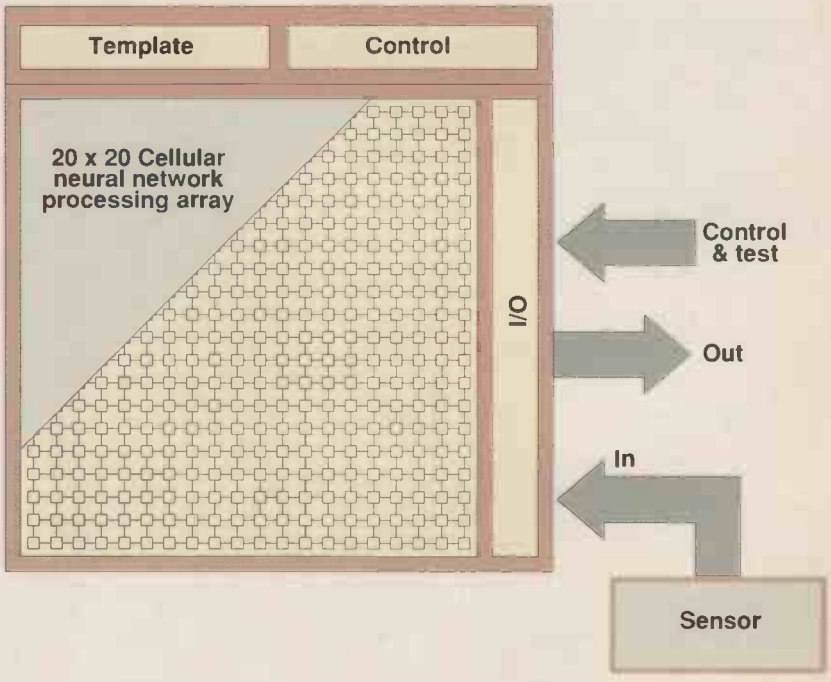
The Katholieke University of Leuven, Belgium has adopted such an approach for telecommunications and analogue signal processing. Taking a cue from biological systems, it has produced a simple multi-cell analogue array suited to image manipulation and sensor data processing for applications such as robot arm control.

The device consist of a 20-by-20 array of simple analogue cells that implements a cellular neural network. Each cell has an input, internal and output node, and is linked to its four nearest neighbours. A set of templates determine the weightings of the signals exchanged between cells. These, coupled with the input data, determine the state of the neural network once processing completes.

The University has developed a library of templates that can be used to program the device to perform a range of applications.

The cells operate in parallel and continuously in time. Moreover, being analogue, the cell circuits work at the full technology bandwidth (f_t).

Processing time is measured in time constants – multiples

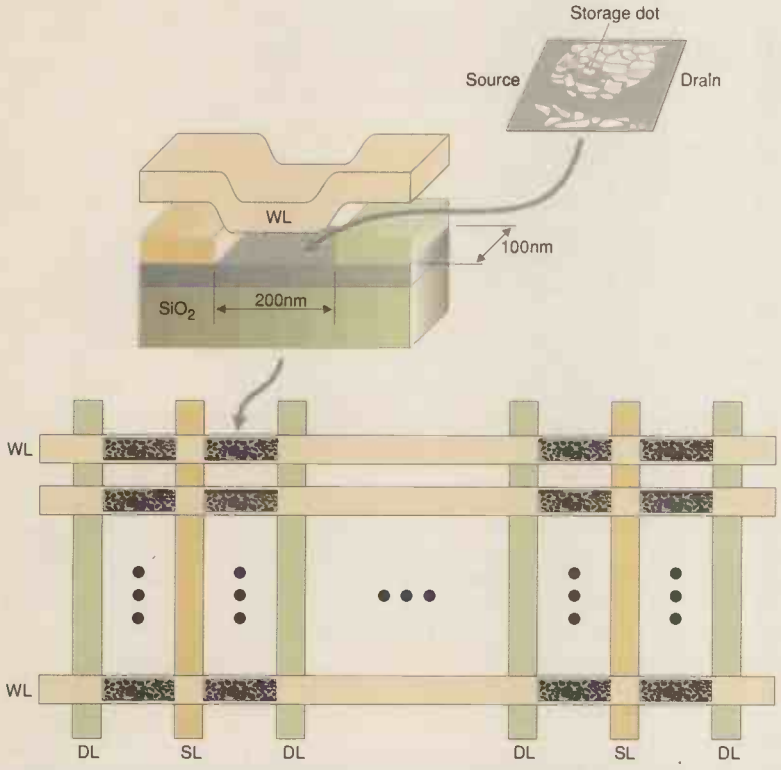


Cellular neural network array. The analogue parallel architecture comprises an array of processing cells arranged in a 20 by 20 matrix. All cells execute in parallel and in continuous time. The device can perform such tasks as edge detection, hole filling and connected component detection.

of 4.8µs. The typical execution time of a non-propagating template is 9.6µs; for the worst case information propagating template it is 145µs.

The device's i/o circuitry can be clocked at 500KHz, enabling the device to process up to 25 image frames/s.

While stressing that a direct comparison with a digital signal processor is not straightforward, the University nonetheless believes the array processor requires up to twenty times less energy (power-delay product) for a given computation.



Single electron memory, proposed by Hitachi, promises terabit storage on one chip. It incorporates a 3nm ultra-thin-film transistor exploiting the Coulomb blockade effect.

Single electron memory cells

The highlight of last year's ISSCC was the emergence of 1Gigabit dynamic rams from Hitachi and NEC. This year Hitachi gave a glimpse of a development which promises storage densities one thousand times greater using single electron memory, or SEM.

Single electron memory has received considerable attention in recent years. First demonstrated at very low temperatures, room temperature has now been attained. The benefit of SEM is its ability to control a small number of electrons, promising reduced power consumption per transistor coupled with significantly greater integration levels due to each transistor's reduced size.

The SEM device uses a 3nm ultra thin-film transistor which exploits the Coulomb blockage effect *Electronics World*, March 1996, p185. The effect works by confining a pool of electrons within a small region such that the stored charge energy is greater than the thermal energy of an external electron. Information is stored by trapping one or more electrons in the pocket and manifests itself in a constricted current.

Hitachi's accomplishment is to be the first to integrate a number of SEM cells to produce an 8-by-8bit array. Moreover, by producing a working device, Hitachi has identified the obstacles to be overcome if volume manufacturing is to occur.

Hitachi's SEM has a 10µs write/erase time. This is faster than flash memory since the number of electrons to be stored or erased is a paltry five compared to 100,000 for flash.

The device's shortfalls include a retention time of between an hour and a day, unacceptably short for nonvolatile store.■

120MHz a-to-d converter in c-mos

Converting a complex envelope signal from rf to baseband, forming in-phase *I* and quadrature *Q* components, is a common requirement for radar and communications applications.

The traditional approach uses cosine and sine heterodynes to separate the *I* and *Q* components before being digitised by matched a-to-d converters, **Figure 1**.

At ISSCC Ericsson and Linköping University detailed a 120Msamples/s a-to-d converter that digitises the baseband components to an accuracy of 10-bits.

The device uses a dual filter approach to separate the components, **Figure 2**. According to Linköping University, implementing the filters using closely matched coefficient values allows its execution within the sampling circuitry of the converter. The consequence is a saving in circuit complexity and power in that the a-to-d conversion is performed at a more leisurely 2MHz rather than at 120MHz.

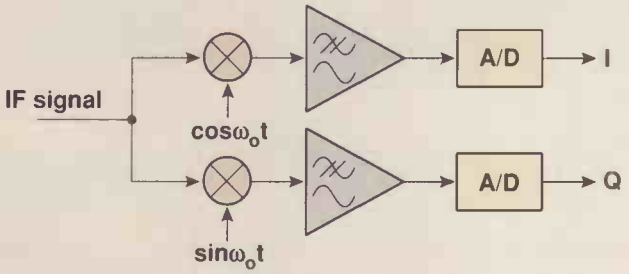


Fig. 1. Classical method for in-phase and quadrature detection. Sine and cosine heterodynes access the complex envelope signal before each arm is low-pass filtered and digitised.

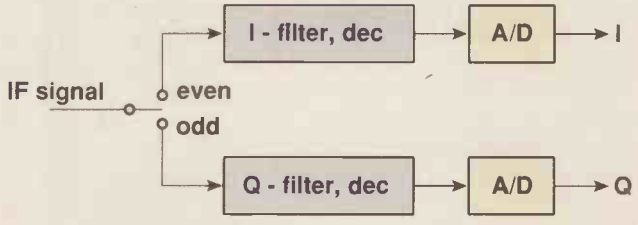


Fig. 2. New a-to-d converter samples at four times the intermediate frequency, undertakes analogue filtering and decimation before digitising the signals at baseband.

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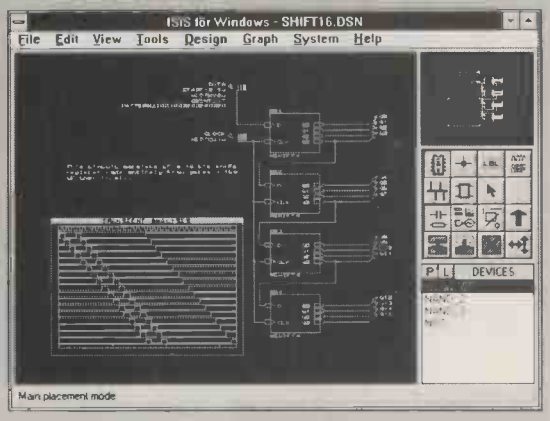

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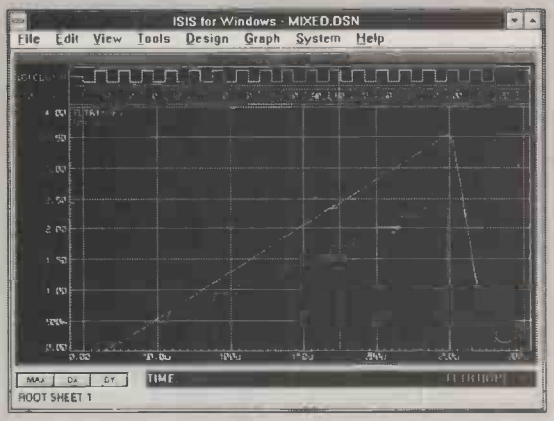

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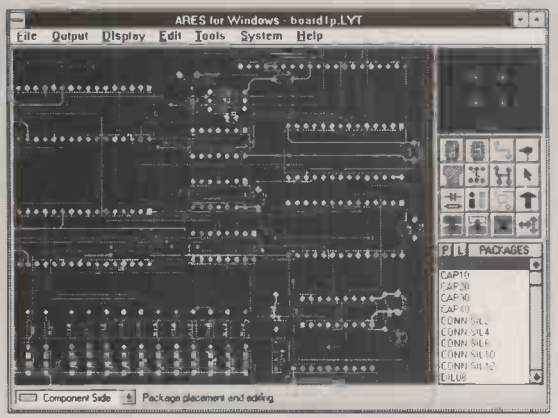

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Relaying **transmission line** principles

Bill Russel demonstrates how rectangular pulses and an artificial delay line simplify the explanation of how transmission lines work.

My previous article outlined a range of simple demonstrative measurements that can be made on an 8µs, 8kΩ artificial line fed from a sine-wave source.

This article examines the effect of applying rectangular pulses to a similar line, using basic test equipment. I constructed a simple battery-powered pulse generator based on a *74HC14* hex schmitt trigger. Since the current drain is only a few milliamps, several hours use can be obtained after each charge.

Layout shown in the upper circuit on page 214 allows for three values of source resistance. The values used give a pulse width of about 2.5µs at a repetition frequency of around 10kHz.

With the source resistance set at 8kΩ, the pulse delivered to a matched line is 3V. Measurements are made with channel 1 on the input and channel 2 on the output, or one of the line taps.

Measurement possibilities of this set-up well exceed the range required for a normal laboratory session. As a result, the examples shown here are limited to recording waveforms at the output or at tap 5. Principles that can be established are as follows

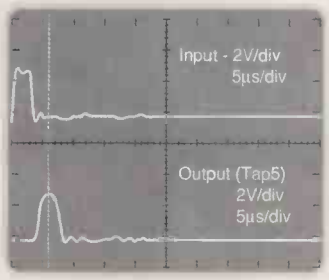
8kΩ source with 8kΩ terminal resistance. Referring to **Fig. 1**, a rectangular pulse of 3V amplitude travels progressively down the line at a speed of 0.84µs per section with little attenuation but some distortion due to the lumped nature of the line. It is accompanied by a current pulse of amplitude 3V/8kΩ, which is 0.375mA.

Some evidence of small reflection reaching the input after 16.8µs, due due to the reactive nature of Z_o.

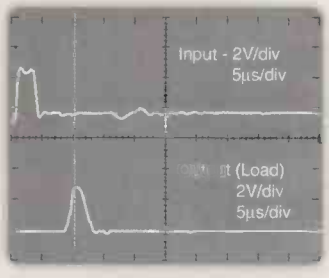
8kΩ source with line open circuit. In **Fig. 2**, complete reflection of the incident 3V pulse takes place at the open-circuit, producing a 6V pulse. The reflected 3V pulse reaches the input 8.4µs later.

Inspection of outputs at taps 1 to 9 shows incident pulses arriving later and reflected pulses arriving earlier until they merge into the 6V pulse at the termination. Note that the display shows only the voltage-time waveform at a particular point in the line, the horizontal axis being time delay in microseconds and not distance along the line. More on this later.

8kΩ source with line shorted. In **Fig. 3**, the incident 3V pulse is completely reflected at the short circuit with reversed polarity. This produces the required zero at the output, and appears at the input 8.4µs later.

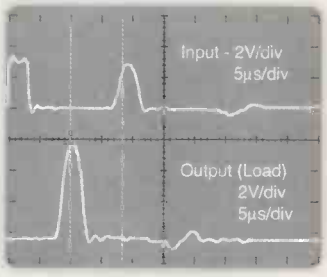


Data position 0 div
dTime 4.20µs
1/dT 238kHz

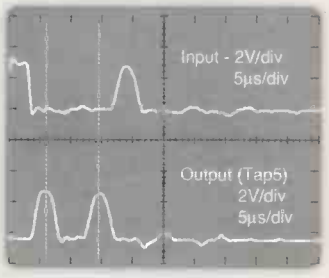


Data position 0 div
dTime 8.40µs
1/dT 119kHz

Fig. 1. The 3V input pulse appears at the output after 8.4µs, with some distortion due the lumped nature of the line, and evidence of a minor reflection at the input after 16.8µs. Output at tap 5 shows the 3V incident pulse arriving after 4.2µs.

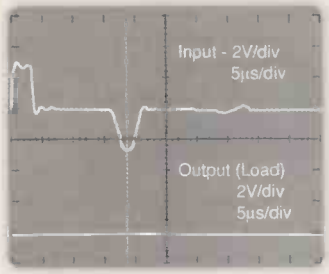


Data position 0 div
dTime 8.40µs
1/dT 119kHz

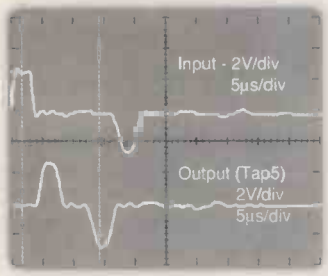


Data position 0 div
dTime 8.40µs
1/dT 119kHz

Fig. 2. Input shows incident 3V pulse, together with a pulse of the same polarity and almost the same amplitude, reflected from the open-circuit termination after 16.8µs. Output shows a pulse of about 6V amplitude, due to the combination of more or less identical 3V incident and and reflected at the termination after 8.4µs. Output at tap 5 shows both the 3V incident pulse after 4.2µs, and an almost identical pulse reflected from the termination after 8.4+4.2=12.6µs. Inspection of outputs at taps 6 to 9 shows incident pulse arriving later and reflected pulse arriving earlier until they merge into the 6V resultant at the termination.

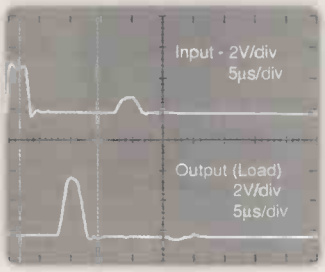


Data position 0 div
dTime 16.80µs
1/dT 59.5kHz

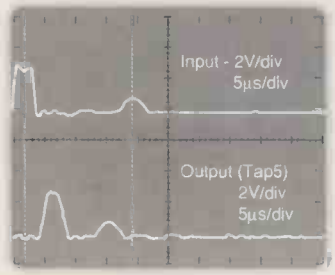


Data position 0 div
dTime 12.6µs
1/dT 79.3kHz

Fig. 3. Input shows inversion of the incident 3V pulse which is reflected at the output and appears at the input after 16.8µs. Output at the short is of course zero, but this can be represented as the combination of a 3V incident pulse with its inverted reflection. This is illustrated by the output at tap 5, which has a 3V incident pulse arriving after 4.2µs together with an inverted 3V pulse reflected from the termination after 12.6µs.



Data position 0 div



Data position 0 div

Fig. 4. Reflected pulse of amplitude 1V appears at the input after 16.8µs. Incident 3V and reflected 1V pulses combined results in a 4V pulse at the termination. Both Incident and reflected pulses are resolved separately in output at tap 5.

Outputs at the tapping points resolve both incident and reversed reflected pulses. The results above show almost complete reflection of an incident voltage pulse at an open or short circuit. They also establish the sign or polarity of reflected pulses.

The following measurements of the magnitude of pulses reflected from loads of $2Z_o$ and $Z_o/2$ can be used to introduce the concept of reflection coefficient, and to deduce its value for any given mismatch ratio.

8kΩ source, 16kΩ load. **Figure 4** shows the 2:1 mismatch at the load end causes the 3V incident pulse to be reflected as a pulse of 1V with the same polarity, producing a 4V pulse at the load.

Output at tap 5 shows incident 3V and 1V reflected pulses. These results indicate that a third of the incident voltage pulse is reflected without change of polarity at a mismatch ratio m of 2:1. This can be shown to agree with the simple formula,

$$\text{Reflection coefficient} = \frac{m-1}{m+1}$$

So far, no attention has been paid to the current pulses implied by the incident voltage pulse on an 8kΩ line. This is because the measurement set up does not allow for their detection. Nevertheless a fair amount of information can be inferred from the known facts.

The current pulse which must accompany the incident voltage pulse of 3V is 3V/8kΩ, which is 3/8mA. At the termination of 16kΩ, the voltage pulse rises to a combination of 3V incident plus 1V reflected without change in polarity giving a 4V resultant pulse. Hence at the termination the resultant current must be 4V/16kΩ, which is 1/4mA.

It would seem reasonable to deduce that, at the termination, a third of the incident current pulse is reflected and inverted to produce a resultant terminal current pulse of 3/8mA–1/8mA, producing the required 1/4mA.

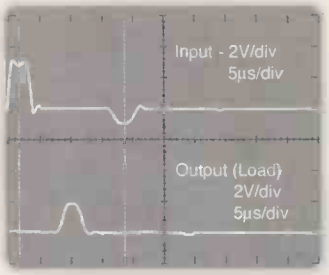
8kΩ source, 4kΩ load. **Figure 5** demonstrates how measurements of input and outputs of a line with a 1:2 mismatch ratio m show a voltage reflection coefficient of 1/3 with reversed polarity. By inference it can be deduced that the current reflection coefficient is also 1/3 but with no change in polarity.

The results obtained can be used to establish some basic rules for a simple treatment of reflections at any resistive termination. As for the current waveforms, the inclusion of a 100Ω current sensing resistor in the return line of both input and termination allows a lot more information to be obtained. However, it is doubtful whether many students would be capable of appreciating the implication of much of this additional data – particularly in the cases where line is mismatched at both input and output ends.

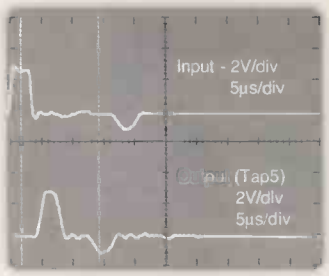
Figures 6, 7 show the voltage waveforms obtained for two of these conditions, and are included with brief comments as examples of situations which would normally be avoided.

4kΩ source mismatch, no load. The 4kΩ source shown in Fig. 6 now delivers a travelling incident pulse of 4V to the line. During the transient phase, this pulse is completely reflected at

Fig. 6. Mismatch at the source results in the incident pulse delivered to the input being about 4V rather than the 3V with a matched source. At the termination the incident pulse is completely reflected producing the pulse of almost 8V at the open-circuit after 8.4µs, and arriving at the input after 16.8µs. Simplified arithmetic of the mismatched input suggests that a third of the reflected pulse will be absorbed – increasing the input amplitude to about 5.3V and two-thirds, or 2.6V, will be inverted and reflected back to the output, arriving after a further 8.4µs. Output waveform shows the increased amplitude at the mismatched input, the large pulse at the open circuit, the 2.6V pulse reflected from the input, plus the first of a series of reflections from output and input.

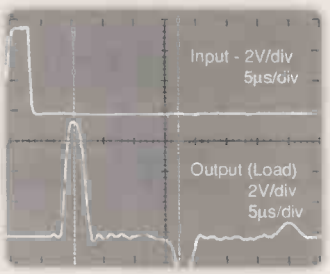


Data position 0 div
dTime 16.80µs
1/dT 59.4kHz

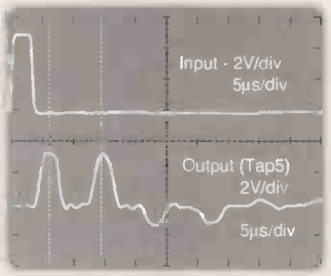


Data position 0 div
dTime 12.6µs
1/dT 79.2kHz

Fig. 5. Reflected 1V pulse, inverted, appears at the input after 16.8µs. Combination of 3V incident and 1V inverted reflected pulses result in a pulse of 2V at the termination. Both incident and reflected pulses appear on the output at tap 5.



Data position 0 div
dTime 16.80µs
1/dT 59.2kHz



Data position 0 div
dTime 8.4µs
1/dT 119kHz

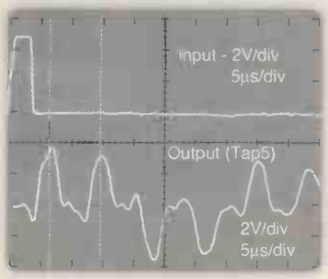


Fig. 7. Worst case condition, where pulses reaching the open circuit are reflected as is, and those reaching the input suffer complete reflection and inversion. Output at tap 5 shows part the series of multiple reflections which then takes place.

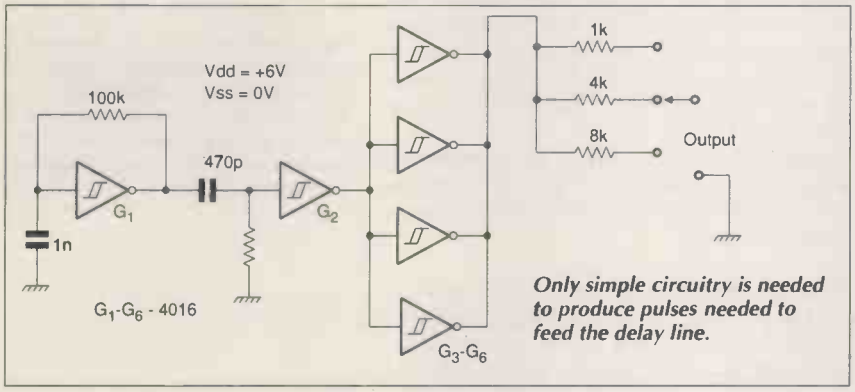
the open-circuit producing an 8V pulse, and the reflected 4V pulse arrives at the mismatched input after 16.8µs.

Waveforms of Fig. 6 are steady state conditions and show no sign of a reflected pulse at the input. Instead, the input shows a final value of input voltage of about 5.3V, plus an inverted pulse of about 2.6V at the output after reflection from the input.

This suggests that when the transient 4V pulse reaches the input mismatch, a third is absorbed increasing the input pulse to 5.3V, and two thirds, or 2.6V, is inverted and reflected back to the output.

Line mismatched at source. In this case, there is a direct connection to the pulse generator via a 50Ω resistor and the load is open circuit, Fig. 7.

Under worst-case conditions, pulses reaching the open-circuit are completely reflected as is. Reflected pulses reaching the input suffer almost complete reflection and inversion. Little, or none, of the pulse energy is absorbed by the generator, or load. The result is that a series of multiple reflections and inversions take place at the generator, accompanied by reflections without inversion at the load. Figure 7 shows part of this series.



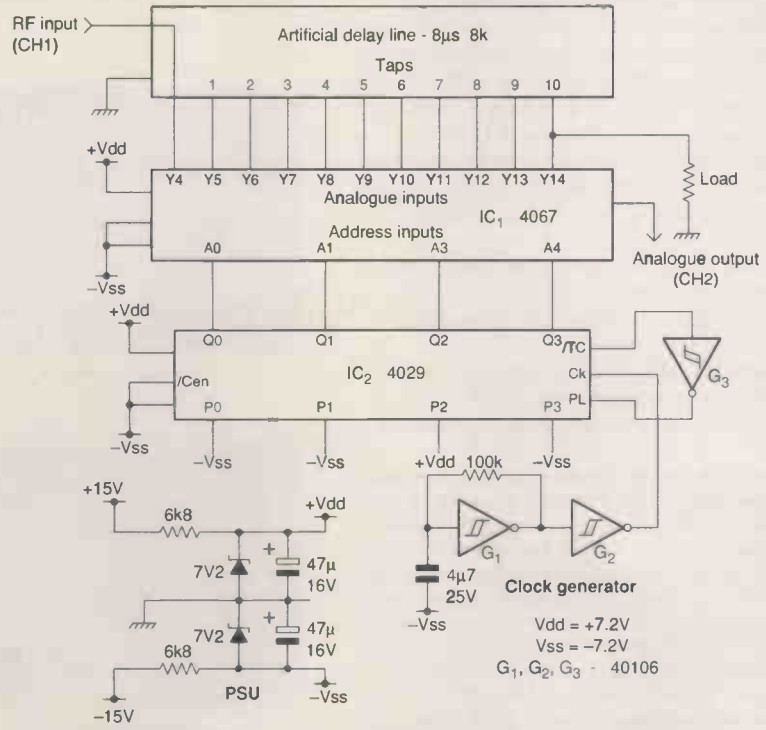


Fig. 8. Adding such a display interface to the delay line allows successive taps to be sampled periodically and displayed as vertical deflections on an oscilloscope .

Extending the idea

Explanatory comments on the above measurements assume a lossless line, and draw on the simple arithmetic of the dc equivalent circuit of the generator, line and load. However, the interest generated encourages many to tackle more rigourous analyses.

For those of you requiring merely a simple introduction to the principles involved, a selection of the more basic measurements should suffice. I have given some thought to the possibility of producing a display in which the horizontal axis represents the voltage at each successive line tap and hence distance along the line.

This problem could be solved by a computer simulation program. But the positive reaction of students who undertook these measurements on an actual line suggested that a hardware solution would be well received.

The main requirement for such a display is that the amplitude of the voltage at the successive taps should be sampled periodically. These voltages should be displayed as a vertical deflection on the oscilloscope. For rectangular dc pulses, the sampled output can be passed direct to the oscilloscope.

In order to cope with dc pulses of both polarities, the sampling device must be operated in the analogue mode.

A prototype circuit along the lines of **Fig. 8,** uses a *4067* analogue multiplex/demultiplexer, driven by a 4029 counter. A 2Hz clock is provided by a *40106* hex schmitt trigger. This device also provides a clock buffer and inverter for the terminal-count output to preset the counter to state 4.

The counter and hence the demultiplexer cycles continuously from states 4 to 14, giving 11 sampled lines. These lines are connected to the artificial line input and the 10 taps.

Channel 1 of the oscilloscope connects to the line input for triggering purposes only. The common output of the *4067* is simply connected to channel 2. The display is really a montage of the voltage time waveforms at a particular tap, updated at half-second intervals to the adjacent tap. It produces the illusion of incident pulses moving from left to right, and reflected pulses moving from right to left.

Where pulses meet, reinforcement or cancellation takes place depending of course on relative amplitude and polarity. The system is operated from a dual 7.2V supply as shown. As a result, it imposes a limit of less than 7.2V peak on the sampled input. This is ample to accommodate all waveforms shown in this article.

Used with a large display oscilloscope, a generator and an artificial line modified so that source and terminating resistors can be altered quickly by switches, this simplified display has proved surprisingly effective in summarising the working principles involved. ■

Further reading

Millman & Taub, *Pulse and Digital Circuits*, Chap. 10.

Oops...

In last month's article please note the following corrections: the caption for Fig. 4. refers to the plots of Fig. 6, the caption for Fig. 5 refers to Fig. 4 and the caption for Fig. 6 refers to Fig. 4. In Fig. 11, input current is 0.5mA, not 1mA. Sorry.

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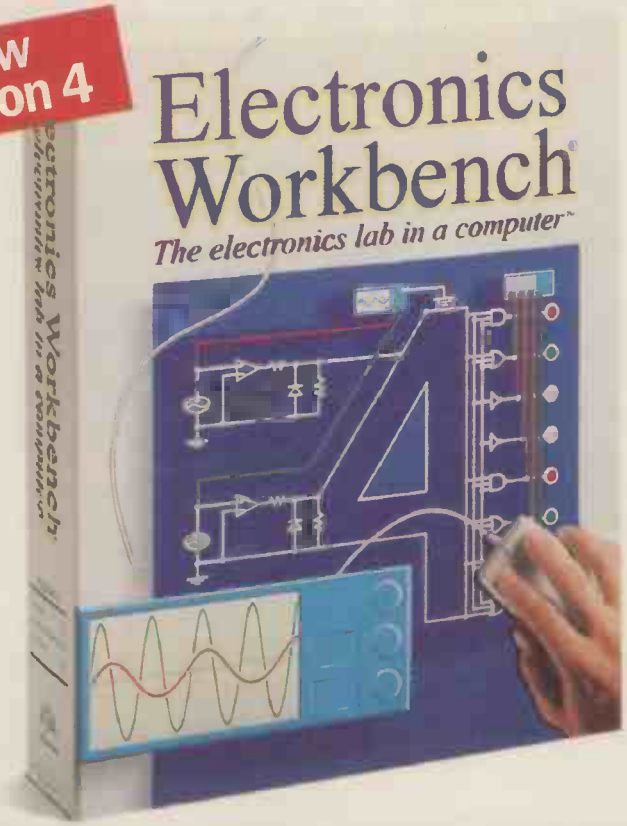

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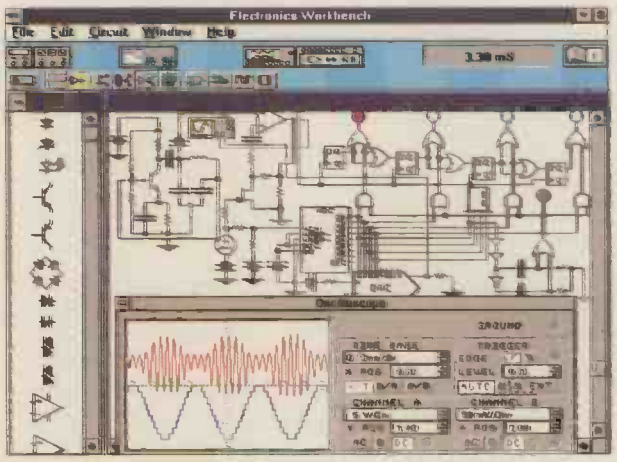

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INTERACTIVE

CIRCLE NO. 115 ON REPLY CARD

Hands-on Internet

Cyril Bateman discusses how Archie and Gopher help you search for files on the Internet.

In order to use the File Transfer Protocol described in the last issue, two descriptions are essential – the location of the required file and the file name.

Internet is huge, and to locate specific files it is necessary to understand and use the established methods and protocols. The desired file can contain anything capable of storage in a computer. Obviously, program software is the most common. But the possibilities are almost endless, from the script of a Shakespeare play or the Dead Sea Scrolls to views from the Hubble telescope or a piece of music[1].

If the file name is known, maybe only vaguely, its location is easy to find; however, the file name is usually unknown.

The one essential document 'Anonymous FTP: Frequently Asked Questions (FAQ) List' is available for down loading from a number of sites.

When you are equipped with FTP and a search tool, every facility becomes possible. As with most computer actions the most difficult part is starting out, which these articles seek to address.

For 'surfers' of the Internet, two different search engines are readily available – Archie & Gopher. These are designed for use as 'local clients' on your personal computer. They are available as starter kits or you can down load them from Internet. By having access to Internet with FTP and carrying out the procedures described here, then all other packages can easily become accessible.

Searching with Archie...

The oldest search tool – Archie – is effectively a card index for FTP files. It was developed at McGill University, Montreal for searching all available Unix based computer archive sources of directories and file names. The name Archie is derived from archive[1].

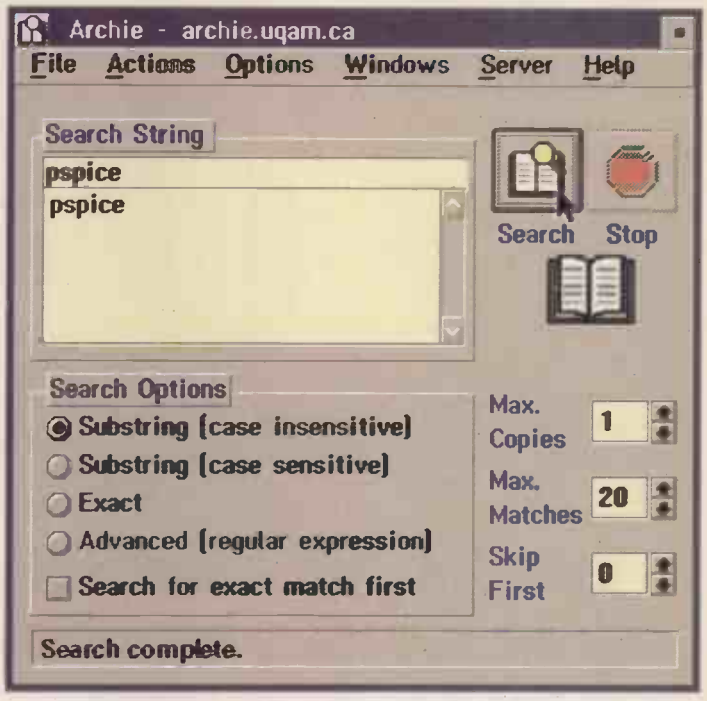


Fig. 1. Using Archie to search for the location of 'PSpice' software file. Search for 'PSpice' using the Archie server located at 'archie.uqam.ca.' in Canada. Note the 'aid memoir' display of used search strings.

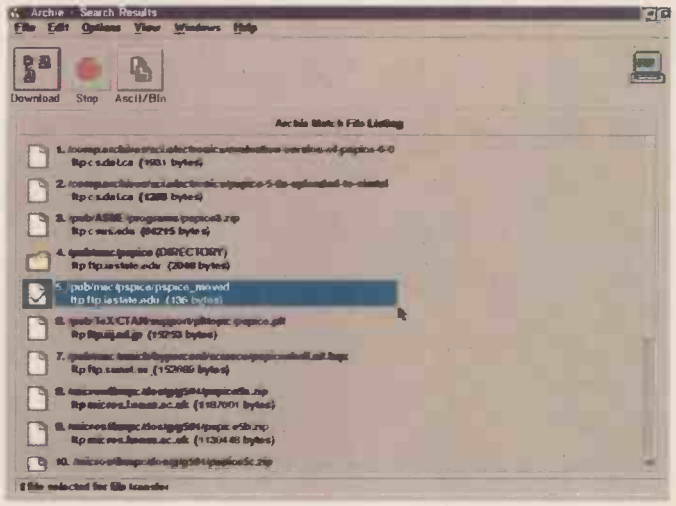


Fig. 2. Using Archie to search for the location of 'PSpice' software file. Result of Archie search for 'PSpice' using the search string 'PSpice'. Interrogation of the highlighted file revealed two locations for the required software. These locations were used for the FTP example in the previous article.

Archie servers search all the 1000 plus Unix-based computers comprising the Archie database archive of FTP files. These servers are periodically automatically updated. In theory, all the servers hold the same information, but due to the updating sequences, this is not absolutely true.

Archie searches are restricted to a directory name or file name. This name can be incomplete, since Archie looks automatically for near matches, and certain 'wildcards' are allowed. Having located the desired file, either Archie or FTP can be used for the download, **Figs 1** and **2**.

All anonymous FTP sites, Unix and non-Unix based, are identified in the Anonymous FTP Sitelist, however since this is an extremely large listing, be prepared for a lengthy download session[2].

... and later with the Gopher

The newer search tool – Gopher – was developed at the University of Minnesota in 1991. While Archie is a single line, single word search at the chosen server, Gopher is menu based, allowing more flexibility and by default searches the contents of all Gopher servers, which is known as 'GopherSpace'. Two variations are included in the search engine, Veronica developed at the University of Nevada and Jughead. Both support Boolean controls and multi word search strings, **Figs 3** and **4**.

To avoid excessive numbers of matches, Veronica and Jughead are best used with multi word search strings. While the desired Boolean controls can be specified, the default for two or more words assumes the implicit 'and'[3].

A Veronica search of the 5000 plus Gopher servers, offers two predefined styles, Fig. 3.

- Find Gopher directories by title word(s) via xxx. This search will find only Gopher directories whose titles contain your specified search words. This is used to find major holdings of relevant information. Having selected a directory it can be 'opened' to show contents.

- Search GopherSpace by title word(s), via xxx. This search will find all types of resource whose titles contain your specified search words.

Jughead searches, like Archie, are restricted to individual locations and are distinguished from Veronica searches by the description 'Search GopherSpace AT xxx' as distinct from 'via xxx'.

Use of the multiword search with implicit 'and', together with the '*' wildcard permitted at the end of a partial word, can provide a tightly focused query and return only the more relevant matches[3].

Equipped with FTP, Archie, and Gopher, any publicly available Internet FTP resource can be located and accessed for file transfer, since it is these protocols which form the basis of the various WWW search engines. ■

References

1. Surfing with intent, *EW&WW*, June '95, pp. 488/492.
2. Anonymous FTP-FAQ. See panel, 'Frequently asked questions'.
3. How to compose Veronica Queries. See panel, 'Frequently asked questions'.

Frequently asked questions

Frequently asked questions articles, called 'FAQs' are readily available for all Internet activities, and should be the first point of reference for any help needed.

For this reason they are widely available, and can be obtained by 'E mail' requests, as well as from the relevant NewsGroups or by anonymous FTP.

Anonymous FTP FAQ
Newsgroups
news.newusers.questions.
news.announce.newusers.
alt.sources.wanted.
comp.archives.
comp.archives.admin.
comp.sources.wanted.
alt.answers.
comp.answers.
news.answers.
FTP
garbo.uwasa.fi pc/doc-net/ftp-list.zip
oak.oakland.edu /SimTel/msdos/info/ftp-list.zip

Archie FAQ
FTP
archie.mcgill.ca archie/pub/archie.faq

Gopher FAQ
Newsgroups
comp.answers.
news.answers.
FTP
rtfm.mit.edu /pub/usenet/news.answers/gopher-faq

Veronica. - how- to- query- veronica
Gopher://Veronica.scs.unr.edu how-to-query-veronica

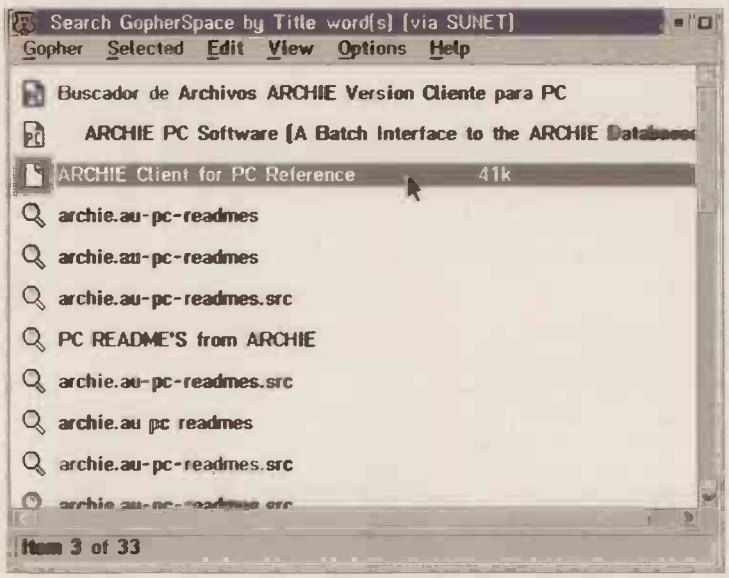


Fig. 4. Using Gopher to search for the location of 'Archie Client' software file. Result of search using the multi word search string 'Archie PC Client'. Further searches using different search strings or different servers will be needed.

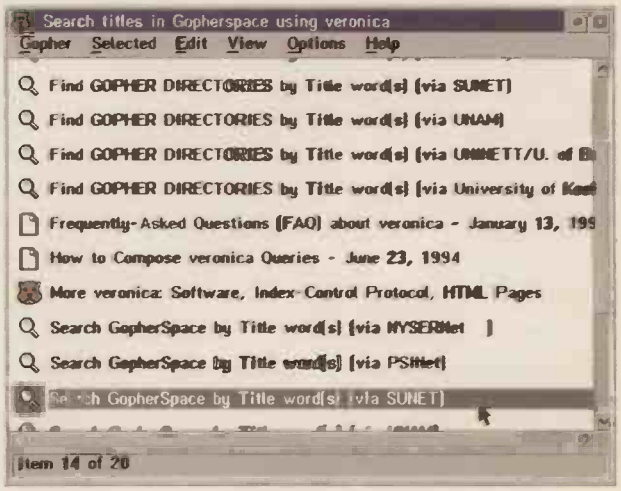


Fig. 3. Using Gopher to search for the location of 'Archie Client' software file. This illustrates just a few of the menu options available for a Gopher search. Note the two main search options discussed and the ready prepared popular Gopher servers. Many other servers throughout the world are also available from other menu selections. Note also the menus provided to supply the two required documents, 'veronica FAQ' also 'How to Compose veronica Queries'. Simply click on the highlighted selection to 'pop-up' the search box.

Designing an SSB outphaser

Outphasers for SSB transmitters demand accurate component values, but analyses of such circuits are rare. David Gibson not only presents such an analysis, but also explains how he has extended the outphaser's scope.

An algebraic analysis of an outphaser, also called a phaser or Hilbert transformer*, is difficult and is not often discussed – even in otherwise comprehensive filter textbooks. The component values are largely folk-lore, passed on from application to application.

You may say that 'if it ain't broke, don't fix it', but an analysis is useful for several reasons – not least because it allows you to check whether circuit values have been transcribed correctly. I have seen examples where this was clearly not the case.

In this article I present networks using op-amps and simple first-order networks. These are easier to adjust than conventional passive second-order networks, as well as being easier to study. This makes it possible to design more accurate networks, or ones with a wider bandwidth for applications in music, audio effects. It also allows frequency shifting, which may required for applications such as spectrum analysis and sonar processing. In addition to presenting analogue networks, I show an example using digital signal processing techniques.

I will not give a detailed mathematical analysis due to its complexity. Most of my work was done with simple Basic programs which plotted phase and amplitude responses. Using this method I was able to tweak the component values to produce some very accurate filters. This method also made it easy to investigate the effects of component tolerances and drifts.

SSB modulation background

The heart of an single-sideband modulator or demodulator is a circuit with the ability to shift a range of frequencies from the audio band to rf, or if. The simplest way to do this is to amplitude-modulate the signal onto a carrier using a balanced modulator.

The unwanted sideband and any residual carrier are removed in a crystal filter. This method has an advantage, namely it is conceptually simple, but also has disadvantages. It can be difficult set up the filters to adequately attenuate the unwanted sideband, and it is inflexible.

A second method is to use an 'outphaser' which is the subject of this article. There is also a third method. Before discussing the outphaser, I will say a little bit about this because, depending on the application, there is sometimes little to choose between these two methods.

The 'third' method

This third method for removing unwanted sideband and residuals was first described by Weaver in 1956, and modified by Turner, writing in *Wireless World* in 1973. In this method, **Fig. 1**, an audio signal is first modulated onto quadrature carriers at a fixed 'intermediate' frequency. The upper sidebands of the two channels are filtered out, leaving the lower sidebands which are in phase quadrature, **Eqn 1**.

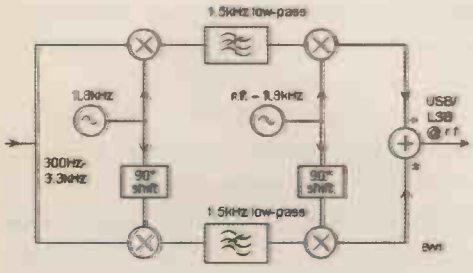


Fig. 1. 'Third method', due to Weaver and Turner. Lower sidebands in phase-quadrature at if are modulated onto an rf carrier, and summed. The unwanted sidebands cancel leaving an ssb signal. The same circuit is used for demodulation, where the salient point is the extremely low if of 1.8kHz which eases the filtering requirements as explained in the text.

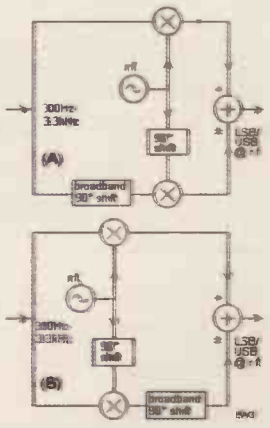


Fig. 2. Outphaser method. Audio input is shifted by 90° and modulated on to two quadrature carriers a). The signals are summed and the unwanted sidebands cancel. The outphaser can alternatively be placed after the modulators b). It is easier to construct a broadband 90° network at rf than at audio (see text) but, when it comes to changing the rf, it is less flexible.

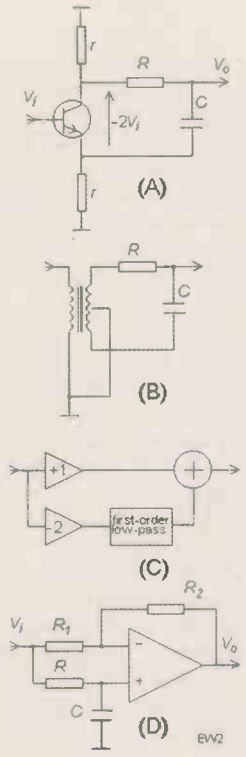
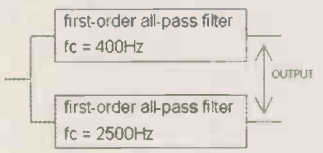


Fig. 3. First-order all-pass filters.
a) Historical filter using transistor; b) Version for use at rf;
c) Functional diagram;
d) Implementation with op-amp.



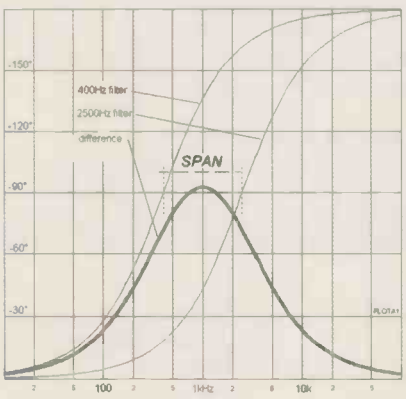


Fig. 4. Difference of one pair of first-order Filters
a) In an attempt to increase the usable frequency range, we utilise the difference between two filters.
b) The phase response of the two filters in a), and the difference of the two. The range of frequencies for which the phase difference is 90° can be improved further by cascading pairs of all-pass filters.

$$\underbrace{\sin\omega_m t}_{\text{audio modulating signal}} \times \underbrace{\sin\omega_i t}_{\text{if carrier}} = \underbrace{\frac{1}{2}\cos(\omega_i - \omega_m)t}_{\text{lower sideband}} - \underbrace{\frac{1}{2}\cos(\omega_i + \omega_m)t}_{\text{upper sideband (removed by filtering)}}$$

and similarly

$$\sin\omega_m t \times \cos\omega_i t = -\frac{1}{2}\sin(\omega_i - \omega_m)t + \frac{1}{2}\sin(\omega_i + \omega_m)t \qquad (1)$$

$$\underbrace{\sin(\omega_i - \omega_m)t}_{\text{lower sideband at if}} \times \underbrace{\sin(\omega_c - \omega_i)t}_{\text{rf carrier}} = \underbrace{\frac{1}{2}\cos(\omega_c - \omega_m)t}_{\text{lower sideband in rf}} - \underbrace{\frac{1}{2}\cos(\omega_c + \omega_m)t}_{\text{upper sideband at rf}}$$

and similarly

$$\cos(\omega_i - \omega_m)t \times \cos(\omega_c - \omega_i)t = \frac{1}{2}\cos(\omega_c - \omega_m)t + \frac{1}{2}\cos(\omega_c + \omega_m)t \quad (2)$$

Equations 1 & 2 describe the 'third' method of ssb generation

The next step is to take the intermediate frequency signals and to modulate them onto quadrature carriers at rf – or more precisely, at the difference between the rf and the intermediate frequencies, **Eqn** 2.

Each of the channels provides an upper and lower sideband at the final rf. The crucial aspect of this is the phase of the signals. From equn 2 you can see that, if the signals are added, the upper sidebands will cancel, leaving only the lower sideband. Likewise, if you subtract the signals you get only the upper sideband.

The advantage of this method is that, by using a fixed intermediate frequency, you ease the problems of filtering the unwanted sidebands. If you choose a very low intermediate frequency, then a simple audio low-pass filter will suffice.

However, the salient point of the Weaver method arises when demodulation is considered. The implementation in Fig. 1 can be used for demodulation simply by swapping the order of the two modulators. Alternatively it would be possible to demodulate directly to baseband, but this would require a highly selective filter to remove the unwanted sideband. The Weaver method uses an intermediate frequency within the audio band, at 1.8kHz.

By choosing the lowest possible intermediate frequency, so that the wanted signal 'wraps round' at zero frequency, the filtering requirement changes from a band-pass filter to a simple low-pass audio filter. Additionally, the low frequency means that the filtering is less stringent, though with an eighth-order filter (48dB/octave) would still only give 24dB attenuation at 2.1kHz, from a cut-off at 1.5kHz.

If the audio band is 300-3300Hz, the low intermediate frequency results in each channel having an upper if sideband at 2.1-5.1kHz, which is filtered out. There is also a lower sideband extending from minus 1.5kHz to plus 1.5kHz.

The concept of a negative frequency can be confusing. Physically, it appears as a 'normal' 1.5kHz, and the information that it is 'negative' comes from the relative phases of the two signal channels. The two channels contain information about the original upper and lower rf sidebands. By adding or subtracting the signals you can cause one or other of the sidebands to cancel out, providing the required information.

One aspect of the Weaver method is that the modulators have to be ac coupled to prevent dc bias from manifesting itself as a 1.8kHz tone. The ac coupling means that there is a notch in the audio response. However, this can be made narrow enough to be un-noticeable.

The modification suggested by Turner in 1973 involved digital modulation techniques. The carriers can be square waves, and the modulators, certainly at low frequencies, can be transmission gates. At vhf it is possible to rely on the harmonic content of the square waves to generate the rf signal. Additional harmonics present throughout the circuit do not cause a problem because they either cancel out, or are filtered.

Sometimes, the audio demodulation is done with a stepped square wave. One implementation is known as a rotary mixer. The third, and some higher, harmonics are absent in a correctly stepped sine wave, which eases the filtering requirements. The size of the steps in the sine wave can be calculated using Walsh functions.

The Weaver/Turner technique was discussed by Hamilton in this magazine in 1993 and was used in a design by Dorey in 1994.

Phasing in SSB designs

As with the Weaver method, the basic idea behind the phasing method is to generate two double-sideband channels where one of the sidebands is in antiphase and can be cancelled out, **Fig.** 2. An rf carrier is modulated directly to produce the sidebands described below.

$$\underbrace{\sin\omega_m t}_{\text{audio signal}} \times \underbrace{\sin\omega_c t}_{\text{rf carrier}} =$$

$$\underbrace{\frac{1}{2}\cos(\omega_c - \omega_m)t}_{\text{lower sideband}} - \underbrace{\frac{1}{2}\cos(\omega_c + \omega_m)t}_{\text{upper sideband}} \quad (3)$$

For the second channel the audio signal is passed through a broad-band phase-shift network which alters its phase by 90° at all frequencies, without altering its amplitude. It is

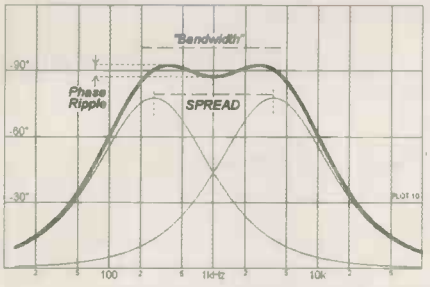


Fig. 5. Difference of two pairs of first-order sections, example 1. Phase ripple is three over a 'bandwidth' of around 200Hz to 5kHz

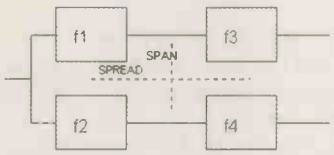


Fig. 6. Two pairs of first order filters. Difference between the two outputs approximates to a 90° phase shift. The filter is described by the span (ratio f_2/f_1, f_4/f_3) and the spread (f_3/f_1, f_4/f_2)

then modulated onto a quadrature carrier to produce a further pair of sidebands.

$$\cos\omega_m t \times \cos\omega_c t = \frac{1}{2}\cos(\omega_c - \omega_m)t + \frac{1}{2}\cos(\omega_c + \omega_m)t \qquad (4)$$

Now, by adding or subtracting the signals it is possible to cancel one or other of the rf sidebands, **Fig. 2a**. It is also possible to swap the order of the components and use the phase-shift network at rf, **Fig. 2b**.

There is not a lot to choose between the Weaver and phasing methods. The Weaver method is slightly more complex in terms of circuitry and frequency control. However, the phasing method needs some accurate components in the rather special phase-shift network.

The phasing method can be used in applications other than 3kHz audio. As I will show, a simple network can be used at rf, and the technique can be used to shift a wider band of frequencies – say 20kHz audio – for music applications. A small shift of 5-10Hz can be used to prevent 'howl-around', while a larger shift can be used for special effects.

Designing the phase-shift network

An integrator or differentiator achieves a 90° phase shift, but has a varying gain with frequency. For 90° phase shift and constant gain, a more complex network is required. It can be proved that a 'perfect' outphaser, which works at all frequencies, is physically impossible to construct†. Thus, any network we construct must be a compromise.

Many outphaser designs of have appeared over the years. It is interesting to look at different designs and to trace their origins by the obscure component values they use – a sort of electronic equivalent of genetic markers. Some designs which have appeared in this magazine are due to Hickman (1991) who reviewed some outphaser and Weaver circuits; Hosking (1994) who described the so-called 'polyphase' network; and, most recently, Green & Hosking (1996) who presented a polyphase receiver design.

The polyphase network is an old solution to the problem. It is something of a sledgehammer approach, which I will not discuss further here. Instead, I will show how an outphaser circuit can be built from simple op-amp filters to achieve varying degrees of sophistication.

First-order network

A simple first-order *RC* low-pass filter has a phase shift of 45° at its –3dB frequency, ω_0. Two networks would result in 90°, but the gain varies with frequency. However, by driving the 'bottom' of a first-order network with an inverted signal, **Fig. 3**, you can get a 90° shift at ω_0 and constant gain. This response is called a first-order all-pass filter. An all-pass filter has a flat amplitude response, but the phase shift varies with frequency.

Figure 3 shows several ways of generating the response. Op-amps are cheap enough, so the method of **Fig. 3d** is the one I prefer. Resistors R_1 and R_2 set the overall gain, whilst R and C set the centre frequency to $\omega_0=1/CR$. I don't want to include too much maths in this article, but it is useful to note that the transfer function, in complex frequency, is,

$$\frac{V_o}{V_i} = \frac{1 - \frac{R_2}{R_1} j\omega/\omega_0}{1 + j\omega/\omega_0} \qquad (5)$$

If R_1=R_2 then this expression shows a unity gain, and phase shift φ defined from,

$$\tan\frac{1}{2}\varphi = -\frac{\omega}{\omega_0} \qquad (6)$$

If the phaser were used at rf, Fig. 2b, in a direct-conversion radio, then its performance might well be satisfactory. The equation above shows that, in the 50m band, at 6MHz, it is possible to maintain a 90° shift to ±3 over a bandwidth of 600kHz. However, a simple all-pass filter is not adequate for use at baseband. With a centre frequency of 1kHz, the variation in phase shift over the audio band of 300Hz to 3kHz would be an enormous –33° to –143°. We need to resort to higher-order sections, or to chains of filters, as I will now describe.

Multi-section filters

Instead of building a single filter with a phase shift of 90° it is easier to build a pair of filters where the difference in phase shift is around 90°. **Figure 4a** shows an example. You could use two first-order filters, as in Fig. 3d, with centre frequencies of 400Hz and 2500Hz. **Figure 4b** shows how the phase shift of each filter varies with frequency.

There is a band, centred at around 1kHz, where the difference in phase shift is close to 90°. With this arrangement an accuracy of ±3° can be achieved from 630Hz to 1600Hz, or 2.5:1. This is still not large enough for speech, where perhaps 20:1 is required, so the principle needs to be extended, as demonstrated in Fig. 4, to higher-order filters.

A common configuration is to use passive second-order filters. It is very rare to see any analysis of such a circuit, though Walters, in 1986, went some way towards explaining the design process.

Occasionally, active second-order all-pass filters are seen. A classic one was presented by Holt & Grey in 1967, and another version given by Gibson in 1992, but these are difficult to set up, and to analyse.

A historical reason for the use of passive second-order filters is that they were easier to construct than passive first-order filters. Fig. 3b gives an example. Nowadays, op-amps are cheap, and make life much easier because active first-order filters are simple and conceptually easier to analyse.

Required accuracy

Before discussing these enhanced filters, you need to obtain some idea of the accuracy required. Phase shift needs to be 90° and the amplitude difference between the outputs of the

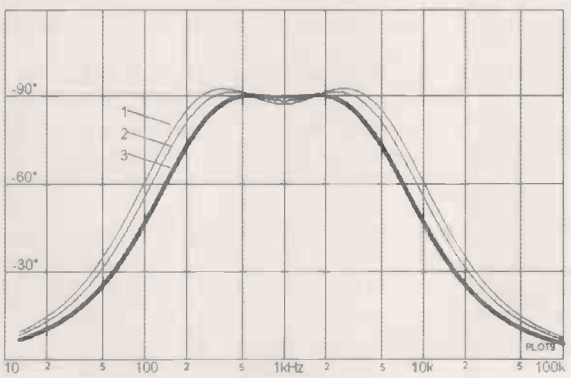


Fig. 7. Difference of two pairs of first-order sections, examples 1-3.

two filter paths should be zero at all frequencies. Unless this is the case you will not achieve perfect attenuation in the unwanted sideband.

Obtaining an expression for the 'leak-through' of the unwanted sideband is straightforward but intricate. You begin by defining the phase error between the two channels to be ϕ degrees. Assuming that one channel has a gain which is a fraction α too high, and the other is α too low.

Clearly it is always possible to represent the gains in this symmetrical way because the absolute gain is less important. Provided that α is 'small', i.e. less than 10%, you can write the ratio of the amplitudes as $1+2\alpha$. Voltage attenuation in the unwanted sideband, V_1, can then be written (Gibson, 1992), relative to the voltage of the wanted sideband V_2 as,

$$\frac{V_1}{V_2} = \sqrt{\frac{\sin^2\frac{1}{2}\phi + a^2\cos^2\frac{1}{2}\phi}{\cos^2\frac{1}{2}\phi + a^2\sin^2\frac{1}{2}\phi}} \qquad (7)$$

Now if ϕ is small too, say less that 10°, it is possible to approximate to,

$$\frac{V_1}{V_2} \approx \sqrt{\left(\frac{\pi}{360}\right)^2\phi^2 + a^2} \qquad (8)$$

For example, if you can maintain the phase error to 8°, and amplitude α to 7%, then both errors contribute equally to the 'leak-through'. The unwanted sideband will be at a voltage level of 1/10.1 of the wanted sideband, or -20dB. An angle of 0.8° and an error of 0.7% would give –40dB.

Note that you will need tight tolerance components in order to achieve this level of performance. Usually, the attenuation is obtained from a combination of rf filtering and outphaser performance. This results in a good overall response with neither item being critical.

A difference of two pairs

Figure 4 showed how you could use the difference of one pair of first-order filters. Extending this to two pairs of filters is straight-forward. Each of the pairs gives rise to a 'hump' in the phase response similar to that shown in Figure 4b. If you place the two humps at the correct separation in frequency, their effects add to give a response with an almost flat top, **Fig. 5**.

Figure 6 shows how the outphaser is configured. There are four first-order all-pass fil-

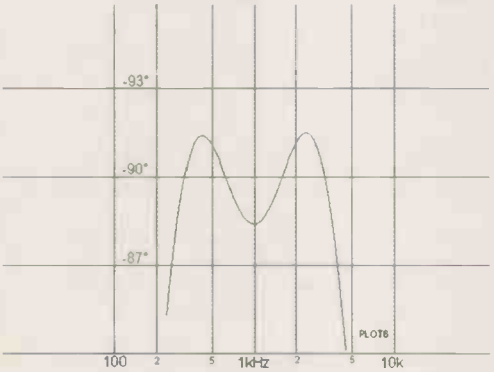


Fig. 8. Difference of two pairs of first-order sections, magnified central portion of response, component values from example 2a.

Notes on the maths

Equation 6, giving the phase shift for a single first-order all-pass filter, can be used as the basis for phase plots. If you are manipulating the equations on paper then equation 9 is a useful short-cut. Its derivation is as follows.

The two filters in the pair, Fig 4, have centre frequencies ω_1 and ω_2. The phase difference, ϕ, comes from

$$\frac{1}{2}\phi = \frac{1}{2}(\varphi_1 - \varphi_2)$$

$$= \arctan\left(\frac{-\omega}{\omega_1}\right) - \arctan\left(\frac{-\omega}{\omega_2}\right) \qquad (A1)$$

Making the substitutions for span and Ω discussed in the main text; taking the tangent of both sides of (A1); and recalling the identity:

$$\tan(a \pm b) = \frac{\tan a \pm \tan b}{1 \mp \tan a \tan b} \qquad (A2)$$

produces equation 9 in the main text. This operation can be applied repeatedly as we chain the filter pairs, but the notation gets rather difficult to follow.

A full analysis should aim to give span and spread in terms of a specified phase ripple and 'bandwidth' rather than simply giving phase as a function of frequency. We can differentiate the expression to find the frequencies of the peaks of the phase response – the turning points of the curve.

By specifying the phase shift at these points to be $^1/_2\phi$ above 90°, and the central trough to be at $^1/_2\phi$ below 90° it is possible to simplify the procedure – although it is still rather difficult. You could differentiate (9) directly, but it is easier to start with (6) and write,

$$\tan\frac{1}{2}\varphi = \Omega \Rightarrow \frac{d\varphi}{d\Omega} = \frac{2}{1+\Omega^2} \qquad (A3)$$

It is now possible to combine expressions for $d\varphi/d\Omega$ for each filter and set to zero to find the turning point. This is tedious and tends to indicate that a computer analysis would save time.

Footnotes

*That is, a device which implements a Hilbert transform. This is one of a number of integral transforms. The Fourier and Laplace transforms belong in this category.

†Take a square wave and look at the phase and amplitude of all its harmonics. If the fundamental has unity amplitude then the amplitude of the resultant square wave is,

$$1 - \frac{1}{3} + \frac{1}{5} - \frac{1}{7} + \frac{1}{9} - \frac{1}{11} + \ldots = \frac{1}{4}\pi$$

Now shift each harmonic by 90° and try to reconstruct the waveform. You end up with a series of the form,

$$1 + \frac{1}{3} + \frac{1}{5} + \frac{1}{7} + \frac{1}{9} + \frac{1}{11} + \ldots \rightarrow \infty$$

The sum increases logarithmically and does not converge, so the resultant amplitude of the waveform is infinite. Thus it is shown that a perfect outphaser cannot cope with this specific waveform. It can be inferred that it cannot cope with a generalised waveform, and so a practical 'perfect' outphaser cannot be constructed.

††Note that finding a flat-top response in the phase domain using multiple all-pass filters is similar to the more conventional problem of finding a flat-top in the amplitude domain when using multiple tuned circuits.

ters with centre frequencies f_1, f_2, f_3 and f_4. These are arranged in two paths, and the wanted signal is the difference between the two.

If only the f_1/f_2 pair were used, the phase difference would be the left hand 'hump' in Fig. 5. Using only the f_3/f_4 pair would give rise to the right-hand curve. Overall phase difference is found by adding the responses to give the third curve on the graph††.

One pair of 2nd-order filters

As I have said, traditional outphaser designs tend to use a single pair of passive second-order filters instead of two pairs of active first-order filters. A first-order network has several advantages over the more complex second-order network,

- It has unity gain, with $r_1=r_2$ so there is no need to set an accurate non-unity gain.
- It only has a single C so this can be chosen for cost and availability. You don't need to choose two accurately matched capacitors in E24 values.
- With C fixed, the only component which affects the centre frequency is R.
- The gain can easily be trimmed to unity by altering r_1 or r_2.
- If – and only if – the gain is trimmed to unity, the phase-shift only depends on R and C. This 'orthogonality' makes simulation and analysis easier, as well as the setting-up.

Using two pairs of first-order filters

To describe the dual first-order filter of Figs 5 and 6, I use two terms. The *span* is the ratio of the centre frequencies of the two filters which comprise a 'hump' in the phase response, i.e. f_2/f_1 and f_4/f_3. The *spread* is the ratio of the centres of the humps themselves.

For a single pair of filters, Fig. 4, you can write the phase shift in a similar way to (6), as

$$\tan\frac{1}{2}(\varphi_1-\varphi_2)=\frac{\sqrt{1/\gamma}-\sqrt{\gamma}}{\Omega+\frac{1}{\Omega}} \qquad (9)$$

where Ω is ω/ω_c is the 'normalised' frequency and $\lambda=\omega_2/\omega_1$ is the span of the pair of filters, with $\omega_c=\sqrt{(\omega_2\omega_1)}$. When you try to extend the analysis to cope with two pairs of filters it becomes difficult to represent them concisely – especially when you want to use the equations to find out what values of span and spread to use.

Fortunately, iterative computer techniques are now possible, and are just as valid. I used a set of small Basic programs to investigate the filters by 'trial and error'.

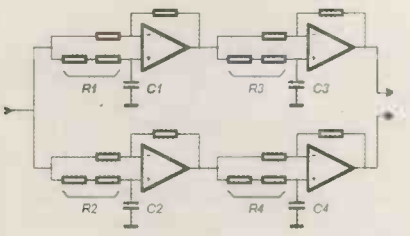


Fig. 9. Circuit of outphaser using difference of two pairs of first-order networks (see text).

Filter examples

Example 1. The filter in Figure 5 is centred around 1000Hz and the spread is 14. Thus the two humps are at f_{12}=267Hz and f_{34}=3742Hz, so that their ratio is 14:1, and the geometric mean, $\sqrt{(f_{12}f_{34})}$, is 1000Hz. In other words, they are at the centre frequency multiplied and divided by √spread. Individual spans are both 4.36 so, similarly,

f_1=128.0Hz f_3=1792Hz
f_2=558.1Hz f_4=7813Hz

where $\sqrt{(f_1f_3)}$=267Hz, f_3/f_1=4.36, etc.

Phase shift at the centre frequency of 1000Hz is 87.48°, i.e. 2.52° below 90°. The peaks are at 92.66°. Response dips to 87° at 216Hz and 4620Hz. This could be loosely called the bandwidth because of the similarity to the useful response of a bandpass filter in the amplitude domain.

Examples 2 and 3. The difference between the peaks of the phase response curve, and the trough at the centre frequency could be termed the 'phase ripple'. It can be reduces by reducing the spread of the filter pairs. As this is done, the phase response becomes flatter, but it is no longer centred at 90°. It has to be corrected by adjusting the span. **Figure 7** shows the effect of reducing the spread to 12 (Example 2) and to 9 (Example 3), while reducing the span appropriately.

Notice that in Example 3, ripple is extremely low – almost within 1/4 degree. Bandwidth however is limited.

Example 2a. Of the above two examples, let us suppose that Example 2 looks like a suitable filter to build. The procedure is as follows. Firstly, note that all the examples used a centre frequency of 1000Hz. The individual sections of Example 2 have centre frequencies of,

f_1=142.9Hz f_3=1715Hz
f_2=583.1Hz f_4=6997Hz

If you want to alter the overall centre from 1000Hz you can scale these frequencies. However, you do not need to do that for this example. Using E24 resistors you can get close to these frequencies:

f_1: 1.0MΩ+110kΩ & 1nF⇒143.4Hz
f_2: 270kΩ+ 3.0kΩ & 1nF⇒583.0Hz
f_3: 91kΩ+ 1.8kΩ & 1nF⇒1715Hz
f_4: 22kΩ+ 750Ω & 1nF⇒6996Hz

Figure 8 shows the central portion of the phase response on an enlarged scale. The slight asymmetry of the curve is due to the errors caused by the resistor approximations. The response is only very slightly different from that predicted by Example 2.

Figure 9 shows a circuit diagram of the complete outphaser. The filters R_1/C_1 to R_4/C_4 use the values from the list above. The resistors should be 1% metal film with a low temperature coefficient. The capacitors should be polystyrene 1% parts.

Unmarked resistors are all equal in value, say 100kΩ. They should be 1% metal film or, possibly 2% thick film resistor packs, for which the temperature tracking will probably be good. The op-amps should have a low input current, for example BiFET types, or you will need to consider the effect of bias currents.

Filter inputs must be driven from a low impedance source so as not to affect the gain or phase response. ■

Next time...

In the concluding part of this article I will look at the effect of component tolerances, which can be significant. I will go on to look at outphasers built from three and four filter sections. These can have an extremely flat top, or a very wide bandwidth. I will conclude by looking at a digital filter implementation of an outphaser.

Further reading

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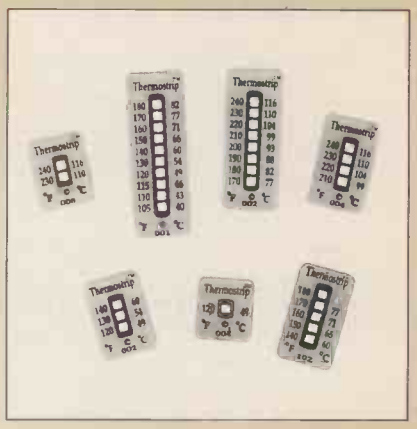

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Two-chip **smart** accelerometer

Benefits of this accelerometer – designed using silicon micromachining – are small size, relatively low cost and repeatable, temperature-stable output. Diedrik de Bruin and Ed Koen of EG&G IC Sensors explain.

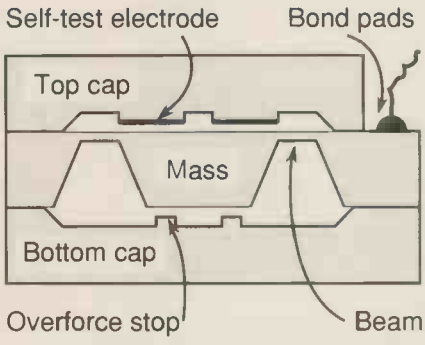

Fig. 1. The accelerometer die's footprint is 3.4mm by 3.4mm. Piezo-resistive transduction provides a relatively high output.

The signal-conditioned accelerometer described here offers many advantages. Manufactured using silicon micromachining, the sensor element has proven reliability. Being wholly monolithic, the signal conditioning circuitry needs no external components and thick or thin film technology.

Both sensor and signal conditioning chips are hermetically packaged together in a ceramic leadless chip carrier. Output parameters are trimmed electrically after packaging. The chip carrier can be mounted in several orientations to allow measurement of acceleration either perpendicular to or in plane with the mounting surface.

Accelerometer overview

Currently the majority of signal conditioned accelerometers are packaged using hybrid technology. Thick or thin-film resistors are used to set parameters such as offset and sensitivity to the desired values.

This approach results in relatively bulky designs with non-uniform mounting configurations. The user is often required to carry out additional mechanical work, such as designing a mounting bracket.

The accelerometer design discussed here is intended to not only lower the cost of the accelerometer, but also the reduce implementation costs. This is accomplished by mating a silicon micromachined sensor die to a signal-conditioning IC in a ceramic leadless chip carrier.

The two-chip approach allows the sensor and signal conditioning chips to be optimised and avoids the yield losses associated with complicated single-chip designs. The accelerometer is compatible with automated pc board assembly while offering multiple mounting options.

Sensor element

The accelerometer structure, **Fig. 1**, measures 3.4 mm square. A seismic mass and four flexures are formed using bulk micromachining processes. Bulk micromachining technology was chosen over surface micromachining because the entire thickness of the silicon wafer can be used for the seismic mass, resulting in a higher sensor output.

Each of the four beams contains two implanted resistors, interconnected to form a Wheatstone bridge. When the device undergoes an acceleration, the mass moves up or down, causing four of the resistors to increase and the other four to decrease in value. This results in an output voltage change proportional to the applied acceleration.

Eight resistors are interconnected such that the effects of any motion other than that caused by an acceleration in the primary direction are cancelled out. Piezoresistive transduction provides a relatively high output level with low impedance and good linearity. As a result, it is not necessary to include signal conditioning electronics on the same chip as the sensor to obtain good performance.

Silicon top and bottom caps attach to the section containing the seismic mass and the beams. These serve several purposes. Precision gaps are etched into the caps to provide air damping to suppress the resonance peak of the structure. Because the part is critically damped, the response is flat up to several kilohertz – independent of temperature.

Small elevated stops on the top and bottom caps limit the motion of the mass to a fraction of the deflection at which fracture occurs. The mechanical structure does not wear and mechanical latch-up cannot occur. The top and bottom cap form an enclosed cavity around the seismic mass, protecting it against contamination which may obstruct its motion.

Because the three sections are bonded together at the wafer level in the clean room the cavity is free of particles and is protected from particulate contamination during the final chip dicing and assembly operations.

Lastly, the top cap is used to enable testing of the accelerometer in the absence of acceleration[1,2]. The over-force stops on the top cap have been enlarged and a metal electrode has been deposited on them. This electrode is connected to a bond pad.

When a voltage is applied between the elec-

trode and the silicon of the seismic mass, an electrostatic force moves the mass toward the top cap. This results in a change in output voltage proportional to the sensitivity and to the square of the applied voltage. It is thus possible to generate an 'acceleration' using an external voltage and to check the functioning of the mechanical structure as well as the electronics.

The accelerometer has been qualified for, and used in, air bag crash detection systems and proven to be very reliable.

Signal conditioning circuitry

Signal conditioning circuitry is made in 1.5µm cmos technology. Signals are processed by differential amplifiers throughout most of the circuit in order to minimise common mode effects and noise.

Switched capacitor circuitry is used to save space and because high accuracy gain stages can be made easily. The –3dB bandwidth of the signal conditioning electronics is about 3kHz. The accelerometer is intended for 5V operation with an output voltage in the 0.5-4.5V range.

Processing the signal

The accelerometer has a differential output with source impedance of around 4kΩ and full scale output voltage of about ±50mV. The offset voltage, i.e. output at zero applied acceleration, may vary a few millivolts over the temperature range of –40 to 85°C. Also, the full scale output decreases over temperature by about –1900ppm/°C.

The signal conditioning circuitry converts the differential signal into a single-ended signal in the 0.5-4.5V range while compensating for temperature-related signal variations. As a result, the accelerometers are interchangeable with a total error of less than 5%.

The signal path is shown in block diagram, Fig. 2. Output signal of the accelerometer die is processed by the following stages:

- The first stage provides a high impedance load for the sensor and amplifies the signal to maximise the dynamic range during subsequent processing. Offset of the sensor is eliminated by adding a voltage generated by a d-to-a converter. This converter is controlled by a digital word representing the programmed offset value.

- The temperature coefficient of offset (tco) of the sensor is compensated by adding a voltage which is controlled by digital words representing the temperature and the programmed tco value. Both the offset and tco voltages are derived from the supply to ensure that the signal remains ratiometric with supply voltage.

- Signal gain can be varied by changing a capacitor ratio using a digital word. The gain can be varied in a 5:1 range to allow for different full scale specifications.

- The sensor's temperature-coefficient of sensitivity, tcs, is compensated in the next stage.

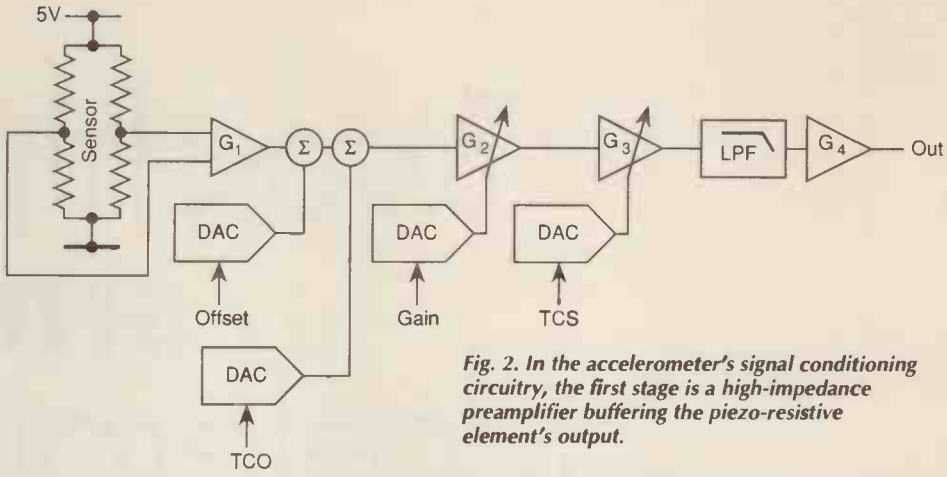

Fig. 2. In the accelerometer's signal conditioning circuitry, the first stage is a high-impedance preamplifier buffering the piezo-resistive element's output.

Sensitivity decrease over temperature is compensated by increasing the signal gain linearly with temperature. This method was chosen over a circuit using constant current excitation of the sensor because of the required voltage overhead of the current source. The sensor is now powered with the entire available supply voltage, maximising its signal.

- Output bias voltage can be set to either 0.5V or 2.5V by connecting an input pad on the chip to ground during assembly of the part. This allows signals to be processed with either a bipolar or unipolar range.

- A two-pole passive filter removes signals generated by the internal oscillator and switched capacitor networks. Switching noise is further minimised by having separate digital and analogue internal supply lines and by the differential signal processing.

- A low impedance output for driving resistive and capacitive loads without influencing the signal is provided by the final stage. The output enters a high impedance state if the device is not addressed.

Error detection functions

Because the accelerometer is intended to be used in safety-critical applications, such as airbag deployment, several features are incorporated to detect a failure of the accelerometer or circuitry.

It is important to prevent floating signals because the resulting output voltage might look like a crash signal and activate the air bag. Such signals could be caused by a discontinuity between the sensor and the circuit or by a malfunction of the sensor itself. Small current sources have been added between each of the signal inputs and the positive supply.

In case one or both of the inputs are open, output voltage is forced to the positive supply. In addition, two window comparators monitor the voltage at both inputs. If the voltage at one or both inputs exceeds the allowed range, an 'alarm' output pin is made high. This output can be monitored by a microprocessor to alert the user to a malfunction of the sensor.

In addition, the sensor has a built-in self-test function which allows the seismic mass to be moved by means of an externally applied voltage. This allows the entire device to be tested, including the mechanical structure of the sensor and the signal conditioning electronics.

By applying a voltage to the bond pad that is connected to the self-test electrode, the output will exhibit a voltage change which is proportional to the full scale output, in contrast to

Number of sensors	Required number of lines	
	Non-multiplexed	Multiplexed
2	4	6
4	10	8
9	20	10
16	34	12
25	52	14

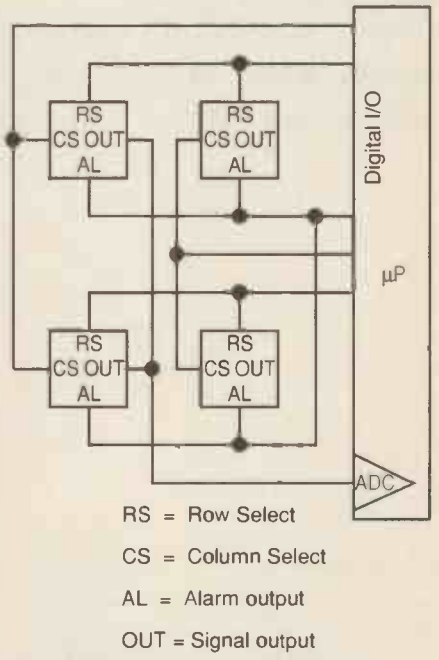

Fig. 3. Control signals and i/o lines are structured so that multiple accelerometers can be accessed via the same bus.

other self-test schemes where the output change is fixed. This makes it possible to verify not only complete malfunction of the device but also a parametric error, giving a better indication of a partial or a developing failure.

Accessing the device

Addressing capabilities have been incorporated in the signal conditioning electronics in the form of row-select and column-select digital inputs.

Both input lines must be high for the accelerometer to be selected. If one or both of the select lines are in the low state, the signal and alarm outputs are in a high impedance 'tri-state' mode. This allows the outputs of multiple accelerometers to be connected together, **Fig. 3**, eliminating the need for analog multiplexers and reduces wiring.

The reduced number of wires is an advantage if four or more devices are needed in a system. The following table shows the number of lines – including supply and ground – required in a measurement system with sensors used in non-multiplexed and multiplexed mode. The digital control lines could be driven by custom designed logic, a card that plugs into a computer, or the i/o port of a microprocessor.

The digital inputs and outputs used during testing and trimming are disabled if the device is not selected, and can therefore be bussed together. This greatly simplifies the test hardware if the accelerometers are characterised and trimmed in an array configuration.

In the case of single-sensor operation, or if multiplexing is not desired, the row and column-select inputs can be left open. Internal pull-up current sources ensure that the accelerometer is selected when these inputs are not connected.

Electrical trimming

Optimal trim values for offset, tco, gain and tcs are different for each sensor. Often a network of thick or thin film resistors is used to set these coefficients. In that case, the desired resistor values are set by laser trimming after characterisation of the untrimmed sensor. This requires a separate trim operation using expensive equipment.

Any additional packaging steps done after trimming, sealing the substrate in a housing, for example, could change the characteristics of the sensor resulting in sensitivity or offset errors. Furthermore, trimmable resistors and the conductive traces connecting them to the electronics take up space and limit the available packaging options.

To avoid these disadvantages the trimming is done internal to the signal conditioning IC. The trim coefficients for offset, tco, gain and tcs are stored in binary registers which are connected to d-to-a converters that manipulate the signal.

In contrast to some designs that require an additional eeprom containing the coefficients, the storage registers are on the same chip as the signal conditioning electronics. The storage registers are made in fuse technology to assure data retention in safety-critical applications.

Before trimming data is permanently programmed into the fused registers, the accelerometer can be operated using data stored in volatile ram registers. This allows for the characterisation of the sensor and electronics during manufacturing in order to extract the required coefficients for offset, tco, gain and tcs.

Fuse trimming is handled by circuitry inside the signal conditioning IC and requires no external equipment. The digital i/o used for characterisation and trim consists of a serial input and a serial output line and a clock input for synchronising the data entry, which uses a 16-bit protocol.

All digital i/o lines are available after final packaging. This allows the accelerometer to be trimmed as the last manufacturing step. Because the data transfer is serial rather than parallel, the pin count is not the limiting factor for the package size.

Packaging details

The package is a leadless chip carrier measuring 0.530in by 0.300in and is 0.150in thick. It is manufactured by screening tungsten interconnect traces onto ceramic layers which are then stacked together and fired.

The accelerometer die and signal conditioning IC are mounted into the package cavity and connections are made from the die to the package with gold-wire bonds. A gold plated Kovar lid is then soldered to the package using a Au/Sn preform. This provides a hermetic seal which will withstand the rigorous environmental requirements of the automotive and military industries.

Reliability is increased with respect to many other designs because of the reduced number of components. No external components such as capacitors are needed for operation. Stiffness and low mass of the package helps to keep its resonant frequency high. Inputs and outputs needed for operation of the accelerometer and for characterisation and trim are brought out to contact pads on the side and on the bottom of the package, **Fig. 4**.

Mounting surface 2 is on the opposite side from the metal lid. Because electrical contact can be made on two surfaces and because of the aspect ratio of the package, it is possible to mount the package either flush with or perpendicular to the board.

In many cases accelerometers need to be mounted at a 90° angle with respect to the circuit board. This normally requires additional brackets and is not compatible with automated manufacturing. The ceramic package allows the accelerometer to be mounted on the pcb using automatic placement equipment, reducing manufacturing cost and saving space.

Another possible application is to make a tri-axial accelerometer by mounting two accelerometers perpendicular to the board and one in parallel, **Fig. 5**. Dimensions of this fully signal conditioned tri-axial accelerometer is only 0.73in by 0.53in by 0.30in.

The accelerometer is available in several g ranges to cover many applications such as ride control, airbag deployment – both frontal and side impact – fusing and arming, vibration monitoring and general instrumentation.

In addition, it is possible to adapt the device to specific customer needs. ■

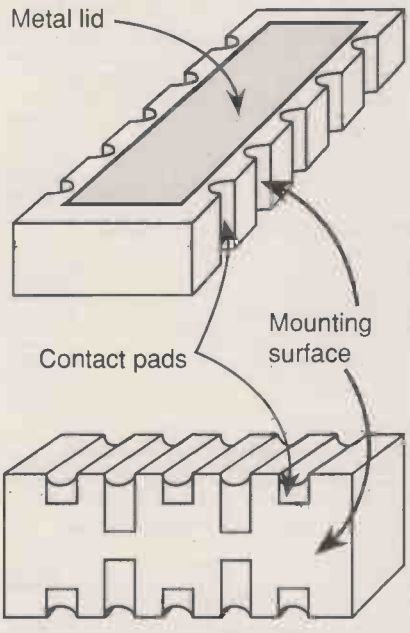

Fig. 4. Stiffness and low mass of the accelerometer's surface-mount package helps keep its resonant frequency high.

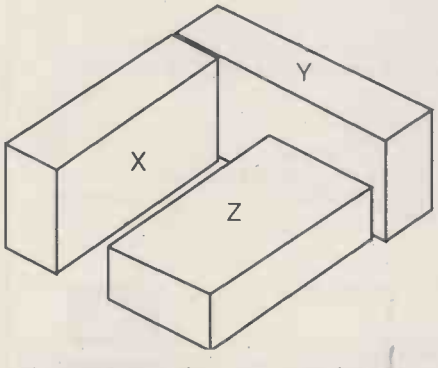

Fig. 5. Orienting three sensors in this way forms a tri-axial accelerometer.

This article is based on a paper presented at Sensors Expo, Cleveland Ohio; contact http:\\www.sensorsmag.com

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Telephone caller ID

Uses, standards, devices and receiver design for Caller ID – the system that allows you to see the number of the person phoning you – are discussed by Seggy Segaran.

The term Caller ID is used to describe the transmission of the caller's telephone number when the telephone rings. This service was introduced by BT at the end of 1994, along with two receiver units.

The *CD50* is a stand-alone battery powered unit with display, that can store details of 50 calls. The *Relate 1000* with combined telephone, is much more sophisticated. Not only can it display the number, but it uses a local directory to look up the name of the caller. It also allows the easy redialling of any of the received numbers.

Currently, the service from BT only delivers the caller's number, the time and date. The enhancement of the service to deliver name has not yet taken place. There is no time scale from BT for this to be available. If the call is from a pay-phone or from abroad, then the text 'payphone' or 'international' is sent. Calls from a significant number of telephones are still delivered as 'number unavailable', presumably because these are connected to older exchanges.

Privacy is an important consideration. Calls from 'ex-directory' lines are delivered as 'number withheld' and so are all calls prefixed with 141. This ensures anonymity for those that require it.

The Caller ID service is only connected on request and there is a quarterly charge. However, the benefits of Caller ID as described below, will surely more than offset this modest charge.

For the domestic user, the service allows screening of incoming calls, which is especially useful during quality family time. Only expected calls or those from close family members need be answered immediately.

The Caller ID device can also be used as a complement or a replacement for an answering machine. It will record the number of those that tend to hang up as soon as the answering machine message starts to play, and also record the number – even if the call is not answered. The *Relate 1000* allows quick redialling of any of the numbers in the calls log.

For the small business user, the Caller ID service is invaluable. Taxi firms and pizza delivery services are regularly abused by pranksters. With the CD50, a simple check of the caller's telephone number with a verbal confirmation can sort these out.

Voluntary organisations can identify malicious callers. They can also identify calls from vulnerable people in trouble, such as emergency calls from disabled or elderly callers. For the tradesman, it allows potential enquiries to be followed up from callers reluctant to use the answering machine.

However, the real benefit to businesses come, when the Caller ID information can be presented to the com port of a pc. This allows the logging of large numbers of calls, instant look up of customer details using the telephone number as a key, and verification of customer identity when releasing sensitive information, such as bank account details.

On another front, the number information can be checked against a stored list of numbers before allowing access to a database, thus providing an effective 'anti-hacking' device.

Companies employing a mobile team, such as cleaning or security staff, can request them to call in from their various sites at the start and the end of their duties. This verifies attendance and time spent at each site. The beauty

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Consumer Microcircuits, 1 Wheaton Road, Witham, Essex CM8 3TD. Tel: 01376 513833, fax: 01376 518247

Useful standards

BT:SIN 227: BT Analogue Caller Display Service- Service description.
BT:SIN242: Calling Line Identification Service: TE requirements.
Available from: Regulatory Services Unit, Room 134, 2 City Forum, 250-258 City Road London EC1V 2TL. Tel: 0800 318601,
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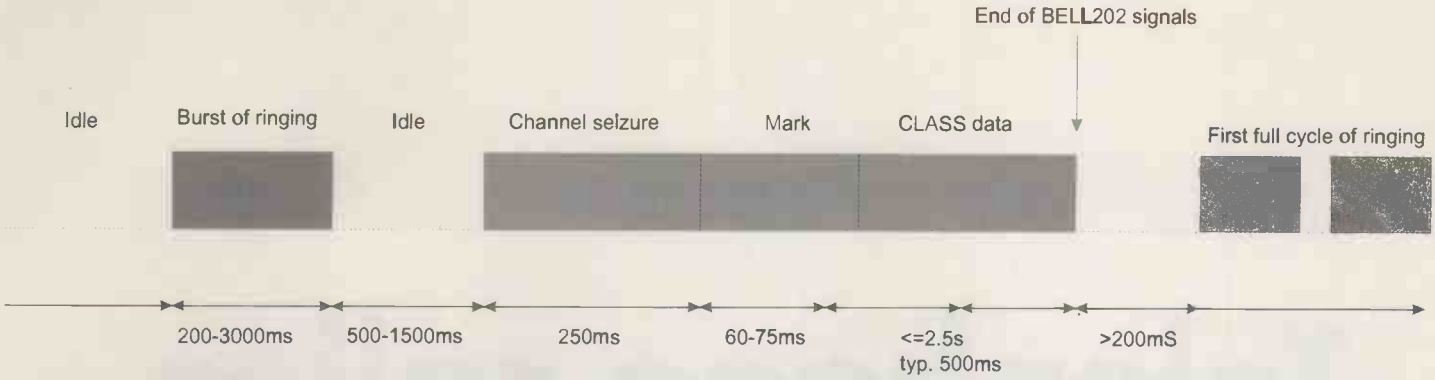

Fig. 1. Timing details for Bellcore's standard for caller ID: data-link layer, on-hook data transmission. Among the first used, this standard was first available in the US.

of this use is that, as the telephone call does not have to be answered, there are no call charges incurred.

Caller ID – history and standards

The Caller ID service was first introduced in the US, based on a series of standards from Bellcore. The information is coded using frequency-shift keying signals, fsk, using the Bell 202 standard. This used a 1200Hz signal for a mark, and a 2200Hz signal for a space.

On call arrival, a single ring burst is sent by the exchange, followed by a burst of fsk signal. The data is preceded by a Channel Seizure signal – comprising alternating marks and spaces – and a mark signal. This allows the fsk receiver to synchronise to the data, and to provide immunity against noise spikes. The sequence of events on call arrival is shown in **Fig. 1**.

The initial ring burst is used by receiving units as a 'wake up' signal. Since these units are battery powered, low power consumption is paramount. The design of the various Caller ID receiver ICs allows the receiver to be placed in a standby mode, with just the ring detector powered. In this mode, current consumption is down to tens of microamps.

On a ring signal being detected, the rest of the IC is powered, and the fsk data is decoded. Since this is done in one or two seconds, and the IC then goes back to low power mode, battery life of a year can be achieved.

BT's Caller ID service

This started off with the Bellcore system as its model but diverged along the way. One new requirement for the BT service was that it should be possible to pass information to the receiving unit, without alerting the phone user. The information was for metering and message waiting status. This precluded the use of the ring signal as being the initial alert signal.

Reversal of line polarity was decided upon as the initial alerting signal. So the 'no ring' call would be presented as line reversal, data, followed by another line reversal. A normal call on the other hand, would be presented as line reversal, data and then ringing.

However, the ringing signal served the purpose of 'wetting' the cable joints, prior to fsk signalling. As there is negligible current flow during a line reversal, the 'wetting pulse' was to be supplied by the receiving unit, before the transmission of the fsk signal.

To ensure that this 'wetting pulse' was applied correctly and in synchrony with other units on multiple installations, another signal was introduced. This was the Tone Alert Signal, or TAS, and was a dual tone of 2130Hz and 2750Hz. After receipt of this, the 'wetting pulse' was to be applied. To ensure good impedance matching during fsk data transmission, the BT standard also calls for an ac impedance during this state.

In addition to the above changes, the BT specification uses V23 frequencies for the fsk signals, which involves 1300Hz for the mark and 2100Hz for the space. The sequence of events for this is shown on **Fig. 2**. The BT specification also allows for a number of new features to be implemented, and has built in some flexibility for future expansion.

The Caller ID service implemented by cable tv companies is closely modelled on the Bellcore service, in that a single burst of ringing is used to initiate the data. However, V23 frequencies are used for the fsk data and there is also some allowance in the application layer for future expansion.

Design of a pc device

To exploit a niche in the market for Caller ID devices, a project was initiated to produce a unit that would meet two key objectives. First it would allow Caller ID data to be decoded from the telephone line, and presented to the com port of a PC. Secondly, it would supply a Windows utility that would,

- Display call details on the screen as the telephone rings
- Allow name look-up from a pre-programmed directory
- Log all calls in a database format for processing later.

The unit had to be compatible with BT and CTA Caller ID standard and would have to be priced at under £50 to reach the home pc user. With these objectives in mind, the design of the product commenced. After a period of study, the following key design decisions emerged.

First was the choice of Caller ID receiver IC: newly available were two ICs that were capable of meeting both the BT and CTA standards. One was the *MT8843* from Mitel Semiconductors and the other was the *FX602* from Consumer Microcircuits Ltd. They both had ringing and line reversal detection capability and also circuits for the detection of tone alert signal.

The *MT8843* was chosen as samples of these were available earlier. Having decided to make the unit compatible with both standards, the actual wetting pulse and ac impedance cir-

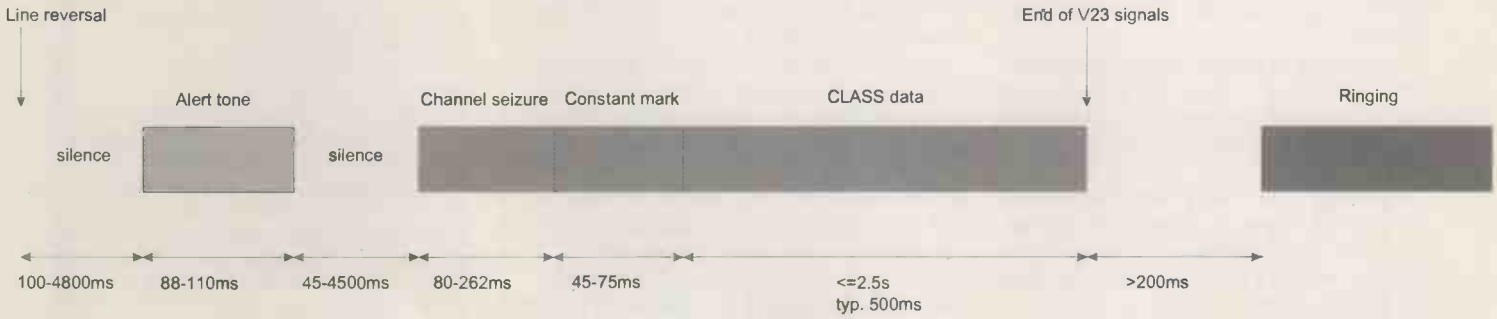

Fig. 2. BT SIN 242: data-link layer, on-hook data transmission, as used for caller telephone number identification throughout most of the UK.

cuits were made optional to save cost. The sensitivity of the receiving circuits were increased to compensate for this.

Powering of the device from a COM port was a key design target, as this would result in lower unit cost. This was made possible by careful design and power management.

To implement the critical timing of the BT standard, to carry out the power management, verification of received data and the serial communication, a Microchip PIC device was used. In addition, a single crystal of 3.579MHz was used as a clock for the PIC and *MT8843* devices to keep costs down.

Visual Basic was chosen for the design of the software as this allowed software to be developed quickly and still allowed very professional screens to be displayed to the user.

A block diagram of the electronics is given in Fig 3. A sample of the Window with call details is shown in Fig 4. Following the above decisions and subsequent detailed design, the project was successfully completed and the device, CID-PC1 is now available from SOLWISE at a cost of £45-00.

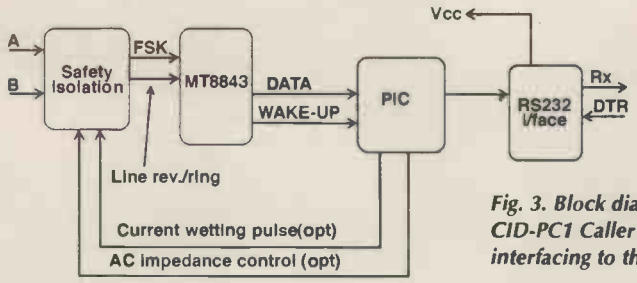

Fig. 3. Block diagram of Tele Products' CID-PC1 Caller ID unit, design for interfacing to the PC.

Summary

Caller ID presents many benefits to domestic and business users. The potential of Caller ID to business users is obvious once the information can be presented to a PC. Above are details of the design of such a unit. ■

About the author

T. Segaran is the founder of Tele-products Ltd. The company specialises in the design and manufacture of telecommunications test equipment and the design and approvals of Telecom products. The company has a Caller ID simulator amongst its range of test instruments. This is capable of simulating most Caller ID standards from around the world.

Before founding Tele-Products T. Segaran worked for Standard Telephones and Cables and at Tunstall Telecom as a Section Leader. He has been instrumental in a number of successful product launches including the early Viscount telephone, the Piper Lifeline, the Minstrel, React, Duet and Converse range of telephones.

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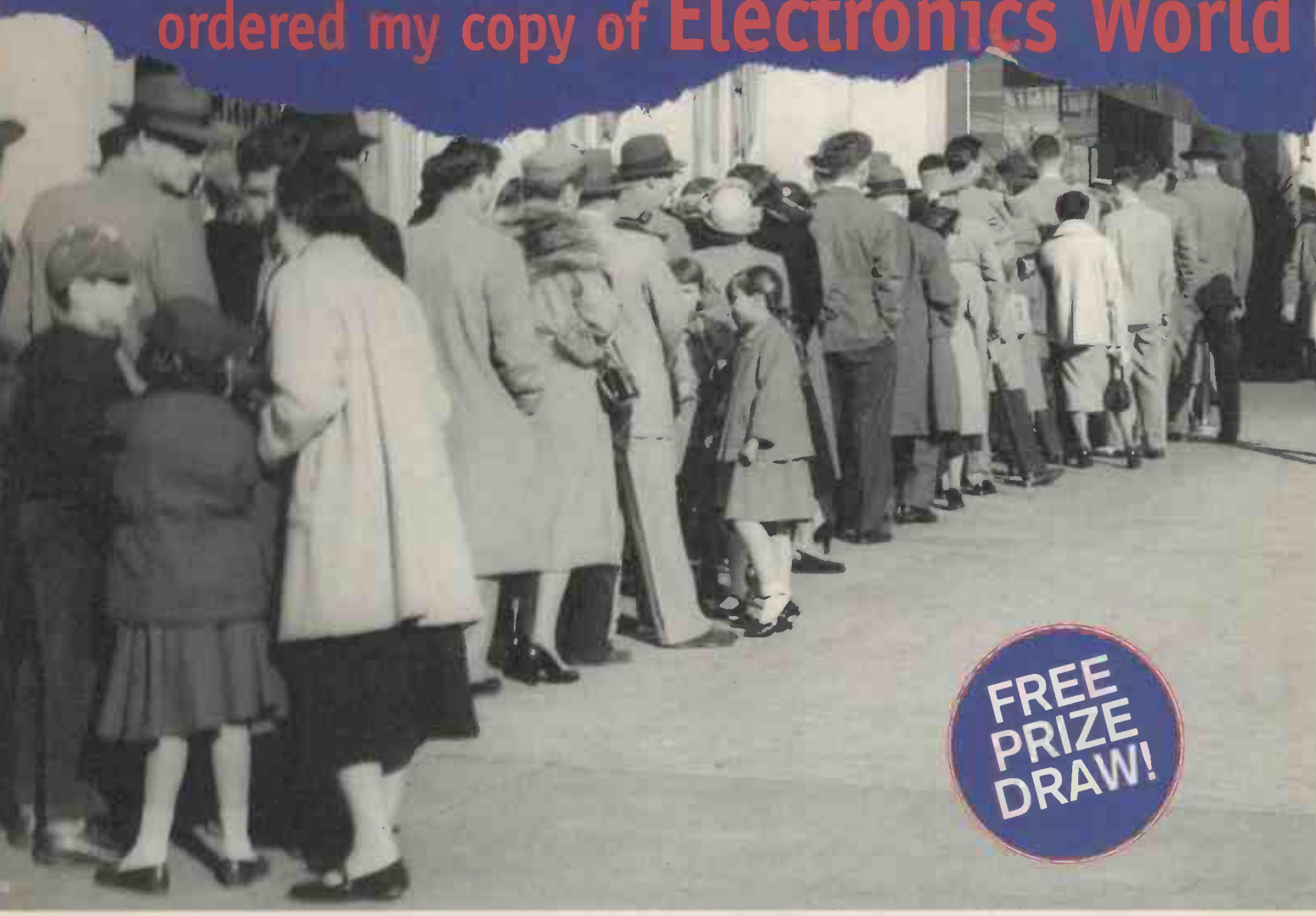

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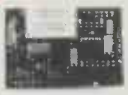

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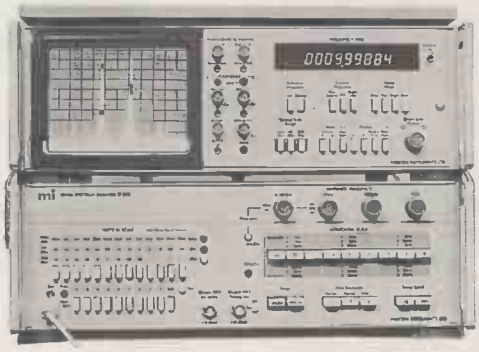

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Night/day light measurement in one range

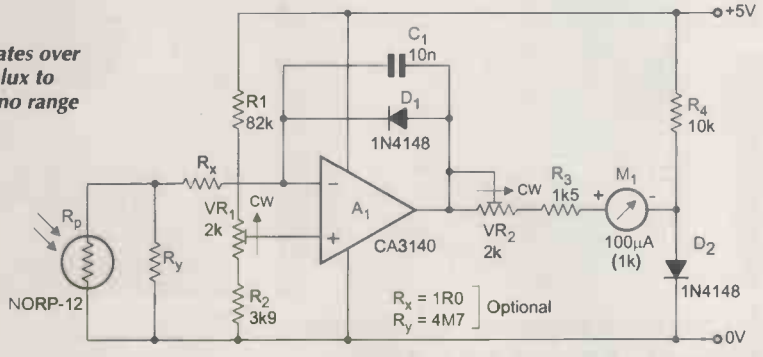
Since the *NORP-12* cadmium sulphide photo-conductive cell obeys a precise log-law ($\log(R_p)$= 4.630–0.6761$\log(L)$, where R_p is the cell resistance at Llux), a low-bias-current op-amp with a log-diode in the feedback loop will give an accurate light reading from moonlight to sunlight in one range. Furthermore, the technique is inherently proof against overload and is inexpensive.

Op-amp A_1 drives a 100μA meter, on which zero is equivalent to 0.1lux and full scale to 10^4 lux. Since log2 is 0.301 and log5 is 0.699, the meter scale may be calibrated in a 1-2-5 sequence in these proportions. If a laboratory standard lamp is available, calibrate the meter at 1lux and 1000lux by the trimmers VR_1 and VR_2; if not, first replace R_p with 42.7kΩ and then 400Ω, these being the resistance values from the *NORP-12* data sheet which does not, of course, allow for tolerances.

Diode D_2 provides a temperature-compensated back-off for the dark-level current at D_1 anode, R_x and R_y trimming the conformance of D_1 to the log-law. Other types of silicon diode such as the *OA202* would improve performance at low current and the *1N4002* at the high end, but the *1N4148*, well shielded from light, is a good compromise.

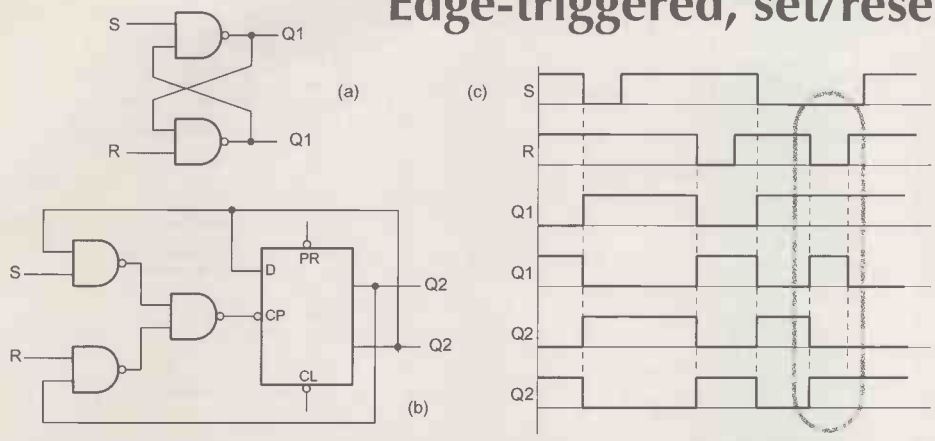
The *NORP-12* has a spectral response similar to that of the human eye, peaking at 550nm; for a response at infrared, a silicon diode, using similar circuit, would be better.

Light meter indicates over five decades - 0.1lux to 10,000lux - with no range changing.



C J D Catto
Cambridge

Edge-triggered, set/reset bistable device



Edge-triggering a S/R bistable device avoids spurious resets when input pulse width is unknown.

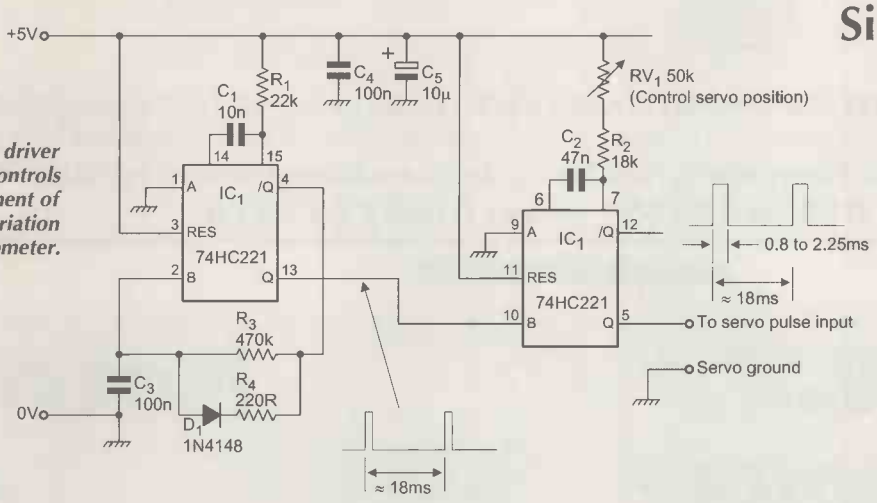
If the widths of set or reset pulses applied to a standard S/R bistable device (a) are unknown, the state of affairs shown in (c) at Q_1 and $/Q_1$ can occur, where the reset pulse arrives during the set pulse; reset has no effect on Q_1, but produces an unlooked-for pulse on $/Q_2$. In addition, the next set pulse will be ignored, since Q_1 is already high.

Since the circuit in (b) responds only to negative edges at the set and reset inputs, the output is as shown at (c) in Q_2 and $/Q_2$.

Giorgio Delfitto
University of Padova, Italy

Simple servo driver

Servo driver controls movement of servo by variation of potentiometer.

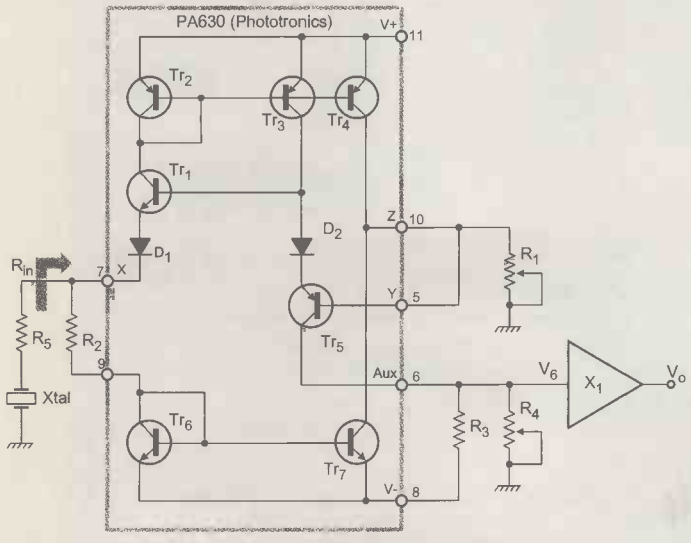


This simple circuit drives model servo motors in response to the turning of a potentiometer.

Half the *74HC221* dual monostable is used as a free-running oscillator, producing narrow trigger pulses for the second half of the monostable, whose output is a train of standard servo pulses about 18ms apart, variable in length by the potentiometer from 0.8ms to 1.24ms. The potentiometer therefore controls the servo.

R G Sutherland
Woking
Surrey

Crystal oscillator using a current-conveyor



Audio current conveyor, in negative resistance configuration, used to drive a crystal at up to 5MHz.

A *PA630* second-generation audio current conveyor, used to provide negative resistance, fulfils all the requirements of a crystal oscillator circuit: high bandwidth, optimum drive level, low damping to retain high crystal *Q* and good input/output isolation. Crystals in the 31.25kHz-5MHz range have been used in the circuit shown.

Transistors $Tr_{1,7}$ form the current conveyor, bias current for all transistors (I_{bias}) being set by R_2, according to

$$I_{bias}=(V_{EE}-2V_{BE})/R_2$$

Resistor R_2 driving the current mirror $Tr_{6,7}$. Positive feedback from the high-impedance output *Z* and the high-*Z* input Y causes the input resistance at the low-*Z* input to become R_{in} is $-R_1$, so that, if the esr of the crystal R_x is equal to R_1–R_5, the circuit oscillates. Resistor R_5 is not essential, but does set the best crystal current.

Output comes from the AUX pin, which gives good isolation and offers a point for level adjustment. Resistor R_4 allows adjustment of V_6, the potential into the buffer stage according to,

$$V_6=(I_{bias}R3-V_{EE})R_3/(R_3+R_4),$$

oscillation increasing until the collector/base junction of Tr_5 becomes forward-biased, reducing the magnitude of the negative resistance at point X.

For a 1MHz crystal with a R_x of 85Ω, $V_{cc}=V_{EE}=15V$; $R_1=100\Omega$; $R_2=R_3=9.1k\Omega$; $R_4=75\Omega$; and $R_5=10\Omega$. Oscillation amplitude is 75mV.

Dan Stiurca
Romania

24V electromechanical counter from 12V

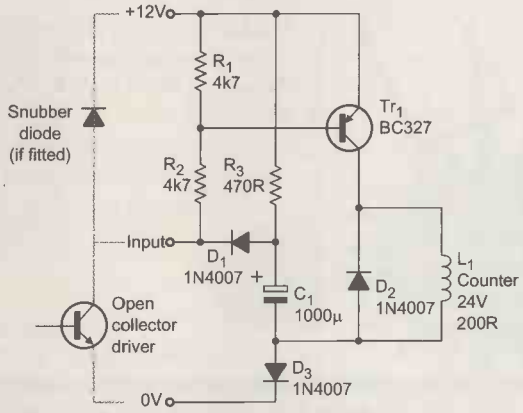
Having a number of 24V counter mechanisms and a 12V controller, it was necessary to produce a suitable interface. This circuit performs that function with no great power dissipation and with less interference radiation than other methods.

With the driver off, the input is at 12V, Tr_1 is cut off and there is no current to the counter coil. Capacitor C_1 charges through R_3 to around 11V via D_3 and no further current flows in the circuit.

As the driver comes on, Tr_1 base current flows in R_2 and the top end of the counter coil goes to almost 12V. Diode D_1 conducts and clamps the top end of C_1 to about 1V, its bottom end and that of the coil going to about –10V, so that the coil sees enough voltage to energise it. As the charge on C_1 decays, the counter still sees about 11V, which should be enough to hold the mechanism in, with reduced steady-state dissipation.

The time constant is chosen to suit a 250ms drive every 3s, but may be varied for any use. A snubber diode is not necessary, since the driver collector never exceeds the supply voltage.

Gerald D Pye
Ipswich
Suffolk



12V-to-24V converter to drive 24V counter coils from 12V, with the incidental advantages of reduced power dissipation and interference.

One op-amp dc motor driver

Used widely in the field of robotics, this current source produces a 2.5A output from a 6.25V input, using only one power op-amp and one power resistor.

Feedback from both ends of the 3W current-sensing resistor R_{sc} got to the op-amp inputs, which is forced to maintain the current through R_{sc}, calculated to be,

$$I_{out}=(V_{in}/R_{sc})(R_2/R_1).$$

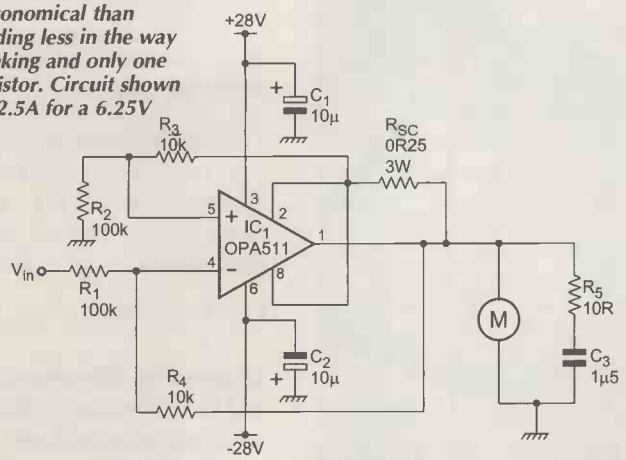
Choosing $R_2=R_4=10k\Omega$ and $R_1=R_3=100k\Omega$, R_{sc} is 0.25Ω to give an output of 2.5A for a 6.25V input.

$$R_{sc}=0.65/I_{out}(A)-0.01.$$

Resistors $R_{1\text{-}4}$ should be 1%, 0.25W types and the op-amp should be on a heat sink; the *OPA511* has an insulated case and needs no isolation.

V Vidyalal, K Rajasree and V Sivanand
Cochin University of Science and Technology, India

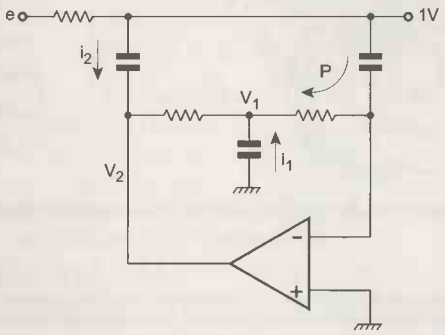
Motor driver for robots. This is more economical than most, needing less in the way of heat sinking and only one power resistor. Circuit shown produces 2.5A for a 6.25V input.



Active, low-pass filters with no dc errors

In the arrangement illustrated, the op-amp in this low-pass, maximally flat Butterworth filter is blocked from the signal path by capacitors, this makes its offset and input current irrelevant. Two-pole, three, four and five-pole versions have been built and offer the further advantage that they use fewer components than more conventional circuits. The op-amps can be operated from a single supply, if required.

No free lunches, though: theoretically, they must work into an open circuit, a requirement that can be met either by including a follower to the output or doubling the value of the input resistor and inserting an equal value to ground. This halves the dc gain and needs a purely resistive, and fairly critical, load.



One of a family of maximally flat low-pass active filters, in which the op-amps are dc-blocked and which are more economical in components than other designs.

To take the three-pole version shown, let p equal j, work backwards from a 1V output to find the input e.

$$v_1=-p \qquad i_1=p^2$$
$$v_2=v_1-(p+i_1)=-2p-p^2$$
$$i_2=p(1-v_2)=p+2p^2+p^3$$
$$e=1+p+i_2=1+2p+2p^2+p^3.$$

So the transmission T is,

$$T=\frac{1}{1+2p+2p^2+p^3},$$

and magnitude $|T|$ is,

$$|T|=\frac{1}{\sqrt{(1-2\omega^2)^2+(2\omega-\omega^3)^2}}$$
$$=\frac{1}{\sqrt{1+\omega^6}}$$

If R and C values are unknown, make one component unity and the two resistors equal. This still leaves four unknowns, so other component values are possible. It turns out that if all three capacitances are unity, so are the resistances.

McKenny W Egerton
Owings Mills, Maryland, USA

EW reader offer

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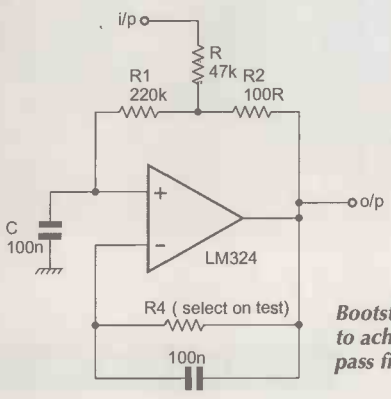
Virtual-capacitance timer and filter

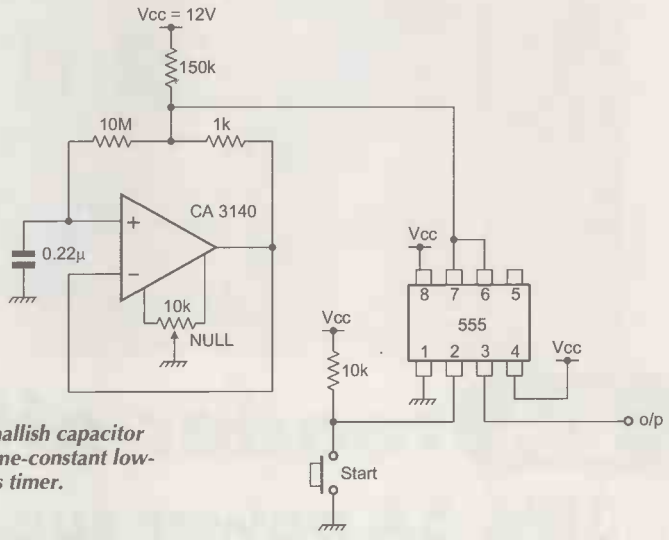
An op-amp and five other components bootstrap a capacitor to make a 200µF component from 0.1µF.

Left is shown a low-pass filter having a time constant of 10s, determined by $R \times C \times R_1/R_2$. Since bootstrapping also increases the effects of op-amp bias current and input offset voltage, R_4 reduces dc following error to around 10mV, its value being greater or less than that of R_1, depending on the sign of the offset; it is bypassed to ensure stability.

In the right-hand diagram, using a fet op-amp allows an increase in the amount of bootstrapping to about 10^3, giving the effect here of a 2200µF capacitor. Used with a 555 timer, and depending on how well the *CA3140* offset voltage can be coped with, a time of 400-600s can be obtained to within ±1% repeatability.

W. Gray
Farnborough, Hants





Bootstrapping a smallish capacitor to achieve a 10s time-constant low-pass filter and 500s timer.

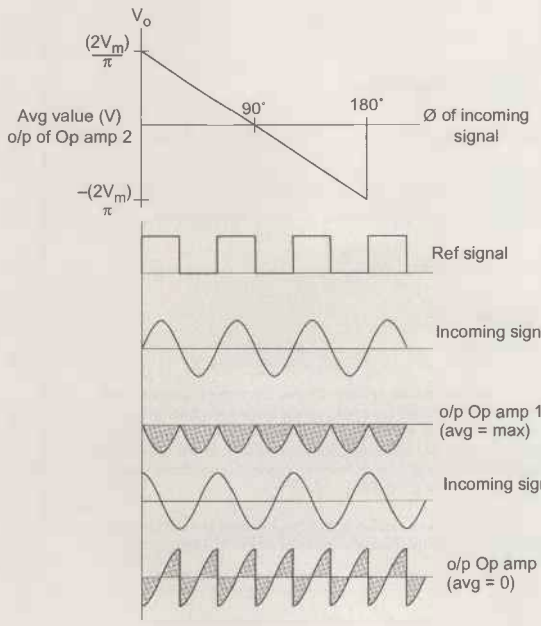
Linear phase detector from two op-amps

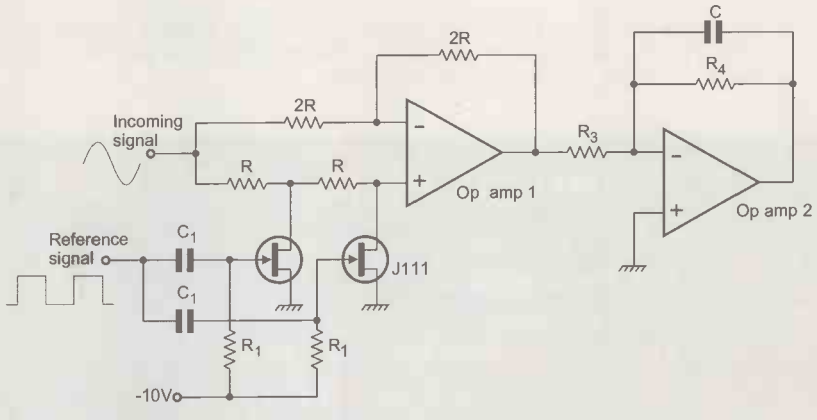
Two op-amps and two fets form an analogue linear phase detector.

An input reference square wave switches on and off the two switching fets, which configure the first op-amp into an inverting amplifier when the fets are on and a non-inverter when they are off, both with unity gain.

If the input signal, shown as a sinusoid, is in phase with the reference, the output of the op-amp is, effectively, a full-wave-rectified version of the input to give the maximum positive circuit output when filtered by the output op-amp. When the input is 90° out of phase, the op-amp output produces an average which is equal to zero and when in antiphase, a maximum negative version of the in-phase output.

Ensure accurate matching of resistors at the op-amp input and also that the on resistance of the fets is low.





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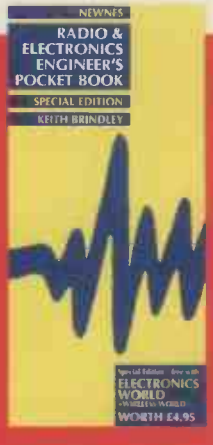

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LETTERS

**Letters to "Electronics World"
Quadrant House, The Quadrant,
Sutton, Surrey, SM2 5AS**

I versus V feedback

Surely current feedback, cfb, is much closer to the correct drive of a loudspeaker voice coil than voltage feedback, vfb, is?

Driving force on the coil is very simply expressed by,

$$F=\beta.i.l.\sin\alpha$$

with wire length l, α a constant, and β flux density, so F is proportional to i.

A similar relationship between voltage and force cannot be written easily, due to the many ill defined terms that compose voice coil impedance.

Apart from the main resonance region, voice coil back emf is negligible, so voltage feedback is also unrelated to voice-coil velocity.

In reply to Mr Allison's query in the December issue, I can say I have tested cfb on two different designs, one low ultra-linear push-pull of *EL84* in 1961 and one 18W with *AD159* transistors in 1964.

Not equipped at the time with sound level measuring apparatus, I had to rely on my ungolden hearing for comparisons with identical amplifiers wired for vfb. No elaborate double-blind protocol was necessary to note the marked differences in tonal responses.

In both modes, 20cm wideband loudspeakers from several French makers were tried.

Speaker designs with curved generatrix cones, such as Supravox *T215* and GeGo '*Supersoucoupe*', lacked treble response with vfb, but gave very clean and sharp treble with cfb.

On the other hand, models with dual cones – Princeps, and Audax *T21PA12* for example – while displaying sufficient – if unnatural – treble response in vfb, sounded very harsh and metallic in cfb.

Clearly the cause of treble roll off in vfb is the voice coil rise in impedance above 2 or 3kHz, mainly due to the inactive coil turns in front of and behind the magnetic gap. This impedance rise limits the drive in vfb, however in cfb it only limits the maximum available power before clipping.

While little difference in medium response between vfb and cfb was audible, the other evident feature of cfb was boomy bass, due of course to a totally undamped main resonance. This is certainly the principal drawback of this mode, as to my knowledge no simple acoustic means allows for efficient damping of the main loudspeaker resonance.

My solution at the time, to try and get the best of both worlds, was to depart from pure cfb, by insuring constant-gain gradual change from vfb at low frequencies to cfb at high medium and treble, **Fig. 3**.

It would be most interesting to repeat these experiments with modern amplifiers and full testing capabilities.

Jean Claude Baumeister
Chantraine
France

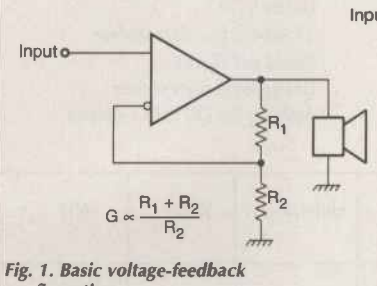

Fig. 1. Basic voltage-feedback configuration.

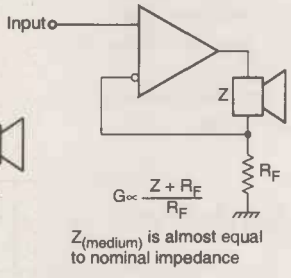

Fig. 2. Current feedback.

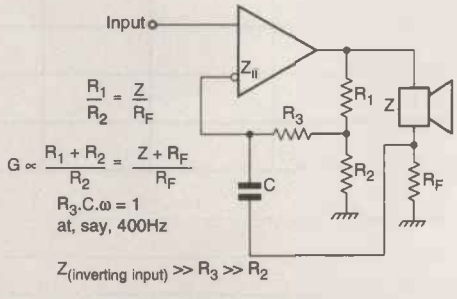

Fig. 3. Composite arrangement combining voltage and current feedback.

Hazy linearity notions?

I would like to comment on Mr Kiyoleawa's hazy notions in the January '96 issue Letters column.

I was glad to see Mr Kiyoleawa confirm that a linear increase of power fet g_m with drain current is a poor basis for making a linear stage. What is really required is linear variation of I_d with V_{gs}. It may be possible to partly cancel fet square-law distortion by push-pull operation. But this can only work in Class-A, when both upper and lower output devices are conducting at the same time.

Economic necessity and energy conservation mean that most amplifiers are Class-B, and to date there is no practicable compromise between these two modes. If fets can only give acceptable linearity in Class-A, then this is not much of a recommendation for them.

I am unable to understand the contention that an fet output stage can have a 'lower' open-loop output impedance, presumably compared with a bipolar version. Field-effect transistor g_m is always much lower than for bipolars, and so this would appear to quite impossible.

A 1Ω output resistance is much too high. It may only have a small effect on loudspeaker damping, but will certainly cause unwanted frequency response variations because of the varying impedance curve of the speaker.

Having done a great deal of practical emc testing recently, I can assure Mr Kiyoleawa that radio-frequency entry via speaker cables is a non-problem – at 3V/m and between 30 and 1000MHz, anyway. The presence of an output inductor may be the critical factor here; at any rate it is no reason to abandon global negative feedback.

I'm afraid that Mr Kiyoleawa has not quite appreciated the action of the voltage-amplifier stage transistor. The impedance at its collector is strongly frequency dependant, halving with each octave as local negative feedback through C_{dom} increases, and crippling its linearity with a dead load of 5kΩ will not alter this fact. I think it will be difficult to find a driver/output pair with a combined h_{fe} of 10,000 at practical current levels; but if the object is, as it appears to be, the avoidance of global negative feedback, then this line of thought is a dead-end anyway.

I have made solid-state amplifiers where the output stage worked open-loop, and the practical result is severe distortion of a unpleasantly jagged kind. I cannot believe that anyone – Subjectivist or otherwise – would find this preferable to the very low thd levels obtainable from a blameless amplifier with global negative feedback.

According to the Toshiba application notes[1], igbts consist of an fet controlling a bipolar power transistor; I have no information on the linearity of these devices, but the combination does not sound promising.

The most discouraging aspect is the presence of a parasitic bipolar-junction transistor that turns the device hard on above a critical current threshold. This inbuilt self destruct mechanism makes overload protection an extremely critical matter; it seems unlikely that igbts will prove popular for audio amplification.

Douglas Self
London

Reference

1. Langdon, S, 'Audio amplifier design-s using IGBTs, MOSFETs, and BJTs', Toshiba Application Note X3504, V.1 Mar 1991.

Does component choice make a difference?

I enjoy *EW*'s audio articles, but the statement by Reg Williamson in his Dec '95 audio preamp article is a little strange to me. I must say that 'audio grade' components are sometimes far too expensive and results are doubtful. I am a technician myself and also sceptical about 'audio grade' components.

But there is a big *but*. Recently I built a d-to-a convertor using Crystal Semiconductors' latest 20 bit device, the *CS4329*. My power supply uses *LT1085CT* and *LT1033CT* regulators. The power supply capacitors are Rubycon Black Gate FK and NX types and Sanyo Os-Con types. These are 'exotic' components and rather expensive.

The overall sound performance with these capacitors is so much better than a *LT1085/LT1033* based power supply using good quality and normal priced Elna RSH capacitors. Using Keith Jarrett's 'Köln Concert' as reference music, you can easily tell which power supply is 'playing'. The soundstage is so much improved. Jarrett's piano really 'sings'. In my opinion there is no doubt that the BGs and Os-Cons improve the sonic overall performance of a system; my ears are me tell me so.

Keep up the good work,
W. de Haan
Leiden, The Netherlands

Agreeable distortion

'Valve sound' is essentially subjectively agreeable distortion. An analogy is the measurable sensation of travelling in a vintage Bentley rather than in a modern mid-range Ford, which is noticeably better in most respects – if not at all.

However, there is one difference – cost. Preferred output valves cost upwards of £20 each.

Morgan Jones' excellently presented article – Jan '96 – exhibits at least one flaw, however, as many of the resistance values arrived at by parallelling are within a fractional percentage of standard values. For example 330kΩ in parallel with 22kΩ is 20.625kΩ. A near value in the E96 range is 20.5kΩ – less than 1% off. If cost is no object, this is the way to do it.

I worked with valves for many years and came to accept that their characteristics varied widely from part to part. Anybody who used the *EF50* will remember this. There is no point in attempting exact design where key parameters can differ by as much as 20%.

In any case, the principal feature of valve amplifiers is that they include an output transformer. If one takes a good solid-state amplifier and includes a 1:1 output transformer within the feedback loop one will achieve much the same effect.

Of course, valve and solid-state amplifiers driven near to or past saturation will sound different, but if one is any sort of a purist this is not a region in which one operates. Vast power capability overkill is an essential feature of hi-fi usage.
Nick Wheeler
Sutton
Surrey

Valve misunderstanding

As a designer of valve amplifiers since 1950 I have read with some disbelief the article by Morgan Jones in *Electronics World* January 1996 and the subsequent correspondence in the February and March issues. Both Morgan Jones and Frank Ogden seem to not understand the operation of the concertina phase splitter.

This circuit does not have the alleged difference in frequency response at the anode and cathode terminals. If the anode and cathode outputs are analysed separately then, of course the anode output resistance is high and the cathode output resistance relatively low as shown by Morgan Jones in his March 1996 letter. However when both outputs are loaded simultaneously with equal capacitances the output voltages remain equal throughout the audio frequency range.

This can be understood intuitively since the anode current is the same as the cathode current so when the two impedances are equal (i.e. equal resistances and equal capacitances) then the output voltages must be equal at all frequencies. It is obvious that the tendency for the anode voltage to decrease more rapidly as the frequency is raised is fully compensated by the by-passing effect of the cathode loading capacitance.

The circuit behaves as if the output resistance at both ports is much the same as the source resistance of a cathode follower using the same valve and cathode load resistance. It can be shown that the effective output resistance used to determine the frequency response at both outputs is,

$$R_o = \frac{r_a R_L}{r_a + R_L(\mu + 2)}$$

Needless to say the 'build-out' resistor spoils this inherent wide-band balance of the concertina phase splitter.

There is another error in the Morgan Jones article in the January 1996 issue where he attempts to balance the signal currents of the input stage and the concertina. The concertina signal current is approximately grid voltage divided by cathode resistance thus the anode load of the input stage should be roughly equal to the concertina cathode resistance and not cathode resistance plus anode resistance as stated.
M.H. McFadden
Belfast

Reference

1. 'Radio Designer's Handbook' F. Langford-Smith p.329 Fourth Edition. Published by *Wireless World* 1953.

Shame about the error

At present I am particularly interested in the subject of valve audio amplifiers. While not having sufficient detailed information on valve characteristics at hand to check all the calculations in the January's valve power amplifier article, I was disappointed to find a clear error in the calculation of the values for the feedback resistor and the input stage cathode resistor. While the circuit diagram indicates a 4Ω output load, the calculation is based on 8Ω.

Speakers with 3Ω or 15Ω coils were common before the advent of the 8Ω speaker. This made a dual secondary winding on the output transformer popular, giving an output impedance of 4 or 16Ω. My calculations show that with a 4Ω load, a cathode resistor of 964Ω is required, and a feedback resistor of 1728Ω; with 16Ω loads they should be 753Ω and 3456Ω respectively. The method of calculating the feedback capacitor was not explained, but this should be less critical than the resistor values, and it should be adequate to adjust this proportionately.

The required values could be obtained in the case of each resistor by using two parallel resistors of standard values as in the article, values as follows:
Stephen Cole
Winscombe
Avon

I can't hear you

For once I find myself in agreement with Ben Duncan, on the issue of the suitability of Windows (Review of Micro-CAP V, *EW+WW* Sep '95). It seems absurd that professional pc users should be saddled with a software package that appears to be a re-invention of an operating system designed in the early seventies for children. Windows is ok for the novice user, but without much doubt anybody with a modicum of experience with a standard keyboard would find it more efficient than a mouse. Windows is, in my opinion, poorly documented, slow, cumbersome and not very logical, and a running joke among my computer literate friends. Unfortunately it is difficult to get by without it, and maintain compatibility.

To load and run Windows at an acceptable speed requires no less than a 486 – most pcs in our department are 386s – at least 8Mb of ram and a large fast hard disk. This hardware is only now becoming acceptably cheap, but Microsoft would like us to move up to Windows 95 with even greater demands on our hardware. To quote one John McCormick, "Why would anyone in their right mind use Windows for anything? You can always buy a slower computer if yours is too fast!" (from "It's not a Bug, It's a Feature!" by David Lubar).

Unfortunately that is the end of good news for Ben. In his article 'Simulated attack on slew rates' (*EW+WW*, April '95) Ben boldly states on p. 307 that "...the headroom is demonstrably safer for drive units and ears alike – no matter how counter-intuitive this seems" in the course of his justification of very high slew rates and the reproduction of "...music transients above 165V...". Ben opened the piece by outlining the high frequency nature of the sound "during an Iron Maiden gig" engineered by a colleague.

New Scientist reports (p5, 27 Jan '96 No. 2014, Australian edition) that "rock concerts are more likely to damage your hearing than listening to a personal stereo or going clubbing", according to French hearing specialist Christian Meyer-Bisch. This conclusion is the result of a study of 1364 people, and

Best rf article '95

Entries for this challenge are currently being evaluated. We hope to be able to make an announcement about the winner in next month's issue.

it is the high frequency content of rock that is identified as the major cause. "Rock is much tougher on the ear at high frequencies than classical music. When played at the same volume on a CD player, the music of heavy metal bands, such as Iron Maiden, is far louder at high frequencies than a piece of Vivaldi" (I think that should be "a piece *by* Vivaldi", I doubt that there would be many pieces of Vivaldi left). The situation is much worse at rock concerts because of the much higher power.

Ben is quite wrong. It is sensible to keep listening levels moderate, particularly for extended periods and especially for high frequencies. There is no good reason to believe that high slew rates are less damaging to the ears. In fact the reverse is more likely to be true. Higher sound levels are more likely to increase that risk of permanent hearing loss. Ben would be well advised to keep some of his 'counter-intuitive' ideas to himself lest he – and his colleague with Iron Maiden – become the target of litigation from deaf concert goers.

Phil Denniss
University of Sydney
Australia

Cable rejection

If I manage to get a common-mode rejection of 3000dB does this mean the end of the universe, and we all get sucked into an audio black hole? (We all know that black noise is the equal absence of noise/√Hz).

On a more realistic note, I find a cable tolerance of a couple of percent to be optimistic; have you measured a cable that has been on the road for six months or so, trodden on, run over, stretched over balconies and generally abused. Have you measured, in real life, such a cable? There is no mention of other cables such as star-quad, or multicore.

Many fixed installations use the Krone IDT method, or similar, involving overall screened cable with say 48 different signal pairs – all with various levels of signal and impedance imbalance. I've used this system a few times. Implemented with care, provides a competent way of installing audio systems.

Just simulating a single cable seems very simplistic. These days you have to consider the whole system, although a basic understanding of common-mode rejection ratio is essential.

Although I have not been involved directly with professional audio for a couple of years I found that:

- In practice you cannot beat the *5534* differential amplifier for a line receiver with a couple of 22pF trim capacitors for trimming common-mode rejection ratio. The single op-amp differential stage is fine for local use.

- The *SSM2142* is a poor device with not very good output common mode rejection, and its relatively noisy. Porter produced a far superior balanced output stage, published in *EW ca* 1989. This had a cm rejection ratio of at least 60dB across the audio bandwidth – even built on veroboard.

Please can Ben Duncan stop pushing Microcap and SSM devices – and stop living in SimCity?

Martin Griffith
Compuserve

Summing up Foster Seeley

I was interested in the article on the Foster Seeley detector in your Dec issue.

I feel the author makes heavy weather of its operation. A qualitative description of the operation of the circuit is as follows:

- The primary voltage & the voltage *injected* into the secondary circuit are in phase – as with all transformer circuits.
- At resonance, current in the secondary circuit is in phase with the injected voltage; this is more easily seen if the secondary circuit is drawn as a series circuit.
- Output voltage across the tuning capacitor lags this current by 90°.
- Thus, the accessible primary and secondary voltages differ by 90° at resonance, as normally drawn in analyses of the circuit.
- Off resonance, the phase of the current to the injected voltage varies, so varying the phase of the output and primary voltages.
- As a side issue, an rf transformer cannot usefully be double-tuned – primary and secondary – if it is tightly coupled. The two capacitors are just in parallel.

Regarding the ratio detector, I prefer to regard it as a sampling circuit. The voltage across the secondary switches the diodes on at its peak; and at that instant, they pass the instantaneous value of the primary voltage to the af output point (where it is stored by the *capacitors*, when the diodes cease to conduct). At resonance, this primary voltage is zero at the peak of the secondary, because the two are in quadrature; off resonance, it varies to give the af output.

I hope these points may help some who find the operation of the circuit difficult to picture from the bare analysis.

J.W.E. Jones
South Australia

Sallen & Key disadvantages

Following recent correspondence on the Sallen and Key filter configuration, I would like to remind readers of a further weakness in the practical implementation of the low-pass configuration. The signal passes through a resistor and then has a path, through the supposed 'feedback' capacitor, to the filter output. If the op-amp output impedance is extremely low – which we assume – then this signal path is effectively shunted to ground.

In reality, however, the output impedance of an op-amp rises with frequency as the open-loop gain falls. It can reach many tens or even hundreds of ohms. Then, high-frequency components of an input signal can leak through to the output.

This failing can be plotted on even the student version of PSpice, where the filter attenuation plot reverses at high frequencies, passing noise and distortion components of the drive signal. It does not occur with the low-pass Rauch filter.

Simon Bateson,
Hutton Rudby
North Yorkshire ■

HELP Wanted

Any queries?

If you have any electronics-related questions that you have not been able to find an answer to, why not see if other readers can answer them? Simply write to me, the editor, at the address on page 267, fax 0181 652 8956, or e-mail martin.eccles@rbp.co.uk.

Can you answer this?

Could one of your readers explain to me a phenomenon connected with the distribution of lines of magnetic flux, of strength,

$$H = \frac{NI}{2\pi r}$$

around a single length of wire carrying a dc current of 1A. With this wire passing through a card at right angles to the wire; if soft iron filings are sprinkled around the wire magnetic lines may be observed which form concentric circles around the wire with spaces between them.

My question is this: has some form of standing wave been set up in the spacing between 'crests'? Being a wavelength the speed of which may be expressed as:

$$\sigma = f_o \lambda ms^{-1}$$

where, were it not for friction would represent the speed of a magnetic field of strength H with frequency f_o where f_o is the frequency of electrons moving around a closed circuit the direction of propagation, as with Huygens wave theory being at right angles to the tangent, of each circular path, i.e. radially. A wire being taken as the simplest and most easily analysed configuration.

Dust tube analogy. If lycopodium powder is placed uniformly within a tube and a pure note of frequency f sent down the tube, disturbances would be set up which if in antiphase with the reflected wave would cause the powder to respond by 'clumping' in heaps at the points of little disturbance, i.e. at rarefactions. This analogy is used to consider the concentric lines of force around a single turn of wire.

I would appreciate any information you may be able to supply me with.

Terence George Heatley
London

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ACTIVE

A-to-D and D-to-A converters

Delta-sigma d-to-as. Two 20-bit, stereo digital-to-analogue converters by Crystal Semiconductor, the *CS4327/9* handle an infinitely adjustable sample rate of 1-50kHz. The *4329* uses switched-capacitor analogue, low-pass filtering to give clock-jitter tolerance and a 105dB dynamic range at 20-bit resolution and 112dB s:n ratio; thd+noise is –97dB. *CS4327* gives 98dB dynamic range, 110dB s:n and –90dB thd+noise. Sequoia Technology Ltd. Tel., 01734 258000; fax, 01734 258020.

Linear integrated circuits

Video amplifier. *EL4093* from Elantec is a complete, dc-restored video amplifier subsystem, providing accurate dc restoration, 300MHz bandwidth and 1500V/µs slewing. It contains a current-feedback amplifier and a s/h amplifier to stabilise video performance, providing signal acquisition in one line scan. If the ttl-compatible hold input is low, the s/h is used to null the dc offset of the video amplifier; when high, it stores the correction voltage on the hold capacitance to maintain dc correction during the next line scan. METL. Tel., 01844 278781; fax, 01844 278746.

375MHz buffers. *MAX4178/4278* (single) and *MAX496/497* (quad) closed-loop buffers have fixed gains of +1 or +2, give a 70mA output minimum, an output swing of more than ±2.5V into 50Ω and tolerate a 70pF load without oscillating. 4178 and 496 have the +1 gain, a 375MHz bandwidth to –3dB (80MHz at –0.1dB) and slew at 1400V/µs, while the other two exhibit 275NHz at –3dB, 120MHz at –0.1dB and a slew rate of 1500V/µs. Supply curent is 8mA/channel and differential phase and gain are 0.01° and 0.01%; input voltage offset is 0.5mV and input noise 5.6nV/√Hz.
Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

Memory chips

32bit flash proms. *EDI* high-speed, high-density flash SIMM and PLCC modules have a 32-bit data bus in capacities to 32Mbit. They are available in single or dual form, organised as 128K by 32 and 256K by 32 in either package; the 512K by 32 devices as SIMMs only. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261678.

1M sram. Toshiba's *TC55V1664/1864* 1Mbyte, 0.4µm cmos srams offer 15 or 12ns access times, wide bandwidth and 3.3V operating voltage, with performance equivalent to that of 5V devices. Toshiba Electronics UK Ltd. Tel., 01276 694600; fax, 01276 694800.

Microprocessors and controllers

Mixed-signal controller. Microchip's *PIC14000* is compatible with the *PIC16/17* architecture, and has 4K by 14 eprom and 192byte of ram. Its 5Mips, 8-bit risc core gives 35 single-word instructions, 20MHz operating speed, six internal and five external interrupt sources, eight levels of hardware stack and 38 special-function hardware registers. The 16ms, 16-bit a-to-d converter is accompanied by two multi-range converters, a low-voltage detector, temperature sensor, voltage control and a 4MHz clock. The device is supported by the *PICMaster* development and emulation system. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

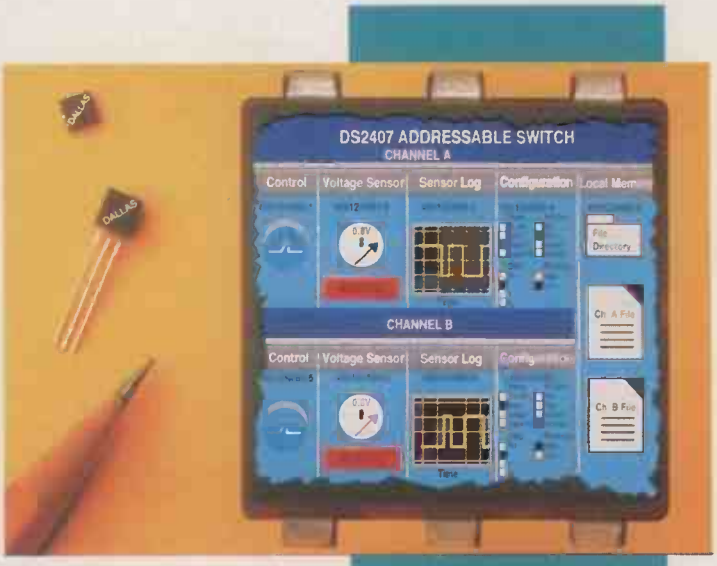
Optical devices

Low-current isolator. *ISP817* from Isocom is an optically coupled isolator that takes a drive current down to 0.5mA, while still producing a high output current – current transfer ratio is 70% at 0.5mA and 100% at 1mA. Forward saturation voltage is 0.4V and i/o voltage isolation is 7.5kV. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

PASSIVE

Transformers. *Clairtronic* transformers in both chassis and pcb mounted types are available from Electrospeed. The units are made in flame-retardant UL94V-0 material, are designed to meet EN60742/60950 safety requirements and are 100% tested for safety factors. Chassis-mounted types have a single primary and dual secondaries, the series comprising units rated at 6VA-50VA. Pcb types have two secondaries rated at 3VA-12VA. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Chip resistors. From the Taiwanese company Yageo comes sub-miniature 0402 chip resistors for both flow and reflow soldering. There is a full range, from 1/16W to 1W types in values from 10Ω to 10MΩ. Easby Electronics Ltd. Tel., 01748 850555; fax, 01748 850556.



Mixed-signal ICs Sensor/actuator. Dallas announces the *DS2407*, addressable switch ic, a sensor/actuator to perform closed-loop control from a PC via an RS-232 link. The sensor responds to a stimulus and inputs it to the PC, which arrives at a decision and instructs the actuator to switch on or off, all over a twisted-pair wire of up to 300m in length, which includes power for the chip; driver software provides GUIs on the PC screen running something like LabView. Many such ics may be connected to the same wire to be controlled by a central PC, since each ic has its own serial number on-chip. Dallas Semiconductor Corporation. Tel., 0121-782 2959; fax, 0121-782 2156.

Chip capacitors. A new range of chip capacitors on 4in diameter reels containing as few as 500 pieces, made by *miniReel*, come in 0805 and 1206 chip sizes in values from 1pF to 2.2µF and using C0G, X7R, Z5U and Y5V dielectrics, depending on value. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Power electrolytics. A useful life of 10000 hours is quoted by Philips for the *PLL-SI 058/059* series of snap-in electrolytic capacitors, which tolerate temperatures from –40°C to 105°C. Capacitance range is 33-47000µF at ±10% and at voltages of 10-100V and 200-400V. These units are charge and discharge proof. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

High C, high voltage, small size. *Wima MKS2* capacitors are particularly useful for decoupling, values available including 0.01µF to 2.2µF, 3.3µF and 4.7µF in a 5mm pitch encapsulated package and rated at 50Vdc or 30Vac. Tolerances are ±20%, ±10% and ±5%. Europa Components & Equipment plc. Tel., 0181-953 2379; fax, 0181-207 6646.

Connectors and cabling

Stackable board connector. To connect boards in a parallel stack at varying distances, Harting Elektronik offers the *har-mik* connector system, which uses male connectors of a constant height and female connectors of different heights. Harting Elektronik Ltd. Tel., 01604 766686; fax, 01604 706777.

Board-mating connectors. Samtec make connectors to join printed boards together at right-angles or parallel to each other. They are available with surface mounting or through-hole terminations and on pin pitches of 1.27mm, 1.27 by 2.54mm and 2mm. Samtec UK Ltd. Tel., 01236 739292; fax, 01236 727113.

Chip carriers. Plastic-leaded chip-carrier sockets by Data I/O, provide reliable mounting for ic programming in middle to high-volume production. The sockets are available for the company's *2900*, *3900* and *UniSite* programmers and fit the receptacle on the programmer, replacing programming adaptors. Data I/O Ltd. Tel., 01734 440011; fax, 01734 448700.

Cable-to-cable connector. Framatome introduces the *Trim Trio SMS Qikmate*, which connects two free cables of widely varying diameters, strain relief being incorporated. Moulded hoods are provided and there is provision for polarising the sockets with extra pins. Framatome Connectors UK Ltd. Tel., 01582 475757; fax, 01582 476203.

Displays

CRT shielding. Magnetic shielding material for colour monitors is produced by Ad-Vance Magnetics to address the requirement for alternating and static field shielding in heavy industry and laboratories where higher than usual fields are experienced, of the order of 10-50 oersted. It is available in 0.64mm sheet and is usually used in two or three layers to reduce a static field of 45 oersted, for example, to 0.16 oersted near the centre of the enclosure. Ginsbury (UK) Ltd. Tel., 01634 290903; fax, 01634 290904.

Filters

Emc filters. Three new ranges of equipment filters by *MPE* are to protect against incoming and outgoing interference at power inputs and cover 1-15A. General-purpose dc types are 100Vdc rated, while the ac filters and mains-input types for switched-mode supplies are rated for 250Vac at 50/60Hz. All use feedthrough capacitors and can be bulkhead mounted to help meet the EMC Directive at high frequencies. A catalogue is available. MPE Ltd. Tel., 01371 875071; fax, 01371 875037.

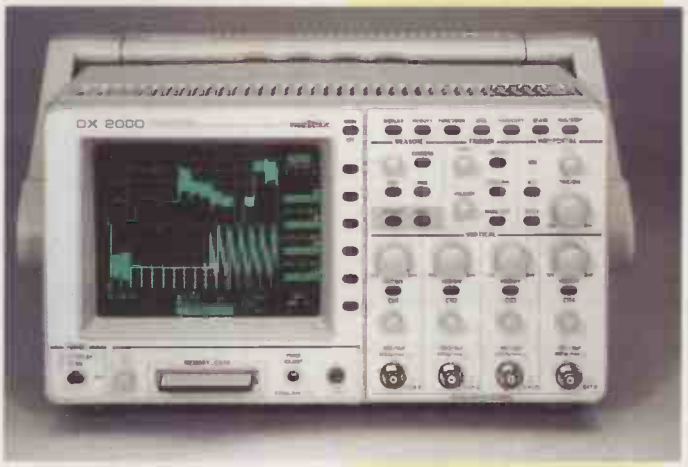
Hardware

Power backplanes. 13-slot C and D sized, 12-layer VXIbus backplanes from Vero can handle powers of more than 3kW. The 9U size has dual 0V ground busbars and four laminated power busbars for low-impedance power distribution or seven voltage rails. The 6U size has two of each. Both conform to the latest VXIbus specification. All sizes have decoupling capacitors in the termination area and additional positions for decouplers at each slot position. Vero Electronics Ltd. Tel., 01489 780078; fax, 01489 780978.

Command panel. Rittal offers the *VIP 6000* housing for machine tool controls and process control stations. It takes all the common control systems and is available with a keyboard housing or tray, the housing being designed to take a machine control panel or a keyboard, the keyboard tray holding a standard keyboard. Screw channels on a 25mm matrix are provided for individual layout design and there is easy access from the rear. Rittal Ltd. Tel., 01709 704000; fax, 01709 701217.

PCMCIA kit. Molex has the *Snapper* kit, which contains all the bits and pieces needed to make a PCMCIA card. The resulting card is compatible with both PCMCIA and JEIDA standards; the Type II kit containing a black plastic frame, a stainless steel snap-on cover, a 68-contact surface-mounted connector and a 15-position 1.27mm-pitch input/output connector. The Snapper cover only needs a small arbor press to close and seal the unit. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Self-assembly emc test chamber. For emc testing, Seaward's *Easi-Screen* is a lightweight emc test compartment in kit form for assembly. Attenuation is better than 60dB. Construction is of polyester/copper/nickel shielding fabric with steel and veneer door and particle board and sheet steel floor. A 16A mains filter and distribution system with an isolator is provided, as is powered ventilation, coaxial inputs and a 60mm waveguide. The chamber is easily dismantled for storage. Seaward Electronic Ltd. Tel., 0191-586 3511; fax, 0191-586 0227.



Test and measurement

Clamps. Northern Design says it has the biggest selection of clamp-on current probes in the civilised world, from the *Micro 2000* finger-operated miniature device for 1mA-200A measurements, to the *P Series* for measurements to 3000A. The range of jaw sizes covers conductors from 15mm cables to 120 by 50mm bus bars. Output can be ac or dc voltage or current to accuracies of 0.25% in the miniature versions or Class 1 for the bigger types. Northern Design (Electronics) Ltd.Tel., 01274 729533; fax, 01274 721074.

Microwave test. MI's *6250 Series* millimetre wave reflectometers extend the insertion and return loss measurement performance of the *6200* microwave test set. Model *6255* multiplies the output of the *6200* to give frequencies in the 50-75GHz range (V band), while the *6256* produces frequencies from 75GHz to 110GHz (W band). Marconi Instruments Ltd. Tel., 01438 742200; fax, 01438 727601.

Network analyser. Rohde & Schwarz introduces the *ZVR* vector network analyser family which, among its other virtues, is modular in form for simple future upgrading. Measurement time is under 120µs per test point, which allows over 25 sweeps/s with 200 points and over 130dB dynamic range with a 10Hz if bandwidth. There are three family members, all with integral generator, test set and multi-channel receiver, and the two lower-priced units can be upgraded to perform as the most expensive one. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

150MHz dso. From Metrix, the *OX2000* 150MHz, four-channel, programmable digital storage oscilloscope, which can capture data at up to 200Msample/s in single-shot mode and to 50Gsample/s for repetitive waveforms. Input sensitivity is 2mV-10V/div and sweep speed 2ns-50s/div. A PCMCIA slot allows long-term storage and a colour VGA output port is provided, as well as interfaces for printing or connection to a PC. Metrix Electronics plc. Tel., 01384 402731; fax, 01384 402732.

GPIB multimeter. *Model 1705GP* from Thurlby Thandar is a GPIB version of the 1705 dual-display multimeter, possessing IEEE-488 and RS-232 interfaces, either of which controls the meter functions and reads back results from the display or the built-in data logger. This 4.5 digit instrument counts to 12000, has a 10µV, 10mΩ, 0.1µA resolution and direct voltage accuracy of 0.04%. Main and secondary displays show two simultaneous readings and the secondary one will show measurement units, the results of calculation, two different parameters of one signal or two different signals. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Audio monitor. Audix's *ARM* audio monitor is now in a new version with 24 stereo inputs instead of twelve; it is meant for on-air broadcast use. There are separate buffered and control outputs for an internal mono cue speaker, an external stereo loudspeaker and stereo headphones connected to the panel's jack. There is an external communications input to inject feeds to the cue speaker. Audix Broadcast Ltd., Tel., 01799 542220; fax, 01799 541248.

Spectrum analyser. Advantest's *R3263* spectrum analyser is intended for use in digital mobile communications. It is small and light,

but provides comprehensive facilities in the 9kHz-3GHz range, with selectable bandwidth from 300Hz to 5MHz. The screen is a 6.5in colour tft type displaying a 100dB range of levels at a horizontal resolution of 1000 points. There is gated and delayed sweep and a timing function to 20µs for burst measurement and one keystroke starts fully automatic test sequences. Two PCMCIA slots allow storage, set-ups and test programs. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Literature

Display panels. Thin-film transistor, active-matrix lcd panels by NEC are the subject of a new brochure, which shows types from a 6.5in unit for instruments to the new 1280 by 1024-pixel, 13in panel for monitors. The brochure contains a section to explain the operation of tft active-matrix displays. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

Valves. A note from Billington Export offers its 1996 catalogue, which contains cross-referencing data, and points out that the company has the *SV811* power triode from Svetlana and the improved Chinese 300B with graphite anode. There is also a separate crt catalogue and both are free. Billington Export Ltd. Tel., 01403 784961; fax, 01403 783519.

Alarms. Roxburgh's complete range of audible alarms and indicator lights is described in the 1996 catalogue, now available. Components included are magnetic buzzers and transducers, piezoceramic transducers, pcb and panel alarms, among which is the *Sonitron* range. There is also a catalogue on the range of *Rafi* electromechanical components – switches, lamps and keyswitches. Roxburgh Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

Floppy catalogues. Minicat Ltd has a compression technique that will put 200 colour images and 1000 pages of text on a 3.5in floppy disk – about 450 times as much as usual. The company also offers an interactive slideshow facility with fade transitions for conferences, running under Windows. MiniCat Ltd. Tel./fax, 01923 823633.

Hitachi on CD-ROM. A new CD-ROM data book from Hitachi covers the *H8* series of microcontrollers and the *SuperH* family of 32-bit risc devices, the disc being effectively equivalent to 19,000 pages of data. Macintosh and Windows users can read the disk. Hitachi Europe Ltd. Tel., 01628 585163; fax, 01628 585160.

Production equipment

Pcb test. Polar's *Toneohm* family of low-cost printed-board short-circuit fault locators is extended to include the *550A* and *850A*, which provide 0-40mΩ ranges for short-circuit tracing on boards with wide tracks, The 850A also has current tracing for shorts on bus-structured boards; both are usable on bare or loaded boards. In use, probes are moved along the tracks while a tone guides the user to within a few millimetres of the fault. All data is presented on a 3.5-digit lcd. Drive is voltage limited to avoid damage. Polar Instruments Ltd. Tel., 01481 53081; fax, 01481 52476.

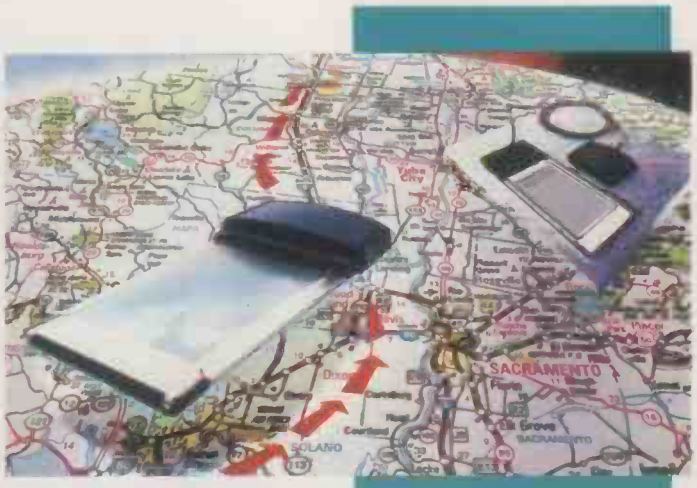
Materials

Liquid resist. Electra announces *Photrak*, which is a liquid photoimageable etch and plate resist for high-resolution pcbs; it can be applied to give 1mil resolution. Using standard 5kW equipment, exposure time is 15-20 seconds and with 7kW, 10 seconds. The material increases developer and stripper bath life by 100%. Application is by screen printing, curtain coating, electrostatic spray or roller and the formula is suitable for use with acid and alkaline etchants, as well as with acid gold-plating solution. Electra Polymers and Chemicals Ltd. Tel., 01732 811118; fax, 01732 811119.

Printers and controllers

Thermal printer. Able Systems has the *Ap1000*, a panel thermal printer in a clear plastic case so that the amount of paper left is visible. It comes in 24 or 42 column form and gives a speed of 96 characters/s, bidirectionally. A full IBM character set is provided. Able Systems Ltd. Tel., 01606 48621; fax, 01606 44903.

Board inspection. *Alpha Hi-Check 500Z* is an accurate, non-contact method of inspection and measurement for printed boards, the workpiece being shown on a high-resolution monitor and/or printed copy. The instrument gives readings down to ±3µm and visual inspection of pads at a magnification of 14 to 270, a 45° attachment allowing all-round vision. Focus is automatic and the measurement readout is shown on screen. Data may be transferred to a pc running Excel. Alpha Metals. Tel., 0181-665 6666; fax, 0181-665 4734.



Navigation systems

PCMCIA GPS. Using only 650mW, Rockwell's *NavCard* LP PCMCIA Global Positioning System receiver is a five-channel unit tracking up to nine satellites to give position, direction and speed, mainly for land vehicles and marine use. It is complete with an integrated antenna, removable to allow the use of an optional remote antenna. Software includes *CityTracker* for urban navigation. If a differential receiver is available, the unit accepts input to improve position resolution to 10m from 100m. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

Power supplies

Switcher controller. Linear's *LTC1430* switching regulator controller converts 5V to 3.3V, 2.5V or other processor core voltages at up to 15A and is for use in equipment based on *Pentium* and *P6* processors. Efficiency is near 95% at high currents and good transient response reduces the size of filter capacitors required. The voltage feedback technique used eliminates the current sensing resistor commonly used. A soft start feature is incorporated. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261678.

SOT-23 voltage reference. *MAX6120* from Maxim is said to be the first micropower, 1.2V three-terminal reference in this package. It is meant for 3V equipment where battery saving is essential and is a low-power alternative to two-terminal shunt devices, since its supply current of 70µA maximum is independent of input voltage. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

10W, open-frame supplies. Toko's *SW10* series of 10W ac/dc open-frame supplies stand only 18mm off the board and take up 65 by 70mm of board space. Input is universal – 85-246V ac – and the units give a single output of 5V, 12V, 15V or 24Vdc, led status indication and a fine output adjustment being standard. Closed-frame types are available. Melcher Ltd. Tel., 01425 474752; fax, 01425 474768.

Rapid-response FORS. If uninterruptible power supplies look likely to be interrupted, Fiskars Power Systems will instantly leap to attention and send in the cavalry. *FORS* (Fiskars On-line Remote Service) is a system whereby the company keeps tabs on its ups units in the field 24 hours a day by way of modems and GSM links, automatically and remotely inspecting all systems, listening for alarms and collecting the relevant data if it thinks it sees a problem. If it does, it calls out the duty engineer and gives him all the necessary data, although he can call for more if he wants to. The service is available for Fiskars PowerServer 30/40, UPS9000/10000 systems. Fiskars Electronics Ltd. Tel., 01734 306600; fax, 01734 305868.

2.5W SOT-23 rectifiers. Microsemi's *Powermite* family of small semiconductor devices now includes a 2.5W, fast 1A schottky rectifier, due in part to the design of the surface-mounted package. Its metal base wraps round each side of the device to increase the heat flow to the board. Its success is demonstrated by its ability to cope with an 8.3ms surge of 70A. Solid State Supplies Ltd. Tel., 01892 836836; fax, 01892 837837.

Switches and relays

Photovoltaic relay. IR has increased its family of photovoltaic relays for Type II PCMCIA fax/modem cards with the *PVO402P*, which is only 2mm high and consists of a double-pole, normally open, solid-state device incorporating both relay and ring

detector. Input/output isolation is 3.75kVrms. Output stage is a Hexfet circuit. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

SM dip switches. Grayhill offers the *Piano-style* and standard-profile spdt, spst and 2pst dual-in-line switches in surface-mount form, made from material to withstand infrared reflow soldering. Roxburgh Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

Windows '95 keyboard. Cherry's Windows '95-compatible keyboard, the *G83-6105*, is a 105-key device with three dedicated keys on the spacebar row: an applications key to pop up the content menu (equivalent to the right mouse button in some applications); and left and right keys for the user interface and its shortcuts. Its membrane switch combined with a rubber sheet, whose domes are individually moulded to provide a uniform response, give an improved action in any position. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Reliable keypads. Oil from the fingers is kept from Lucas rubberised keypads by means of a layer of polyester in *Duralith* barrier switches. These are polyester half switch consisting of a contact layer, screened contacts and a spacer layer. A range of options includes a choice of tactile response and pcb substrates. Lucas Control Systems Products. Tel., 01535 661144; fax, 01535 661174.

Quiet, solid-state relays. Solid-state relays, by Laser Energy, in the *ECO* range 'totally eliminate' additive radio frequency interference, meeting VDE 0871 well enough to class the device as noise-free. Current handling is 10-100A and forward voltage is reduced to enable a reduction in heat sink size. No additional filters are needed. Campbell Collins Ltd. Tel., 01438 369466; fax, 01438 316465.

Keylock switches. Grayhill's *Series 03* range of low-cost keylock switches is now available in the UK, a range that includes multi-level security switches and basic on-off types. There is an on-off model measuring 0.6in in diameter, a two-position dpdt switch, a three-position progressive-contact switch and a multi-level type giving four-operator security and limited access to switch positions. Switches have ratings of 1-2A at 240V ac. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Octal bus switch ics. Bus switches in the *FST3xxx/32xxx* families serve to solve the problems associated with shared memory and multiple processors in common buses without additional propagation delay, timing skew, noise or power consumption. Quiescent current is typically 0.1µA. Integrated Device Technology. Tel., 01372 363734; fax, 01372 378851.

Pot. switches. *Eco* switches by Omeg come in rotary and push-push varieties and are meant to mount directly onto the company's 16mm ECO potentiometers. The rotary switches are produced in ratings of 1A and 4A at 250V, in single and two pole types and terminated in pcb pins or tags. Push-push models are 10A, 250V units and are also available as modules for other manufacturers. Power rating of both types is 0.25W in linear ranges of 1kΩ to 1MΩ and 0.12W for non-linear types from 4.7kΩ-470kΩ. Omeg Ltd. Tel., 01342 410420; fax, 01342 316253.

Attenuator relays. Teledyne's *RF300* relays are small (7mm high), are emi-shielded and handle high frequencies and are therefore suitable for use in uhf attenuators. Rf signal repeatability is 241±0.1dB from zero to 3GHz. Teledyne Electronic Technologies. Tel., 0181-571 9596; fax, 0181-571 9637.

Keyboard switches. Providing a snap action and a satisfying feel, NSF *Keylite* keyboard switches come in various colours and designs and posses momentary or latching action. They accept one or two leds and are fitted with lugs resistant to solder creepage and gold/silver-plated contacts. Designs in the range include half key, stepped, paddle, sloping and illuminated types. Lucas Control Systems Products. Tel., 01535 661144; fax, 01535 661174.

Trip amplifiers. Providing relay contact at preset ac and dc levels, *UltraSlim Pak* trip amplifiers from Weidmuller Klippon are easily configured, with setpoints from 10mV to 200V; input currents are 1mA to 100mA ac or dc. Isolation between input, output and power supply is provided and the two output relays are in spdt form and rated at 120Vac or 24Vdc. Weidmuller (Klippon Products) Ltd. Tel., 01795 580999; fax, 01732 844444.

Transducers and sensors

Slotted sensors. Omron has added to its range of optoelectronic switches a number with increased slot widths of 8mm. *EE-SX1070/3070/4070* are configured as phototransistor, photo-ic (light off) and photo-ic (light on) respectively, all with resolution to 0.5mm. The photo-ic versions have an amplifier and Schmitt to give high output for direct drive of other circuits; frequency response allows 3000 operations per second. Omron Electronics Ltd. Tel., 0181-450 4646; fax, 0181-450 8087.

Displacement transducers. Monitran's new linear differential displacement transducers are for use in applications where they must withstand pressures up to 6000lb/in^2, or 400bar. *MTN/P* units can be used inside hydraulic and pneumatic cylinders to act as feedback devices for actuator control. They are in stainless steel and come in measuring ranges of ±25mm to ±500mm, giving dc or current-loop output. Monitran Ltd. Tel., 01494 816569; fax, 01494 812256.

Magnetic field sensor. Designed to detect and measure a changing magnetic field, the Zetex *ZMY20M* now tolerates disturbance fields up to 30kA/m. It takes the form of thin-film magnetoresistive permalloy in a Wheatstone bridge arrangement to give an output proportional to the field. An internal magnet in the E-line or SOT223S package counteracts unwanted external disturbances to allow measurement down to 0.1kA/m. Bridge resistance is 1.7kΩ and output is 12-22mV/V at 0-1MHz. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Rotary sensor. Control Transducers's *WPM* absolute rotary position sensor is completely self-contained and uses the MystR conductive plastic for long life with excellent linearity (±0.075%) and resolution. It is contained in a 22-51mm anodised aluminium housing for servo mounting. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

COMPUTER

Data communications

V.34 modem. Rockwell's *RCV288ATFW/SP* modem chip is a complete V.34 design offering 115.2kb/s data and Group 3 fax, voice and speakerphone facilities; it needs no external controller. Adpcm coding and decoding allows digital storage using 2-bit or 4-bit compression and 7200bit/s decompression, while the voice mode supports business audio and Rockwell's integrated communications system programme for digital phone answering, voice annotation and audio file play and record. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810

Little transceiver. *STD-300* from Circuit Design is a 50 by 28mm narrow-band radio data transceiver intended to add telemetry to portable data terminals such as data loggers and card readers. Its high selectivity programmable pll-synthesised transmitter stage and a sensitive double superhet synthesised receiver allow a reliable range of 1km at a data rate of 2400b/s. It operates in the 434MHz band and is compatible with ETS-300-220; spurious emissions are at less than –60dBm and under 200nW to adjacent channels. Low Power Radio Solutions Ltd. Tel., 01993 709418; fax, 01993 708575.

Modem modules. *SocketModems* are a pin-compatible range of modem modules, including a low-power 2400bit/s data-only type up to a V.32bis type providing data, fax and voice. Also in the range is the *TDE-D300* parallel interface, an ISA-bus card to go in a pc's 8-bit card slot, hosting any parallel socketModem. A demonstration board has a speaker and a socket for DAAs. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

Industrial notebook. A joint GE/Lockheed/Martin Marietta/Mitac project produced the *MNB* series of heavy-duty notebook computers for use in unfriendly surroundings. It has either a 486DX2 66 or a 486DX4 100 processor with 4Mb dram and a 520Mb, 2.5in removable hard disk. Lcd displays of various types can be provided, with provision to connect an external VGA monitor. There are two PCMCIA slots for Types I or II cards and a standard ISA or two PC104 cards can be used internally. The whole thing is in a cast aluminium chassis and enclosure. Kerry Technology Ltd. Tel., 01825 766776; fax, 01825 768020.

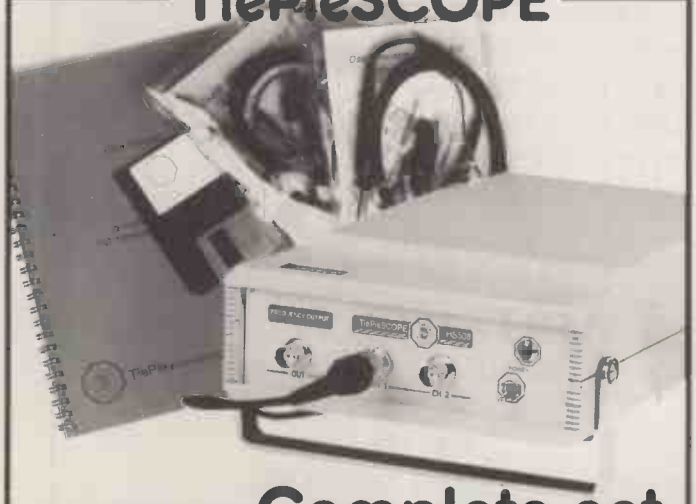
Data logging

Portable logger. A new, portable data logger, the *SA32* from Martron, has on-board data-processing functions, takes 33 input channels and measures voltage, current, resistance and temperature. Sampling speed is 50 measurements per second on each channel to a resolution of 1µV and with an accuracy 0f 0.01%. The instrument will create up to 68 mathematical data channels from the original data. Software runs with Windows, dos, Modbus and J-bus. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

Mass storage systemsSolid-state file cards. IBM's cards provide users of portable computers with an alternative to magnetic disks for PCMCIA memory modules. Two forms of card, PCMCIA Type I and II both have a standard PCMCIA-ATA interface and capacity up to 40Mbyte. They use a single 5V supply at less power than disk drives, an advantage over the drives being that there are no delays. Disadvantages of first-generation flash memory cards are avoided by the use of a controller chip and dram buffers to avoid the need to erase memory before storing data. DIP Systems. Tel., 01483 202070; fax, 01483 202023. ■

STOP! Just Take Two Steps to do Your Measurement

TiePieSCOPE



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You can simply plug the new TiePieSCOPE - HS508 into the parallel port of your portable or desktop PC. With the advanced software, you can use this two channel, 8 bits, 50 MHz measuring instrument as a fast digital storage oscilloscope, including a lot more features than a single oscilloscope! Moreover, the TiePieSCOPE - HS508 contains a multiple display voltmeter (up to 5 MHz true RMS), a spectrum analyzer with an harmonic distortion meter and a transient recorder for recording a variety of signals.
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CIRCLE NO. 139 ON REPLY CARD

Chopping on a bridge

Darren Heywood's chopping approach to measuring bridge design results in an unusual combination of low cost and high stability.

I was challenged by a friend to design an high-sensitivity amplifier circuit for a transducer. My choice was to connect the transducer in a Wheatstone bridge configuration.

Output span from the transducer was just 0 to 5mV. This meant that the signal would have to be amplified by at least 1000 in order to bring the signal to workable levels, ie 0-5V.

I started the design by simply setting the resistor ratios R_f/R_i on a *741* op-amp to yield the required gain, **Fig. 1**. But the configuration was unstable and would not null. Furthermore, I noticed that by simply blowing a little air over the circuit, the output would suddenly drift towards either supply rail and saturate.

Consulting the data sheets revealed that the drift gradient for a *741* was in the region of approximately 20µV/°C. A simple calculation exposes the problem. Assume a change of say 5°C referred to the op-amp input. This means a ΔV_{offset} of 100µV (5×20) or 0.1mV. Multiply this figure by 1000 and you get 0.1V at the output due solely to temperature change.

Another contributory factor to drift in the circuit is the type of resistors used. Carbon types for instance have a drift of approximately 300ppm while metal film types exhibit approximately 50ppm. Moreover, when soldering the resistors onto a circuit board, a thermocouple is created due to Seebeck effect and noise levels inherent in the circuit change with temperature.

The obvious solution to the temperature drift problem would seem to be to obtain an op-amp with a very low drift figure. The *OP27* has a drift rate of just 1 to 2µV/°C, depending on the part-number suffix.

Inserting the new op-amp into Fig. 1 reduced the drift problem, but the output still varied to unacceptable levels. I began to realise that a totally different circuit concept was required – namely a chopper amplifier*.

Designing a circuit exhibiting near-zero drift is one challenge, but is it possible to incorporate chopper technology into a 4-to-20mA system? Signal transmission relying on current change is superior to an equivalent based on voltage because current operation minimises line loss. Current loops are widely used in both instrumentation and digital transmission systems, **Fig. 2**.

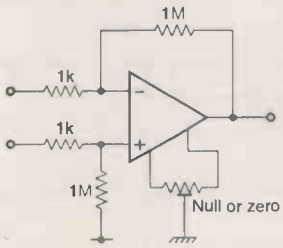


Fig. 1. Designing a transducer amplifier with a gain of 1000 should be easy, given an op-amp with a high input impedance and a gain of a few million.

Implementing the chopping bridge

You should first decide on an overall feedback system. I chose voltage-to-current feedback, **Fig. 3**. Assuming V_{is} drops slightly due to the resistance-temperature transducer increasing, the op-amp responds by increasing its output.

*At £1.33 in 100 offs, the MAX427 op-amp has a drift of 0.8µV/°C. In similar quantities, the ICL7650 chopper amplifier is £2.30 while the MAX420 chopper is £3 – Ed.

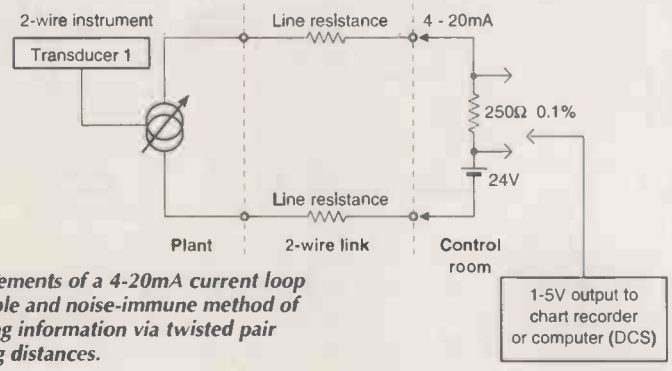


Fig. 2. Elements of a 4-20mA current loop – a reliable and noise-immune method of conveying information via twisted pair over long distances.

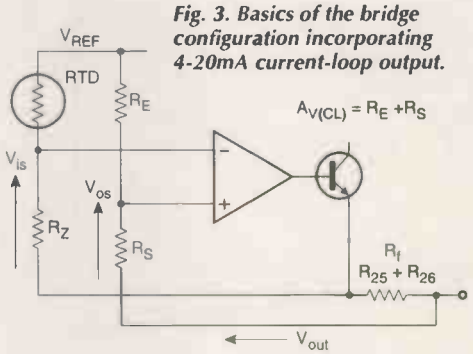


Fig. 3. Basics of the bridge configuration incorporating 4-20mA current-loop output.

In turn, voltage across R_f starts to increase and V_{os} decreases until V_{is} equals V_{os} and equilibrium is reached.

For the op-amp, it is desirable to have high gain and lowest possible drift. This ensures temperature stability and improves resolution. Selecting a high gain 'off-the-shelf' op-amp achieves good resolution, but not temperature stability.

This dilemma forces the use of chopper amplifiers, which normally means added complexity, extra components and increased costs.

Supplying the bridge

To provide a reference, a temperature compensated voltage source is needed with low output impedance and low current consumption. The *LM723* voltage stabiliser, IC_1 of **Fig. 4**, is very cheap, widely available, and contains a 7.2V temperature compensated voltage reference capable of sourcing up to 20mA. In addition, it has a high gain op-amp, a pass transistor capable of sinking 150mA, a current limit transistor and a zener diode – all for approximately 40p.

A 24V supply is needed while the *723* voltage reference is about 7.2V. If the 7.2V reference is used as the op-amp psuedo ground, then IC_2 can swing approximately ±7V. This leaves approximately 10V for external line and measurement resistance. Hence approximately 15V divided into 20mA equals 750Ω and 1200Ω minus 750Ω leaves 450Ω for external resistances.

The feedback system around IC_2 is a hybrid type. You may think that the gain is set with this feedback system,

$$A_v(CL) = \frac{R_6}{R_3} \times \left(1 + \frac{R_7}{R_8}\right)$$

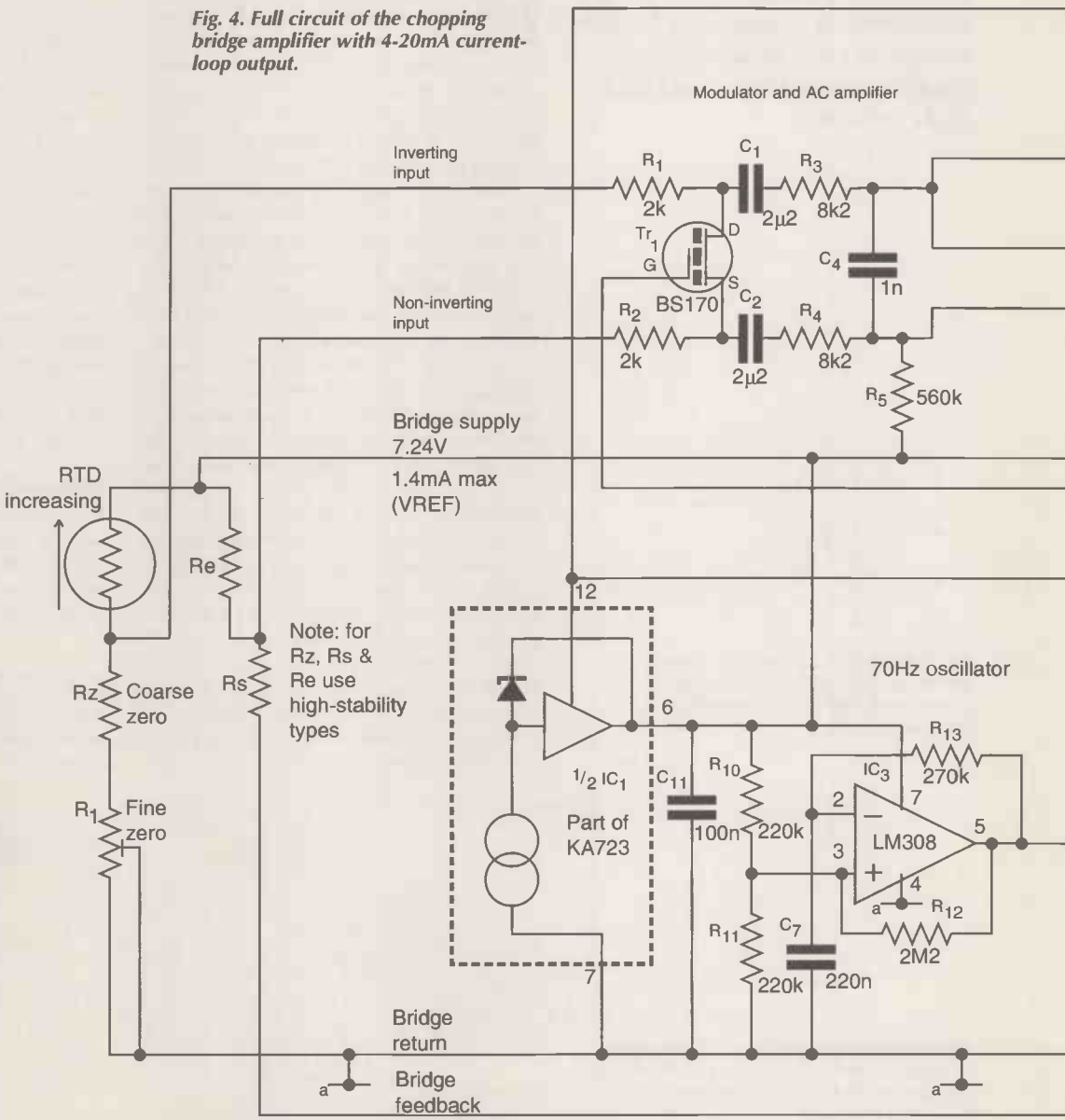
But at the gain demanded from IC_2 the above equation fails. This problem occurs because,

$$A_v(CL) = \frac{A_v(OL)}{(1 + BA_v(OL))}$$

ie $A_v(CL)$ is approximately $A_v(OL)$.

If you check out the gain/frequency response curves as given by the manufacturers, they reveal that in open loop mode, the *LM308* outputs 110dB gain at approximately 10Hz and rolls off at the first order rate of

Fig. 4. Full circuit of the chopping bridge amplifier with 4-20mA current-loop output.



18dB/decade thereafter. Why control gain with just resistors? You can control it with frequency as well. The above then demonstrates that the chopping frequency is most important.

Oscillator design choice

This application needs an oscillator with a low current consumption and that remains at a stable frequency even if the 24V supply is varied from 24V down to say 15V. It must also swing from the supply to ground to ensure IC_2's common mode input range is maximised. It must also have a 180° complement output.

The simplest choice is to use another 308 since it consumes only 300µA. Notice that IC_3 is powered by V_{REF}. This clamps IC_3 to maintain fixed stable frequency. Output of IC_3 is then fed into Tr_3 and Tr_2, the latter being driven by Tr_3. Both drains are connected to the positive supply rail.

At 24V, the two zener diodes limit the common mode range to about 16V to reduce stress on the mosfet gates. Note that bipolar transistors connected in astable mode with 390kΩ load resistors as Tr_3 and Tr_2, take too long to switch off.

One improvement that may possibly be made here is to connect Tr_3 and Tr_2 in bistable mode, using IC_3 as the driver. In this way, Tr_3 and Tr_2 outputs would have ideal overlapping switching times.

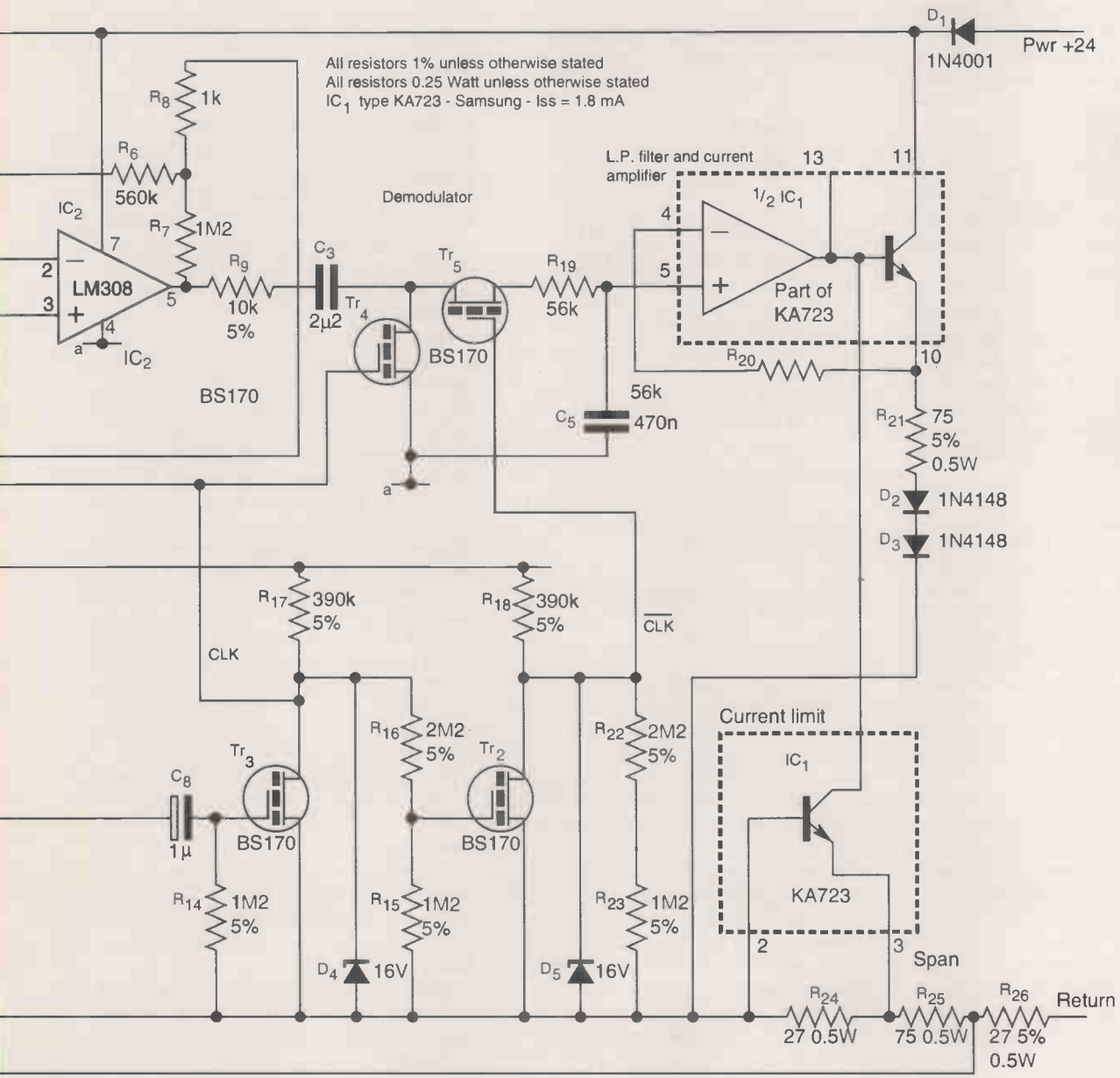
Modulation and demodulation

First, the modulation system used is synchronous. This simplifies the circuitry and maintains excellent restoration of the amplified signal.

Assume Tr_1 and Tr_4 are both off, Tr_5 is on, and there is slightly less potential at the inverting input than the non-inverting input if IC_2. This means that C_2 will have a slightly greater charge stored than C_1.

Now, Tr_1 and Tr_4 are both on, Tr_5 is off, C_1 and C_2 are both rapidly shunted together and because C_1 has slightly less charge than C_2. A small difference charge is forced into IC_2 inverting terminal. This causes IC_2's output to swing negative and equalises at some point via the feedback resistors. At the same time Tr_4 shunts C_3 to ground which negatively charges C_3 from V_{REF} point of view.

At this point, Tr_1 and Tr_4 are both off, Tr_s is on and C_1 now has a negative difference



charge and as such, current is pulled from IC_2 inverting terminal. This causes IC_2's output to swing positive until equilibrium is again reached via feedback.

Both positive and negative output pulses are equal in magnitude but opposite in polarity. During the positive pulse, Tr_5 is about 5Ω and thus the positive pulses from IC_2 are sampled and stored in C_5.

Due to the previous negative cycle, C_3 was charged from ground and thus positive only amplified pulses which are referenced to ground are passed onto or into R_{19}/C_5. By charging C_3 from ground, level shift from V_{REF} to ground is accomplished. Remember that IC_2 output swings around its psuedo ground V_{REF}.

Notice that C_1, C_2 and C_3 isolate IC_2's quiescent point so IC_2 is allowed to drift. Also, increasing the dc signal on the inverting terminal of IC_2 to above that of the non-inverting terminal causes a phase change at the output of IC_2. This produces dsb suppressed carrier modulation!

Current amplifier

To produce the current amplifier, IC_1 is simply connected as a unity gain voltage buffer. Current gain, however, is determined by the current flowing through R_{21}. The smaller R_{21}, the higher the current gain.

Note the internal pass transistor within the *723* is providing the current gain and not the amp. The amp simply controls the current very accurately. Diodes D_2 and D_3 lift the turn on level to 1.2V. This is done because the *KA723* op-amp does not saturate at exactly ground. All the above means 4mA, or zero, begins at around 1.2V and ends at around 2.4V, i.e. 20mA.

The bridge system

Referring to Fig. 3, assume for a range of 0°-100°C, the rtd's resistance changes from 100Ω to 139.02Ω. Also assume that 100Ω represents 4mA and 139.02Ω represents 20mA.

In my bridge configuration, an increase in rtd resistance causes V_{is} to fall. Due to feedback, the amplifier increases current output across R_{27} and R_{23} until the selected feedback resistances R_S/R_E equalise the change. Thus V_{is} is always approximately equal V_{os} and is true for any feedback system.

The higher the open-loop gain the less the error between V_{is} and V_{os}. Again assume that the rtd is 100Ω, V_{is} equals V_{os} and the system draws 4mA. Now, the rtd begins to increase in value so voltage V_{is} starts to fall. Voltage V_{os} follows V_{is} because the system is closed loop, **Fig. 5**.

If the rtd carries on increasing then at some point the system will reach 20mA. In theory, any zero/span ratio can be achieved. Here are the equations governing the system calibration under static conditions are,

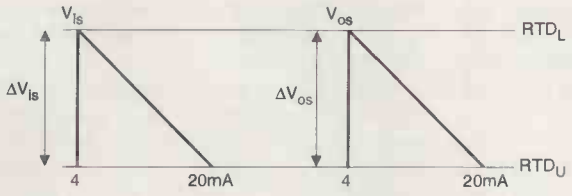


Fig. 5. Ideally, the voltage output slope of the 4-20mA converter should follow the voltage input slope in the configuration of Fig. 3.

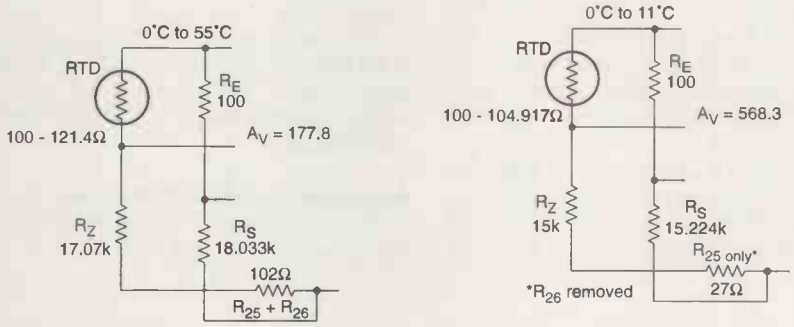


Fig. 6. Examples of bridge connections for two resistance-temperature transducers with different characteristics.

$$\Delta V_{os} = \frac{1.632 R_E}{R_S + R_E}$$

$$\Delta V_{is} = \frac{V_{REF} R_Z (RTD_U - RTD_L)}{(RTD_L + R_Z)(RTD_U + R_Z)}$$

$$V_{os} = \frac{R_S V_{REF} - 0.408 R_E}{R_S + R_E}$$

$$V_{is} = \frac{V_{REF} R_Z}{RTD_L + R_Z}$$

and the limitation equations are,

$$\Delta V_O = \frac{1.632 (V_{REF} - V_{OS})}{V_{REF} + 0.408}$$

$$\frac{R_S}{R_E} = \frac{1.632 V_{OS} + 0.408 \Delta V_O}{\Delta V_O V_{REF}}$$

Note that system span is controlled by R_{25} plus R_{26}, R_S and R_E are span alignment resistors only and zero is controlled with R_Z. For any given calibration, ΔV_{os} must equal ΔV_{is} and V_{ostart} must equal V_{istart}. Also, ΔV_{is} must not exceed approximately 9mV. This is due to the maximum current that can be drawn by the bridge.

For any given zero/span range, ΔV_{is} should always be as large as possible – why attenuate then amplify? Reducing resistor R_{23} narrows the span, however the equations supplied have to be amended slightly. I have provided two calibration scenarios. Bridge **Fig. 6a)** is 0°C=4mA to 55°C=20mA, while the bridge illustrated in **Fig. 6b)** is 0°C=4mA to 11°C=20mA.

Dynamic loop performance

Unfortunately, I did not have the equipment needed to maximise speed via damping. However, you must remember that we are trying to amplify thermocouples and rtds which have an inherently slow response speed of approximately 10 to 15 seconds. So if the system is slightly overdamped, performance is not downgraded.

The system loop's dynamics and bandwidth are set via R_{19} and C_5. I chose these values to coincide with a –3dB of 7Hz. This is ten times less than the switching frequency. This is well within the criteria of the sampling theorem.

At very narrow spans IC_2 has to produce higher gains and as such becomes too slow to respond to the induced error caused by *KA723* pin 5. Thus no overshoot occurs at narrow span demands. Switching frequency was selected upon the above criteria.

The loop is guaranteed to be conditionally stable. The only unstable condition that can occur is if the input signal approaches 70Hz and is in phase with the switching (chopping) frequency. This is highly unlikely to happen.

Capacitor C_4 was inserted between the inputs of IC_2 to limit overshoot, slowing IC_2 down slightly during wide span conditions.

Diode D_1 protects against reverse polarity supply connection.

Summary

Components for the bridge amplifier are well under £5 yet open-loop gain is in excess of 48000 and temperature stability is excellent. Noise is also low since the circuit is narrowband.

I have shown here what can be achieved with an alternative bridge topology and that high performance need not mean expensive components. ■

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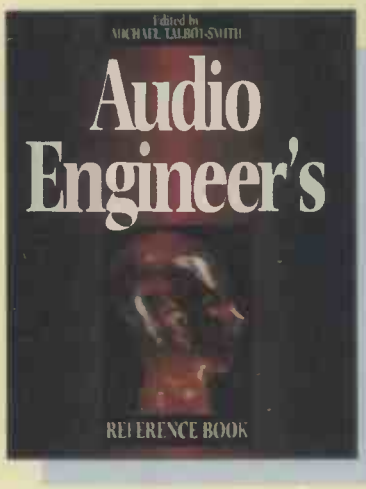

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4.6mw LASER POINTER. BRAND NEW MODEL NOW IN STOCK!, supplied in fully built form (looks like a nice pen) complete with handy pocket clip (which also acts as the on/off switch.) About 50 metres range! Runs on 2 AAA batteries. Produces thin red beam ideal for levels, gun sights, experiments etc. just £39.95 ref DEC49 TRADE PRICE £28 MIN 10 PIECES

BULL TENS UNIT Fully built and tested TENS (Transcutaneous Electrical Nerve Stimulation) unit, complete with electrodes and full instructions. TENS is used for the relief of pain etc in up to 70% of sufferers. Drug free pain relief, safe and easy to use. can be used in conjunction with analgesics etc. £49 Ref TEN/1

RUSSIAN MONOCULARS Amazing 20 times magnification, coated lenses, carrying case and shoulder strap £29.95 REF BAR73

PC PAL VGA TO TV CONVERTER Converts a colour TV into a basic VGA screen. Complete with built in psu, lead and s/ware.. Ideal for laptops or a cheap upgrade. Supplied in kit form for home assembly. SALE PRICE £25 REF SA34

EMERGENCY LIGHTING UNIT Complete unit with 2 double bulb floodlights, built in charger and auto switch. Fully cased. 6v 8AH lead acid req'd. (secondhand) £4 ref MAG4P11.

YUASHA SEALED LEAD ACID BATTERIES Two sizes currently available this month. 12v 15AH at £18 ref LOT8 and 6v 10AH (suitable for emergency lights above) at just £6 ref LOT7.

ELECTRIC CAR WINDOW DE-ICERS Complete with cable, plug etc SALE PRICE JUST £4.99 REF SA28

AUTO SUNCHARGER 155x300mm solar panel with diode and 3 metre lead fitted with a cigar plug. 12v 2watt. £8.99 REF SA25.

ECLATRON FLASH TUBE As used in police car flashing lights etc, full spec supplied, 60-100 flashes a min. £6.99 REF SA15B.

*SOME OF OUR PRODUCTS MAY BE UNLICENSABLE IN THE UK

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24v AC 96WATT Cased power supply. New. £9.99 REF SA40

MICRODRIVE STRIPPERS Small cased tape drives ideal for stripping, lots of useful goodies including a smart case, and lots of components. SALE PRICE JUST £4.99 FOR FIVE REF SA26

SOLAR POWER LAB SPECIAL You get TWO 6"x6" 6v 130mA solar cells, 4 LED's, wire, buzzer, switch plus 1 relay or motor. Superb value kit SALE PRICE JUST £4.99 REF SA27

RGB/CGA/EGA/TTL COLOUR MONITORS 12" In good condition. Back anodised metal case. SALE PRICE £49 REF SA16B

PLUG IN ACORN PSU 19v AC 14w , £2.99 REF MAG3P10

13.8V 1.9A PSU cased with leads. Just £9.99 REF MAG10P3

UNIVERSAL SPEED CONTROLLER KIT Designed by us for the C5 motor but ok for any 12v motor up to 30A. Complete with PCB etc. A heat sink may be required. £17.00 REF: MAG17

PHONE CABLE AND COMPUTER COMMUNICATIONS PACK Kit contains 100m of 6 core cable, 100 cable clips, 2 line drivers with RS232 interfaces and all connectors etc. Ideal low cost method of communicating between PC's over a long distance utilizing the serial ports. Complete kit £8.99. Ref comp1.

VIEWDATA SYSTEMS made by Phillips, complete with Internal 1200/75 modem, keyboard, psu etc RGB and composite outputs, menu driven, autodialler etc. SALE PRICE £12.99 REF SA18

AIR RIFLES .22 As used by the Chinese army for training puposes, so there is a lot about! £39.95 Ref EF78. 500 pellets £4.50 ref EF80.

PLUG IN POWER SUPPLY SALE FROM £1.60 Plugs in to 13A socket with output lead. three types available, 9vdc 150mA £1.50 ref SA19, 9vdc 200mA £2.00 ref SA20, 6.5vdc 500mA £2 ref SA21.

VIDEO SENDER UNIT. Transmits both audio and video signals from either a video camera, video recorder, TV or Computer etc to any standard TV set in a 100' range! (tune TV to a spare channel) 12v DC op. Price is £15 REF: MAG15 12v psu is £5 extra REF: MAG5P2

***MINATURE RADIO TRANSCEIVERS** A pair of walkie talkies with a range up to 2 km in open country. Units measure 22x52x155mm. Including cases and earp'ces. 2xPP3 req'd, £30.00 pr. REF: MAG30

***FM TRANSMITTER KIT** housed in a standard working 13A adapter!! the bug runs directly off the mains so lasts forever! why pay £700? or price is £15 REF: EF62 (kit) Transmits to any FM radio.

***FM BUG BUILT AND TESTED** superior design to kit. Supplied to detective agencies. 9v battery req'd. £14 REF: MAG14

TALKING COINBOX STRIPPER COMPLETE WITH COINSLOT MECHANISMS originally made to retail at £79 each, these units are designed to convert an ordinary phone into a payphone. The units have the locks missing and sometimes broken hinges. However they can be adapted for their orginal use or used for something else?? SALE PRICE JUST £2.50 REF SA23

GAT AIR PISTOL PACK Complete with pistol, darts and pellets £12.95 Ref EF82B extra pellets (500) £4.50 ref EF80.

6"X12" AMORPHOUS SOLAR PANEL 12v 155x310mm 130mA. SALE PRICE £4.99 REF SA24.

FIBRE OPTIC CABLE BUMPER PACK 10 metres for £4.99 ref MAG5P13 ideal for experimenters! 30 m for £12.99 ref MAG13P1

MIXED GOODIES BOX OF MIXED COMPONENTS WEIGHING 2 KILOS YOURS FOR JUST £6.99

4X28 TELESCOPIC SIGHTS Suitable for all air rifles, ground lenses, good light gathering properties. £19.95 ref R/7.

RATTLE BACKS Interesting things these, small piece of solid perspex like material that if you try to spin it on the desk it only spins one way! In fact if you spin it the 'wrong' way it stops of its own accord and go's back the other way! £1.99 ref GI/J01.

GYROSCOPES Remember these? well we have found a company that still manufactures these popular scientific toys, perfect gift or for educational use etc. £6 ref EP70

HYPOTHERMIA SPACE BLANKET 215x150cm aluminised foil blanket, reflects more than 90% of body heat. Also suitable for the construction of two way mirrors! £3.99 each ref O/L041.

LENSTATIC RANGER COMPASS Oil filled capsule, strong metal case, large luminous points. Sight line with magnifying viewer. 50mm dia, 86gm. £10.99 ref O/K604.

RECHARGE ORDINARY BATTERIES UP TO 10 TIMES! With the Battery Wizard! Uses the latest pulse wave charge system to charge all popular brands of ordinary batteries AAA, AA, C, D, four at a time! Led system shows when batteries are charged, automatically rejects unsuitable cells. complete with mains adaptor. BS approved. Price is £21.95 ref EP31.

TALKING WATCH Yes, it actually tells you the time at the press of a button. Also features a voice alarm that wakes you up and tells you what the time is! Lithium cell included. £7.99 ref EP26.

PHOTOGRAPHIC RADAR TRAPS CAN COST YOU YOUR LICENCE! The new multiband 2000 radar detector can prevent even the most responsible of drivers from losing their licence! Adjustable audible alarm with 8 flashing leds gives instant warning of radar zones. Detects X, K, and Ka bands, 3 mile range, 'over the hill' 'around bends' and 'rear trap facilities. micro size just 4.25"x2.5"x.75". Can pay for itself in just one day! £79.95 ref EP3.

SANYO NICAD PACKS 120mmx14mm 4.8v 270 maH suitable for cordless phones etc. Pack of 2 just £5 ref EP78.

3" DISCS As used on older Amstrad machines, Spectrum plus3's etc £3 each ref BAR400.

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COINSLOT TOKENS You may have a use for these? mixed bag of 100 tokens £10 ref LOT20.

PORTABLE X RAY MACHINE PLANS Easy to construct plans on a simple and cheap way to build a home X-ray machine! Effective device, X-ray sealed assemblies. can be used for experimental purposes. Not a toy or for minors! £6/set. Ref F/XP1.

TELEKINETIC ENHANCER PLANS Mystify and amaze your friends by creating motion with no known apparent means or cause. Uses no electrical or mechanical connections, no special gimmicks yet produces positive motion and effect. Excellent for science projects, magic shows, party demonstrations or serious research & development of this strange and amazing phychic phenomenon. £4/set Ref F/TKE1.

ELECTRONIC HYPNOSIS PLANS & DATA This data shows several ways to put subjects under your control. Included is a full volume reference text and several construction plans that when assembled can produce highly effective stimuli. This material must be used cautiously. It is for use as entertainment at parties etc only, by those experienced in its use. £15/set. Ref F/EH2.

GRAVITY GENERATOR PLANS This unique plan demonstrates a simple electrical phenomena that produces an anti-gravity effect. You can actually build a small mock spaceship out of simple materials and without any visible means- cause it to levitate. £10/set Ref F/GRA1.

WORLDS SMALLEST TESLA COIL/LIGHTENING DISPLAY GLOBE PLANS Produces up to 750,000 volts of discharge, expenment with extraordinary HV effects, 'Plasma in a jar', St Elmo's fire, Corona, excellent science project or conversation piece. £5/set Ref F/BTC1/LG5.

COPPER VAPOUR LASER PLANS Produces 100mw of visible green light. High coherency and spectral quality similar to Argon laser but easier and less costly to build yet far more efficient. This particular design was developed at the Atomic Energy Commision of NEGEV in Israel. £10/set Ref F/CVL1.

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PULSED TV JOKER PLANS Little hand held device utilises pulse techniques that will completely disrupt TV picture and sound! works on FM too! DISCRETION ADVISED. £8/set Ref F/TJ5.

BODYHEAT TELESCOPE PLANS Highly directional long range device uses recent technology to detect the presence of living bodies, warm and hot spots, heat leaks etc. Intended for security, law enforcement, research and development, etc. Excellent security device or very interesting science project. £8/set Ref F/BHT1.

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ULTRASONIC BLASTER PLANS Laboratory source of sonic shock waves. Blow holes in metal, produce 'cold' steam, atomize liquides. Many cleaning uses for PC boards, jewllery, coins, small parts etc. £6/set Ref F/ULB1.

ULTRA HIGH GAIN AMP/STETHOSCOPIC MIKE/SOUND AND VIBRATION DETECTOR PLANS Ultrasensitive device enables one to hear a whole new world of sounds. Listen through walls, windows, floors etc. Many applications shown, from law enforcement, nature listening, medical heartbeat, to mechanical devices. £6/set Ref F/HGA7

ANTI DOG FORCE FIELD PLANS Highly effective circuit produces time variable pulses of accoustical energy that dogs cannot tolerate £6/set Ref F/DOG2

LASER BOUNCE LISTENER SYSTEM PLANS Allows you to hear sounds from a premises without gaining access. £12/set Ref F/LLIST1

LASER LIGHT SHOW PLANS Do it yourself plans show three methods. £6 Ref F/LLS1

PHASOR BLAST WAVE PISTOL SERIES PLANS Handheld, has large transducer and battery capacity with external controls. £6/set Ref F/PSP4

INFINITY TRANSMITTER PLANS Telephone line grabber/ room monitor. The ultimate in home/office security and safety! simple to use! Call your home or office phone, push a secret tone on your telephone to access either A) On premises sound and voices or B) Existing conversation with break-in capability for emergency messages. £7 Ref F/TELEGRAB.

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ELECTROMAGNETIC GUN PLANS Projects a metal object a considerable distance-requires adult supervision £5 ref F/EML2.

ELECTRIC MAN PLANS, SHOCK PEOPLE WITH THE TOUCH OF YOUR HAND! £5/set Ref F/EMA1.

PARABOLIC DISH MICROPHONE PLANS Listen to distant sounds and voices, open windows, sound sources in 'hard to get' or hostile premises. Uses satellite technology to gather distant sounds and focus them to our ultra sensitive electronics. Plans also show an optional wireless link system. £8/set ref F/PM5

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SWITCHED MODE PSU'S 244 watt, +5 32A, +12 6A, -5 0.2A, -12 0.2A. There is also an optional 3.3v 25A rail available. 120/240v I/P. Cased, 175x90x145mm. IEC inlet Suitable for PC use (6 d/drive connectors 1 m/board). £10 ref PSU1.

VIDEO PROCESSOR UNITS?/6v 10AH BATTS/12V 8A TX Not too sure what the function of these units is but they certainly make good strippers! Measures 390X320X120mm, on the front are controls for scan speed, scan delay, scan mode, loads of connections on the rear. Inside 2 x 6v 10AH sealed lead acid batts, pcb's and a 8A? 12v torroidial transformer (mains in). Condition not known, may have one or two broken knobs due to poor storage. £17.50 ref VP2

RETRON NIGHT SIGHT Recognition of a standing man at 300m in 1/4 moonlight, hermatically sealed, runs on 2 AA batteries, 80mm F1.5 lens, 20mw infrared laser included. £325 ref RETRON.

MINI FM TRANSMITTER KIT Very high gain preamp, supplied complete with FET electret microphone. Designed to cover 88-108 Mhz but easily changed to cover 63-130 Mhz. Works with a common 9v (PP3) battery. 0.2W RF. £7 Ref 1001.

3-30V POWER SUPPLY KIT Variable, stabilized power supply for lab use. Short circuit protected, suitable for profesional or amateur use 24v 3A transformer is needed to complete the kit. £14 Ref 1007.

1 WATT FM TRANSMITTER KIT Supplied with piezo electric mic. 8-30vdc. At 25-30v you will get nearly 2 watts! £12 ref 1009.

FM/AM SCANNER KIT Well not quite, you have to turn the knob your self but you will hear things on this radio that you would not hear on an ordinary radio (even TV). Covers 50-160mhz on both AM and FM. Built in 5 watt amplifier, inc speaker. £15 ref 1013.

3 CHANNEL SOUND TO LIGHT KIT Wireless system, mains operated, separate sensitivity adjustment for each channel, 1,200 w power handling, microphone included. £14 Ref 1014.

4 WATT FM TRANSMITTER KIT Small but powerful FM transmitter, 3 RF stages, microphone and audio preamp included. £20 Ref 1028.

STROBE LIGHT KIT Adjustable from 1-60 hz (a lot faster than conventional strobes). Mains operated. £16 Ref 1037.

LIQUID LEVEL DETECTOR KIT Useful for tanks, ponds, baths, rain alarm, leak detector etc. Will switch 2A mains. £5 Ref 1081.

COMBINATION LOCK KIT 9 key, programmable, complete with keypad, will switch 2A mains. 9v dc operation. £10 ref 1114.

PHONE BUG DETECTOR KIT This device will warn you if somebody is eavesdropping on your line. £6 ref 1130.

ROBOT VOICE KIT Interesting circuit that distorts your voice! adjustable, answer the phone with a different voice! 12vdc £9 ref 1131

TELEPHONE BUG KIT Small bug powered by the 'phone line, starts transmitting as soon as the phone is picked up! £8 Ref 1135.

3 CHANNEL LIGHT CHASER KIT 800 watts per channel, speed and direction controls supplied with 12 LEDS (you can fit triacs instead to make kit mains, not supplied) 9-12vdc £17 ref 1026.

12V FLOURESCENT LAMP DRIVER KIT Light up 4 foot tubes from your car battery! 9v 2a transformer also required. £8 ref 1069.

VOX SWITCH KIT Sound activated switch ideal for making bugging tape recorders etc, adjustable sensitivity. £8 ref 1073.

Check out our WEB SITE

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PREAMP MIXER KIT 3 input mono mixer, sep bass and treble controls plus individual level controls, 18vdc, input sens 100mA. £15 ref 1052.

METAL DETECTOR KIT Range 15-20cm, complete with case, 9vdc. £8 ref 1022.

SOUND EFFECTS GENERATOR KIT Produces sounds ranging from bird chips to sirens. Complete with speaker, add sound effects to your projects for just £9 ref 1045.

15 WATT FM TRANSMITTER (BUILT) 4 stage high power, preamp required 12-18vdc, can use ground plane, yagi or open dipole. £69 ref 1021.

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PC TMER KIT Four channel output controlled by your PC, will switch high current mains with relays (supplied). Software supplied so you can program the channels to do what you want whenever you want. Minimum system configeration is 286, VGA, 4.1,640k, serial port, hard drive with min 100k free. £24.99

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HUGE BUBBLE MAKING KIT You'll be amazed at the the size of the bubbles you can acheive with this bubble making kit. Once you have got the knack it is possible to make bubbles of up to 40 feet long. £11.99 ref E/9.

FM CORDLESS MICROPHONE This unit is an FM broadcasting station in minature, 3 transistor transmitter with electret condenser mic+fet amp design result in maximum sensitivity and broad frequency response. 90-105mhz, 50-1500hz, 500 foot range in open country! PP3 battery required. £15.00 ref 15P42A.

MAGNETIC MARBLES They have been around for a number of years but still give rise to curiosity and amazement. A pack of 12 is just £3.99 ref GI/R20

STETHOSCOPES A fully functioning stethoscope for all those intricate projects. Enables you to listen to motors, pipes, heartbeats, walls, insects etc. £6 ref MAR6P6.

NICKEL PLATING KIT Proffesional electroplating kit that will transform rusting parts into showpieces in 3 hours! Will plate onto steel, iron, bronze, gunmetal, copper, welded, silver soldered or brazed joints. Kit includes enough to plate 1,000 sq inches. You will also need a 12v supply, a container and 2 12v light bulbs. £39.99 ref NIK39.

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HY24060m, 240v adjustable from 0-60 mins. £6.99

BUGGING TAPE RECORDER Small voice activated recorder, uses micro cassette complete with headphones. £28.99 ref MAR29P1.

POWER SUPPLY fully cased with mains and o/p leads 17v DC 900mA output. Bargain price £5.99 ref MAG6P9

9v DC POWER SUPPLY Standard plug in type 150ma 9v DC with lead and DC power plug. price for two is £2.99 ref AUG3P4.

COMPOSITE VIDEO KIT. Converts composite video into separate H sync, V sync, and video. 12v DC. £8.00 REF: MAG8P2.

FUTURE PC POWER SUPPLIES These are 295x135x60mm, 4 drive connectors 1 mother board connector. 150watt, 12v fan, iec inlet and on/off switch. £12 Ref EF6.

VENUS FLYTRAP KIT Grow your own carnivorous plant with this simple kit £3 ref EF34.

6"X12" AMORPHOUS SOLAR PANEL 12v 155x310mm 130mA. Bargain price just £5.99 ea REF MAG6P12.

FIBRE OPTIC CABLE BUMPER PACK 10 metres for £4.99 ref MAG5P13 ideal for experimenters! 30 m for £12.99 ref MAG13P1

ROCK LIGHTS Unusual things these, two pieces of rock that glow when rubbed together! belived to cause rain! £3 a pair Ref EF29.

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9-0-9V 4A TRANSFORMERS, chassis mount. £7 ref LOT19A.

2.5 KILOWATT INVERTERS, Packed with batteries etc but as they weigh about 100kg CALLERS ONLY! £120.

MEGA LED DISPLAYS Build your self a clock or something with these mega 7 seg displays 55mm high, 38mm wide. 5 on a pcb for just £4.99 ref LOT16 or a bumper pack of 50 displays for just £29 ref LOT17.

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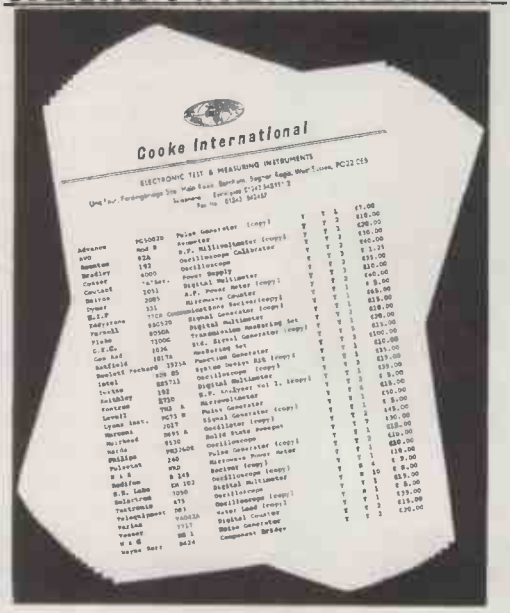

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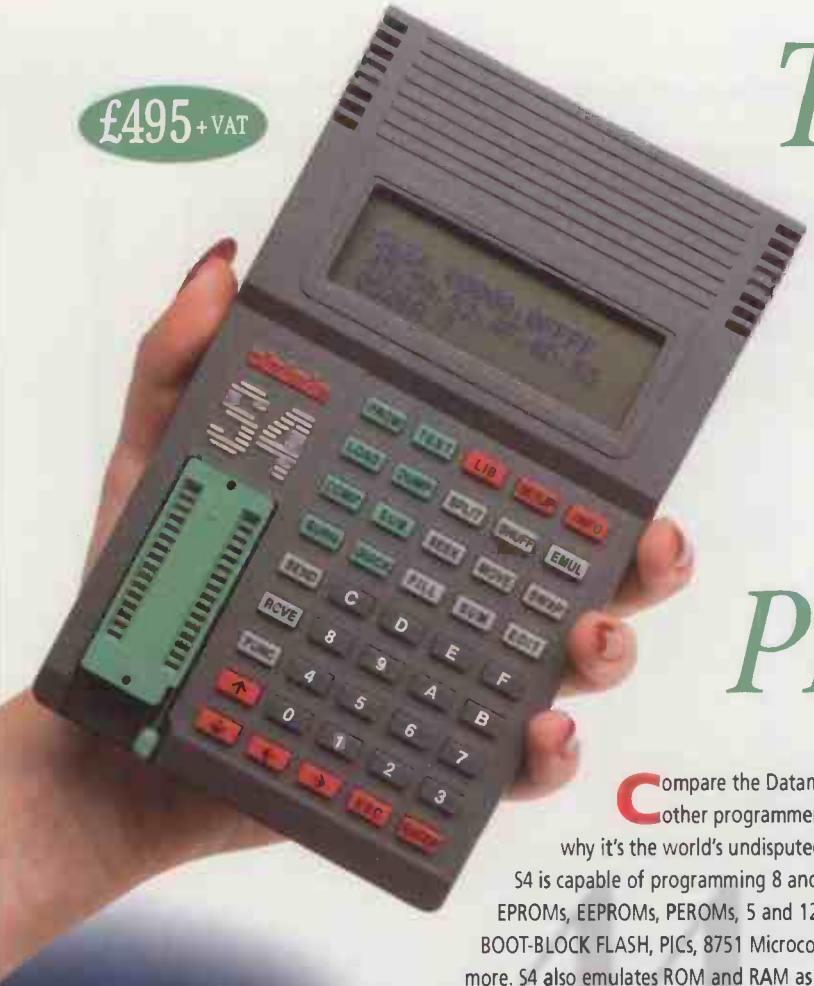

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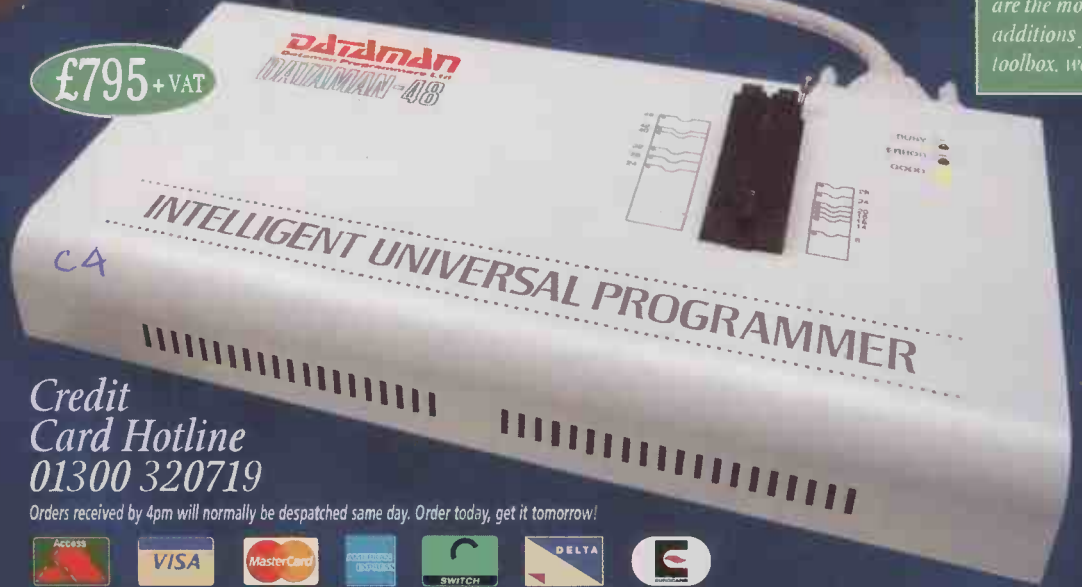

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