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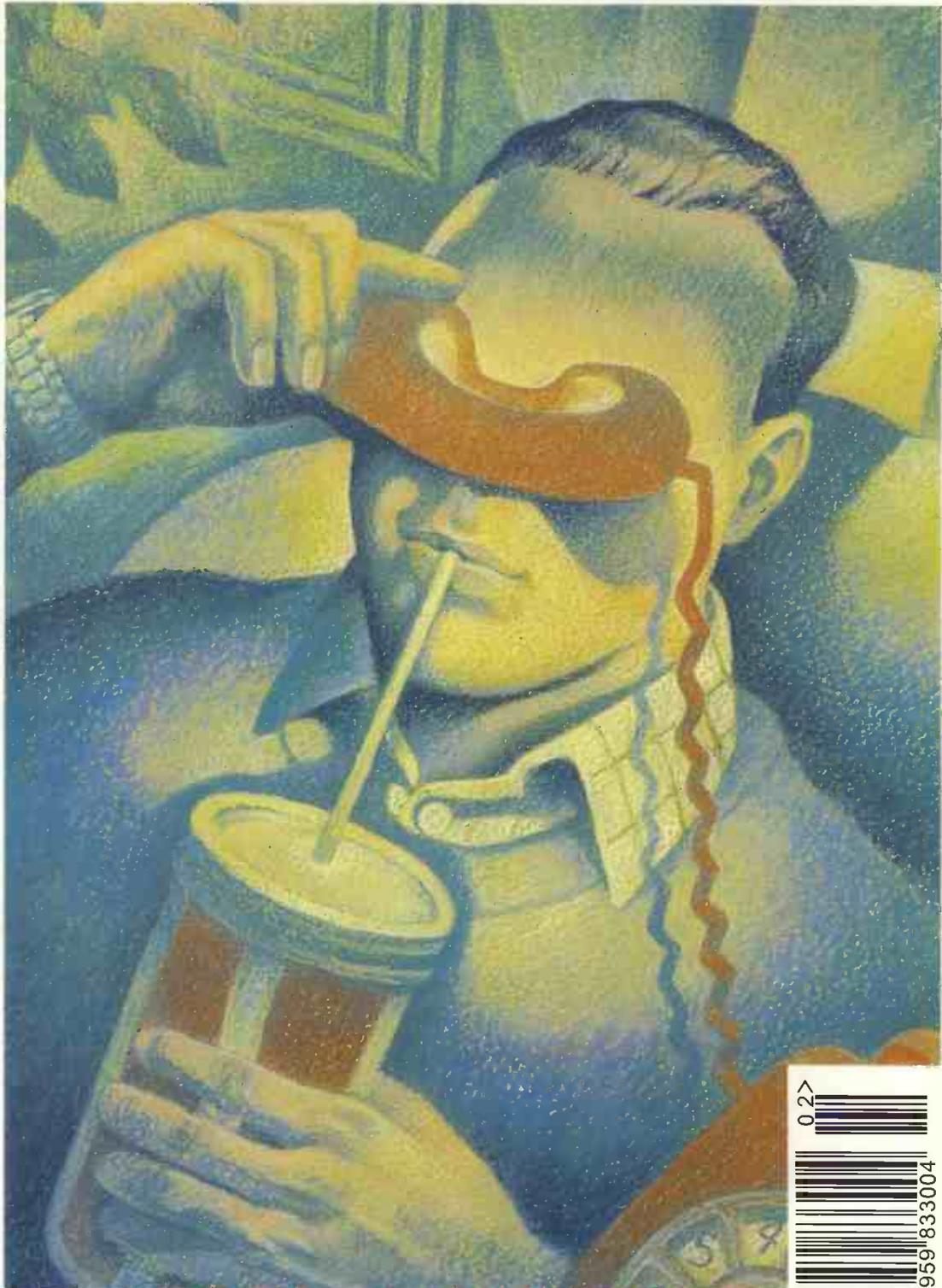
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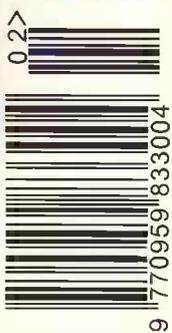
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Cover: Hashim Akib



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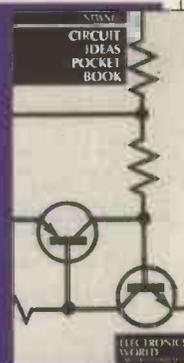
### In next month's issue:

Free Circuit Ideas Pocket Book – a selection of circuit ideas from *EW+WW* brought together in one handy volume.

- Doug Self discusses building power amplifiers for supply immunity.
- RS232 communication is faster using psk, as Lanka Kumar *et al* show.
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Fill in the newagents' order form between pages 144 and 145.

MARCH ISSUE – ON SALE February 23



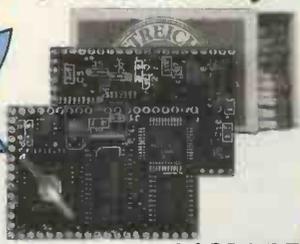
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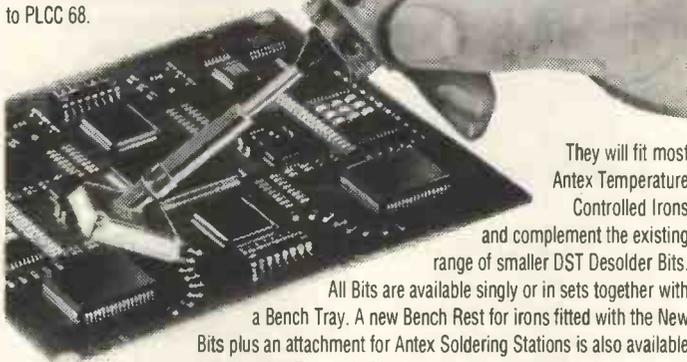
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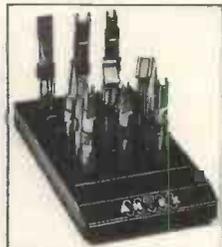
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**B**T's Head of Research, Dr Alan Rudge, – formal title MD Development and Procurement – stands out like a sore thumb amongst BT's other bosses. He talks with the frankness and honesty of an engineer, not the venerated smarm of an accountant or professional manager.

In early December, BT celebrated ten years of privatisation. At one event, 'A glimpse of the future', Rudge described his job. Because he cannot personally oversee all research projects at Martlesham, he gets most closely involved in the ones that go wrong. "It's like being at the end of a sewer pipe", said Rudge, before demonstrating a few of the projects that are going well.

One is 'remote foetal scanning'. Guests at the top of the BT Tower watched high-quality moving ultrasound images transmitted in real time by telephone wire from a hospital on the Isle of Wight. The pictures of a 15 week-old foetus came down the wire in a two-megabit stream.

Earlier Rudge had talked of BT's plans for a Video On Demand trial this summer (1995), using 2Mbit ADSL and covering 2500 homes in the Ipswich area. "It is a market trial," stressed Rudge, "not an experiment". Presumably then, the foetal images were streaming from the Isle of Wight thanks to VOD technology?

No. BT had ganged together 30 ISDN channels. Why? Because even though the VOD market trial was still only six months away, BT could not muster a single pair of ADSL codec/modems and found it easier to sledgehammer the problem with 30 ISDN lines.

Barely able to disguise his frustration at the Government's refusal to promise BT the right to earn money from piping entertainment through its network after 2001, Rudge became even more frank about the practical limitations of VOD.

"ADSL is not economic for 30 million customers. The cable companies are digging up the roads and putting in copper. That is the technology we are trying to get away from.

We have £15 billion, the price of two Channel Tunnels, earmarked for an optic fibre highway. We don't want to invest too much in an interim solution. But we need to know where we stand after 2001 – and how we can earn revenue from our highway".

The consumer electronics industry is already using the same MPEG compression technology that makes ADSL VOD possible, to release movies of better-than VHS quality on pairs of one hour five inch CDs. The recently announced Philips/Sony High Density Digital Video Disc puts a full length movie onto a single five inch CD, with broadcast quality.

At pressing costs of under 50p per disc, Video CD provides a very cheap way of distributing movie material by retail sale, mail order or corner store rental. To compete, VOD must offer similar quality, on the same or more modern movies, at similar price points.

Prototype VOD codec/modems currently cost around \$10,000. At the recent CD-i conference in Dusseldorf, panelists from Philips and Motorola were arguing over whether consumer box prices could fall to below \$750 by the end of the century.

Rudge is right. However exciting VOD and the enabling technology may be, and however excited BT's publicity machine may get, using old-fashioned twisted pairs to carry VHS quality pictures is an interim solution which sucks investment from the long-term strategy of replacing copper with fibre.

BT feels confident that the Labour leaders-in-waiting have grasped this point. The best leaving present the current Government could give Britain before it collapses into a lost election would be to move fast with a clear challenge to BT and the Labour party – 'OK, BT; if you put your money where your mouth is, and pledge to 'fibre' the country by the end of the century, we pledge that you can use that fibre in the next century, in any way you like. And we will put it to a Commons vote so that Labour can't reverse that policy.'

Barry Fox

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## Aware proposes ADSL improvement

M-band wavelet transforms are being used by Aware, the Massachusetts-based signal processing software developer, to provide radical improvements to the performance of the adsl modulation technique used by BT in its video-on-demand service.

Adsl, or asymmetric digital subscriber loop signalling, is a technique which enables a broadband signal, theoretically up to 6Mbit/s, to be transmitted over copper telephone lines from the exchange to the subscriber's home. BT in the UK and Bell Atlantic in the US are using adsl systems supplied by Westell International to transmit a video channel carrying video on demand and home-shopping services to subscribers.

Current adsl systems use carrierless amplitude and phase (cap) modulation which effectively divides the 2.048Mbit/s channel data rate between two carriers of different phase and amplitude. BT believes that cap will support a

2Mbit/s: video channel over 6km of copper pair which represents 91 per cent of its subscriber lines.

A multi-carrier version of adsl, although more expensive than cap-based systems, is able to support data rates up to 6.9Mbit/s and is more immune to line interference.

Aware, which along with fellow US software developer Amarti, as well as Analog Devices and Westell, has been developing the technique known as discrete multi-tone (dmt), is proposing a new approach based on wavelet rather than Fourier signal processing techniques. Drawbacks with dmt have led the US developer to investigate the use of the alternative discrete wavelet multitone (dwtm) modulation technique.

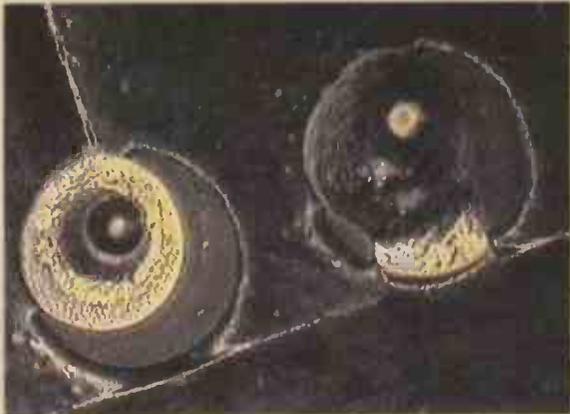
The attraction of dwtm, using an M-band wavelet transform, is that it provides both a longer symbol duration and lower sidelobes than the fft-based dmt. The wavelet methodology provides a more practical implementation of

classical Fourier transforms.

Aware claims that use of M-band wavelet transforms instead of fft in the classic discrete multi-tone technique produces a 4dB improvement in signal margin in a practical adsl multi-carrier transceiver. According to Aware, the computational complexity of the wavelet transform is around one half that of the Fourier transform.

Aware quantifies performance improvement in terms of the interference between sub-channels and symbols for dmt and dwtm systems of similar complexity. Average improvement is around 6dB in signal sensitivity and Aware quotes a minimum improvement figure of 4dB.

Aware also says that the use of Wavelet instead of Fourier transforms does not threaten compatibility with the other elements of the adsl transceiver design, as proposed by Amarti which is working with Motorola. Richard Wilson, *Electronics Weekly*



Cambridge Technology Systems claims that deep field video microscopy, a system it markets in the UK for Scientific Instruments, can be used to reveal micro-defects in solder adhesion and integrity – particularly at the contact points between miniature components and pcbs. The right hand side of the photo shows a sound micro-pin soldered connection; on the left is a poor one which, due to lack of solder volume, shows early signs of failure. (The solder appears black in the photo.)

This photo was taken at 80x magnification. The magnification range can be extended to 1,000x if necessary, to resolve fine surface detail.

## Double science increase not followed through

Although there are now more young people in the UK studying science to the age of 16, fewer of them are carrying on to take science A levels. This is one of the main findings of a report, 'The Impact of Double Science' published by the Engineering Council and written by two authors from the University of Manchester.

There has been a marked growth in entries for science GCSE since compulsory science to the age of 16 was introduced. The increase in entries is particularly marked in double science; these have risen from 22.3% in 1989 to 82%. The term 'double science' indicates the amount of curriculum time allocated to science: the equivalent of two subjects.

However, at A level, physics entries have fallen by 18.7% over the last four years and chemistry entries by 8.1%. The most likely

reason for this is the gap between GCSE and A level. Other possible reasons are the upheaval in science education that takes place between 14 and 16, poor teaching, signals from the labour market and the rigidity of the three A-level system. All are likely to have contributed.

The lack of follow on to science A levels has to be seen in the light of an overall rise in A level entries since 1962, from 11 to 31% of the age group. The report's authors suggest bridging the gap between GCSE and A level by recording performance in the double award separately for biology, chemistry and physics or by aligning the two better so that A levels begin where the GCSE leaves off. It is also suggested that students should take five subjects at A level, rather than three, giving them more choice of subject as well as leading on more naturally from GCSE.

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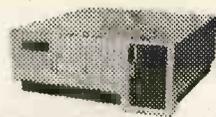
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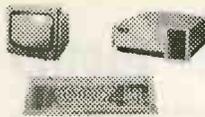


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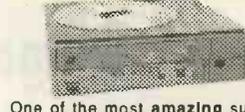
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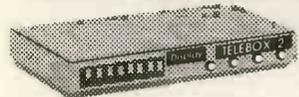
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CIRCLE NO. 107 ON REPLY CARD

## Intelligent agents under development

Researchers at BT's Martlesham laboratories in Suffolk are leading the world in the development of intelligent agents for use in tomorrow's 'super highway'. These agents could take on everything from organising the cheapest route for a long distance phone call to buying a car.

Part of the motivation for developing intelligent agents is the fear that dealing with the sheer volume and complexity of services available on the superhighway will be too complicated for most human beings. Intelligent agents will act as friendly intermediaries between us and the network allowing us to interact with the network in a natural way.

Software used in BT's intelligent agents differs from ordinary pieces of code in two key ways. It is not a set of detailed instructions on how to respond to every situation, but a general description of the way the agents should behave. They mimic

the way humans behave, working by trial and error rather than planning in detail. Secondly, the software can learn from experience and adapt the way it behaves to perform better.

Although BT is keeping the details of how its intelligent agents are programmed to behave and learn a closely guarded secret, Dr Chris Winter, project leader for its intelligent agent programme, says the techniques are borrowed from work on neural networks and game theory.

BT believes an early application of intelligent agents could be to plot the cheapest routes for telephone calls. Winter's group has already demonstrated intelligent agents performing this function. The call could be routed through several different networks representing the cheapest path to its final destination. Rates could be negotiated based on knowledge of the current costs and demand for

bandwidth on the networks.

BT is not the only company working on intelligent agents. Other major telecoms firms such as Northern Telecom and AT&T are running similar programmes, although BT believes it is leading the way in agents that can negotiate.

While BT's agents live on their 'master's' network terminal all the time, other companies are developing agents that are more mobile. "Our agents stay in one place at the moment and communicate across the network," Winter says, "but AT&T's agents actually move around the network, copying themselves". He is not yet convinced of the value of such roaming agents.

More important, in Winter's eyes, is to make the way we interact with agents more and more like the way we deal with other humans, with meaning conveyed through facial expression or hand motions.

Karl Schneider, *Electronics Weekly*

## £25m R&D project

British semiconductor equipment companies are poised to participate with other European firms in the largest single microelectronics R&D project to date, funded by the EU's £10bn Fourth Framework Programme. If accepted, the proposal, worth £25m will examine ways of extending photolithography technology beyond the visible light spectrum, including the possible use of X-rays. Potential UK participants include Rutherford Laboratories, Lumonics and Leica Cambridge. The project is being jointly funded by the EC and Jessi, the pan-European collaborative microelectronics venture.

## World's smallest power amp for mobile comms systems

Toshiba has developed what it claims to be the world's smallest power amplifier module. The amplifier is aimed at the digital mobile communications systems currently under development and is approximately one third the size of current power amplifier modules, at 5.5 x 5.5 x 2.0mm. It also operates

at 2.7V, reducing demands on the battery.

The amplifier IC design has been simplified, making it relatively easy to mass produce. The drain offset self-alignment is simpler than that of other Mesfets.

Total assembly area is minimised through application of a land grid

array (lga) structure that puts the thermal bumps and i/o on the underside of the module. This enables all of the i/o electrodes on the motherboard to be covered, and supports highly integrated packaging.

Peripheral components consist of six passive components with a size of 1 x 0.5 x 0.5mm.

## MOD spending cuts will shake up industry

Britain's defence electronics exporters could lose out to foreign competitors as a result of government plans to reduce the funding of Ministry of Defence procurement activities by £6m. Defence electronics suppliers are lobbying the MoD to lift the threat, which is the result of rationalisation of its electronics components standards review body. The electronics standards arm of the Defence Research Agency (DRA) is due to have its role drastically reduced next March.

While accepting the need for cuts, the components industry is concerned that its competitiveness abroad will be jeopardised if the cost-cutting programme is mishandled. "It is vital that a common standard of quality-assessed component procurement is adopted - equipment

makers and component companies both want that", said a spokesman for the Federation of the Electronics Industry.

At the centre of the row is the future of the DRA's electronics standards watchdog, known as the PCS. This oversees the MoD's priority and quality programme for the ordering of electronics components as laid out in DEFCON 17 document. The PCS, which accounts for almost half of the MoD's electronics standards budget, also ensures that the interests of UK suppliers are represented on international standards committees such as CECC and NATO's priority list, known as MUAHAG.

Removal from Mil Standard lists would threaten exports of UK suppliers. According to one industry source, 75 UK companies currently have components

qualified for the internationally-recognised Mil Standard. "All these things could go into jeopardy," said the source.

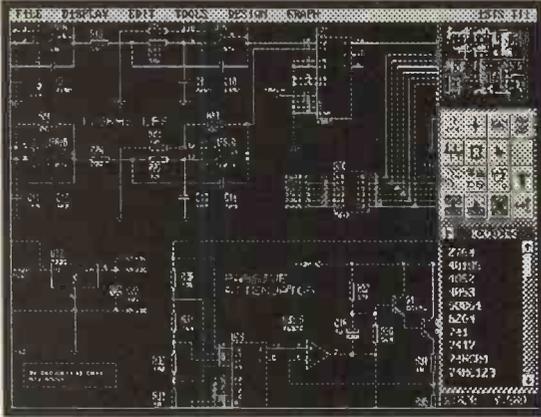
The government issued its plans for the PCS in a discussion document published in October. The scale of the hostility from industry has surprised the government, which has been forced to rethink its original proposal to close PCS in March with the loss of as many as fifty staff. "As I understand things, it is all back in the melting pot," commented a spokesman for the FEL.

One option being considered is replacing the out of date DEFCON 17 protocol with self-auditing of component quality by the suppliers. "You don't need the PCS to police industry with self-audit," said the FEL spokesman.

R. W.

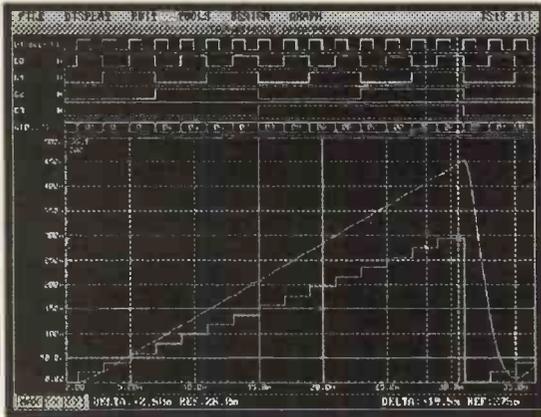
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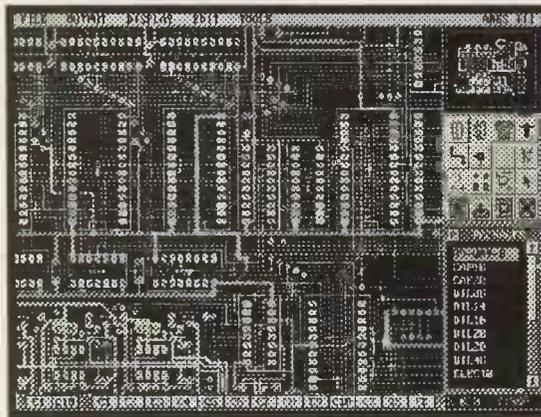
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CIRCLE NO. 108 ON REPLY CARD

# War escalates on video cd format

Fifteen years ago, after the VHS, Beta and V2000 video wars, the electronic companies vowed they would never cripple themselves again with a format battle. They remembered this when they launched cd, after agreeing in advance on a single system. Then they forgot their lesson and launched DCC and Mini Disc as rivals. So far both have paid the price.

Now the companies are making another mistake and lining up for a fresh format war over a new standard for High Density Video cd. This battle is even more dangerous because it will unsettle the trade and public who are currently being asked to support the one hour single density Vcd system which is just reaching the shops.

Two years ago Philips and Sony developed Video cd – a 12cm cd which stores an hour of movie or karaoke pictures of VHS quality. Video cds are just reaching the shops. The first public demonstration of HD-DVD, a high density digital video disc that needs a completely new player, was given recently.

Because DVD stores up to ten times more digital data, it can carry a full length feature movie, with widescreen pictures of quality that equals broadcast television and six tracks of stereo or multilingual sound. One disc can carry both original and expurgated versions of the same movie. DVD can work with a personal computer to deliver a multimedia mix of reference and games material. Philips and Sony say they hope the new technology

“will become the basis of optical media for the coming multimedia era”.

Because the potential rewards are so huge, Toshiba is promoting a rival system for ‘the 21st century’ which it claims the Hollywood movie studios support. The balance of power sits with Matsushita, maker of Panasonic equipment and the largest consumer electronics manufacturer in the world.

All today’s cds, whether used for audio, video or multimedia, rely on a finely tuned optical system. A spiral track of microscopic pits is read by a finely focused laser beam. The pits are spaced by 1.6 micrometres, which is around one fiftieth the width of a human hair. The laser emits infra-red light, with wavelength of 780 nanometres. The single-sided disc can store around 650 megabytes of data which is read at a constant rate of around 1.5 million bits a second. This lets a single-sided disc store just over an hour of moving pictures which have been very heavily compressed to the MPEG-1 digital standard.

The new high density cd will use a laser which emits red light with a shorter wavelength: 635nm. This lets it focus into a tighter spot and thus read smaller pits in a half width spiral. This gives the disc a storage capacity of 3.7 gigabytes. The rate at which the bits are read also varies, between 1 and 10 million a seconds, depending on whether the picture contains moving detail or stationary objects.

Together, these tricks let the new

disc store 135 minutes of pictures coded to the superior MPEG-2 standard. IBM, Apple, Compaq and Microsoft are already working with Philips and Sony to set a multimedia standard for using the new disc with a new generation of pcs.

In a surprise move Philips and Sony have worked with US optics company 3M to propose a future option. This will double the storage capacity to 7.4 gigabytes and 270 minutes of continuous movie playback. The double-high density disc is pressed with two spiral tracks instead of one, each at slightly different depths in the plastics material.

The effect is like writing different messages on both sides of a window pane. The laser focuses on one track for one play through the disc, and then re-focuses on the other track for a second play through.

Toshiba’s proposed disc will use similarly small pits, but glue two separately pressed discs together, back to back. The laser reads one side of the disc first, then moves round to the other side of the disc and reads that to double overall playing time.

Philips and Sony say it will be difficult to mass-produce Toshiba’s glued discs. Toshiba says it will be difficult to make and read the double layer: single sided discs used by Philips and Sony.

The companies promoting HD-cd now have both to support the future high density market without destroying the fragile market for the existing single density systems – a tricky problem of their own making.

## C&W commits to multimedia

Cable & Wireless says its purchase of a small stake in General Magic, the California-based multimedia technologies consortium, signals the company’s commitment to developing multimedia services.

“We believe that the Telescript technology developed by General Magic will enable us to offer multimedia services to consumers and businesses alike,” said Mike Harris, CEO of C&W Federal Development.

C&W was one of four new companies to take a stake in General Magic last week – the other three were Japanese electronics firms Mitsubishi, Oki Electric and Sanyo. Existing stakeholders include Apple, Northern Telecom, Matsushita, Philips, NTT, Fujitsu, Toshiba and Motorola.

## VLSI goes for MPEG

VLSI Technology has acquired MPEG 1 and 2 video decompression design blocks for its Asic library through a licensing deal with two year-old Californian video compression start-up, Mediamatics.

VLSI, which has previously said that it would not offer a standalone MPEG decompression device, intends to use the technology in its cable and satellite tv set-top box chips.

Mediamatics uses a combination of software and hardware processing for the MPEG 2 video decompression algorithm. The deal includes its MPEG 1 and 2 hardware decoder designs as well as a suite of design tools which will enable VLSI to

implement the technology as functional blocks in its asic library.

Earlier this year VLSI acquired a stake in French data transmission specialist Comatlas, as part of its plan to introduce silicon for the set-top box market. It has recently introduced a QPSK satellite tv demodulator and forward error correction device jointly developed with Comatlas.

● Digital media compression specialist, Divicom, has selected LSI Logic’s ATMizer architecture as the enabling technology for its digital broadcast system that will transport MPEG-2 compressed video streams over ATM. The MPEG-2 to ATM mapping

scheme used in Divicom’s encoder has been proposed to the ATM forum for adoption as the standard for MPEG-2 transport.

● C-Cube Microsystems has introduced chips for compressing video systems using MPEG-2, dramatically reducing the size of the encoder.

● Pioneer Electronic Europe nv has signed a contract with French DBS and pay tv company, Canal+, that will enable the company to offer digital broadcasting by the middle of this year. Panasonic will supply digital satellite receivers to Canal+ subscribers. It was selected as a supplier as the result of a competition launched in April 1994. ■

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# RESEARCH NOTES

Jonathan Campbell

## Helping hand cuts robots free – almost

Robots are still too stupid to be left to their own devices, while teleoperating them can be tedious for an operator. But the newest technique being hurriedly bolted-together in automation laboratories around the world is teleassistance, where a robot busies itself about its task, receiving only vital strategic information from its human master.

Teleassistance is quite different from the familiar teleoperation, where an operator remotely guides a robot to investigate a bomb, explore a volcano or even repair an orbiting space shuttle. Time delays can play havoc with sensitive control, and operators can end up feeling like their system is floating in treacle.

Obviously packing enough electrical intelligence into a robot so that it can act independently in complex situations is still science fiction rather than fact. But researchers, such as a team at the University of Rochester, are finding that a few simple hand signals from a human can dramatically increase both a robot's usefulness and independence.

*Robots are not yet clever enough to be given a completely free hand. With Teleassistance, the robot has a great deal of autonomy, and the operator only needs to intervene with a few crucial commands.*

Rochester graduate student Polly Pook, who recently presented her work at a conference on intelligent robots and systems in Germany explains:

"Any time you're dealing with an unknown environment, or with other people, the robot needs judgement and common sense – abilities that are extremely difficult to program".

In teleassistance, just a few strategic cues from a human operator – and a little programming – allow the robot to accomplish tasks autonomously, she says.

Pook describes today's robots as good at responding to local feedback, but still needing someone to provide context and set their agenda.

"In teleassistance, the robot has control over local problems like balancing and avoiding obstacles, and humans provide the goals and high level direction, such as where to go," says Pook.

He uses hand signals to communicate with the robot. The signals tell the robot what to do next, and sometimes, where or how

to do it. The robot detects hand motions, through sensors mounted on a glove Pook wears.

"People frequently communicate with non-verbal signs," says Pook. "We point when giving directions, hold up a hand to mean stop and gesture to say come along. I can do the same with the robot".

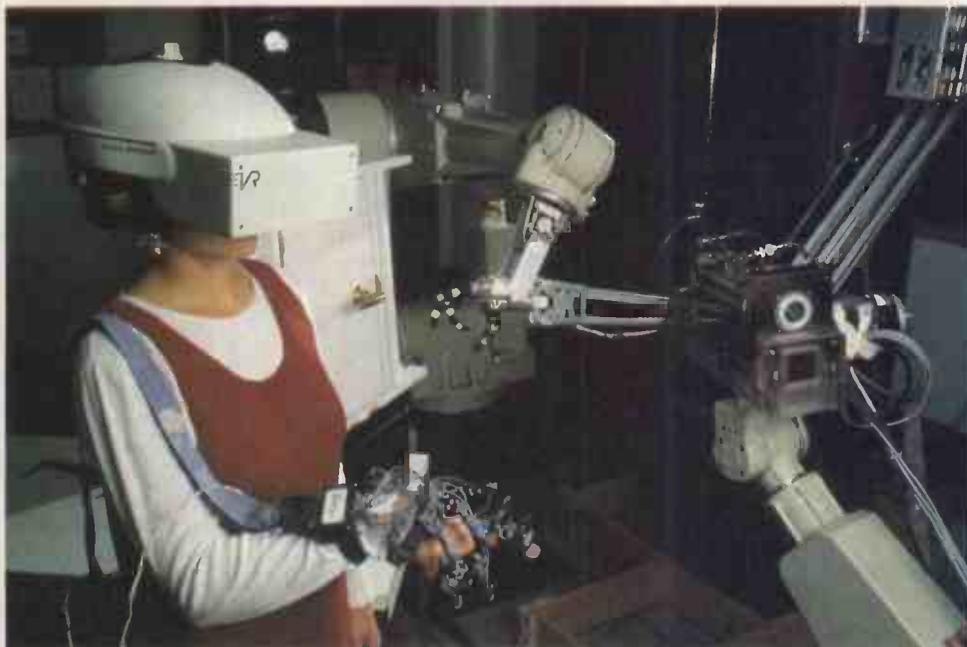
So far, Pook has guided the robot in two tasks: opening a door and flipping a plastic fried egg. They may sound easy, but insignificant changes such as substituting a different door knob or spatula, or rearranging the positions of objects, can easily confound the robot.

In the door-opening exercise, the robot reaches for the door after Pook points to it, and checks how Pook shapes her hand to discover the type of handle. Then the robot takes over using pre-defined programs to grasp and turn the handle and swing the door open.

In Pook's experiments, the tele-assisted robot opens the door much more reliably than an autonomous robot, while relying on just a fraction of the human effort compared to traditional teleoperation.

Within each context, it is able to respond quickly and appropriately to sensory feedback, without having to consider all possible world scenarios. "Once the robot knows its context, the behaviour flows smoothly and we can skip a lot of mathematics," says Pook.

Computationally-assisted telemanipulation is a science that is growing extremely fast. Pook recalls that when she began her work in teleassistance just a few years ago, hers was one of a handful of such presentations. But at the Iros conference, dozens of groups presented work in the area, with possible applications spanning robot assistance for the elderly and disabled, mining, exploration of remote sites and hazardous waste clean up.



Picture courtesy James Montanus/University of Rochester.

## Submarine that swims like a fish

Scientists at Massachusetts Institute of Technology have moved one step closer to their goal of relating the physics of how a fish swims to development of an autonomous underwater vehicle. Following extensive testing of a 'robotuna' fish that swims up and down a tank, attached to a structure containing all its electronics, the MIT team has now started design of a more complex 'robopike', that will manoeuvre attached only by a tether.

David Barrett at MIT says that testing of robotuna has gone so well that the second model could be available this summer. The tether attachment should enable robopike to make hairpin turns, while accelerations of 10g are being targeted. A real fish could experience 25g.

The robofish project began about three years ago, prompted by the need for a better propulsion system for autonomous under-water vehicle (auvs). Small robotic submarine

auvs could be used to map the ocean floor or find sources of underwater pollution. Unfortunately, conventional auvs can not carry enough batteries to allow them to explore areas such as the mid-Atlantic ridge. The obvious solutions are either to develop better batteries or improve the propulsion system. The MIT group has opted for the latter.

At present, the researchers are still concentrating on getting robotuna to swim as efficiently as possible in a straight line, to test whether their system is actually better. But other studies are also being carried out. For example, MIT is finding out more about the fundamental fluid mechanics of swimming. The robot is also acting as a test bed for possible controllers and sensors, and the researchers are using it to find a computer control system that will make a robot swim most effectively.

The project has involved a lot of watching, both real tuna at the New England Aquarium and the



electronic robotuna. The mechanical fish has been built, as far as possible, to resemble a real one and its 2843 parts include over 40 ribs, a set of tendons, a segmented backbone with vertebrae, and a Lycra skin. The robot swims down the tank propelled by a tail that gently swishes back and forth as its flexible body follows suit.

*MIT's robotuna could lead to a new method of propulsion for autonomous underwater vehicles.*

## Does electronics make us worse drivers?

As automobile manufacturers fight to pack ever more electronics into their cars, two studies recently carried out cast doubt on whether drivers can actually cope with any more technology.

The first concerns driver experience with relatively simple anti-lock braking (Allan Williams and Joanne Wells, *Accid. Anal. and Prev.*, Vol 26, No 6, pp.807-811).

In anti-lock systems, sensors monitor wheel rotational speed and as brakes are applied and the vehicle decelerates, a control unit determines when any wheel is about to lock up. If locking occurs, the control unit sends a signal to valves to reduce the brake pressure just enough to allow the wheel to start rotating again. The cycle is repeated until the driver's foot is removed from the brake or traction is restored. We all expect shorter stopping distance and improved stability should be the result.

But data suggests that cars equipped with four-wheel anti-lock brake systems are no less likely to crash than those with conventional brakes.

Previous studies have suggested that one reason could be because drivers perform more aggressively and carry out more

dangerous manoeuvres if they believe technology is going to come to their rescue.

In addition, Williams and Wells have found that many drivers simply do not know how to use anti-lock systems in an emergency. The most effective way to activate anti-lock systems is to brake hard and keep the foot down on the pedal – the opposite of how people are taught to avoid skids using conventional brakes. In the Williams and Wells work in the US, 33-50% of drivers of cars fitted with anti-lock systems said they would pump the brakes in a skid, negating much of the anti-lock effect.

In a second unrelated but relevant study appearing in the same journal (pp. 703-717), Nicholas Ward and Andrew Parkes from Loughborough University have been investigating head-up displays (HUDs), now beginning to find their way into cars.

HUDs have obvious advantages. Information can be displayed in a driver's field of view, eliminating the need to shift the gaze from the outside world, and the proliferation of intelligent vehicle systems such as collision avoidance, route navigation and vision enhancement systems would seem to make them ideal

display units.

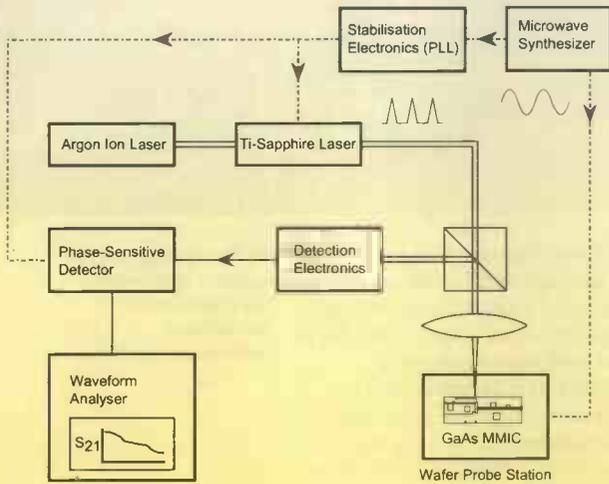
But the main objective of the HUD has to be an increase in safety of vehicle operation. Unfortunately, much of the understanding of how HUDs are used comes from aeronautical applications, and the requirements of driving a car along a road are quite different. For example, background complexity adversely affects target visibility – and complex backgrounds are certainly a normal feature of every-day motoring. Studies suggest too that for some people, the HUD may be much less effective than traditional head-down displays. Effects on speed and distance judgement could also have serious effect on safe traffic negotiation, and overall the safety consequences of such changes in attentional behaviour and information processing are not clear.

Their results convince the researchers that the HUD should not be seen as just another display opportunity.

Taken with the findings on anti-lock brakes, a worrying conclusion seems to be that without very careful thought, electronics designed to make motoring safer could actually be making the roads a lot more dangerous.

## Testing time for microwave circuits

NPL in the UK, the Fraunhofer Institute for Applied Solid State Physics in Germany and Dassault Electronique, France, have joined forces to develop an electro-optic probe station for measurement of rf and microwave integrated circuits to 20GHz and beyond.



*Electro-optic probe station for GaAs microwave integrated circuit measurement.*

*Researchers at NPL investigating electro-optic sampling for measurement of fast electrical pulses and microwave signals. NPL has joined with the Fraunhofer Institute and Dassault to develop a useable test technique.*

Demand for increasingly dense components, where it is difficult to lay down test points, are pushing current test techniques to their limits. But the NPL/Fraunhofer/Dassault work, sponsored by the European Commission, should improve circuit measurement capabilities by extending existing wafer-probe techniques.

Recent advances in the fields of electronics and opto-electronics have had a significant impact on communications, information systems and military markets. This has led to development of components and systems with bandwidths of many tens of gigahertz, capable of handling optical or electrical pulses and transients in the pico to nanosecond range. Electro-optic probing could become an important tool in the validation of new microwave integrated circuit designs.

As its name implies, the probe station will operate by using the electro-optic effect to sense microwave electrical fields at points inside an integrated circuit. The effect is exhibited by certain classes of non-centrosymmetric crystals which, in the presence of an applied electric field, acquire an induced anisotropic refractive index. Such

materials are able to act as transducers for the detection of microwave signals by optical means, because the time-varying refractive index can be used to modulate an optical signal used in a probe. Change of phase of the optical signal can be related to the local microwave E-field strength in the circuit.

By scanning a sharply focused laser beam across a circuit, a high spatial resolution two-dimensional map of the field strength inside the circuit can be produced.

Two distinct techniques are possible for the electro-optic characterisation of microwave integrated circuits. Firstly, for III-V semiconductors which exhibit the electro-optic effect, such as GaAs, it is possible to use the phenomenon in the circuit substrate itself to sample the microwave field.

If the substrate is not electro-optically active – silicon or alumina for example – a small crystal of an electro-optically active material such as LiTiO mounted close to the circuit under test can be used as an external field probe.

The goal, according to Dr Alan Roddie of NPL, is a non-invasive technique that will be able to obtain results where conventional technology falls short. ■





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*With BT announcing extensive video-on-demand trials in the middle of this year, interest in the technology needed to control television pictures via the telephone is high.*

*Geoff Lewis believes that the technology in question – adsl – could even become a world-wide standard.*

# ADSL

## an easy route to video on demand

In the UK, legislation currently banning British Telecom (BT) from carrying broadcast tv signals over its telephone network, does not preclude the provision of a video-on-demand, or VOD, service. With suitable technology, this large installed subscriber network of more than 30 million twisted pair cables between local exchanges and domestic subscribers, means that BT is well placed to be able to provide a VOD service, in a relatively short space of time and at low cost. The telecommunication companies should thus be able to surmount the typical £200 consumer barrier for a system upgrade. Furthermore,

such a system might well be extendable to support talking books for the blind, educational needs for open colleges, and the exchange of medical X-ray and various scan information for the Health Service. One asymmetrical digital subscriber loop software vendor is already thinking of introducing interactive video games for multiple players, each competing with the others for some particular goal.

The unshielded twisted pair cables that were originally designed to provide analogue voice services in the frequency range 300 to 3400Hz, use copper or aluminium wires in various gauges, ranging from 0.32 to 0.9mm

### Who needs adsl?

Before a new technology becomes acceptable, the general public has to accept that it provides a solution to a recognised problem at an acceptable add-on cost. This can be seen by comparing the adoption rates of satellite and cable delivered television programmes with the rapid up-take of video recorders.

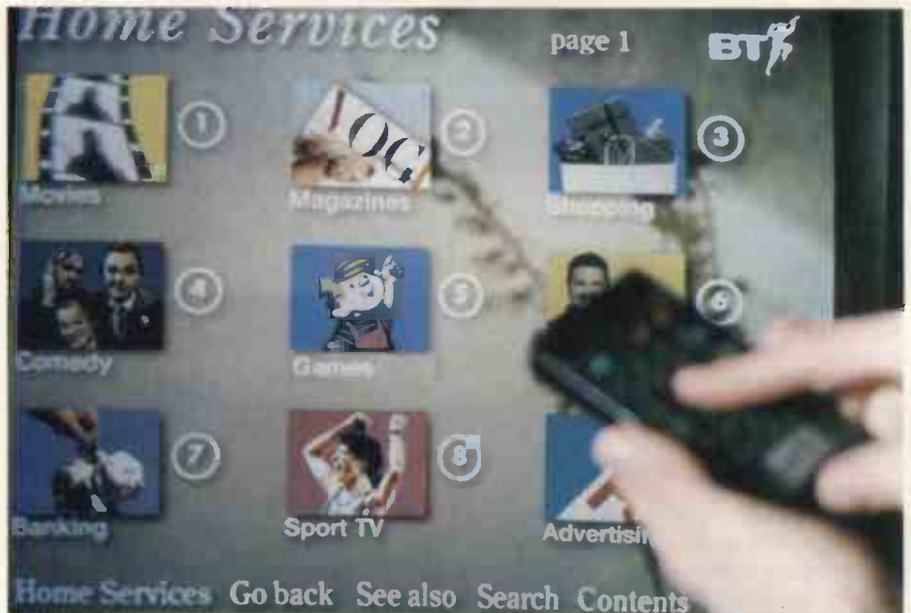
After the standards battle had been settled, with plenty of software available, the video recorder quickly became part of the entertainment system. First the recorder was used as a time shifting device to combat channel conflicts, then as a primitive form of video on demand, VOD, following the introduction of the local video shop.

It offered a more immediate form of video on demand, removing the need for the journey to the shop, the viewer would be presented with a solution to a problem that had probably not been previously recognised. Adsl promises to meet this need – and at a competitive price.

Although adsl is a product of the 1990s, internationally there are many players investigating its possibilities. Every month a new one joins the band of would-be service providers. Among the latest is Australia, where their sub-continental super-highway includes plans to incorporate adsl as a way of providing new services for areas of low population density.

in diameter. Since any particular subscriber's final link may consist of a combination of these, the line attenuation may well vary from link to link.

About five years ago, it was recognised that unshielded twisted pair cables were in use with local area networks, lans, and carrying digital data at rates greater than 10Mbit/s.



Video on demand gives customers the sort of control over selected programmes and services that they would normally have on a video recorder – play, fast forward, rewind and freeze frame.

Research work in the USA at Bell Laboratories and in the UK at British Telecom then took a new look at the rf characteristics of the local loop. Figure 1 indicates how the line attenuation and group delay varies for a typical section of unshielded twisted pairs up to 1MHz.

While attenuation over the major part of the range is high at about -0.075dB/kHz, the characteristic is fairly linear and easily equalised. But most importantly, the group delay is practically constant so that any pulse distortion will be small, with consequently few bit errors in a digital transmission. These features allow the simultaneous transmission of high bit rate digital signals and conventional telephony over the same local loop.

Figure 2 shows the general allocation of the frequency spectrum above the voice band. A low bit rate control channel operating at either 9.6, 16 or 64kbit/s provides for the subscribers' demands, together with a high bit rate channel for the delivery of selected tv and

other programmes. It is this asymmetrical bit-rate allocation that gives the system its title of asymmetrical digital subscriber loop or adsl.

Organising the system in this way is cost effective. The hardware costs for the individual user end is relatively low, while the high costs of the high bit rate end are shared by many subscribers. Adsl can provide video with video-recorder type characteristics – fast forward, fast rewind, freeze frame, pause, etc functions – on demand.

Using MPEG-1 digital compression with suitable channel coding, video signals of various grades can be transmitted direct to the home. A bit rate of 2.048Mbit/s can provide VHS quality video together with digitally coded stereo sound. At 6.144Mbit/s with MPEG-2 compression, the video quality is equal to that of a broadcast transmission. The introduction of a new consumer-based demand would provide the necessary impetus to reduce the high cost of current MPEG-2 chip sets.

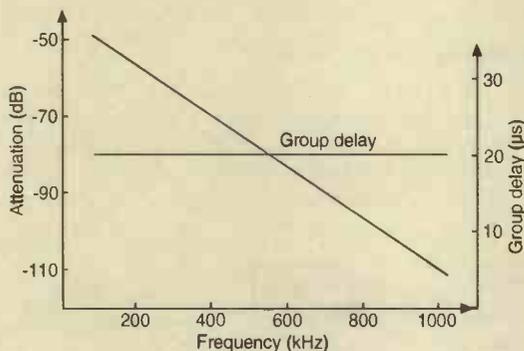


Fig. 1. Rf characteristics of unshielded twisted-pair telephone. While the attenuation might be anticipated, the constant group delay characteristic is fortuitous for a digital system.

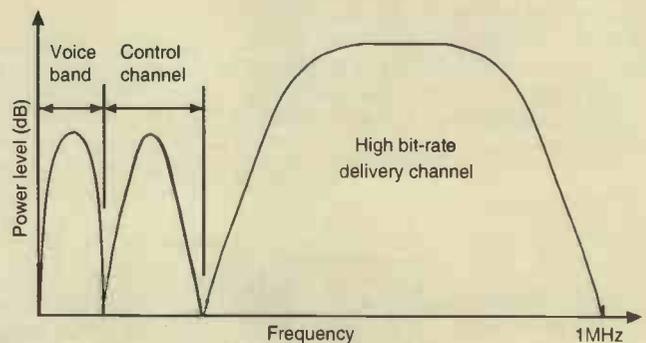


Fig. 2. Frequency-division multiplexing for adsl. This shows clearly adsl's ability to support simultaneous voice, telephony and the subscriber's demand and programme delivery channels.

There are various ways of allocating the high-bit-rate down-stream channel; four times 1.544Mbit/s and 6.176Mbit/s suit North America while three times 2.04Mbit/s and 6.144Mbit/s cater for Europe; other combinations are possible. Note that this makes the bit rates compatible with integrated-services digital network, isdn. As you would expect, the higher the bit rate systems, the shorter the local-loop penetration distance.

Analogue filters with high stop-band attenuation are used at each end of an adsl network in order to combine or separate the telephony and the extra services with the minimum of mutual interference. In practice, it is necessary to be able to separate adsl signals of a few millivolts in amplitude from a few tens of volts of ringing tone.

Because of the make-up of the local loop, many of which include overhead sections, there are many connections that can give rise to resistive variations which tend to be temperature and moisture dependant. In particular, overhead line vibrations and other variables can give rise to signal drop-outs that last for several milli-seconds. Adsl networks are designed not to reset under these conditions but to ride out such breaks.

Cross talk between the individual pairs in a multi-pair cable is not significant and neither is cross talk between other isdn users. Impulsive noise from ringing tones and switching signal could be troublesome, but by using a suitable transmission technique with forward error correction, fec, this has little significance in a well maintained network.

To minimise the effects of both radiated and induced radio frequency interference, adsl employs a balanced feed of signal to the line. By taking suitable preventative measures, adsl signals of less than 1mV amplitude can be recovered even under noisy conditions and in the presence of high level telephony signals. The specification includes forward error correction in the form of Reed-Soloman coding with the possibility of adding interleaving. Parameters for these features are incorporated in the American National Standards Institute (ANSI) documentation.

Digital adaptive equalisers adjust automatically to suit individual line characteristics. These cater for the problems including the temperature and moisture dependent parameters and continuous interference. Provision is made so that at each new line acquisition, the adaptive equalisers are reset under the control of a line-probing bit sequence that establishes the system signal-to-noise ratio. To maximise

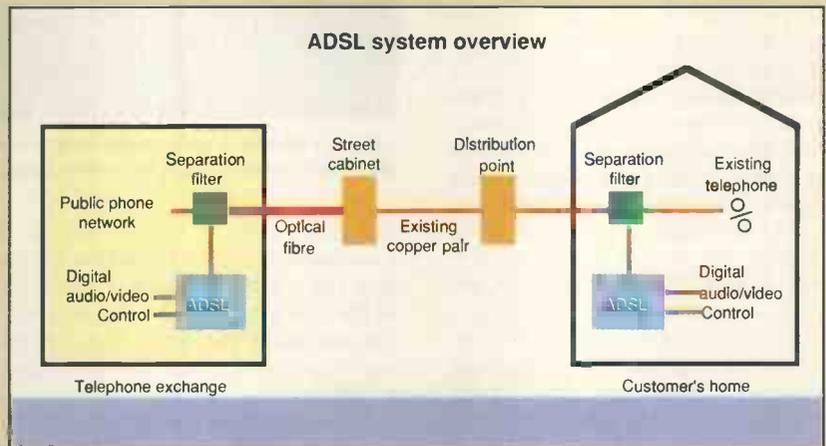
## Interactive tv trials

Towards the end of last year, BT announced that it is to start consumer trials of interactive tv – which includes video on demand. Starting in the middle of this year the trial is to involve 2500 households in Colchester and Ipswich. The move follows successful technical trials, held near Ipswich, involving sixty BT employees.

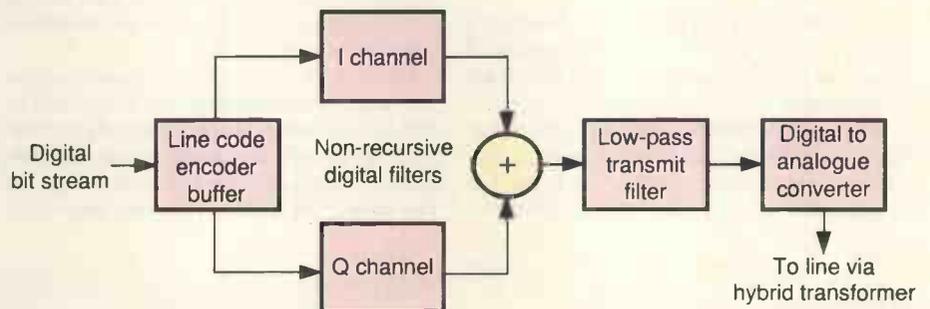
The new service will bring together the telephone and television to enable customers to choose from a range of services

and material. Once selected in the home, video information is transmitted from a central database over the telephone network to the television, without affecting the normal telephone line.

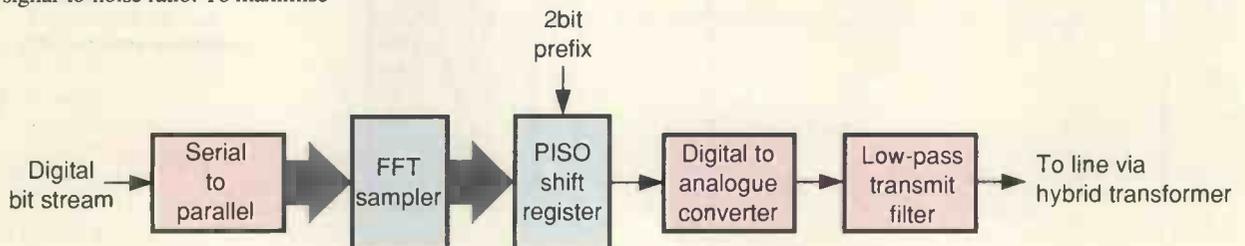
During the trial, BT aims to offer shopping on demand, a range of educational programming for homes and schools, movies and television programming (video on demand), a home banking service, a magazine service and a community link local information service.



Telephone television. With adsl, 16kbit/s data from the subscriber is used to access a one-way broadband video signal from the exchange.



The basic cap modulation system depends on relatively low cost, readily available vlsi chip sets.



Discrete multi-tone modulation system. While there is nothing new in the processors, vlsi device development for dmt is still at an early stage of development.

## Video-on-demand technical trial results

Objectives of BT's technical trial were to test the feasibility of delivering video and multimedia services, with no market information objectives.

The set top box used was based on an Apple Macintosh LC475 running a Mac operating system modified to support MPEG and a 2Mb/s network interface. At start-up it downloaded the operating system and Oracle Media Objects (OMOs), the run time version of the authorware tool in which the services are created.

The network connected to the set-top box was a 2Mbit/s stream delivered over either copper, using asymmetric digital subscriber loop (adsl), or fibre. Over an ordinary telephone loop, adsl technology delivers 2Mbit/s in one direction, a 9.6kbit/s bidirectional control channel, and the analogue telephone service.

The variety of adsl used in the trial was discrete multitone, which divides the spectrum into a number of bands and

spreads the audio transmission over them in a such a way as to minimise interference and noise. For the fibre delivery a passive optical network was used.

Users were directly connected to the server system without any concentration or switching stage. The server system consists of an nCube massively parallel computer and the system ran on Oracle database software with the video content compressed to MPEG 1 standards.

Results of the trial were:

- Adsl technology is robust and has wider than expected range – of about 6km. Resistance to noise and to installation wiring anomalies is excellent, corresponding to 92% network penetration.
- The combination of MPEG coding and adsl works well. As a result, adsl is a viable technical solution to the problem of delivering high-bandwidth asymmetrical services.

BT will be using adsl for the market trial along with fibre in the ratio of about four copper to one fibre. In the technical trial discrete multitone was used, but carrierless amplitude and phase modulation adsl is to be used for the market trial. This is because it is available earlier in a more integrated and therefore cost effective form, says BT.

For the market trial, the same combination of server and set top box technologies supplied by nCube, Oracle and Apple Computers will also be used. However, there will be a switching/concentration stage which will be asynchronous transfer mode, atm, based. Users will be connected to an atm switch with a concentration ratio of roughly two to one. This means that there are about twice as many users as there are server ports, and about half of them can be on line to the system at any one time.

the channel capacity, bit rate reduction coding may be used. For example, with 2B1Q coding, each pair of binary input digits is converted into one of four quaternary output symbols.

### Modulation format options

A number of different modulation formats have been tested and these include, qam, cap and dmt.

#### Quadrature amplitude modulation, qam.

First, the binary data stream is split into two sub-streams and each separately modulated on to orthogonal versions of the same carrier frequency. The two modulated signals are then added and low pass filtered before transmission to network.

#### Carrierless amplitude-phase modulation, cap.

The basic implementation of this technique is shown in Fig. 3. Again, the bit stream is split into two components. Each is then passed through a pair of Hilbert non-recursive digital filters that have an impulse response differing in phase by  $\pi$ . The outputs are then added, passed through a digital-to-analogue converter and filtered before being passed to the transmission network.

**Discrete multi-tone modulation, dmt.** This preferred method has a lot in common with cofdm, which is short for coded orthogonal frequency division multiplexing, in that the main channel is divided into many sub-channels.

Each serial digital input signal is first encoded into a parallel form and then passed through a fast-Fourier-transform processor. This converts the frequency domain samples

into time domain values with a sliding time-window effect. These values are then transcoded into a serial format and digital to analogue converted before transmission.

Basic transmission processing is shown in Fig. 4, which shows that two or more prefix bits are added to the bit stream in order to minimise inter-block interference. When these bits are dropped in the decoder, the bit stream is typically broken up into 512 bit blocks to improve the definition of the FFT window. The line characteristic for each narrow sub-channel is practically constant so that the minimum of pulse smearing is created thus improving signal quality. Any impulse noise that appears as interference is spread over many sub-channels by the FFT processor window so this form of interference is less likely to create data errors.

Within a sub-channel, the number of bits transmitted can be varied adaptively depending on the signal to noise ratio in the sub-channel. This not only improves signal quality on a particular line, but it also minimises the effect of cross-talk from other lines.

While the dmt modulation scheme is more robust under noisy conditions, the processors are more expensive than those for cap modulation. However, at least one chip manufacturer has stated that chip sets could be available by the end of 1996 for under \$100. Other proponents of adsl have variously estimated that if the take-up by subscribers is greater than about 10%, the system would be profitable to both chip set manufacturer and video-on-demand providers.

Whichever modulation scheme is used, the digital data is converted into analogue format, low pass filtered and coupled to the balanced

line network via a hybrid transformer. In each case, the down-stream receiver functions in a complementary manner.

Hardware development is aimed at a single board approach with extensive use of standard vlsi chips wherever possible and with consumer prices very much in mind. The concept, which is a product of the 1990s, is compatible with the optical fibre networks that are currently in the process of being installed. When fibre is eventually installed in the final loop, both the quality of the images and the extent of the information services available in the home will be vastly expanded. Data rates in excess of 6.144Mbit/s have already been tested over relatively short voice-grade transmission lines.

Although the technology has been largely been driven by the laboratories such as BT and Bell, hardware manufacturers are starting to develop and offer systems to implement adsl. While BT is currently running trials in the UK, it is interesting to note that the telecoms organisations of Hong Kong, Korea and Singapore all plan to start trials before the end of 1994. With such Far Eastern interest, the costs of installing adsl must surely drop dramatically within the next three years.

Standards for adsl are currently being developed by ANSI by the working Group TIE1.4, under Document No TIE1.4-007 1993. The European Telecommunication Standard Institute (ETSI), group STC TM3 RG12, is currently engaged in agreeing the details of this annex. Since it is not expected that an independent European standard will develop and because of isdn compatibility, adsl could offer a possible world standard distribution system for television signals. ■

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Interfacing via the pc's printer port, this 16-channel data-capture subsystem with 12-bit resolution can be moved between machines without having to open the computer's lid. In addition to its analogue inputs, the subsystem also provides an eight-bit digital output port, as Pei An explains.

# CATCHING DATA VIA LPT1

The present precision data acquisition system has a 16-channel analogue multiplexing facility and a 12bit a-to-d conversion accuracy. It incorporates an on-board digitally-controlled variable gain amplifier, offering gains of 10, 100 and 1000.

Input voltage is in the range from  $-5V$  to  $+5V$  with an accuracy of about  $1\mu V$ . The conversion rate is 7Hz. In addition, the system has an 8-bit digital output port. This allows the device not only to read analogue signals into the computer, but also to send data out of the computer for controlling external equipment.

There are two main modules in the data log-

ging system – an a-to-d conversion board and an interfacing card. The former deals with analogue signal multiplexing, amplifying and analogue-to-digital conversion; the interfacing card manages communication between the a-to-d board and the computer.

I have designed two versions of the interfacing cards. The first is intended for interfacing with the centronics port of the computer and the second for interfacing with the RS232 ports. Both cards incorporate an 8255 programmable peripheral interfacing (ppi) IC. This provides three programmable eight-bit i/o ports through which the a-to-d converter board is connected. In this article, only the design of the centronics i/o card is discussed.

The a-to-d board can be connected to any i/o ports, provided that two 8-bit outputs and one 8-bit input are available. A complete data logging system connected to a pc or a laptop computer is illustrated in Fig. 1.

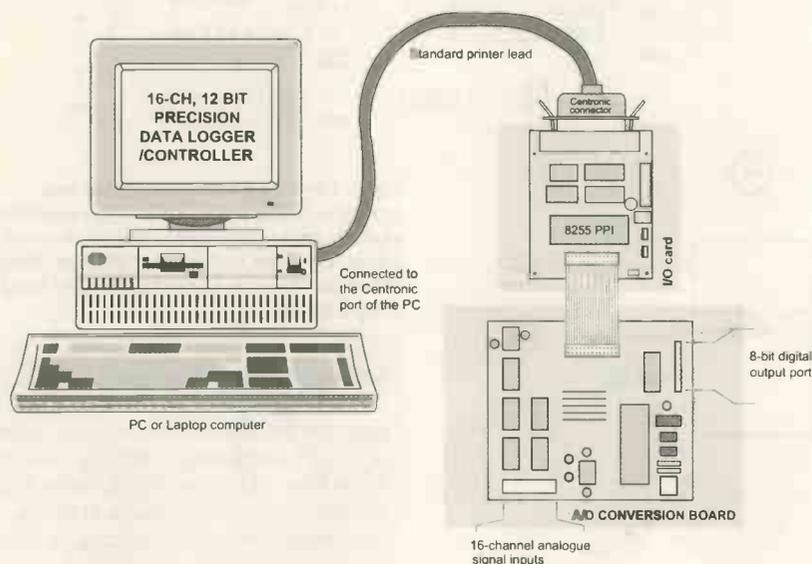
The description of this system is divided into two parts. The first one concentrates on the works of the a-to-d conversion board and the second on the centronics i/o card.

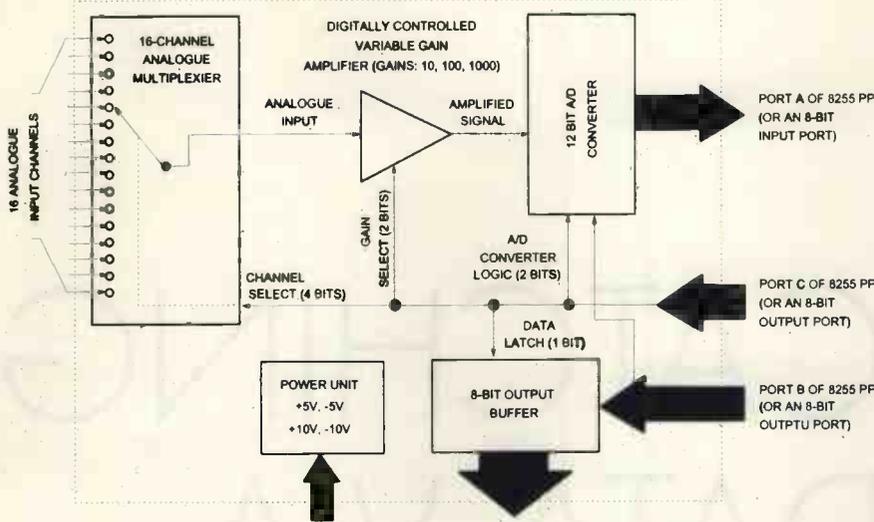
## A-to-d conversion module

The system comprises five blocks. These are the analogue multiplexer unit, the digitally-controlled variable gain amplifier unit, the a-to-d conversion unit, the digital output unit and power supply unit. Figure 2 shows the block diagram of the a-to-d board while Fig. 3 gives the circuit diagram.

**Analogue multiplexer.** The analogue multi-

Fig. 1. Precision data logger/controller comprises an a-to-d converter board providing 16 analogue input lines and 8 ttl outputs. It is controlled by a pc or laptop computer via the centronics printer port using appropriate i/o card.

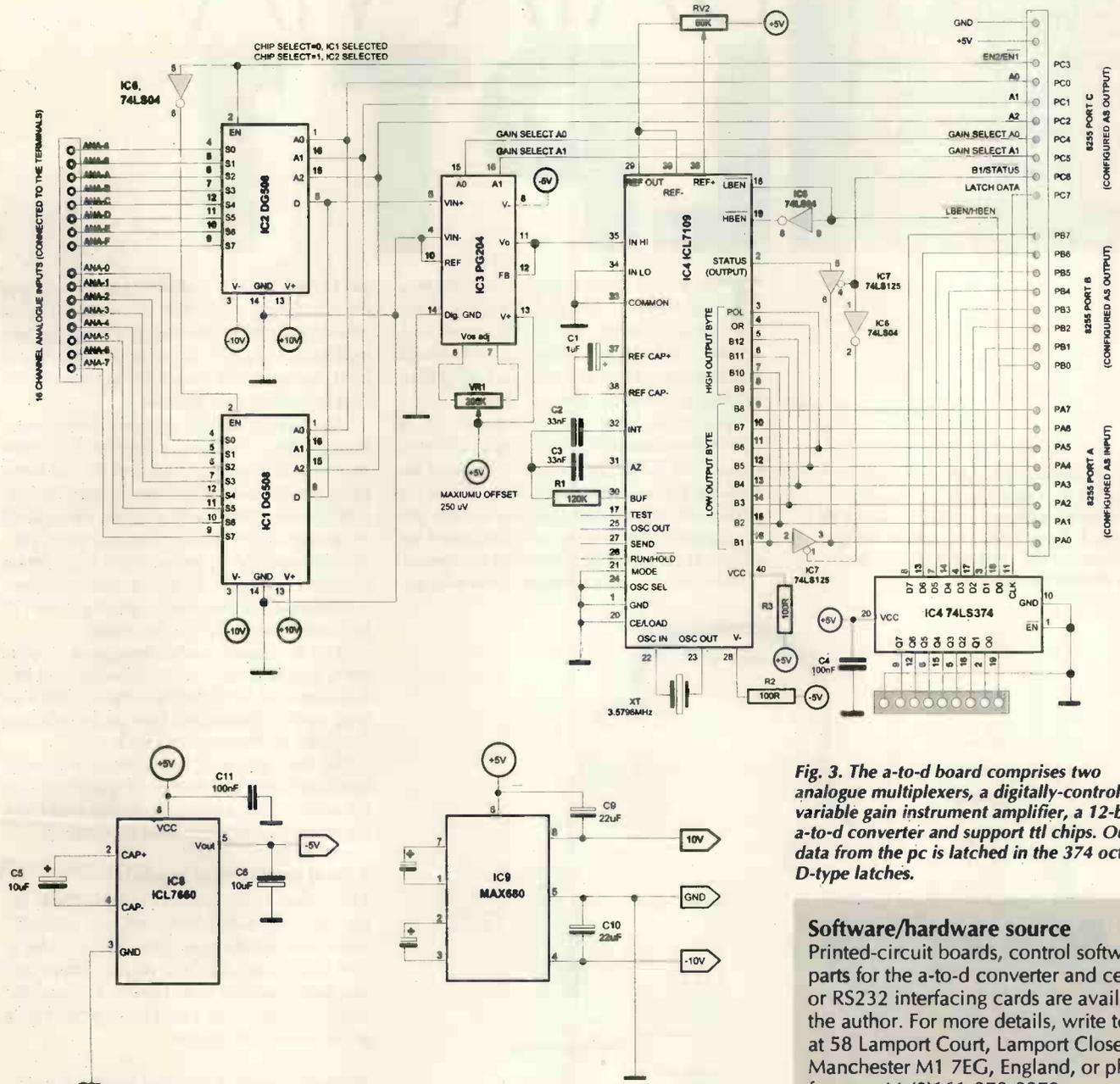




plexer unit is built around two DG508 eight-channel analogue multiplexer ICs, namely IC<sub>1,2</sub>. These provide 16 channels of analogue inputs.

The 508 is a high quality analogue switch exhibiting very low on resistance and high off resistance. On resistance is virtually constant over the entire analogue signal range. The IC can handle analogue input voltages up to  $\pm 40V$ . A dual balanced power supply is needed. Supplies V+ and V- connect to the +10V

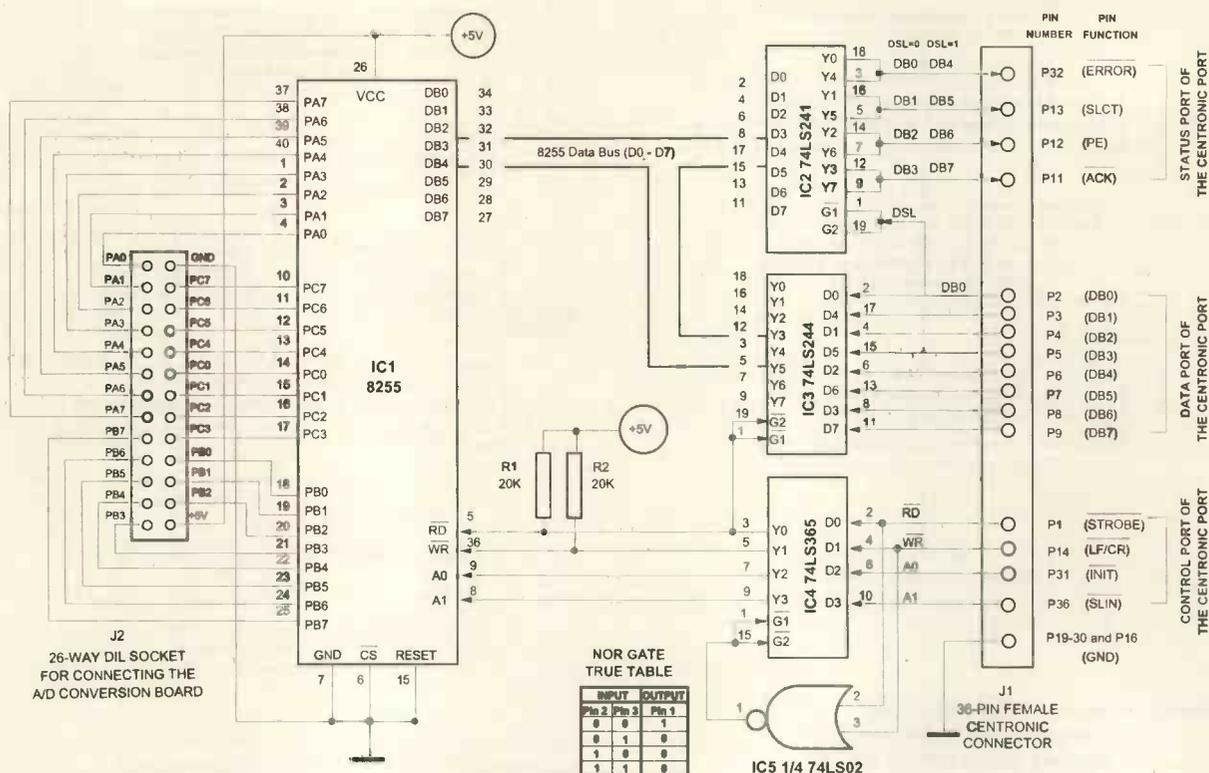
**Fig. 2. Block diagram of the a-to-d conversion board.** One of the 16 analogue inputs is selected by the analogue multiplexer. The selected signal is first amplified by the digitally-controlled variable gain amplifier, then fed into the 12-bit a-to-d converter. Data from the computer is stored in the output buffers. The board communicates with the pc via one 8-bit input port and two 8-bit output ports.



**Fig. 3. The a-to-d board comprises two analogue multiplexers, a digitally-controlled variable gain instrument amplifier, a 12-bit a-to-d converter and support ttl chips. Output data from the pc is latched in the 374 octal D-type latches.**

**Software/hardware source**

Printed-circuit boards, control software and parts for the a-to-d converter and centronics or RS232 interfacing cards are available from the author. For more details, write to Dr An at 58 Lampport Court, Lampport Close, Manchester M1 7EG, England, or phone or fax on +44-(0)161-272-8279.



**Fig. 4. Centronics port to 8255 peripheral interface chip.** Communication between the computer and the 8255 is achieved by several ttl chips. Data and status ports centronics interface are used for sending data to and reading data from the 8255 peripheral IC. The control port manages operation of the 8255 peripheral chip.

and -10V rails of the power supply. Pins S<sub>0</sub> through S<sub>7</sub> are the analogue inputs while D<sub>0</sub>, on pin 8, is the analogue output.

Analogue input is selected by applying an address at A<sub>0-2</sub> address lines. During the operation, the enable input should be pulled high. Address lines A<sub>0-2</sub> are connected to the lines DB<sub>0-2</sub> on port C of the 8255 peripheral IC. Enabling for IC<sub>2</sub> connects directly to DB<sub>3</sub> of port C and the enable for IC<sub>1</sub> connects to DB<sub>3</sub> via inverter IC<sub>6</sub>. This ensures that only one multiplexer works at a time.

Analogue outputs of the ICs are wired together and then connected to the input of the amplifier, Fig. 3.

**Variable-gain amplifier.** Analogue outputs from the multiplexers connect to the input of the PGA204 amplifier. This device is a high performance, low cost, general purpose instrumentation amplifier with programmable gain. It is laser trimmed for very low offset voltage and drift combined with high common-mode rejection. In addition, it will operate with supplies from 4.5 V to 18 V and its quiescent current is about 5mA.

Inputs V<sub>+</sub> and V<sub>-</sub> of the PGA204 connect to the +5V and -5V rails of the power supply. Gains of 1, 10 100 and 1000 are digitally selected by two ttl/cmos-compatible address lines, A<sub>0,1</sub> on pins 15 and 16.

There is no latching for the address lines. A change in the address inputs immediately selects the new gain. However, a delay of around a microsecond is needed for the amplifier to settle to a new output voltage in the newly selected gain.

Input connects between V<sub>in-</sub> and V<sub>in+</sub>. Internal input protection allows overloads up

to ±40V on the inputs without damage. Output V<sub>0</sub> is referred to the output reference REF, which normally connects to the analogue ground via a low impedance.

The PGA204 has an output feedback connection at pin 12 which must be connected to the V<sub>0</sub> output terminal for proper operation. Normally, a constant current of approximately 1.3mA flows through the digital ground pin. This makes it necessary to return the digital ground through a separate connection path so that the analogue ground is not affected by current in the digital ground.

Pins 6 and 7 allow offset of the input stage to be trimmed. Address lines A<sub>0,1</sub> connect to BD<sub>4,5</sub> of the 8255's port C, which is configured as an output port.

**A-to-d conversion unit.** Amplified analogue signal is finally fed into the ICL7109 12-bit a-to-d converter. This IC is a high-performance, low-power, integrating a-to-d converter. It needs +5V and -5V power supplies on pins 40 and 28 respectively. Pin 1, GND, connects to the ground.

References REF<sub>IN+</sub> and REF<sub>IN-</sub> connect to a bandgap voltage reference. The 7109 provides an on-board voltage reference which is normally 2.8V below V<sub>+</sub> and has a typical temperature coefficient of ±80 ppm/°C. This refer-

ence voltage is at REF<sub>OUT</sub>.

The IC needs some capacitors and resistors for the analogue side of the converter, values of which should be calculated according to the manufacturer's data sheet. An on-chip oscillator operates with an inexpensive 3.5795MHz tv crystal producing 7.5 conversions per second.

Digital communication with the 7109 is configured in direct mode, achieved by making the mode input pin open or low. In this mode, output data is directly accessible under control of the chip. The RUN/HOLD pin is unconnected and thus pulled high by the 7109's internal pull-up resistor. This allows the converter to perform a-to-d conversions continuously and to update the 14 tri-state outputs B<sub>1-12</sub>, OR and POL. Lines B<sub>1-12</sub> present the 12 bits of the conversion data. Input over-ranging is indicated on the OR pin while POL indicates polarity of the signal.

During a conversion cycle, the status output goes high then low after new converted data has been stored in the output latches. It is used as a 'data-valid' flag for monitoring the status of the converter. In this design, STATUS and data bit B<sub>1</sub> share the same line, which connects to DB<sub>0</sub> of port A of the 8255 peripheral interface. Data line DB<sub>6</sub> of port B selects whether STATUS or B<sub>1</sub> is read into the pc.

Chip enable line -CE/LOAD connects to the ground to enable the IC. When -LBEN is low, the low eight bits B<sub>1-8</sub> output the data while high bits B<sub>9-12</sub>, OR and POL are high impedance. When -HBEN is low, B<sub>9-12</sub>, OR and POL will output data and the low bits are in high-impedance state.

In Fig. 3a, some outputs in the two groups are connected together to form an eight-bit

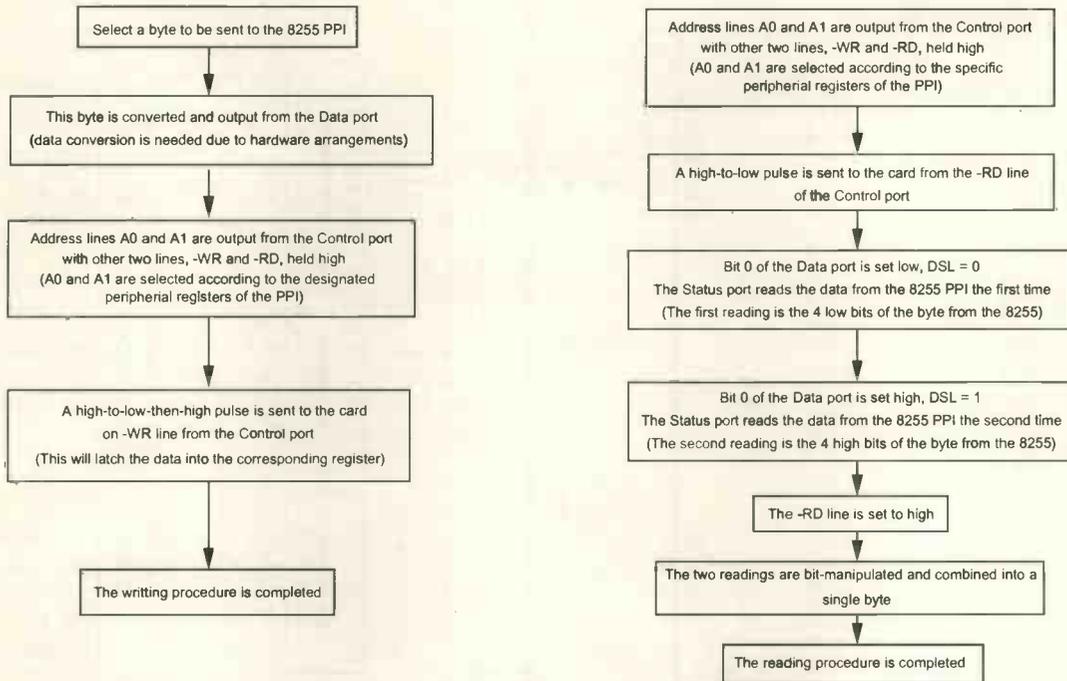


Fig. 5. Software flow for writing data to, and reading data from, the peripheral registers of the 8255 parallel interface. When writing, the data port of the centronics interface supplies data to the 8255 and the control port controls transfer. During reading, data from the 8255 is read into the pc via the status port. As only four inputs are used by the status port, the eight-bit data is loaded into the pc in two consecutive readings, which is controlled by the first lsb of the data port. Operation of the 8255 is again managed by the centronics control port.

output instead of 14 bits. Lines *-HBEN* and *-LBEN* are used to select the lines either in the high byte or in the low byte. These eight-bit data buses connect to port A of the 8255 peripheral interface which is configured as an input port. Line *-HBEN* connects to *DB<sub>0</sub>* of port B while *-LBEN* connects to *DB<sub>0</sub>* of port B via inverter *IC<sub>6</sub>*.

**Digital output unit.** An *LS374* octal latch is used. Inputs to the IC are supplied by port B of the 8255, which is configured as an output. This data is latched to the outputs by taking the clock line from low to high. The clock is connected to *BD<sub>7</sub>* of the 8255's port C.

**Power unit.** The power unit derives *-5V* and *±10V* dc supplies from a 5V dc rail via two voltage conversion chips, Fig. 3b. The *ICL7660* converts +5V to *-5V* while a *MAX680* converts +5V to +10V and *-10V*.

**Interfacing to the pc**

Interfacing via the centronics printer port was discussed in an article entitled 'Real world control via LPT' in the September issue of *EW+WW*. Briefly, the centronics port consists of three independent ports, namely, the data

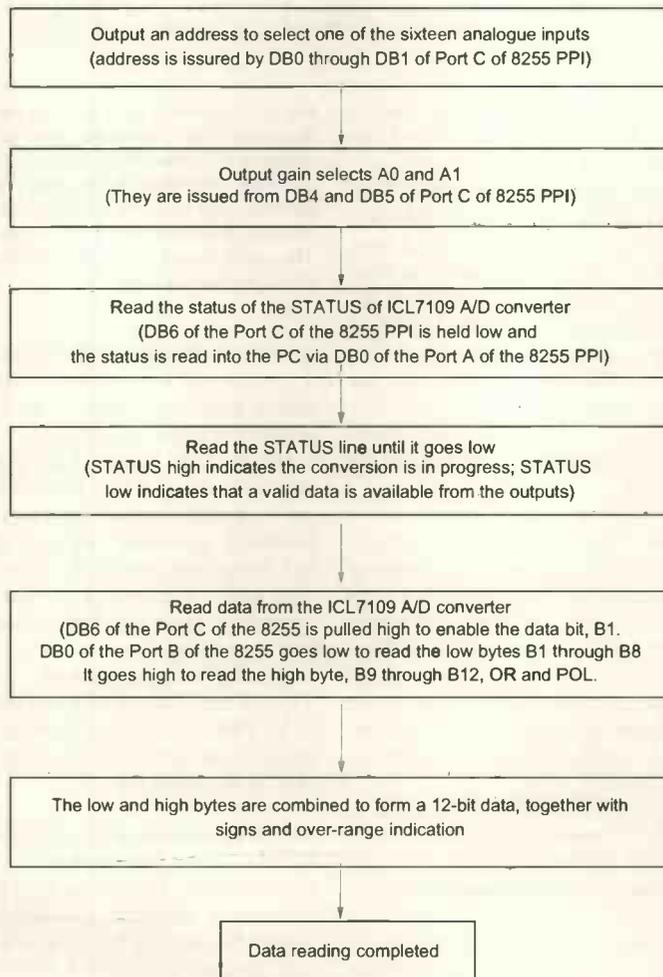


Fig. 6. Analogue-to-digital conversion begins with selecting one of the 16 analogue inputs. It is followed by sending a command to select the gain of the amplifier. After this the status of the a-to-d converter is polled continuously until it indicates that conversion is finished. The 12 bit data - plus polarity and over-range flags - are read into the pc twice. They are finally combined to form a single 12-bit word.

output port, the control output port and the status input port.

For the data logger, the centronics data port is used to send information to the interface card while the status port is used to read data from the IC. The control port manages reading and writing operations of the 8255 interface chip.

The 8255 is an industry standard programmable peripheral interface with four internal registers. Three of these are called peripheral registers and are associated with ports A, B and C. The fourth is the control register.

All three peripheral registers are used for data transactions between the 8255 and external circuits while the control register is used to initialise the operation modes of the parallel interface.

There are eight bidirectional data lines,  $DB_{0-7}$ , through which data is written to or read from the internal registers under the control of read and write lines. Address lines  $A_{0,1}$  select a particular register.

Data transfer is facilitated by  $IC_2$  and  $IC_3$ , which are tri-state buffers. Control over the 8255 is made by  $IC_4$  and  $IC_5$  which are a tri-state buffer and nor gate respectively.

You can see from Fig. 4 that two lines of the centronics control port, pins 31 and 36, are

connected to address lines  $A_{0,1}$  of the 8255 peripheral interface via  $IC_4$  - a tri-state buffer.

The other two lines of the control port, pins 1 and 14, connect to  $-RD$  and  $-WR$  of  $IC_1$  via  $IC_4$ . A problem could occur when these two lines are both low, since the 8255  $-RD$  and  $-WR$  lines cannot be set low simultaneously. To prevent this, a nor gate is used. Its two inputs connect directly to the two control lines from the centronics port, pins 1 and 14, and its output connects to the enables of  $IC_4$ .

When the two lines from the control port are both low, output of the nor gate goes high. This disables the buffers on the  $LS365$  and sets all the outputs at high impedance. Resistors  $R_{1,2}$  pull the  $-RD$  and  $-WR$  lines high.

To write data to an 8255 register, firstly the required data is written to the centronics data port together with an address to the control port, then a high-to-low-then-high pulse is issued from the control port write line. This enables data on the inputs of  $IC_3$  buffers to be transferred to the 8255 data bus and in the same time written into the selected register.

Reading data from the 8255 is slightly complicated, since the centronics port only has five input lines. In order to read eight bits of data, the computer has to read at least twice. This is done by  $IC_2$ , an  $LS241$  tri-state octal buffer.

In Fig. 4, when the first enable, pin 1, is low, the four left hand side buffers work, i.e. the outputs follow the inputs. When pin 19, the second enable, goes high, the four right-hand side buffers operate.

By connecting pins 1 and 19 together to form a data-select line and by putting the line low and then high, the status port can read the four bits connected to the left and right hand buffers in turn. These two readings are then bit-manipulated and combined to form a single eight-bit byte.

Operating in such a manner, the 8-bit data appearing on the input lines of  $IC_1$  can be read into the centronics port. Referring to Fig. 2, the data-select line is controlled by  $DB_0$  of the centronics data port.

The card incorporates a 7085 1A, 5V regulator. An external 8-12V dc supply capable of delivering about 100mA is needed. In my design, the regulated 5V supply also feeds the expansion socket for conveying power to the a-to-d conversion board. An 800mA on-board fuse is used.

### Programming

I have written a control program in Turbo Pascal 6 for this data logging system. Flow charts for the centronics i/o card and data logging system are shown in Figs 5 and 6. ■

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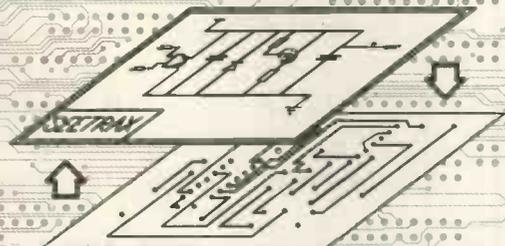
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In audio systems, there is usually more than just a length of cable between the power amplifier circuit and the loudspeaker; additional components are there because the amplifier designer has no control over the load that the amplifier will have to feed. They are intended to isolate load variations from the amplifier circuit so that it operates as intended for all reasonable loads.

Usually, a series resistor-capacitor network is across the amplifier's output terminals and an inductor of a few microhenries goes between the output of the circuit and the live speaker terminal. The RC network provides an effective low-impedance load at high frequencies where, for example, an electromag-

netic loudspeaker will have an impedance much larger than its nominal value. The inductor removes the effects of significant capacitive loads on the amplifier, such as may be encountered with long leads. The amplifier 'looks into' the inductor, whose impedance rises with frequency and buffers the capacitance loading at these frequencies. It sounds very simple, but analysis reveals that the circuit's operation is rather complex.

Why are amplifiers sensitive to reactive loads? Most employ some negative feedback to linearise their behaviour, the stability of the feedback loop being all-important. To ensure adequate stability margins for a multi-stage amplifier with a resistive load is difficult

enough, but to do it for unspecified reactive loads is virtually impossible. An inductive load may cause the open-loop gain of the amplifier to rise at high frequencies and prevent satisfactory stability margins being obtained. A capacitive load will introduce additional phase lag into the forward gain path and this may also lead to unsatisfactory stability margins.

Negative feedback has had a bad press over the last few years, some designers making a feature of using as little as possible. There is nothing wrong with the technique in itself, but its use without appreciating and avoiding its limitations can – and does – cause problems<sup>1</sup>. Most importantly, open-loop bandwidth of the

amplifier must cover the whole audio frequency range. Outside these frequencies the gain must be rolled-off in a controlled manner to ensure satisfactory gain and phase stability margins; behaviour made difficult to achieve by unspecified reactive loads. Since most designs employ dc-coupled feedback loops, serious stability problems are normally confined to high frequencies above 10kHz, extending to around 1MHz<sup>2,3</sup>.

**Series inductor**

To start, assume a load consisting of a resistive loudspeaker plus some shunt capacitance. Inductor *L*, shunt capacitance *C* and the load resistor *R* form the damped resonant circuit of Fig. 1. In practice, the capacitance is mainly due to the speaker cable, but a well designed amplifier should be stable with appreciably larger values than those expected from this source. This implies that the design has good stability margins and component tolerances between samples are unlikely to be a problem. Also, although having an amplifier remote from the speaker is not a good idea, it may be operated in this way and must not be unstable even with very long cables.

Equations in the diagram describe the impedance of the network. When the imaginary parts in the numerator and denominator of the right hand impedance expression are equal, it is resistive. The equation for the zero-phase frequency, *zpf*, shows it is real only for capacitance above a minimum value (*C<sub>min</sub>*); with *C<sub>min</sub>* in circuit, *zpf* tends to zero, the significance of *zpf* being that, below it, the impedance of the network is capacitive and, in spite of the inductor, will capacitively load the system. Above it, the impedance is inductive with leading phase.

Figure 2 is *zpf* shown plotted against capacitance for two values of inductor and a 8Ω resistor; also shown is the frequency calculated from the *LC* product. With large values of capacitance, circuit *Q* is high and the two frequencies are virtually identical. As the capacitance is reduced towards the minimum value, *Q* is lowered and *zpf* differs appreciably from the *LC* frequency. The larger inductor lowers the *zpf* for a given value of *C* and hence the frequency range where the amplifier experiences capacitive loading.

With a larger load resistor, *C<sub>min</sub>* is lower and the curves are shifted upwards and to the left.

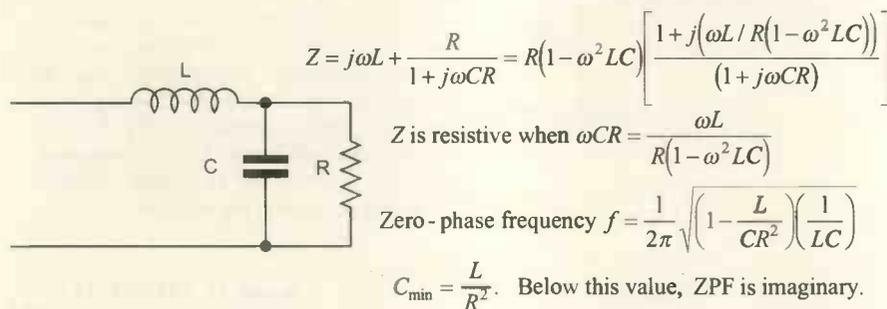


Fig. 1. Impedance and frequency for zero phase of the deceptively simple circuit with series *L*, cable capacitance and resistive speaker load. Above and below zero-phase frequency, impedance is inductive and capacitive respectively.

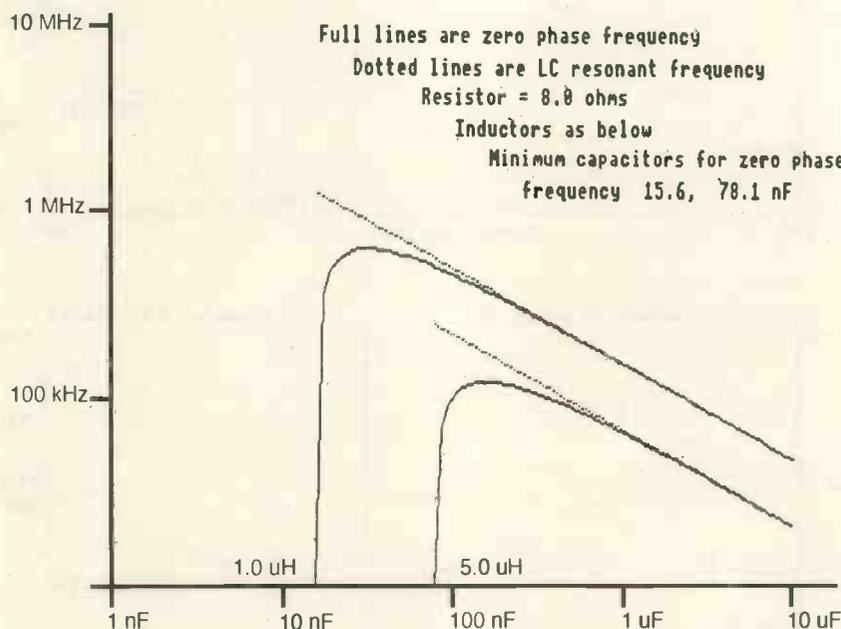


Fig. 2. Zero-phase frequency against *C*, with *L* of 1μH and 5μH and an 8Ω resistor in the circuit of Fig. 1.

**“Negative feedback has had a bad press over the last few years, some designers making a feature of using as little as possible. There is nothing wrong with the technique in itself, but its use without appreciating and avoiding its limitations can – and does – cause problems.”**

The *zpf* curve follows the *LC* line to lower values of *C* and reaches higher frequencies, as expected because circuit *Q* is increased. Before proceeding, the series *RC* network must be added across the input, 10Ω and 100nF being typical values. In the introduction I suggested that its purpose was to prevent the inductive speaker presenting a high impedance load at high frequencies, but now a physical inductance is connected to the amplifier output.

Figure 3 shows the circuit as so far described and two simulated responses of its input impedance for shunt capacitances of 200nF and 2μF, showing the lower *zpf* and the larger maximum phase lag occurring at a lower frequency for the larger *C*. The input *RC* network increases the maximum phase lag slightly but, more importantly, reduces the magnitude at higher frequencies. Without the network, the phase would tend towards +90°.

Connecting a load of 8Ω and 2.2μF in parallel is a common test to assess stability margins. It will almost certainly cause a square-wave output to ring severely on the transitions, but the ringing should die away quickly, indicating that the amplifier has adequate margins. Most of the ringing may be due to the transmission characteristic of the *LCR* circuit with the waveform at the input side of the inductor looking much better. In this case the additional phase lag at relatively low frequencies is not causing a problem.

However, with an appreciably lower capacitor, more representative of cable capacitance, things may not be so good. Maximum phase lag is less, but it occurs at higher frequencies; active devices in the amplifier are also causing appreciable lags and the circuit may not be able to tolerate the additional loading.

When looking at the loading impedance, one must consider both magnitude and phase. A large magnitude with phase tending to -90° represents a small capacitor which may have little effect. A small magnitude with a smaller

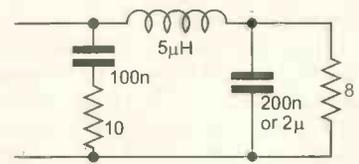
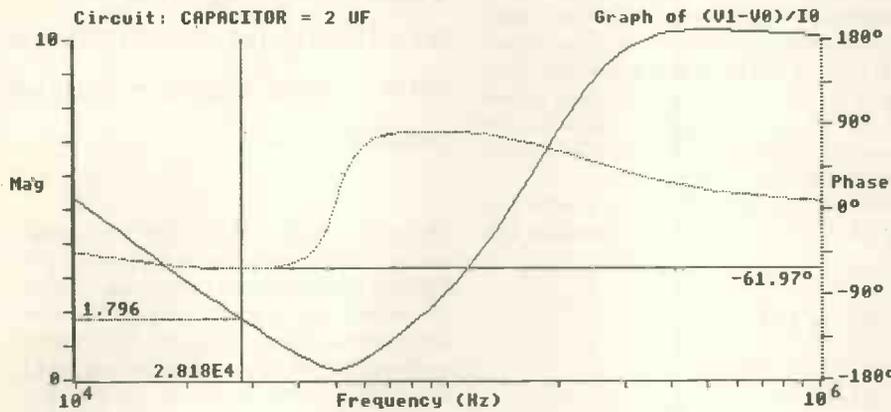
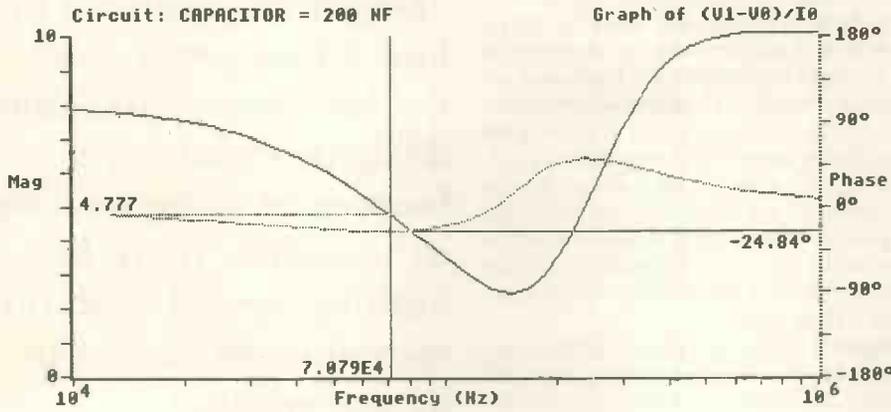


Fig. 3. Additional CR across input of circuit in Fig. 1 reduces impedance magnitude swing at high frequencies. Performance curves are shown on the left.

seeking great accuracy unless an amplifier is being designed for use with one specific speaker system; hardly a practical proposition.

Input impedance responses in Fig. 4 show the result of using the three-component compensation circuit with the 8Ω load replaced by the loudspeaker model. Note the wide variation in magnitude with the scales going up to 25Ω as opposed to 10Ω in Fig. 3.

Comparison of the 200nF responses in Figs 3 and 4 reveals how much the loudspeaker model has changed matters for the worse. Maximum phase lag is much greater than before and it occurs at a higher frequency; the magnitude at maximum phase lag is considerably reduced; just the effects we have been trying to avoid!

Figure 1 shows that, for 8Ω and 5μH, the value of  $C_{min}$  is 78.1nF. With the simulated speaker, capacitance values below  $C_{min}$  do result in an impedance with a lagging phase angle. The upper responses in Fig. 5 show what happens with just 50nF: a low magnitude with over 60° of lag at around 300kHz. Curves are not shown, but with only 10nF, the

phase angle may take an appreciably greater quadrature current component.

**Practical loudspeaker loads**

At audio frequencies, the impedance of loudspeakers varies widely from the nominal resistive value; however, in the context of this article, it is the impedance above 10kHz that is important. Measurements have been made on a number of 8Ω, two-unit systems and the results suggest that the impedance curves for such systems are not too dissimilar: the magnitude rises gradually from about the nominal value at 10kHz to above 100Ω at 1MHz.

Figure 4 is a reasonable model with typical component values shown. There is no point in

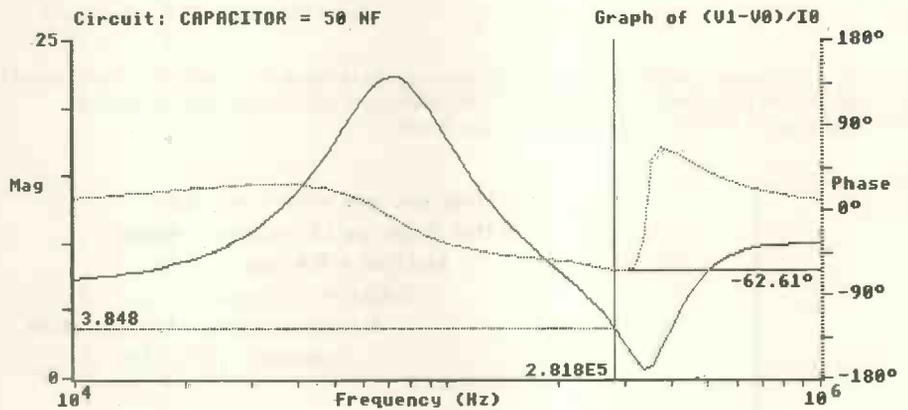
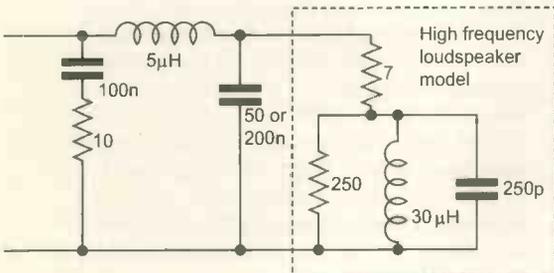
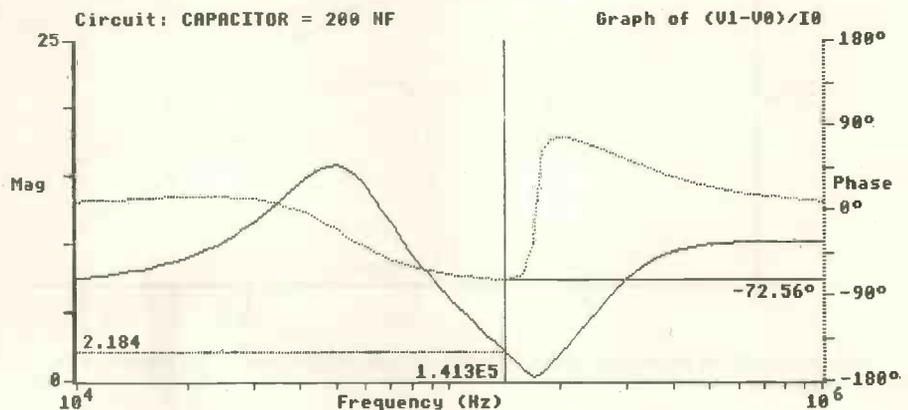


Fig. 4. Three-component circuit of Fig. 2, with a speaker instead of an 8Ω resistor. Both magnitude and phase swings are much worse, even though magnitude scale is now 25Ω. Performance curves are shown on the right.



lag approaches 45° over the range 100kHz-700kHz.

Inductive speaker impedance is again causing a problem and second series RC network must be added to ground from the load side of the inductor. Considering the other impedances in the circuit and the likely low output impedance of the amplifier, values of 10Ω and 500nF form a negligible shunt across the speaker at audio frequencies, and largely overcome the problem. Figure 5 is the complete circuit and simulated input-impedance plots for three shunt capacitor values, two the same as in Fig. 4; the reduced magnitude variation, on a 0 to 10Ω scale, and the smaller lag angles are clearly shown. Responses for the larger capacitors are not very different from those shown in Fig. 3.

Investigation of the five-component isolation circuit with the loudspeaker model replaced by a 8Ω resistor gives responses that are also not very different from those in Fig. 3. There is only a small phase lag for capacitors less than  $C_{min}$ . With larger values, the magnitude and phase at the maximum-phase-lag frequency are within 20% and 10° respectively, although this condition does occur at a slightly lower frequency. To make the test more stringent, a larger load resistor could be used to increase the lag angle somewhat.

A point that needs watching is the dissipation in the 10Ω resistor of the second RC network. A sinusoidal output of 100W at 1kHz into a 8Ω load will give less than 100mW in the 10Ω resistor but, at 10kHz, the power increases to about 8W. Under normal use with music signals, this will present no problem, but if high-frequency, high-power testing is attempted, a suitably rated resistor must be used. Similar considerations apply to the first RC network, but the smaller capacitor reduces the 10kHz dissipation to less than half a watt.

**Conclusions**

High-frequency impedance of real loudspeakers makes the three-component load isolation circuit with shunt CR and series L perform relatively ineffectively.

Connecting a second CR network in series across the output terminals to the speaker largely compensates for the speaker impedance, the resulting five-component isolation circuit presenting a more easily driven load to the amplifier. When loaded by a loudspeaker, it performs in a similar manner to the three-component circuit with a resistive load. Provided a range of capacitors are used with the five-component circuit, testing with a load resistor and capacitor in parallel is meaningful. It provides a loading not too far removed from what may be encountered in practice.

This article is intended as a general discussion of the effects of loading on amplifiers. Computer analysis and simulation cannot be performed in general terms, so component values have had to be introduced to illustrate the points discussed. I do not suggest that they are suitable for general application and it is left for designers to evaluate the component values most suited to their particular amplifier. ■

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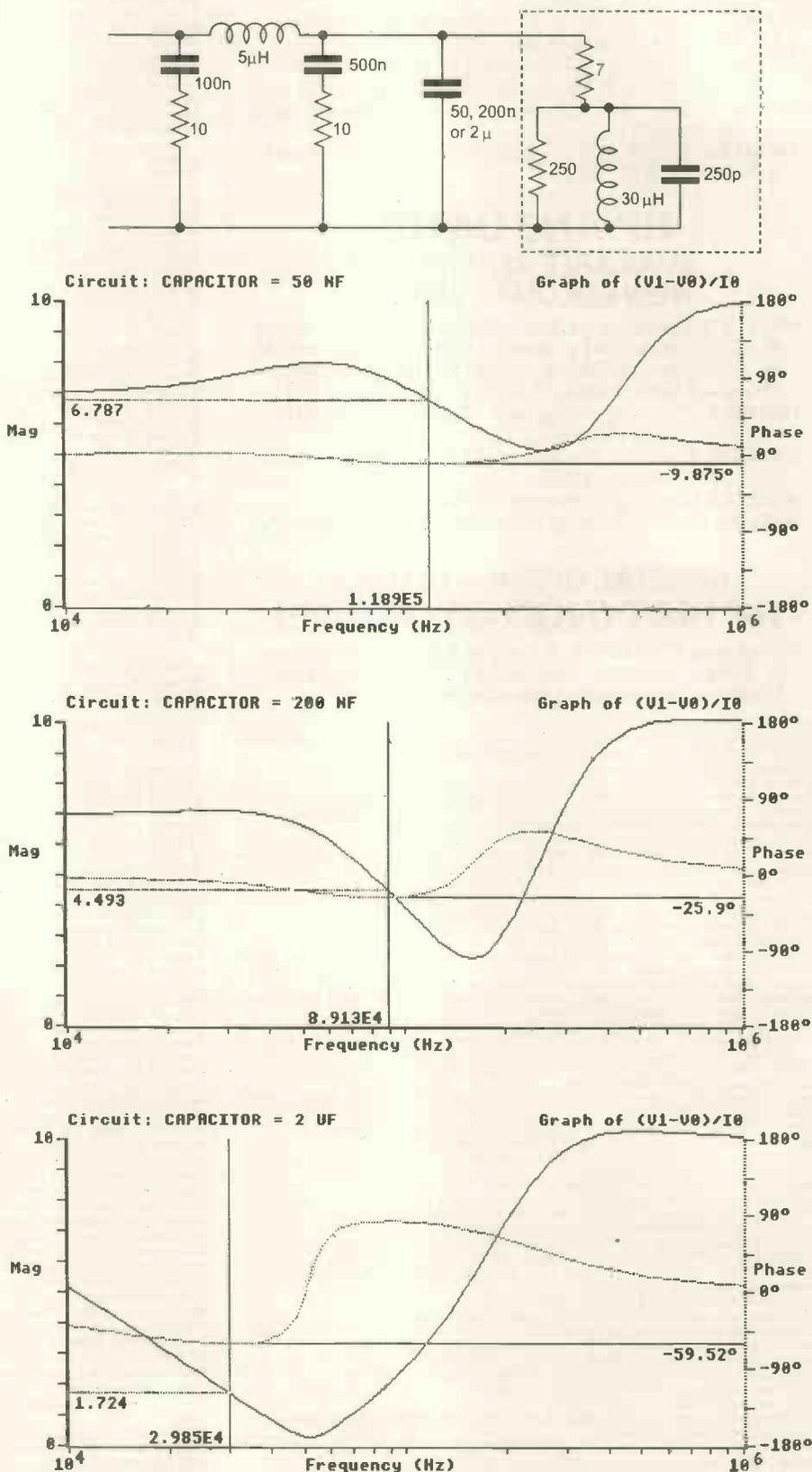


Fig. 5. Second RC after the L solves the problem and this is the complete circuit. Magnitude scales on graphs are back to 10Ω.

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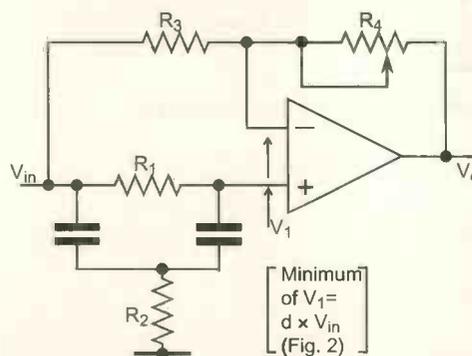
This filter is very simple: it uses a single op amp and provides a buffered output. The basic circuit, Fig. 1, consists of a minimal network, with a resonance dip  $d$  that is amplified to a complete null by the op amp. The filter is zeroed in two, preferably orthogonal, directions: gain and phase (frequency). These are separated: the gain adjustment is made with  $R_4$  and the phase is tuned by adjusting the network frequency. To maintain orthogonality both  $R_1$  and  $R_2$  have to be changed proportionally, assuming the two capacitors are equal.

**Selectivity**

Like the twin-T, this circuit has a low Q, Fig

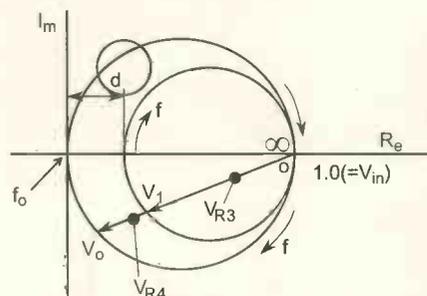
3, and ought to be improved by positive feedback. This involves feeding back part of the op-amp output to  $R_2$ , Fig. 4. At resonance the output is zero, as if  $R_2$  were grounded, which means that the balance condition at resonance is as it was without feedback. For all other frequencies the gain will be closer to one, resulting in a narrower null.

According to Thévenin, the feedback circuit  $R_2$ ,  $R_5$  and  $R_6$  can be replaced by only  $R_5$  and  $R_6$ , where  $R_5$  is slightly larger than the original  $R_2$ . In this way, power dissipation in the feedback network is minimised to practically zero. In spite of this, the equivalent voltage divider has zero impedance. A typical feedback factor is 85%, which gives a second harmonic atten-



$$\begin{aligned} \text{If } \sqrt{R_1 R_2} &= R_0 \\ C_1 &= C_2 = C_0 \\ 2\pi f_0 &= \frac{1}{R_0 C_0} \\ \frac{1}{d} &= 1 + \frac{R_1}{2R_2} \\ \frac{R_4}{R_3} &= \frac{d}{1-d} = 2 \frac{R_2}{R_1} \end{aligned}$$

Fig. 1. Basic circuit of the null filter. The resonance dip,  $d$ , is amplified to a complete null by the op-amp.



$$\begin{aligned} d &= \frac{2R_2}{2R_2 + R_1} \\ \frac{R_4}{R_3} &= \frac{d}{1-d} \end{aligned}$$

Fig. 2. Plotting  $V_1$ , a circle in complex notation, and multiplying by  $1+R_4/R_3$  in the amplifier gives a zero for  $V_0$  at frequency  $f_0$ .

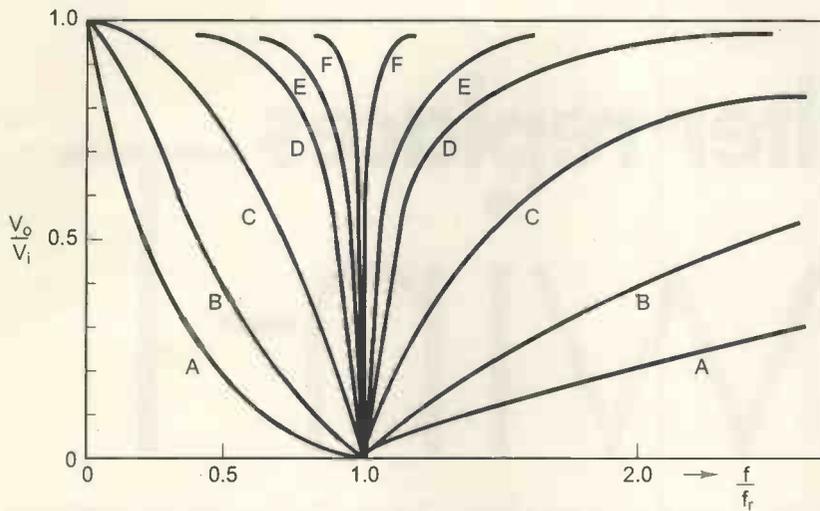


Fig. 3. Positive feedback improves Q of the null filter. These curves show measured frequency response with various amounts of feedback, K.

uation (amplitude error) in a thd-analyser of approximately 25%, or 15% for the third, etc). These harmonics can be corrected in a subsequent, slightly peaked, second order high-pass filter, with a cut-off at approximately  $1.8f_0$ , which will also be used to amplify the distortion products.

**Common-mode distortion**

At resonance, the low signal level *d* reduces the common-mode-distortion proportionally. Use of a low-distortion op-amp with sufficient slew-rate will not cause deterioration of the residual signal.

With this method, it is possible to make a simple analyser with a resolution of 10ppm, or better. Each plug-in circuit board may comprise one state-variable oscillator and one null-filter. Operating at frequencies of, say 1, 10 and 30kHz, such an analyser would be useful for audio amplifier testing.

Another interesting feature is that the signal input to the non-inverting terminal is equal to *d* multiplied by the signal amplitude. Selecting *d* at 0.1, the op amp can analyse a 200Vp-p sine-wave directly (sine-wave only!). It is therefore possible to measure the output of any size of power amplifier without needing an attenuator.

**Wide-band tuning**

Figure 5 shows how to provide orthogonal tuning in a wider range using a ganged potentiometer,  $R_5+R_7$ . Since the original  $R_2$  is equal to the parallel connection of  $R_5$  and  $R_6$ , an equivalent value for  $R_1$  is designed in the same way with  $R_7$  and  $R_8$  in parallel. Ratio  $R_1/R_2$  is constant when tuning, which means that  $R_4/R_3$  (gain) will be constant (see the second equation in Fig. 2).

For thd-measurements at one selected frequency,  $R_5$  and  $R_7$  can be low-resistance trimmers in series with larger metal-film resistors. These, and  $R_1$ , are set. When automatic zeroing is preferred, the ground connection of  $R_6$  is then used as the input for two quadrature sine-wave voltages.

**Summary**

The null-filter can be used in a wide range of applications, from a simple hum-suppressing filter to an advanced thd-meter with high resolution. It features low noise. Tests with a 1V input show that the noise is typically 3µV rms. One can clearly read 7µV thd, that is 0.0007% at 20kHz (7ppm). At higher voltages or lower frequencies the resolution is even better.

It is possible to tune the circuit manually using coarse and fine tuning potentiometers, but at lowest distortion level an automatic balancing circuit is preferred, Fig. 5. It can be applied to  $R_6$  via the output from an op amp summing the gain and phase error sine-waves.

For improved resolution, it is advantageous to connect an op-amp with 20-40dB gain directly to the output of the circuit. The op amp can be wired as a peaked high-pass second-order filter with gain, which compensates for the response of the null-filter at the second, third, etc, harmonics.

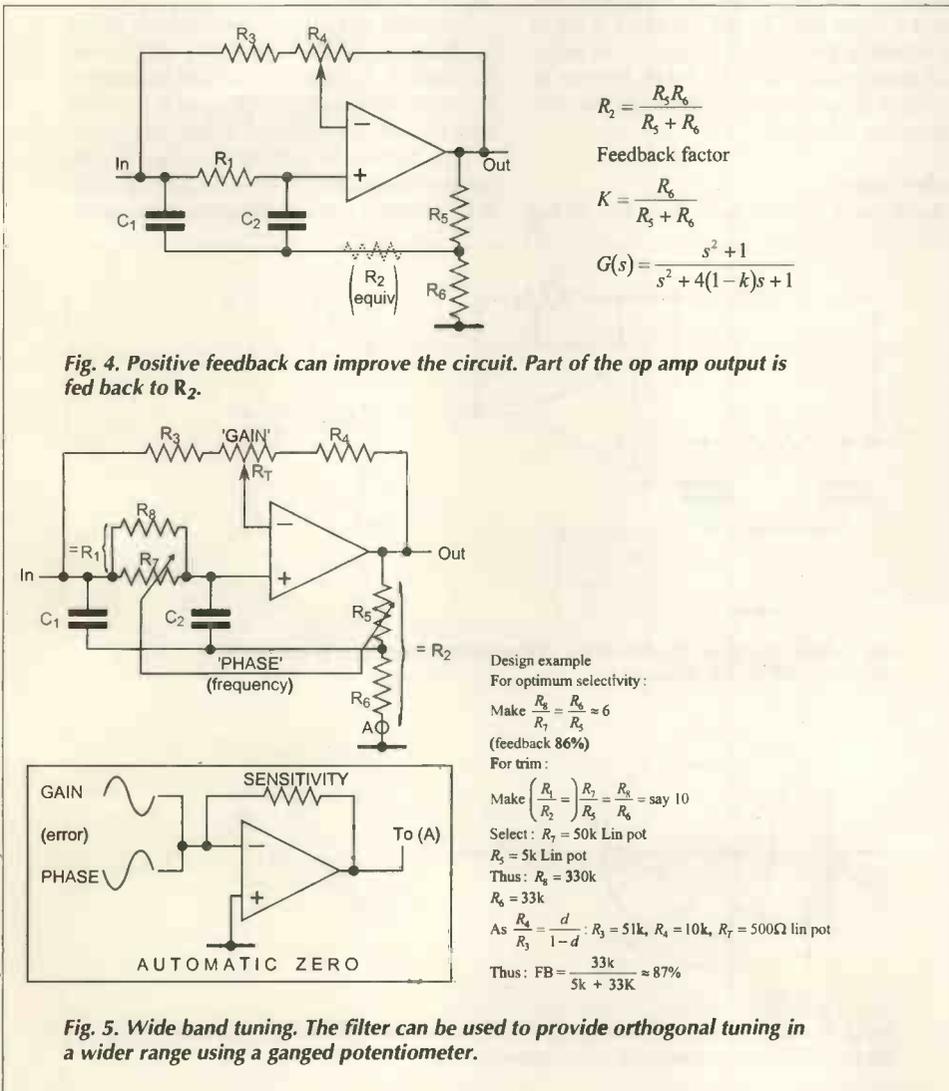


Fig. 5. Wide band tuning. The filter can be used to provide orthogonal tuning in a wider range using a ganged potentiometer.

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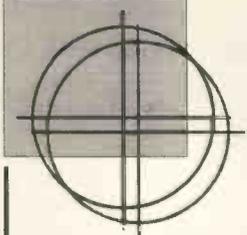
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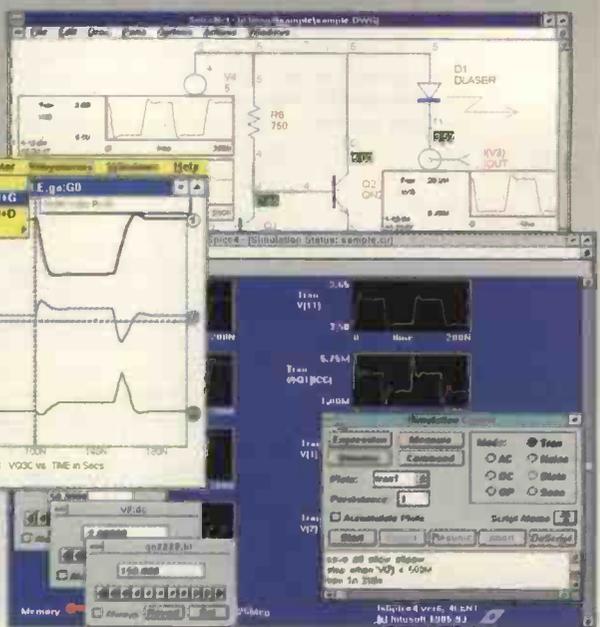
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# Tweaking the diode detector

Important applications for rf detectors of wide dynamic range include the various schemes for linearising radio transmitters. Simple diode detectors can be used, but suffer from a limited dynamic range and – a point which can be important – although amplitude response in the large signal range is linear, it is *not proportional to the signal amplitude*. This is a result of the diode's forward voltage,  $V_f$ .

## Infinite-impedance detector

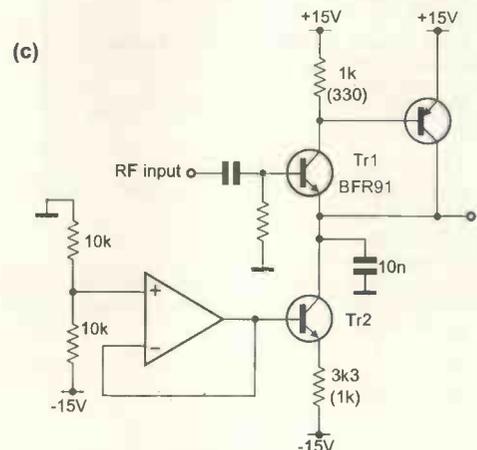
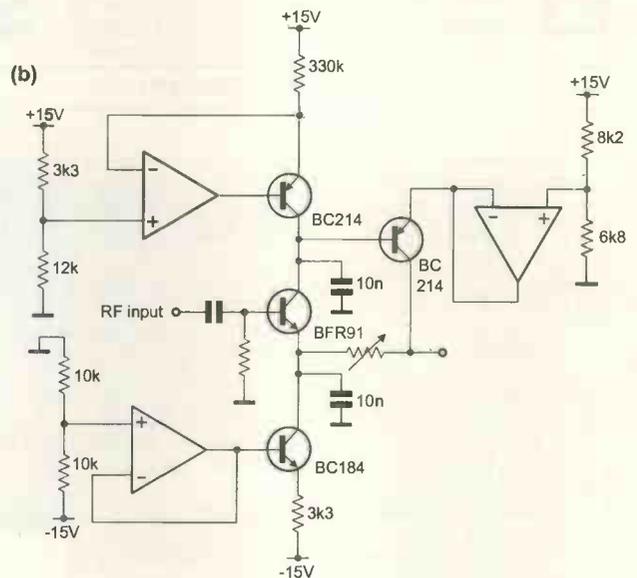
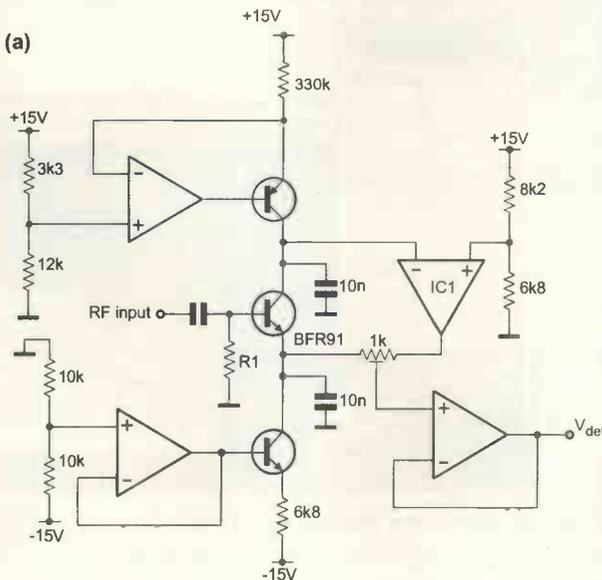
Since my interest while I was making this study was in level measurement rather than just detection of the

presence of a signal, my thoughts turned to the infinite impedance detector<sup>1</sup>. In this, the detector is a transistor or, in earlier times, a valve. It is in its active range at all times, so one might think that the troublesome curve at the foot of the characteristic in Fig. A(d) could be avoided. (See panel on page 123)

My plan was to surround the infinite impedance detector with a high-gain servo loop which would jack up the dc voltage at the emitter of the detector to restore the no-signal dc conditions. The transistor could be a type with a very high frequency cut-off  $f_t$ , and both the emitter and the collector could be bypassed at rf, confining rf solely to the

After working through a number of variations on the infinite-impedance detector, Ian Hickman returns to the diode – but finds a wide dynamic range and a way round forward-voltage offset.

Fig. 1(a). First try at infinite impedance detector. It is embedded within a servo loop maintaining constant collector current. Variant at (b) has a lower (more manageable) amount of loop gain. Circuit at (c) uses both transistors in the servo loop operate at rf.



## Diode performance

Figure A(a) shows an ideal diode which has an infinite slope resistance when the voltage at its anode is negative with respect to the cathode, and a slope resistance of zero when forward biased. Such diodes don't exist, but they can be closely approximated, over a limited frequency range, by a combination of one or more real diodes and op-amps. Figure A(b) shows a slightly less unrealistic representation of a real diode: the slope resistance when forward biased is still zero, but an anode voltage positive by an amount  $V_f$  relative to the cathode must be applied to cause current to flow.

Figure A(c) is one step nearer reality, showing as it does a finite diode slope resistance when forward biased. Unfortunately, the resistance does not drop instantly from infinity at forward voltages below  $V_f$  to a low constant value at  $V_f$  and above, but makes the transition gradually, as the curve at the bottom of the characteristic in Fig. A(d) shows. Projecting the characteristic in the forward biased regime back until it cuts the voltage axis gives a value which may be taken approximately as the diode's  $V_f$ , the voltage which must be applied before some arbitrary small current flows.

When a practical Fig. A(d) diode is used as an amplitude detector, the detected dc voltage output increases with rf input up to the point where the peak reverse voltage equals  $V_{br}$ , the diode reverse breakdown voltage. With decreasing rf input, the detected voltage falls linearly at first, then reaching a point where it falls faster than the input, since the latter no longer comfortably exceeds  $V_f$ . This is the square-law region, where the detected voltage is proportional

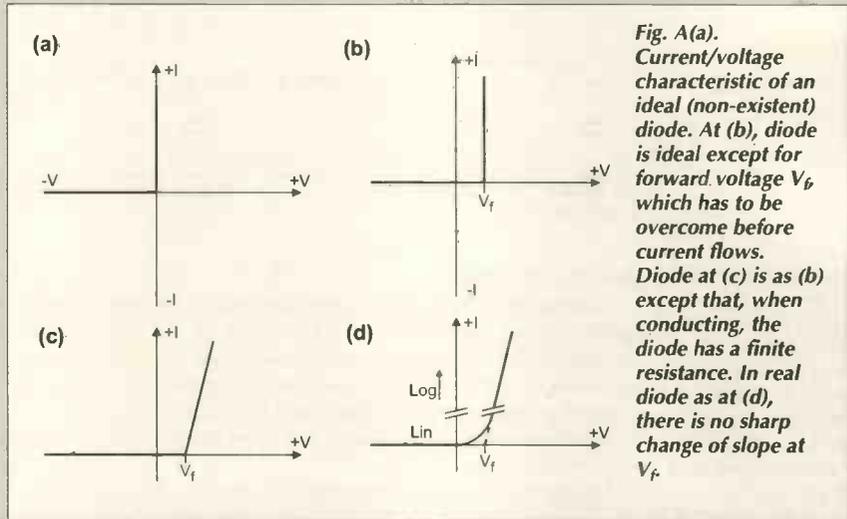


Fig. A(a). Current/voltage characteristic of an ideal (non-existent) diode. At (b), diode is ideal except for forward voltage  $V_f$ , which has to be overcome before current flows. Diode at (c) is as (b) except that, when conducting, the diode has a finite resistance. In real diode as at (d), there is no sharp change of slope at  $V_f$ .

to the applied power rather than to the applied voltage. Where it is simply desired to detect the presence of a signal, rather than accurately to measure its amplitude, the diode can still be successfully used in the square-law region down to the level where the detected voltage starts to disappear into the circuit noise.

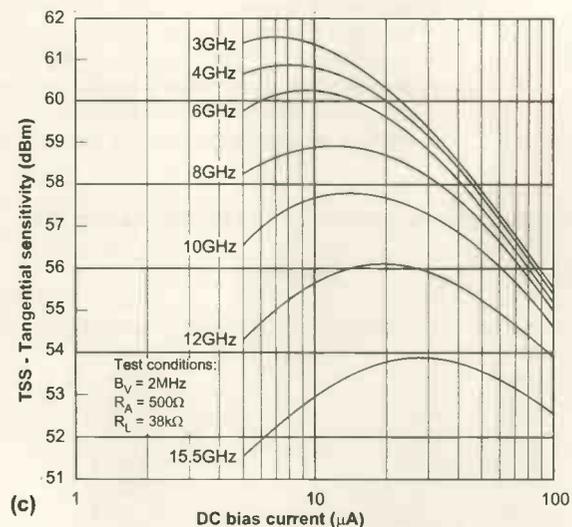
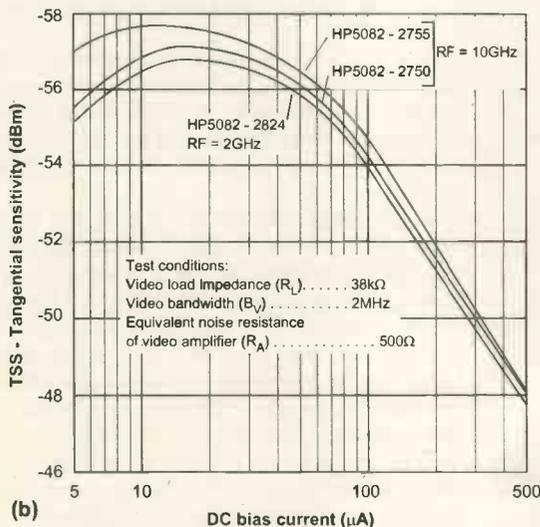
This level of rf input to the detector is known as the tangential sensitivity, and can be determined by displaying the detected voltage on an oscilloscope whilst driving the diode with an rf signal 100% on/off modulated by a square wave. Tangential sensitivity is the rf level at which the level of the top of the 'grass' in the off periods just coincides with the level of the bottom of the grass in the on periods, as in the middle illustration in Fig. B(a). Results obtained depend to some extent on the oscilloscope

intensity setting and upon the operator but, being a quick, simple and easy test, it is widely employed.

Tangential sensitivity can be improved by applying a small forward bias current to the detector diode as shown in Fig. B(b) for various diodes. Optimum bias and the improvement that results depend upon frequency. This is illustrated in Fig. B(c) for a particular diode, where you can see that the tangential sensitivity can exceed -60dBm. This means that the sensitivity of a simple diode video receiver is only 35 to 40dB less than that of a superheterodyne receiver. Not only is the diode video receiver much cheaper, simpler and easier to maintain, but it can easily be designed to offer a greater rf bandwidth than a superhet, and it is useful in a variety of applications, both military and civil.



Fig. B(a). Tangential sensitivity of a diode, taken when output is as in middle case. Tangential sensitivity can be improved by applying a small forward bias current to the detector diode, as shown at (b) for various diodes. Optimum value of bias and the improvement that results depends upon frequency, as illustrated at (c) for a particular diode. ((b) and (c) reproduced by courtesy of Hewlett-Packard Ltd, from H-P Application Note 923.)



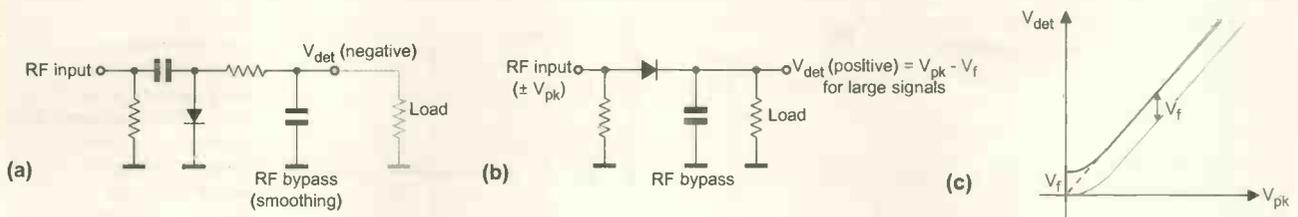


Fig. 2(a). The shunt diode detector needs a dc path to ground at its output to enable it to follow a decreasing rf input level. It produces the same dc detected output for any input level as circuit at (b), the conventional peak detector circuit. By adding  $V_f$  to the detected output, as at (c), it becomes proportional to the signal level, except in the low-level square-law area.

base/emitter circuit.

Identical dc conditions, regardless of whether rf was present or not, could be ensured by the use of constant-current generators in the emitter and collector circuits. When two near-perfect constant current generators – one a source and the other a sink – fight each other, their junction is a point of very high voltage gain.

Figure 1(a) shows the resulting pipe-dream, where the op-amp supplies the difference between emitter's constant tail current  $I_e$  and the constant collector current  $I_c$ . Total gain within the loop includes a contribution equal to the slope resistance of the collector circuit divided by ( $R_1$  plus  $r_e$ ), where  $r_e$  is the common-base input resistance of the transistor. Some quick mental arithmetic showed this to be so great that the roll-off of loop gain had to start at such a low frequency that the circuit's response to a change of input level was inordinately slow. So the op-amp was replaced by a p-n-p transistor,  $IC_1$  being redeployed to provide a voltage source for its emitter, as in Fig. 1(b).

While this circuit worked, it represented little advance in low-level sensitivity; in fact, better results were obtained by removing the collector decoupling and closing the loop around  $Tr_1$  and  $Tr_2$  at rf as in Fig. 2(c), which is similar to a circuit which appeared in Ref. 1. This offered a useful improvement in low-level sensitivity compared to a simple diode detector. However, the impression was that all these circuit variations were basically more or less complicated ways of extracting such dc output as was obtainable from a basic diode detector circuit.

**Back to the diode**

My attention was redirected to the shunt diode detector circuit of Fig. 2(a), which conveniently has one end of the diode grounded. The diode dc restores the input rf negative-going with respect to ground, the smoothing circuit then picking out the mean level of the waveform at the diode's anode. Detected dc is thus equal to the peak value of the rf, or would be with an ideal diode, where  $V_f$  is zero. The dc detected

output of this circuit versus input rf level is the same as the usual diode peak detector circuit of Fig. 2(b).

At high levels, the curve is linear, but does not pass through the origin when projected backwards; the detected dc output is not proportional to the rf level. It could be made so by adding a constant offset, equal to  $V_f$  to the detector's output, which would simply have the effect of raising the whole curve by  $V_f$ , as in Fig. 2(c).

Although the output is now – at least at higher input levels – not only linear but also proportional to the input, there is a standing dc output  $V_f$  when no rf is applied. This is an unfortunate state of affairs since  $V_f$  is temperature dependent. But with a little lateral thinking, it is possible to add the  $V_f$  offset to the detected output in such a way that, as the detected output falls to zero, so does the added offset.

After a few iterations, a circuit designed to do just this finished up with the comparatively simple arrangement of Fig. 3(a). It was built on copper strip-board, except for diode  $D_1$ , with  $R_1$ ,  $R_2$ ,  $C_1$  and  $C_2$ , which were all mounted on the back of a bnc socket, the body of which was soldered to a sheet of copper-clad board. To increase the sensitivity of detector diode  $D_1$ , a small amount of forward bias is applied via a 10M $\Omega$  resistor  $R_1$ , and the resultant offset balanced out by the corresponding drop across  $D_2$ . The 2.2M $\Omega$  resistor is returned to the positive or negative 15V rail as necessary, depending on whether  $D_1$  or  $D_2$  exhibits the larger forward voltage at the current defined by  $R_1$  and  $R_6$ .

These diode voltages are, of course, temperature dependent but, provided they track, they represent purely a common mode signal and  $IC_{1a}$  is so connected as to reject any common mode input. Resistor  $R_2$  with  $C_2$  provides the rf smoothing to extract the detected dc component, while  $R_2$  plus  $R_3$  matches the value of the three other bridge resistors  $R_7$ ,  $R_8$  and  $R_9$ .

Op-amp  $IC_{1a}$  exhibits an inverting gain of unity to detected output of  $D_1$ , and its positive going output is applied to inverting amplifier  $IC_{1b}$ , whose gain can be set to a gain of unity, or more or less as required. With the wiper of  $R_{16}$  set to ground, the detector law is exactly as in Fig. 3(a).

Positive-going output of  $IC_{1a}$  also goes to inverting amplifier  $IC_{1c}$ . With  $R_5$  correctly set up, the outputs of  $IC_{1b}$  and  $IC_{1c}$  will both be zero in the absence of any rf input, at least assuming all the op-amps ideal. With an increasing level of rf applied, the output of  $IC_{1a}$  is initially small since it operates in the square-law region, but the output of  $IC_{1c}$  will rise much faster as it operates at a gain of 40dB, defined by  $R_{14}/R_{13}$ . However, with increasing detector output  $D_3$  starts to conduct, progressively reducing the gain of  $IC_{1c}$ . Ultimately, the gain of  $IC_{1c}$  falls below unity, indeed almost to zero; in fact, the ratio of the forward resistance of  $D_3$  in the milliamps range (about 20 $\Omega$ ) to

Table 1. Measured performance of test set-up Fig. 4(b).

Input dBm	Output				
	mVpk	Ideal normalised	mVpk	Actual normalised	Error %
50	3180	1.0	3180	1.0	-
+20	1005	0.316	946	0.297	-6
+10	318	0.1	299	0.094	-6
0	101	0.0316	85	0.027	-15
-10	31.8	0.01	31	0.0097	-3
-20	10.1	0.0032	10	0.0031	-1.9
-30	3.2	0.0001	1.3	0.00041	-59

the value of  $R_{13}$ , or roughly  $-40\text{dB}$ .

By advancing the wiper of  $R_{16}$  from ground, a proportion of this voltage can be injected into the non-inverting input of  $IC_{1b}$ . Note that to an input at its non-inverting input, the gain of  $IC_{1b}$  is around  $6\text{dB}$ , due to  $R_{10}$  and  $R_{11}+R_{12}$ . This permits a useful degree of dynamic range extension, by ensuring that the output of the whole circuit starts to rise appreciably at a much lower rf input level than would otherwise be the case.

**Test results**

I tested circuit operation at rf levels of  $+20\text{dBm}$  downwards using the set-up in Fig. 3(b), and the results are recorded in Table 1. Originally, I planned to use the video oscillator's maximum output frequency of  $10\text{MHz}$ , but the spectrum analyser showed that there was significant second harmonic distortion at this frequency – bad enough to be clearly visible even on the oscilloscope. I therefore carried out the tests at  $5\text{MHz}$ , at which frequency the second harmonic was over  $35\text{dB}$  down and all other harmonics much lower still. Harmonic distortion – especially even-order distortion – is an important consideration here, as the diode detectors of Figs 3(a) and 4(b) sense the peak of the rf waveform, although the indicated result is conventionally presented as the rms value of the signal. This is assumed to be a pure sinewave.

Generator output at  $5\text{MHz}$  was  $+20\text{dBm}$  using the indication on the spectrum analyser, after setting the gain of the latter using its internal calibration output. As a cross check, I used a newly calibrated oscilloscope to measure the peak-to-peak voltage at the output of the  $50\Omega$  through-load, i.e. at the input to the detector circuit. The result was  $3.2\text{V}$  pk-pk, perhaps a shade under. This agrees with the expected value of  $3.18\text{V}$  pk-pk for  $+20\text{dB}$  relative to  $1\text{mW}$  in  $50\Omega$ .

Circuit setup to take the results in Table 1 was as follows. With the wiper of  $R_{16}$  at ground and  $R_{12}$  set to mid travel,  $R_5$  was adjusted for zero reading on a digital voltmeter. To check that there was negligible hum pick-up, in view of the high circuit gain and the division of the circuit between two different boards, the oscilloscope was also connected to the output of  $IC_{1c}$ . A  $5\text{MHz}$   $+20\text{dBm}$  input was then applied to the  $50\Omega$  through-termination and  $R_{12}$  adjusted to give the expected theoretical output of  $3180\text{mV}$ . Input level was then reduced to  $-20\text{dBm}$  and  $R_{16}$  adjusted to give an output of  $31.8\text{mV}$ .

Results were not repeatable, due to the need for frequent resetting of  $R_5$ . As a result I relocated  $D_2$  from the circuit board to a position adjacent to  $D_1$ , well decoupled to prevent it picking up any rf. The rear of the BNC socket, with the two diodes and related components, was then enclosed in a box, shielding the diodes from both draughts and light. This largely cured the drift problem, although a more modern quad op-amp with improved dc characteristics, in place of the  $TL084$  used, would be even better.

Following final adjustment of  $R_5$ , I repeated the  $+20\text{dBm}$  and  $-20\text{dBm}$  adjustments, iterating them alternately until no further adjustment was needed. Later, I took a set of measurements and recorded the results shown in Table 1. Plotting them on log/log

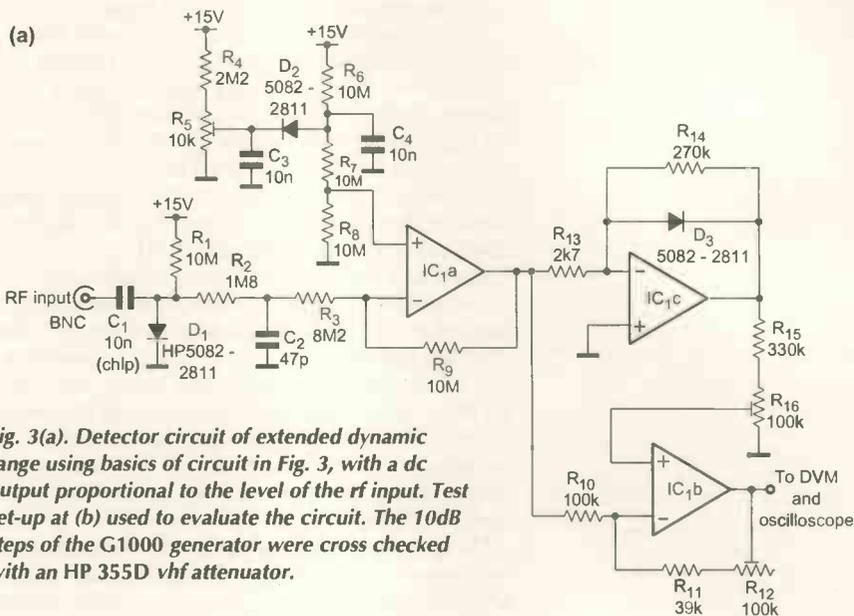


Fig. 3(a). Detector circuit of extended dynamic range using basics of circuit in Fig. 3, with a dc output proportional to the level of the rf input. Test set-up at (b) used to evaluate the circuit. The  $10\text{dB}$  steps of the G1000 generator were cross checked with an HP 355D vhf attenuator.

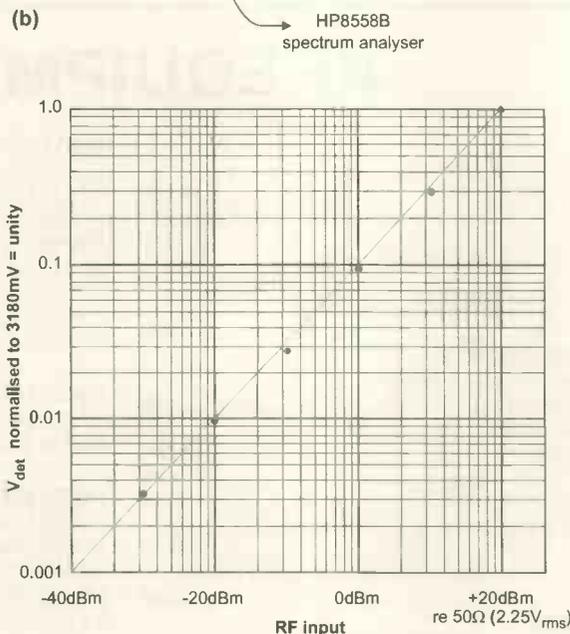
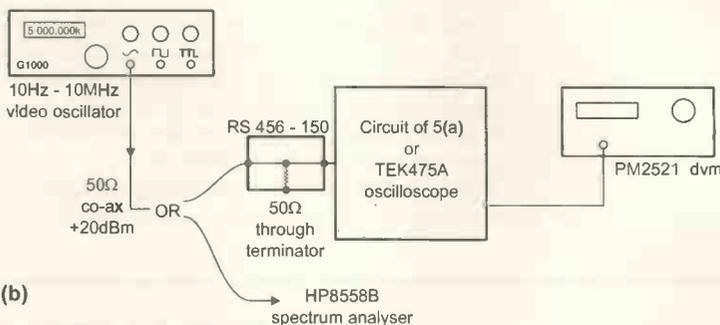


Fig. 4. Results from Table 1 over a  $50\text{dB}$  range, plotted on log log graph paper. Vertical axis is actual detected dc output voltage and the horizontal indicates peak input voltage, measured points being indicated on the graph. Sloping line at  $45^\circ$  indicates output that would be provided by perfect detector. For convenience in plotting, both peak input voltage and dc voltage have been normalised to  $3.18\text{V}$  ( $2.25\text{V}_{rms}$ ) at unity.

graph paper over a 50dB range gave Fig.4.

Results obtained to date are, as can be seen, very encouraging, though certain aspects await further investigation, notably the large negative error at around -10dBm. The three diodes were *HP5082 2811* schottky barrier types intended for general purpose applications. They were not a matched set and were used simply because a number of them were in stock, although they are available as unconnected matched quads under the type number *5082 2815*.

How much improvement using matched diodes would achieve is unknown, as are a number of other possibilities. These include setting up at, say, +20 and -10dBm, instead of +20 and -20dBm, or choosing a different value for the forward bias applied to  $D_1$  and  $D_2$ . Other parameters available are varying the ratio of  $R_{14}$  to  $R_{13}$ , and also their absolute values, and the inclusion of resistance in series with  $D_3$ . These might help to reduce the effect of a mismatch between two separate characteristics.

The detector law is a square law at low levels, while the compensation diode law, determining the voltage across  $D_3$ , is basically a log law relative to the voltage applied to  $R_{13}$ .

Where maximum sensitivity is required from the circuit, one of the low  $1/f$  (flicker) noise diodes from the *5082 2xxx* series might be a better choice. For example, the 30V  $V_{br}$  of the *5082 2301* would allow the detector to accept inputs up to almost +34dBm against only +27dBm for the *5082 2811*.

For the widest possible dynamic range, the *5082*

*2800 (1N5711)* with its  $V_{br}$  of 70V would accept an input of up to +40dBm. Assuming that it too would work down to -30dBm in the circuit of Fig. 4(a), it would provide a detector with a dynamic range of over 70dB.

Detectors of wide dynamic range can be used in the linearisation of transmitters using a non-constant envelope type of modulation, e.g. am, dsb-sc and all varieties of ssb from compatible am, through pilot carrier types to pure J3E. A negative-feedback loop from the detected audio recovered from a coupler at the transmitter's output, feeding back to the audio input to the exciter, can suppress third-order intermodulation products in a solid state transmitter to 60dB below peak envelope power<sup>2</sup>, provided the transmitter is designed to hold am to pm conversion in all stages to a low level. This can provide a much cheaper solution than complex arrangements such as a polar loop or cartesian loop scheme. ■

References

1. Hickman I. *Measuring detectors*, EW + WW, November and December, 1991.
2. UK Patent 2209639A, Single Sideband Transmitters, 1991.

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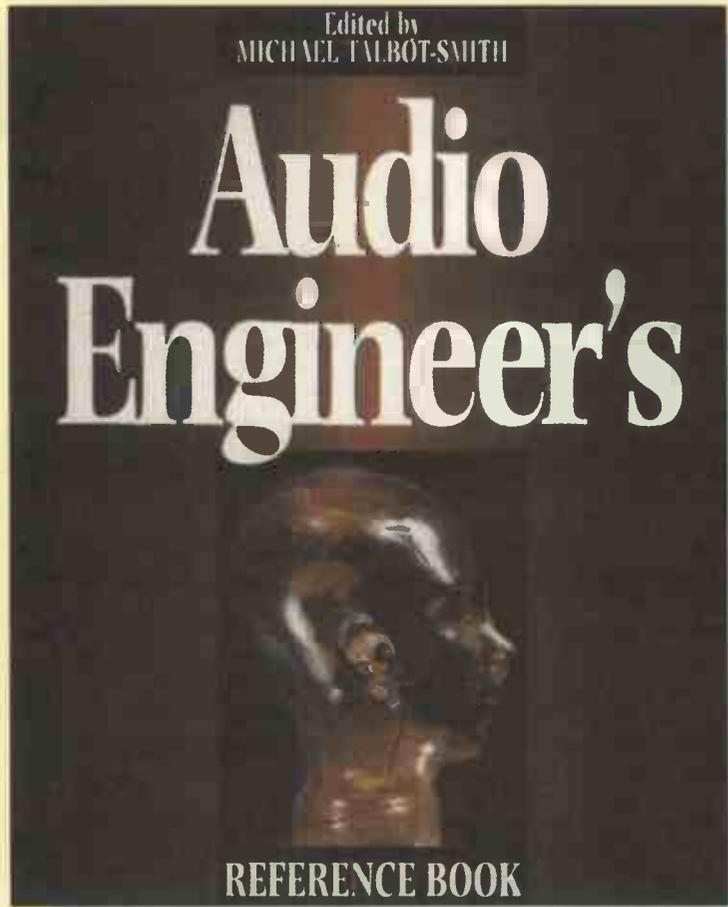
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## Jump-starting a rom-less microcontroller

Configuring a test clip as shown allows the use of a single sram for both program and data storage by providing a means of loading program code into ram.

With the test clip in place over the ram chip, data reads come from the eeprom and writes go to ram and eeprom. The following procedure prevents bus contention.

The eeprom contains a copy of the required ram data and has a routine at the start which writes 01 to a location 00 in the eeprom, then reading it back. If 01 is read, the application code starts, since it now runs from ram without the clip. If not, ram copies the eeprom by reading each byte and writing that value back and then stops until the clip is removed and the controller reset. After this procedure, new code can be loaded via the serial port, this code being held in the internal controller ram to avoid overwriting itself. The clip is only then needed in the event of code corruption or a crash.

The system shown uses a Mitsubishi 50747 with the controller and ram permanently powered to avoid the need for ram power-down protection; the

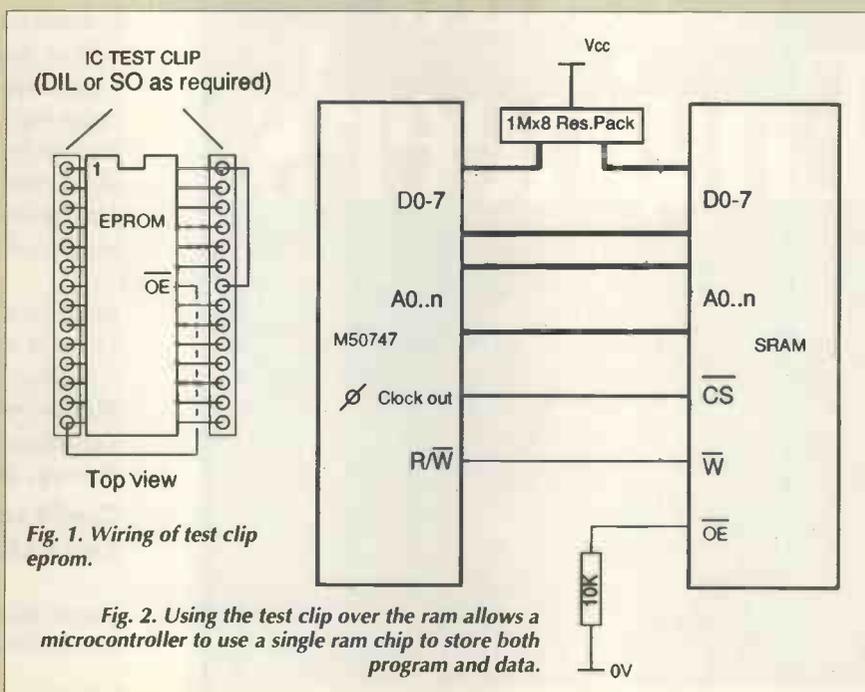


Fig. 1. Wiring of test clip eeprom.

Fig. 2. Using the test clip over the ram allows a microcontroller to use a single ram chip to store both program and data.

controller can use sleep mode. Do not leave i/o pins floating, as that increases current drain. The  $1M\Omega$  resistor pack stops the data bus floating.

**Mike Harrison**  
White Wings Logic  
Loughton  
Essex

## Quadrature-output oscillator for sine waves

As an alternative to the use of rom look-up tables to generate stored sinusoids, Szymanski<sup>1</sup> described a circuit using steep-cut switched-capacitor filters. In the version shown here, the outputs are two sinusoids at  $90^\circ$  to each other.

Clock signals go straight to the National Semiconductor MF6-100 switched-capacitor filter, which is a Butterworth low-pass type. After division by 40 in the 4017 decade counter and two stages of the 4024 binary counter, the  $f/40$  clock drives two halves of a dual JK bistable device, being inverted to one of them. Inputs to the MF6s are now square waves at  $f/80$ , separated in phase by  $90^\circ$ , and the outputs of the low-pass filters

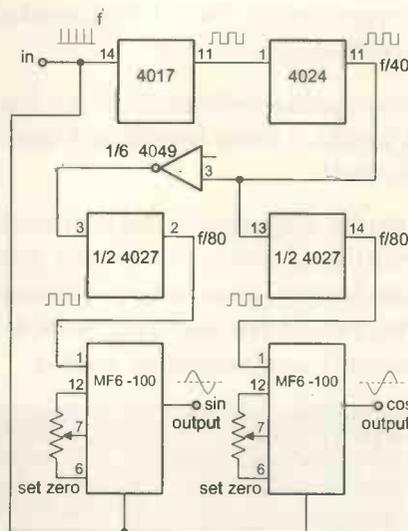
are sinusoids.

Some clock signal appears at the output, but is easily removed by a simple RC section without introducing undue phase shift. Improvement can be obtained by increasing the clock division ratio, while ensuring that the last counter stages divide by two.

**Harold W Shipton**  
St Louis  
USA

1. EW+WW, July 1994.

Two divider chains, driving switched-capacitor filters, produce sinusoidal outputs at  $90^\circ$  over three decades of  $f_{in}$ .



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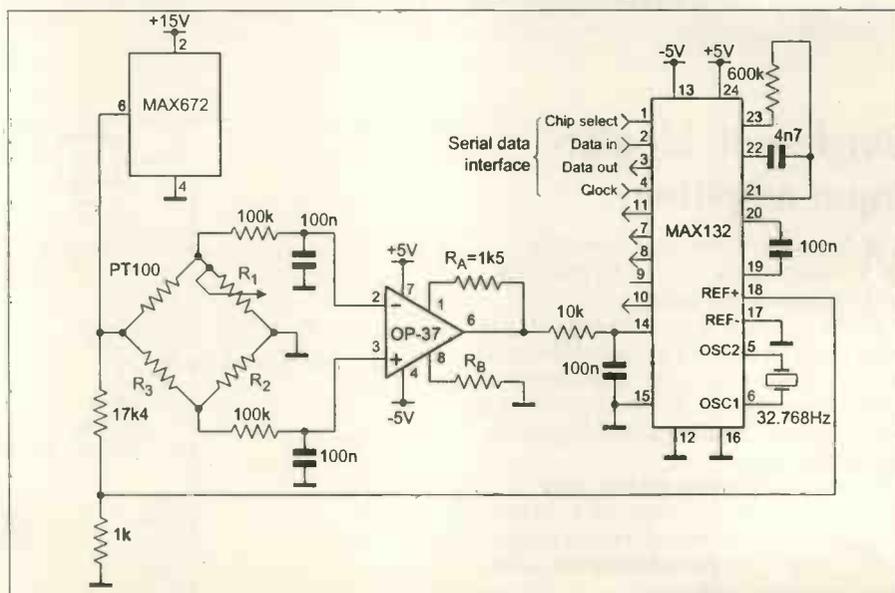


## Digital thermometer uses low-noise bridge amplifier

To avoid the problem of conventional feedback impairing the low-noise characteristic of an instrumentation amplifier, connect the feedback to the amplifier's trimming pins. The OP-37 differential amplifier in the circuit shown provides a gain of  $3 \times 10^3 / R_B$ , resistance  $R_B$  being in kilohms, with no effect on noise level. In this case, bridge output is 30mV and the gain to drive the MAX132 a-to-d converter is 17;  $R_B$  is therefore 180 $\Omega$ .

**K Kraus**  
Rokycany  
Czech Republic

*Avoid ruining the noise performance of a differential bridge amplifier by taking the feedback to the trim pins. Circuit shown is a digital thermometer, but the principle applies for any bridge amplifier.*

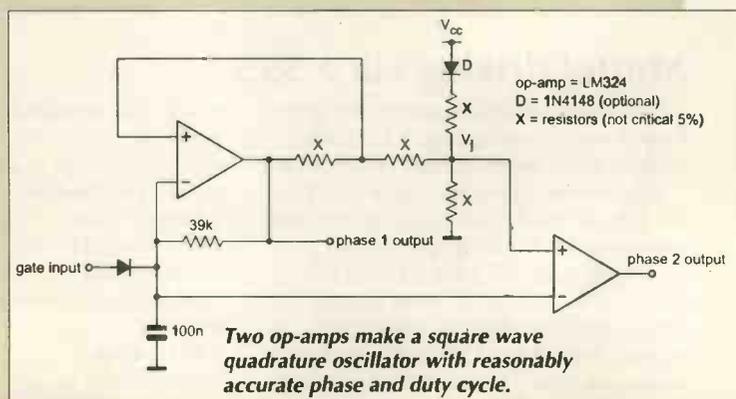


## Quadrature oscillator produces square output

Originally intended to drive a stepping motor, this RC oscillator produces 90Hz square waves with a 90° phase difference and 1:1 mark:space ratio.

Start operation by taking the gate input low. Diode D compensates for op-amp asymmetry, since the op-amp has the same 5V supply, the mark/space ratio and phase difference being accurate to within about 5%. Frequency is around  $0.36/CR$  and is hardly affected by variations in  $V_{cc}$ , since all thresholds are proportional to the supply voltage. Voltage  $V_I$  varies during the cycle and compensates for the exponential RC charging curve.

**W Gray**  
Farnborough,  
Hampshire



## Active photodiode source

Diode current to a photodiode receiving pulsed infrared signal commonly comes by way of a resistor, as shown in Fig. 1. Drawbacks to this simple arrangement are that the resistor shunts signal current, injects thermal noise and may set too low a limit on the current for high light levels.

One solution is to use a gyrator, but it is simpler to use the n-channel fet in Fig. 2. Capacitive coupling between source and gate gives the required high ac series impedance and the direct connection between the gate and amplifier input provides bias at any diode current up to  $I_{ds}$ .

Corner frequency of the resulting high-pass function is  $g_m/2\pi C$ , the best hf response being obtained when the load is a low impedance, such as the virtual earth of the CA3140 fet-input inverting amplifier. Transistor  $Tr_1$  can be a general-purpose n-channel jfet such as a 2N3819, selected for adequate  $I_{dss}$  and maximum pinch-off

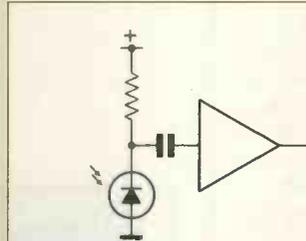


Fig. 1. Simple resistive bias to a photodiode may limit diode current at high light levels and inject noise.

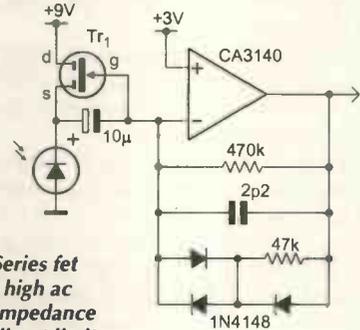


Fig. 2. Series fet gives a high ac series impedance and will not limit current.

much less than  $V_{dg}$ . Two diodes in the feedback path provide progressive limiting for large signals.

A New Fishponds Bristol

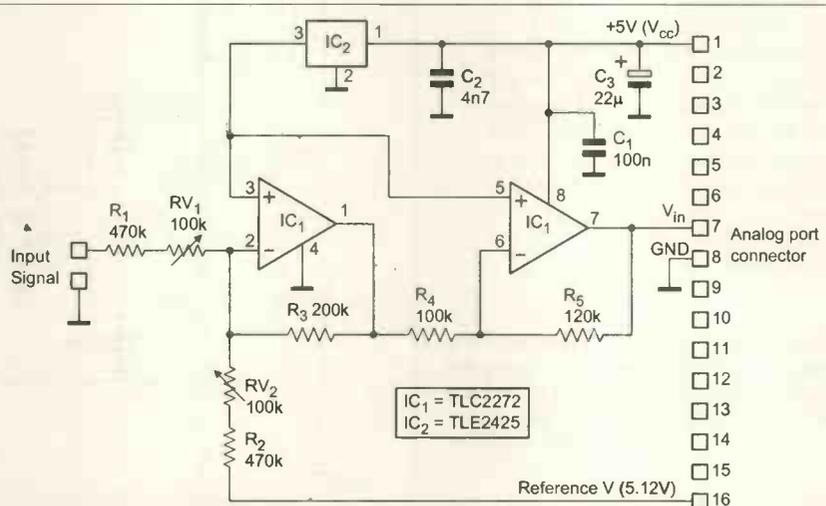
## Single-rail, bipolar-input amplifier

Without much trouble, an amplifier supplied from only a +5V rail can be made to process  $\pm 5V$  inputs.

A 2.5V reference derived from the 5V rail often found on interface connectors biases the non-inverting inputs of the op-amps. Fixed current through  $R_2$  and  $RV_2$  to the first stage inverting input should be supplied via a stable 5V supply.

To avoid saturation at the first-stage output, gain of the first stage is 0.4, set by  $RV_1$ , and 1.2 for the second. Output voltage swings from  $V_{cc}$  to ground, using the entire a-to-d converter input range.

Giorgio Delfitto  
University of Padova  
Italy



This single-rail amplifier processes  $\pm 5V$  inputs, using a 2.5V reference and the 5V supply from an interface connector.

## Mosfet driving via a 555

If you need to drive one mosfet and have a space-saving requirement, a 555 timer replaces half a dozen discrete components.

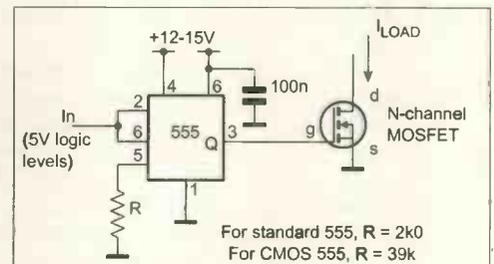
If the mosfet supply lies within 12-15V, the timer drives the gate to at least 10V and, because of the 555 totem-pole output, switching on and off takes 100ns or less.

Input impedance to the timer is high, output impedance low and setting the voltage on pin 5 to +3.5V by  $R$  ensures level conversion from 5V logic to that needed by the mosfet gate. Pin 4 is the reset input,

which can, if required, be used for an inhibit input. For pulse control of the mosfet, configure the 555 as a free-running or monostable flip-flop.

Supply voltage for cmos 555s should be 14V or less and switching speed may be slightly reduced. Decouple the timer supply close to the chip and mount the device near the mosfet.

S H Dolding  
Carlingford  
NSW, Australia



One more use for a 555: as an inverting gate driver and level shifter for mosfets.

## Preventing battery over-discharge

To prevent failure of rechargeable batteries from over-discharging, this circuit switches the load out of circuit at a preset battery voltage. Depending on Hexfet type, it handles loads of over 20A, needs no heat sink and in a no-load condition takes only 370 $\mu$ A.

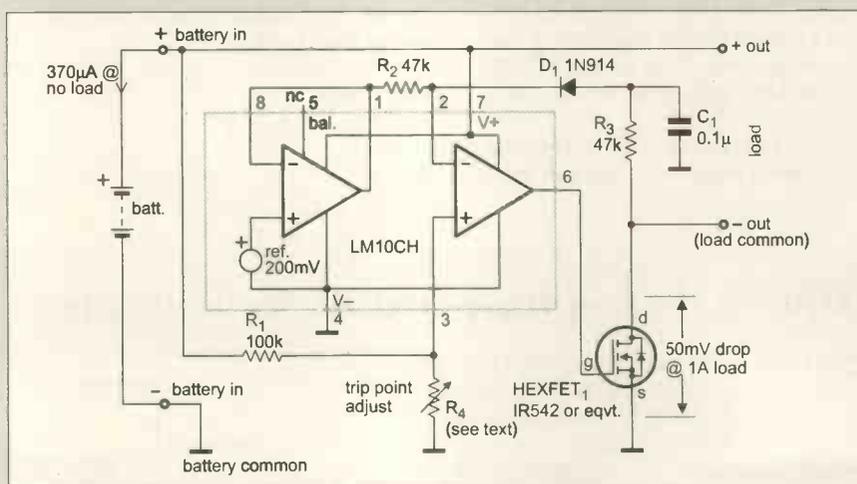
An LM10CH contains an op-amp, buffer and voltage reference. The buffered reference is taken via  $R_2$  to the op-amp, which is used to compare the reference with the fraction of battery voltage set by  $R_{1,4}$ , its output switching the Hexfet on and off.

If battery voltage falls below the trip point which, for 12V cells, is about 10.5V, the fet switches off and the negative output terminal rises towards the battery voltage, latching itself off through  $R_3, D_1, C_1$  and the comparator. If the battery voltage increases after the load is removed, the switch stays off due to the voltage through  $D_1$  from the rail. If either the load or battery is removed and

reconnected, the circuit restarts if the battery voltage is above the trip point.

V Labuc  
Hudson  
Quebec, Canada

Switching circuit avoids damage to rechargeable cells, which must not be allowed to discharge to below a given terminal voltage. Circuit handles over 20A load.



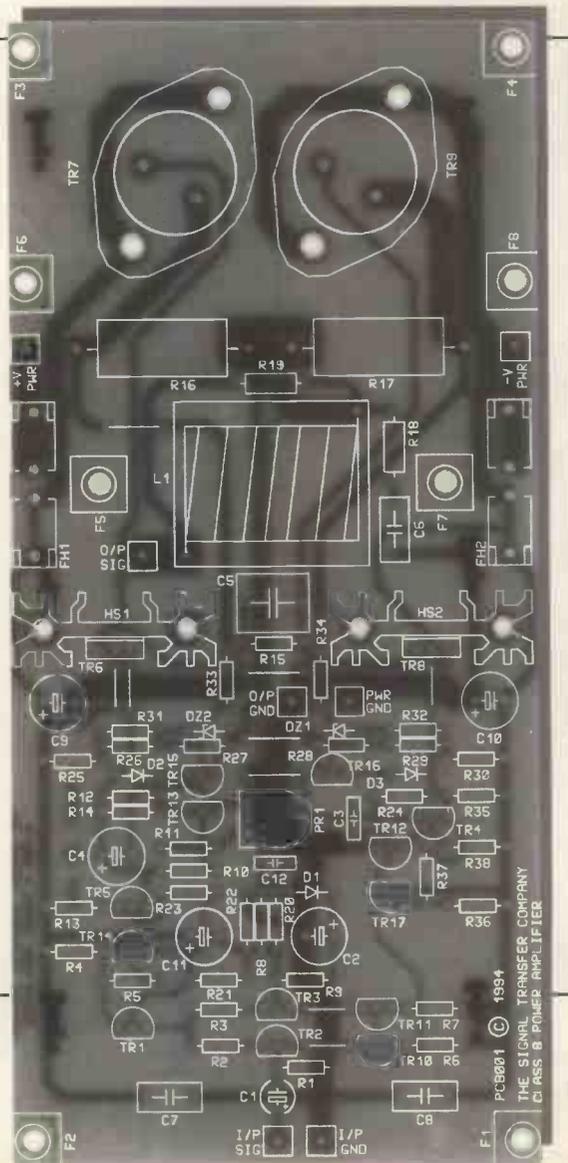
## PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via EW+WW.

Detailed on page 139 of the February 1994 issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8 $\Omega$ , the amplifier features a distortion of 0.0015% at 50W and follows a new design methodology.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 081-652 3614. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.



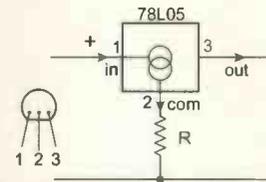
## Tricks with a voltage regulator

It is worth pointing out that between the input and common pins of a Motorola 7805ACP 5V voltage regulator is a useful current regulator. A batch of these devices showed a current of 2mA regardless of the applied voltage over 7V or load. A TI8306 (78L05) still had the constant current, but of 3.5mA.

To increase the output of the IC regulator, one can insert a diode in the common leg

but, in view of the constant current of 2mA, a resistor of 500Ω/volt is satisfactory. There seems to be no loss of conventional regulator performance when the 7805 is used in these ways.

**John Cronk**  
Prestatyn  
Clwyd  
Wales



**Unorthodox use of a Motorola 7805 voltage regulator to obtain a current regulator and to increase voltage output.**

## Driving three-phase brushless dc motors

Figure 1 shows a circuit designed to test the functioning of a range of three-phase brushless motors obtained from disk drives.

A 4018 divide-by-*n* counter generates the three-phase drive waveforms<sup>1</sup> shown in Fig. 2, feedback from Q<sub>3</sub> causing the first three Johnson counters to generate a six-step sequence, which is sequentially stepped out via Q<sub>4</sub> and Q<sub>5</sub>. Waveforms from Q<sub>1</sub>, Q<sub>3</sub> and Q<sub>5</sub> each take six clock pulses per cycle, the phase lags being 120°.

Direction control is provided by an input to the 4070 Ex-Or, which inverts phases 1 and 2 when the input is high, and the 4001 simply switches off the motor drive transistors by grounding their gates.

N-channel IR 110 mosfets rated at an amp were used in the prototype, but for higher current, Fig. 3 shows a circuit using TIP 122 Darlingtontons.

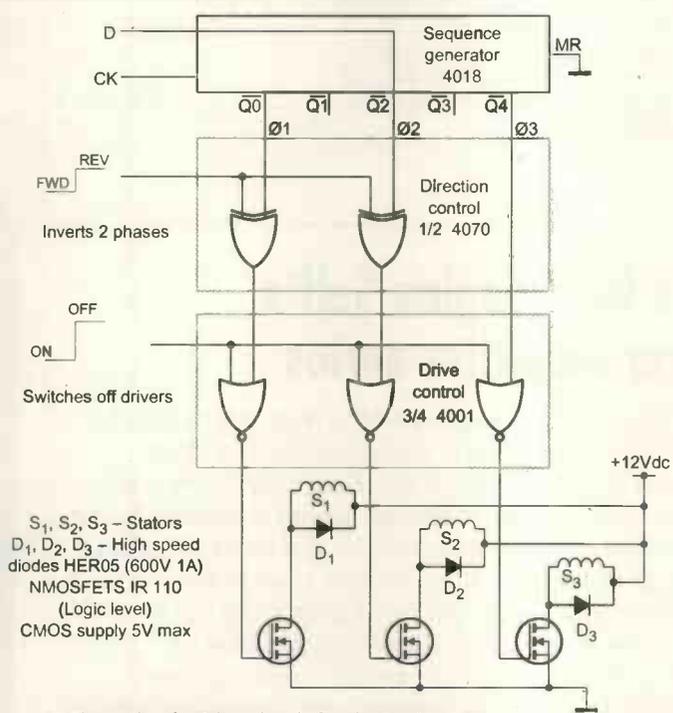
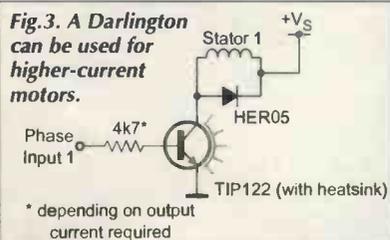
Motors tested appeared to be rated at 24V, 2A, and maximum clock frequency was 200Hz, giving a motor speed of 2000rev/min. At very low clock frequencies – 2Hz – the motor stutters before rotating smoothly, an effect that disappears at higher speeds.

**W A Russell**  
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Tauranga  
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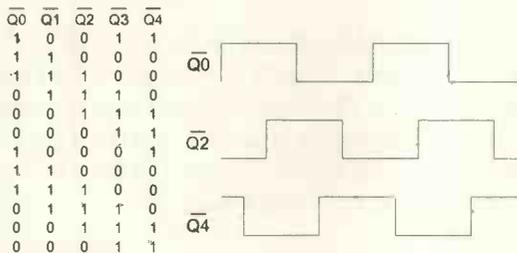
### Reference

1. Lancaster D. *Cmos Cook Book*, 2nd edition.

**Fig. 3. A Darlington can be used for higher-current motors.**



**Fig. 1. Simple drive circuit for three-phase, brushless dc motors, with reverse and inhibit inputs.**



**Fig. 2. Drive waveforms from circuit of Fig. 1.**

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# Improving analogue conversion

*Getting the best high frequency performance from a clocked d-to-a converter normally needs a complex post-conversion filter designed using cad. Michael Batty's  $\sin(x)/x$  correction filter provides an easy to implement yet effective means of compensating for a clocked data converter's natural roll off.*

Most d-to-a converters are effectively sample and hold devices. They latch each successive incoming value to provide a first-order 'staircase' approximation of the output signal.

Conversion is normally followed by a low-pass reconstruction filter. This is typically a high-order elliptic design giving a relatively flat response up to just below half the sampling frequency,  $F_s$ , followed by a sharp cut-off. But this arrangement provides no correction for the sinc function ( $\sin(x)/x$ ) fall-off in amplitude naturally produced by the sample and hold function.

As shown in Fig. 1, the roll off reaches the considerable value of  $-3.9$  dB at  $F_s/2$ , falling to zero at  $F_s$ . Correction for this roll off can be designed into the reconstruction filter if computer-aided design tools are available, but the simple separate correction filter described here yields excellent results and allows more flexible design.

The correction filter is a simple second-order low-pass section, with frequency and  $Q$  optimised to have a response almost exactly inverse to that of the sinc roll off function up to around  $0.4F_s$ . If required, a single additional second-order all-pass stage can be cascaded to improve delay equalisation of the correction filter.

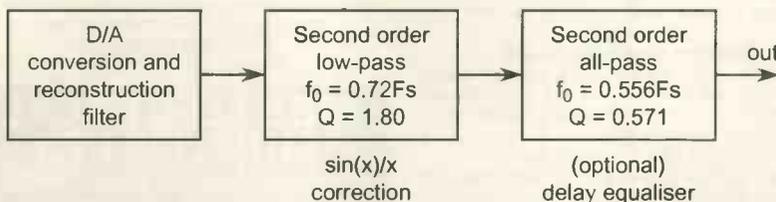


Fig. 2. Amplitude correction and optional delay correction stages are simply cascaded at the output of the d-to-a converter. These two stages consist of second-order low-pass and all-pass filters respectively, with parameters as shown. The delay correction stage should only be needed if it is desired to approximate linear phase (constant delay) behaviour from the whole system.

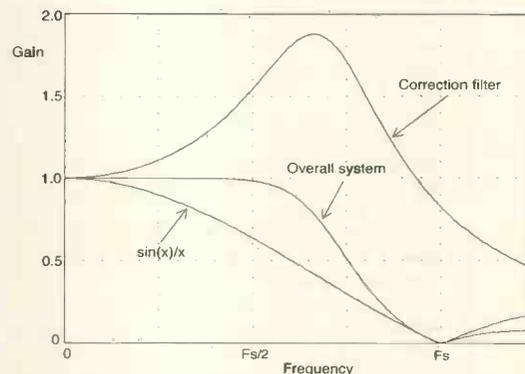


Fig. 1. Amplitude correction scheme uses a filter with a response which is almost exactly complementary over a wide range of frequency to the  $\sin(x)/x$  roll off inherent in the d-to-a converter. This cancels the roll off to a surprising degree of accuracy, considering the simplicity of the scheme, although the response deviates markedly above half the sampling frequency. However, by this stage, the stopband response of the normal analogue reconstruction filter should give adequate rejection.

Note that in most practical situations, the delay equalisation stage may be omitted, since it is unlikely that the accompanying reconstruction filter will itself have linear phase. Also, the correction filter alone is *not* useful as a reconstruction filter, due to its rather poor roll off characteristics.

Figure 2 shows the general configuration of the correction scheme, with optimised filter parameter values. Frequency and  $Q$  for each stage were arrived at by a simple but partly computer-assisted optimisation procedure, with a goal of maximum flatness and linear phase up to  $0.4F_s$ .

Results are shown in Fig. 3. The low-pass stage produces a surprisingly flat equalisation of the sinc function: to within  $\pm 0.01$  dB from

0 to  $0.4F_s$ , with an error of  $-0.12$  dB at  $0.5F_s$ . The optional delay equalisation stage results in an overall phase linearity to within  $\pm 0.5^\circ$  from 0 to  $0.4F_s$ , or a group delay variation of less than  $\pm 0.7\%$  within this range. In practice these figures will almost certainly be swamped by analogue component tolerances.

Figure 4 shows a practical implementation for an  $F_s$  of 20kHz using single op amp stages, quite adequate for the low  $Q$ s involved, although the choice of op amp may depend on the value of  $F_s$  if high accuracy is required. In any case precise components of 1% tolerance or better should be used.

Preceded by a unity-gain buffer stage to provide a low source impedance, the low-pass section uses a standard unity-gain Sallen and Key design. The optional delay equaliser section includes a final gain stage to compensate for 12.2dB loss in the all-pass stage. Component values in the two filter stages may of course be modified for other values of  $F_s$  by scaling either resistor or capacitor values in the standard manner.

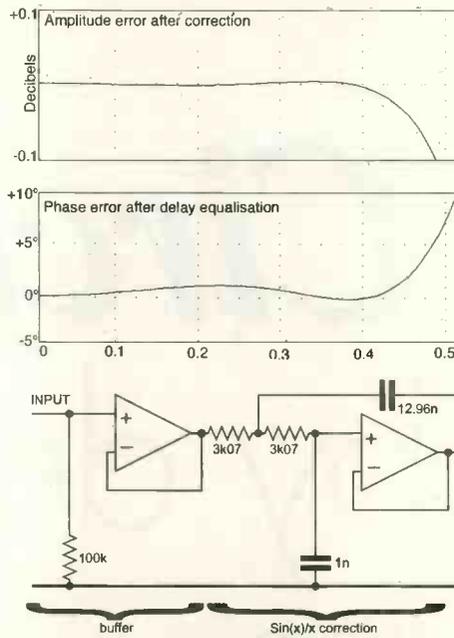


Fig. 3. Residual amplitude ripple is extremely small by most standards, and would suit, for example, instrumentation applications such as digital waveform generators. With the optional delay correction, the residual phase deviation from linear behaviour is also fairly small (less than  $0.5^\circ$  up to  $0.4$  times  $F_s$ ). In reality, it is likely that real component tolerances will introduce larger variations than these.

Fig. 4. The correction scheme may be realised at audio frequencies using simple second-order op-amp filter circuits, this diagram showing a practical circuit designed for a sampling frequency of 20kHz. For precision applications a full circuit simulation is suggested, to examine the effects of passive component tolerances and real op-amp limitations. For other sampling frequencies, all  $R$  (or  $C$ ) values in the second and third op-amp stages may be scaled inversely with frequency in the usual way. Details of the all-pass filter design may be found in the reference.

Further reading

Analog filter design by M. E. Van Valkenburg, 1982, Sect. 6.5, 18.4, pub. Holt Rinehart & Winston.

D-to-a converters and  $\sin(x)/x$  roll off

Spectral roll off produced by a d-to-a converter which always holds (latches) its last value is a well known effect in signal processing systems, and may be briefly (although by no means trivially) understood as follows:

Prior to sampling, a signal spectrum can consist of frequency components up to half the sampling frequency,  $F_s$ , as shown in Fig. A (if they are any higher we run into under-sampling and hence aliasing problems – another issue).

Ideally, sampling a signal can be quite correctly considered as multiplying it by a train of impulses, Fig. B, thus producing a spectrum containing an infinite number of copies of the signal band, each symmetrically 'folded' about harmonics of the sampling frequency. This may perhaps be understood more simply in the frequency domain by considering sampling as a mixing process; the various copies of the spectrum formed correspond to all the sum and difference products formed by the signal and the uniform comb of harmonics of the sampling signal.

To accurately reconvert the signal into analogue form, it must, of course, be smoothed; this is just equivalent to low-pass filtering. In the frequency domain, this reconstruction filter does its job perfectly if it removes all copies of the spectrum except the one near zero frequency, as shown in Fig. C. The analogue signal is then perfectly recovered. Clearly, an ideal reconstruction filter would have zero attenuation (and phase shift) up to  $0.5F_s$ , but infinite loss above this.

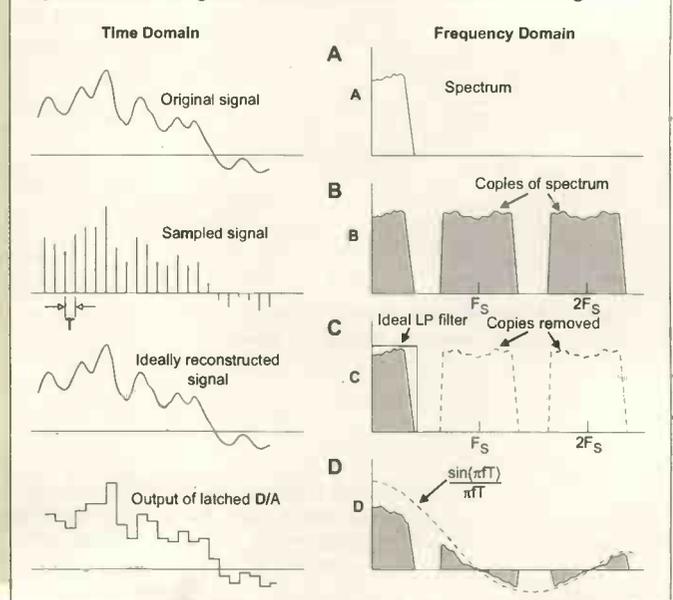
The effect of holding the last output of a d-to-a converter is to provide a degree of (stepwise) smoothing, which does give some low-pass filtering, but is far from ideal. At this point, we need a little signal processing theory to help. In the time domain, the hold function is equivalent to convolving the sampled waveform with a rectangle function of width one sampling period. Now Fourier theory (in particular the convolution theorem) tells us that in frequency terms, this is exactly equivalent to multiplying the spectrum by the Fourier

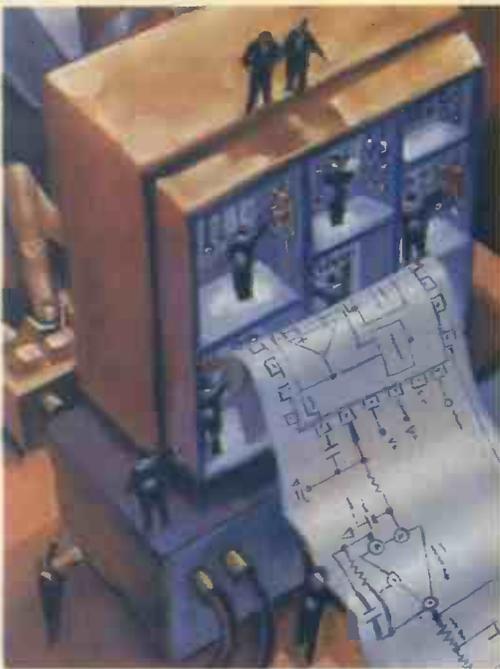
Transform of the rectangle function, which just happens to be of the form:

$$\frac{\sin(\pi f T)}{\pi f T}$$

(commonly known as a sinc function), where  $f$  is the frequency and  $T$  the sampling period,  $(1/F_s)$ . This is illustrated in Fig. D, and produces a zero value at  $F_s$ , with a  $-3.9$ dB response at  $0.5F_s$ , thus distorting the spectrum considerably from the original. Clearly if the sampling frequency is very high compared to the signal (ie highly oversampled), then the effect on the signal may be small, and this is often used to advantage.

Spectra illustrating the need for  $\sin(x)/x$  corrective filtering.





# Circuits by design

## 4: Filters

**Due to the number of calculations involved, filter design in particular is greatly simplified by using cad software.**

**Owen Bishop describes how to get the best out of modern circuit simulation and analysis tools.**

A first-order low-pass passive filter is easily realised with a pair of components, Fig. 1, or as a Spice netlist, Fig. 2. With the selected values of the capacitor and resistor, the -3dB point - cut-off frequency - is  $f_c = 1/2\pi RC = 1592\text{Hz}$ . I will use this simple example to demonstrate some of the frequency response analyses available in *SpiceAge for Windows*. For all of these analyses, a voltage probe is set between 'vout' and 'gnd', so that it monitors voltage across the capacitor.

The first run is a straightforward Bode Plot, input being a 1kHz sine wave at 1V. Frequency range is set for 10Hz to 1MHz in 100 steps. Under the Frequency drop-down menu, select Y-display mode, then tick dB plot and Phase plot.

Under Analyse, selecting Frequency response now produces Fig. 3. On the left is a decibel scale, instead of the usual voltage scale, with 0dB being equivalent to 1V, the amplitude of the voltage generator. On the right is a phase scale running from +180° to -180°.

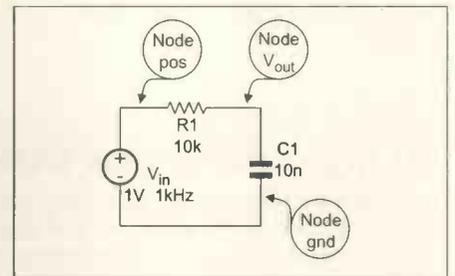


Fig. 1. Nodes of this simple low-pass filter are named, ready for entering in as a Spice netlist.

The graph has two curves, one for output signal amplitude and the other - in a different colour on the display - for the phase difference between the input and output signals. The graphs show that amplitude begins to fall above about 50Hz.

The cross-hairs are placed on the -3dB point, as indicated by the value displayed below the graph. At this point the frequency

```
* CBD04 - Low-pass passive RC filter, 1.6kHz
V Vin      -out:gnd  +out:pos  v=1.000000 Ex=Sine Fr=1.00000k
R R1      p1:pos   p2:vout   v=10.0000k
C C1      p1:vout  p2:gnd    v=10.0000n
```

Fig. 2. The netlist above defines the values of the components in Fig. 1, and the connections between them.

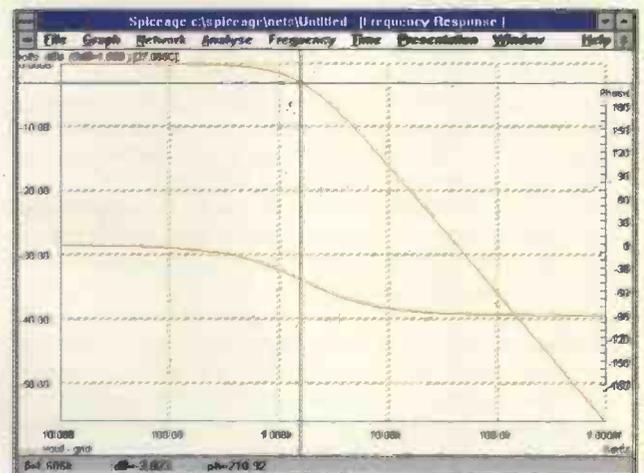


Fig. 3. A frequency sweep shows a typical low-pass amplitude curve, right. Phase of the output signal lags when frequency is greater than about 1kHz, tending toward 90° as frequency approaches 1MHz.

reading is 1.606kHz, which is the closest reading to the calculated value of  $f_c$ . Beyond this point, response falls steadily at  $-6\text{dB}$  per octave.

Phase changes are shown on the lower curve. At low frequencies, the output signal is in phase with input. As frequency increases, the output lags further and further behind input, eventually finishing with a phase lag of  $90^\circ$ .

Another display mode, Fig. 4, is obtained by selecting Real and imaginary in the list of Y display modes. In this, the real and imaginary components of the voltage across the capacitor are plotted separately against frequency. The real component, shown in the upper curve, falls from 1000mV at frequencies below about 80Hz to zero at frequencies greater than about 200kHz. At  $f_c$ , it reaches its half-way value of 500mV, equivalent to  $-3\text{dB}$ , as indicated by the vertical cross-hair. The imaginary part, lower curve, has its greatest negative value of  $-500\text{mV}$  at this frequency. At this frequency the phase angle is  $-45^\circ$ .

A third display mode is Nyquist Plot, obtained by running the same simulation, Frequency response, but with Complex plane ticked as the Y-display mode. Display, Fig. 5, is a plot in the complex plane, the x-axis being real and the y-axis being imaginary. Both axes are graduated in a linear voltage scale though, on the screen, the scales are not equal.

Note that in this example all the values on the y-axis are negative and the origin is in the top left corner. The curve passes through a set of points labelled with the frequency.

A graph of this type is most easily interpreted by reference to a set of phasors. Beginning at top right of the graph with the

lowest frequency, the output is in phase with and has the same amplitude as the input, dashed line of Fig. 6. The real component has the value 1000mV and the imaginary component is zero.

As frequency increases up to 1592Hz, the imaginary component increases but is negative, while the real component decreases. At 1592Hz they have equal magnitude (500mV and  $-500\text{mV}$ , see Fig. 3) and the phase angle is  $-45^\circ$ . This is the point at the bottom centre of the graph in Fig. 5, the curve being the locus of the tip of the phasor as frequency changes.

As frequency increases further, we follow the curve around to the left, both components decreasing. You will notice that the real component decreases at a faster rate and thus, at the limit, the phase angle is  $-90^\circ$ .

### Designing active filters

As a change of approach, I will use *Mathematica* to assist with the designing of a first-order low-pass active filter, based on an operational amplifier, Fig. 7. It is to have a dc gain of 25, and a cut-off frequency of 10kHz.

Decide on a possible value for the capacitor, say 2.2nF. Calculating the values of the resistors follows a routine. As in the programs described last month, set up a Notebook that begins with the definition of a function LowPass, Fig. 8. This needs 3 values: dc gain ( $a$ ), cut-off frequency ( $f_c$ ) and capacitor value ( $c$ ). It calculates the values of the input resistor ( $r$ ), the amplification resistor ( $ra$ ), and the feedback resistor ( $rf$ ).

Equations for these calculations appear in the definition, and balance the bias current by making  $ra$  and  $rf$  in parallel equal to  $r$ . These calculations would normally be done by hand, on paper, but here we can run through them simply by clicking the Action menu and selecting Evaluate notebook. As explained last month, you can also click on and evaluate one stage at a time when testing and debugging the routine.

When the function is called, it returns the value of  $rf$ , given by the final equation in the definition. Commands ' $ra$ ' and ' $r$ ' following this causes the calculated values of  $ra$  and  $r$  to be displayed. These are then converted to standard resistor values (see panel). In this way, the three resistor values required for building the filter are obtained, assuming an ideal op-amp.

Figure 8 has a final line needing more explanation. The function BodePlot is one defined in a package specially intended for engineers. This is the *Electrical Engineering Pack*, obtainable as an optional extra from the publishers of *Mathematica*. It consists of a number of Notebooks, describing various ways of using *Mathematica* in electrical and electronic applications. These are presented as a printed book and as a series of files on disk ready to load when running *Mathematica*.

More interestingly for us, it also contains a package of functions relevant to electrical and electronic engineering. The Bode Plot is one of the routines included in the *EE* package and

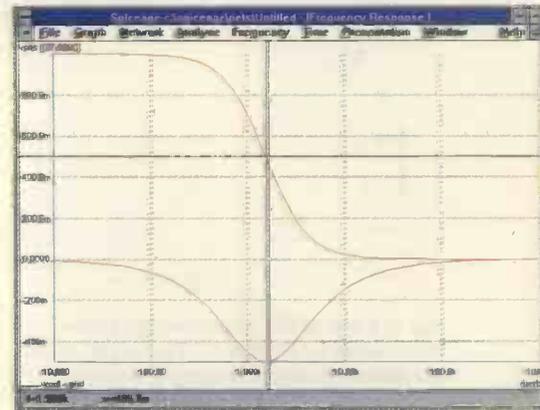


Fig. 4. In this curve, frequency response of the low-pass filter is plotted in terms of the real and imaginary parts of the output voltage phasor.

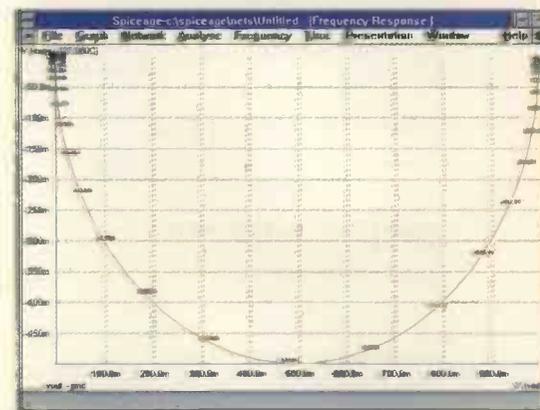


Fig. 5. A Nyquist Plot in the complex plane defines the locus of the apex of the output voltage phasor as frequency rises from 0Hz (top right) to 1MHz (top left).

### Packages

A number of standard packages are included with *Mathematica*. These provide some of the routines which are less often required than those in the kernel. Examples are statistical functions, and calculus functions such as the Laplace transform. These are read into the program by a command such as

```
<< Calculus`LaplaceTransform`
```

or

```
Needs["Calculus`LaplaceTransform`"]
```

It can be arranged that a package is loaded automatically when one of its functions is called. With the engineering package, the easiest method is to begin the session by typing:

```
Needs["EE`Master`"]
```

This loads all the functions that are present in the *EE* package.

### Standard resistor values

The calculations return the values of resistors to several significant figures, usually at least five. In practice, a circuit is built from resistors of the E12 or E24 series and only rarely from E48 or higher-precision resistors. The *EE* package has a routine *SelectResistor* to pick out the standard resistor nearest to any given resistor value. The syntax is:

```
SelectResistor[resistance, tolerance]
```

In Fig. 8 this routine is used for finding practical values for  $R$ ,  $RA$  and  $RF$ , assuming 5% tolerance. In this Notebook, the selected value is assigned to the variable before calling BodePlot. Because of this, the plot shows the behaviour of the filter using standard resistors.

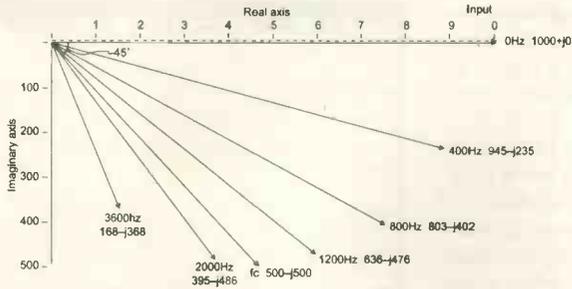
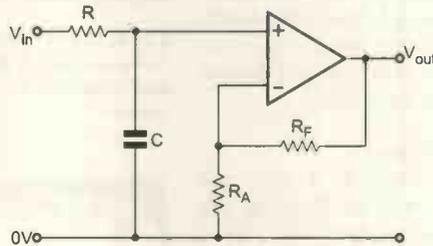


Fig. 6. A few of the output voltage phasors represented by the Nyquist Plot of Fig. 45 show how the almost semicircular curve is traced.

Fig. 7. A low-pass active filter based on op-amps, to be used as a basis for analysis by Mathematica.



CBD04 - Low-pass RC filter

```

LowPass[a,fc,c]:= (r=1/(6.28319 fc c);rf=r (a-:
Input["DC gain"];
a = #;
Input["Cut-off frequency"];
fc = #;
Input["Capacitor"];
c = #;
LowPass[a, fc, c]
r
SelectResistor[rf,0.05];
rf = #
SelectResistor[r,0.05];
r = #
BodePlot[(a/(2 Pi r c))/(I w+1/(2 Pi r c)),
{w,100,10^7}]
    
```

Fig. 8. Mathematica procedure for obtaining a Bode Plot for a low-pass active filter such as that in Fig. 7.

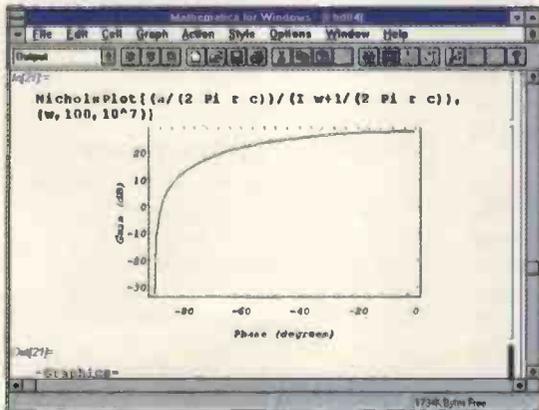


Fig. 9. This procedure modification gives a Nichols Plot. Gain is related to phase angle. Curve shows gain reducing and phase lag increasing at higher frequencies.

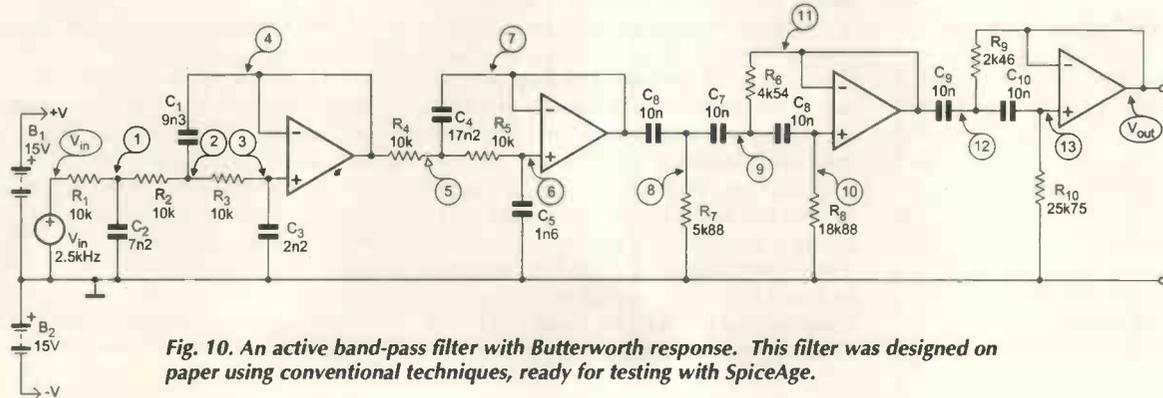


Fig. 10. An active band-pass filter with Butterworth response. This filter was designed on paper using conventional techniques, ready for testing with SpiceAge.

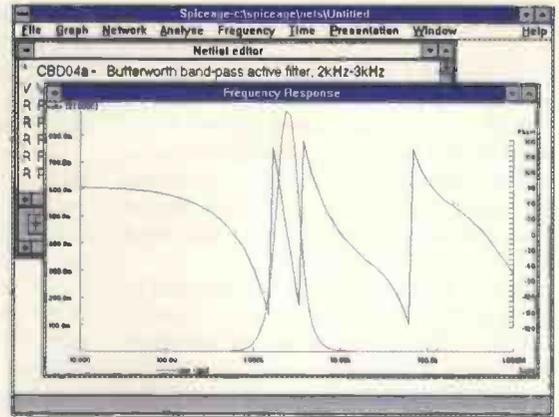


Fig. 11. Transient analysis shows peak response at 2.5kHz, and a fall-off on either side of this at the specified rate.

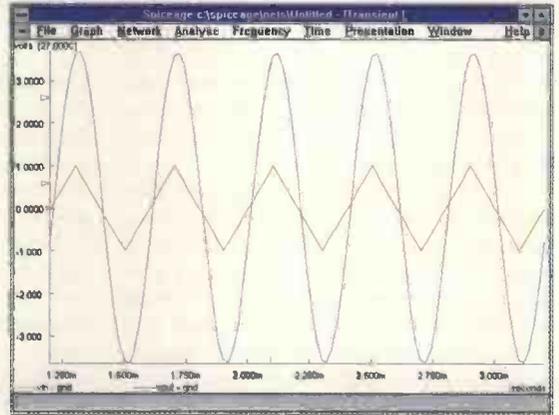


Fig. 12. Band-pass filter of Fig. 10 in action. A 2.5kHz triangle wave is filtered to a 2.5kHz sine wave, all harmonics are filtered out because they are well above the pass band.

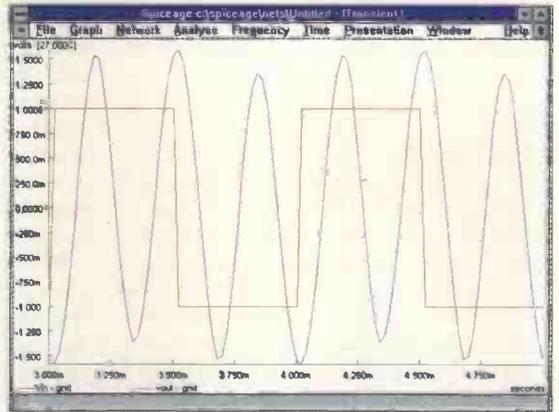


Fig. 13. When filter input is a 1kHz square wave, well below the pass band, only the 3kHz sine wave component comes through strongly. This transient analysis plots the filter's 1kHz square input and 3kHz sine output.

## Enlarging the plot

Although the graphs are too small to read when first plotted, they are easily enlarged. Place the cursor anywhere in the cell – except for the cell bracket – and click the mouse. This makes a box appear around the graphs. The box has handles comprising small shaded squares; clicking on one of these and dragging it toward the edge of the screen causes the display to be enlarged. Click in an adjacent cell to remove the box.

I will discuss several others in due course.

Syntax of BodePlot is:

BodePlot[ transfer-function, {frequency, minimum, maximum}]

In the case of the first-order filter, the transfer function in the frequency domain is:

$$H(s) = \frac{A/2\pi RC}{s + 1/2\pi RC}$$

This is incorporated into the function in Fig. 8, substituting  $j\omega$  for  $s$ , but using  $I$  for  $j$ ). In curly brackets we have  $\omega$ , the frequency variable, with 100 and  $10^7$  setting the plot to range from 100Hz to 10MHz.

Evaluating the Notebook, input '25' as the dc gain, '10000' as the cut-off frequency and '2.2 10^-9' (note the space between the '2.2' and the '10') as the capacitor value. The program responds by displaying 180858 as the value of  $rf$ , and 7234.31 for  $R$ . These are converted to 180k $\Omega$  and 7.5k $\Omega$  respectively. Then it displays the Bode plot in the form of two graphs.

The upper graph shows amplitude against frequency. It is plotted on a decibel scale, with 28dB ( $=20\log 25$ ) as the reference; the amplitude falls by 3dB, to 25dB, at the cut-off frequency, as indicated by the cross-hairs. The lower graph is phase against frequency, falling from 0° to -90° over the displayed range. Cross hairs indicate that the output lags the input by 45° at the cut-off frequency.

## Other plots

The *Engineering Pack of Mathematica* has functions for two more useful plots, one of which is the Nyquist plot. The syntax is the same as for the Bode plot; just substitute 'NyquistPlot' for 'BodePlot' in Fig. 8 and evaluate the Notebook.

The other function is the Nichols Plot, in which the gain of the filter in decibels is plotted against the phase angle. Substitute 'NicholsPlot' for 'BodePlot' in Fig. 8 and evaluate the Notebook. In Fig. 9, gain rises from a low level when the phase angle is -90° at high frequency to 28dB when the phase angle is close to 0°, i.e. at low frequency.

## Filter design

The task is to design an active band-pass filter

## Creating .WAV files

Readers whose simulation software does not have the facility to generate .WAV files may make use of this simple technique. You need a program for generating a .WAV file from an equation. A suitable program, which is a complete sound file editor, is *GoldWave*, by Chris Crain. This is available as shareware from Chris at PO Box 51, St. John's, NF, Canada A1C 5H5.

Internet: chris3@garfield.cs.mun.ca. Registration fee is \$30 Canadian. It is also on the diskette packaged with *Soundblaster: The Official Book* by Peter M. Ridge, David M. Golden, Ivan Luk and Scott E. Sindorf, 2nd Edition, published by Osborne/McGraw-Hill.

When you have performed a transient analysis, preferably of five or more cycles, obtain a Fourier transform. Use the cross-hairs to measure the amplitude of the fundamental and each harmonic. Also measure their phase angles. Given this information, write an equation to reconstruct the waveform.

As an example, here are the data for a 1kHz triangular wave with a 1V amplitude:

Frequency	Amplitude (V)	Phase
1kHz	0.8063	-90°
3kHz	0.09019	+90°
5kHz	0.03247	-90°
7kHz	0.01647	+90°
9kHz	0.01013	-90°

Written as a Fourier series:

$$y = -0.8063\sin x + 0.09019\sin 3x - 0.03247\sin 5x + 0.01647\sin 7x - 0.01013\sin 9x$$

The angle should include the phase angles but, since all angles are either +90° or -90°, it is simpler to alter the angle of reference and give the terms the sign of the phase angle.

The right-hand side of this equation is typed into the *GoldWave* Expression Evaluator, substituting 't\*1000' for  $x$  to obtain a 1kHz wave. The expression was evaluated for  $t=0$  to  $t=2$  to give a two-second file. You could substitute any other factor instead of 1000 to obtain sounds of other frequencies.

The same triangle wave is filtered through a three-pole RC low-pass filter with a -3dB point of 2kHz. The output transient is evaluated from 2ms to 12ms to eliminate the starting-up transient. After a transient analysis as above, the Fourier analysis gives:

Frequency	Amplitude (V)	Phase
1kHz	0.2773	178°
3kHz	0.00758	-60°
5kHz	0.00103	+90°
7kHz	0.00025	-108°

Here it is necessary to take full account of the phase angle, after converting it into radian (all terms positive):

$$y = 0.2773\sin(x+3.11) + 0.00758\sin(3x-1.05) + 0.00103\sin(5x+1.57) + 0.00025\sin(7x-1.88)$$

Type this in, substituting 't\*1000' for  $x$ , as before. The whole expression is bracketed and multiplied by three to compensate for the attenuation of the filter. The result is a .WAV file sounding like a 1kHz sine wave, which is virtually what it is after such heavy filtering.

with the Butterworth response, using 741 op-amps. The -3dB frequencies are 2kHz and 3kHz; response is 30dB down at 1kHz and 6kHz. The filter consists of a low-pass filter cascaded with a high-pass filter.

The standard routine for deciding how many poles are required and for calculating resistor and capacitor values involves very little calculation. Most of the values are obtained by consulting tables. It is hardly necessary to invoke the aid of *Mathematica*.

You will find that both the low-pass and high-pass sections must be five-pole filters.

Each consists of a three-pole filter followed by a two-pole configuration. Figure 10 shows the circuit and is marked with the calculated values. Having completed the design, we use *SpiceAge* to verify it and to investigate some of its properties.

Power supply to the capacitor is provided in the netlist by two 15V batteries connected in series, and to the positive, negative and ground lines. Input is a sine wave voltage generator, amplitude 1V, first set to 2.5kHz, which is the mid-frequency of the filter.

The procedure is the same as if the filter has

been assembled with real components. I will demonstrate how it works. The advantage with using a simulator is that it is not possible to burn out expensive components because of wrong connections. Additionally, there is no problem with electrostatic charges when using CMOS components.

A probe is set to monitor *vout*. Call up the Time selector box and enter a start time of 0s, a stop time of 400µs and a step time of 8µs. This gives time for one complete cycle

The first test, a Transient analysis, gives an output of 400V amplitude; obviously something is wrong. Instead of desoldering connecting wires, it is possible to 'isolate' parts of the circuit and test them individually for faults; simply type an asterisk at the beginning of any statement to remove the component from the circuit.

Check that the remaining components have the correct temporary connections and test again. Isolating the first op-amp still produces an excessively high output, so obviously something more fundamental is wrong. Careful scrutiny of the netlist reveals a typing error; there is 'v-' where there should be '-v'. Once this is corrected the circuit works perfectly. The output amplitude is about 1/5 that of the input and appears to be almost in phase.

To confirm that the filter behaves as intended, set the Frequency range to extend from 10Hz to 1MHz and run a Frequency response analysis. The result, Fig. 11, shows that the response is almost exactly as required. At the low end it is -30dB at 1kHz, and -3dB at 1kHz. At the high end it is -3dB at 3.2kHz and -30dB at 6.1kHz.

The pass-band is narrow and fall-off is steep on both sides so that the typical flat pass-band of a Butterworth filter is not shown. Also shown in the graph is how phase angle varies with frequency. The angle decreases from 90° at low frequencies to -180° at about 2kHz. It continues to fall, but having completed half a circle, now appears as +180° on the graph falling to 0° (really -360°) at 2.5kHz. It continues to fall as frequency increases.

**Filtered waveforms**

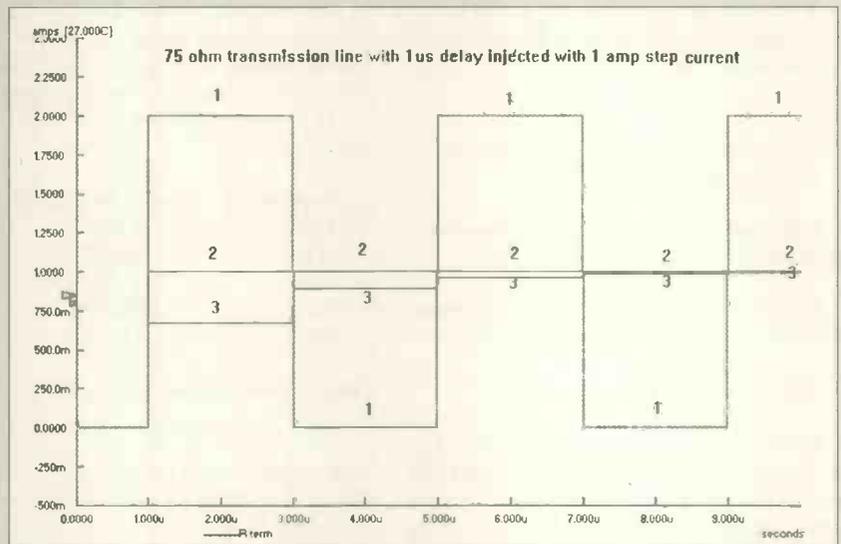
*SpiceAge* can show the effect of the filter on various waveforms. Change the excitation of the voltage generator to 'triangular' and set the Time selector to start at 0s, stop at 2ms with steps of 20µs. This gives time for 5 complete cycles.

Transient analysis shows that the output amplitude increases gradually during the first three cycles and is constant thereafter. This is because it takes about that length of time for the capacitors to become charged to their operating levels. Such a result is not a good basis for a Fourier analysis.

Re-set the timing to begin at 1.2ms and stop at 3.2ms. The first three cycles are ignored and the fourth to eighth cycles displayed, Fig. 12. In the figure, the output appears to have a greater amplitude than the input, but this is because a scale factor of times five is set for the output waveform.

Input waveform is clearly triangular but the

# Win Spiceage for Windows



Two readers answering the questions below correctly will win a copy of *SpiceAge for Windows* Level 3 – base professional level – worth £395+VAT. Simply answer the following questions and complete the tie break, posting your reply to arrive at the offices of Those Engineers before 25 February 1995.

Q1. *SpiceAge for Windows* is written in which country: USA, Germany, UK, Israel, Australia, Japan?

Q2. In a lossless circuit, resonance is given by  $\omega=1/(LC)^{0.5}$  rad/s. Write down the analogous expression for a mechanical spring plus mass system. Please define completely the terms within your expression.

Q3. Which of the following traces which show the current in the

terminating resistor is the correct response of a 75Ω transmission line terminated with a 150Ω resistor to which a current of 1A is suddenly applied?

Q4. You are on site repairing a board for which you need a 20kΩ 5% resistor. You have just dropped your ohm meter in a puddle and your tool kit contains only one each of 5% values fitting the 10, 12, 15, 18, 22, 33 and 47 decade multipliers between 1kΩ and 47kΩ. What do you do to keep within the original value to 3 significant figures?

Q5. Tie breaker. Complete the following in no more than 100 words. "I would be interested in using *SpiceAge* to..." Your reply will be judged (i) on its feasibility and (ii) on how interesting your application is.

Your answers must arrive before 25th February 1995. The winners will be announced in the April 1995 edition of *Electronics World + Wireless World*. No purchase is necessary. The judges' decision is final. The competition is open to EC residents only but is not open to employees or suppliers of Those Engineers Ltd. No correspondence will be entered into. Send your answers by post please to: Those Engineers, Sawcomp, 31 Birkbeck Road, London NW7 4BP.

output appears to be a sine wave, almost in phase with the input – actually about 360° out of phase, as explained above. Fourier analysis shows a single spectrum line at the fundamental frequency, 2.5kHz. Harmonics are well above the pass-band so are considerably attenuated. The output is an almost pure sine wave.

Try the effect of the filter on a square wave, by changing the excitation of the generator. Also reduce its frequency to 1kHz, to bring it below the pass-band. Display the transient analysis from the second to the fourth cycles. As before, the output appears to be a pure sine wave, but its frequency is 3kHz, not 1kHz, Fig. 13. A square wave has only the even harmonics and an examination of the subsequent Fourier analysis has only three significant spectrum lines:

- 1 kHz, amplitude 40mV – the fundamental, but much reduced by filtering
- 3 kHz, amplitude 289mV – second harmonic, within the pass-band
- 5 kHz, amplitude 15mV – fourth harmonic, reduced by filtering

**Making sounds**

Although a simulator can tell you all you need to know about the voltages and currents in a

circuit – and display graphs showing the shapes of the waveforms – there are occasions on which it would be informative to hear a waveform. This is important to anyone designing audio equipment, in which subjective judgements play no small part.

There are several ways in which you can make your circuits 'emit' sounds. *SpiceAge* is able to produce a file called a log file which is an ascii file containing a table of the output voltage at regular intervals of time. This is the data which is used by *SpiceAge* to plot graphs, such as Fig. 12.

Log files are of no direct use for producing sound, but their data is convertible into the format of a .WAV file. A .WAV file is one of the standard file formats used by computer sound cards. When the card reads the file, it produces the corresponding sound.

Most recent versions of *SpiceAge* have a routine for converting a log file to a .WAV file, which is playable on a sound card such as those in the *Soundblaster* range. The file is monophonic, so the inexpensive *Soundblaster* V2 is adequate. The files play on stereo sound cards too, including 16-bit cards.

When creating a .WAV file in *SpiceAge*, the first stage is to set up the netlist. Next select Log file in the Presentation menu. In the Time selector box enter a suitable start and stop time. Only one cycle is required but, with reac-

tive circuits, allow a few cycles to run before the analysis period to allow values to settle.

Step time must not be less than 16µs. In the most recent version, the time selector accepts 'samples per second' as an alternative to sample times. Set Probe 1, to measure input or output voltage. Run a Transient analysis and note the hard disk activity as the circuit is analysed and the data is stored in the log file. The graph is displayed as usual after the analysis is complete.

In the File menu select Convert log file to .WAV format. A dialogue box asks you to name the log file to be converted (for example, the one just created), then a box asks you to provide a name for the .WAV file.

Assuming that you have a *Soundblaster* or similar sound card installed, play the file by selecting Play sound. A dialogue box lists the .WAV files available on the current drive; select one of these and the sound begins immediately. Another box appears inviting you to 'Stop sound' by clicking on a button.

The files produced by *SpiceAge*, derived from the transient analysis, last only for the period analysed. This period is unlikely to be more than a few milliseconds. Correspondingly, the files require few bytes. The play routine of this program repeats the file until told to stop – unlike most file-playing programs which play a .WAV file only once. ■

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Hewlett Packard 3437A System voltmeter	£350
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Hewlett Packard 8901B – Modulation Analyser AM-FM 150KHz-1300MHz	£3750
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Marconi 2831 Channel access switch	£400
Marconi 8938 A/F power meter	£325
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Rohde & Schwarz SMF2 Mobile tester	£4000
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Schlumberger 4923 – Radio Code Test Set	£2000
Schlumberger 2720 – 1250MHz Freq. Counter	£600
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Thurlby Thandar TG1304 Programmable Function Generator (13MHz)	£600
Time 9811 Programmable resistance	£600
Time 9814 Voltage calibrator	£750
W&G MU3 Test point scanner	£500
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Watanabe WTR211 3 pen plotter	£250
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Weller D900 Desoldering station	£150
Wiltron 352 Low freq. differential input phase meter	£350
Wiltron 560 Scalar Network analyser	£800

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CIRCLE NO. 120 ON REPLY CARD

# A question of *balance*

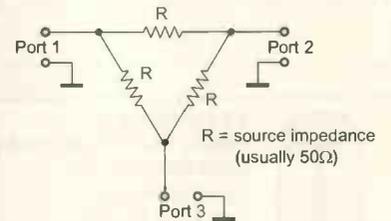
*Two transmitters into one aerial system, two signal generators connected to a circuit under test, one generator feeding multiple outputs... All these applications require load balancing for correct operation. Steve Winder offers a thumbnail tutorial on obtaining the perfect balance.*

**P**ower combiners and splitters find use in rf whenever there is a need to bring sources or loads out to a single port. Combiners are physically the same as splitters: one device can serve two functions. Inputs become outputs and outputs become inputs.

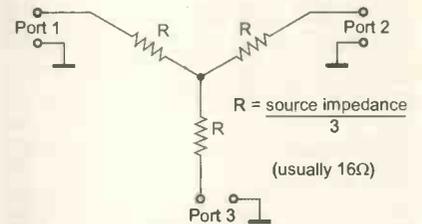
Simply joining sources together will not do; impedances must be matched and isolation between the two sources may be required. A typical use would be in intermodulation distortion testing where two signal generators are coupled to a single output connected to the unit under test. Another relates to single side-band circuits, where two phase shifted signals are combined to eliminate one of the sidebands.

Power splitters are used whenever one signal source is used to drive two or more circuits. For example, this could be when two radio receivers use the same antenna.

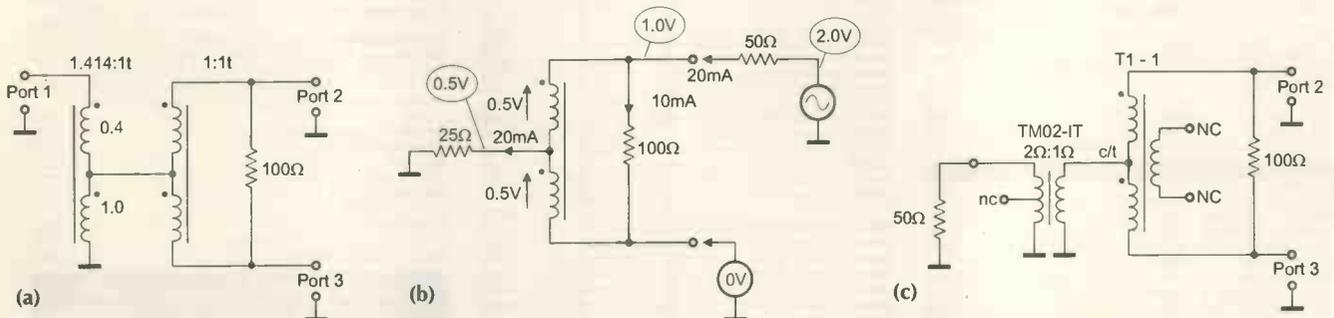
The design coupler naturally depends on the application. Combiners used for two-tone testing require isolation between the two input



**Fig. 1a. Delta resistive splitter network. Each resistor should equal the source resistance for perfect matching.**



**Fig. 1b. Star resistive divider. This is functionally the same as the circuit shown in Fig. 1a.**



**Fig. 2. Derivation of the transformer coupled power divider. a) shows the basic circuit. The first transformer changes the source impedance from 50Ω to 25Ω while the second provides the balancing and isolation action. When ports 2 and 3 see an equal load, there will be no power dissipated in the 100Ω resistor. Fig. 2b) is the circuit model while c) is a practical implementation with readily available prewound components.**

ports and transformer coupled circuits give this isolation.

Where the two signals widely differ in frequency, a diplexer could be the best choice, for instance to separate a vhf radio signal and a uhf television signal transmitted over the same co-axial cable. The following descriptions of power splitters and combiners assume a 50Ω source and load impedance is assumed.

**Resistive power splitters**

Resistive power dividers can be realised in either of the two forms, shown in Fig. 1. Form *a* uses a delta network of resistors, each equal in value to the source impedance, 50Ω (51Ω will do). Form *b* is a star network of resistors, each resistor has a value equal to one third of the source impedance. In a 50Ω system, the three resistors in the star network will have a perfect value of 16.67Ω. In both cases, input impedance matching is dependent on the impedance of both loads.

When a signal is applied to any port of the delta network, the other two ports will deliver a level equal to half the applied voltage. This is one quarter of the input power. Since both outputs have the same voltage on them, no current flows through the resistor connected between the output ports. Each of the other resistors in the delta network is in series with a load resistor of equal value. Half the applied voltage will be dropped across resistors in the splitter, leaving half of the voltage across the load.

Looking at the input and output ports, the star network behaves in an identical way to the delta network. The voltage at the two output ports is half the applied voltage. Both star and delta networks deliver just 3dB of isolation between ports.

**Transformer coupled dividers**

Using transformers in a power splitter, or combiner, enables isolation of the two output ports. This is important in some applications, such as at the output of a diode mixer where signals could mix to generate unwanted products. Isolation between the two outputs is often about 35dB. It also eliminates the losses inherent to the resistive splitter. Half the power goes to each output resulting in a 3dB reduction in signal level. Some transformer losses are inevitable however.

The transformer coupled splitter comprises two transformers, as shown in Fig. 2a. One is an autotransformer, the other transformer has a single centre tapped winding. The auto-transformer reduces the input impedance by half, to present a 25Ω source impedance to the centre tap of the second transformer. Splitters of this type are available commercially, in modular form, from manufacturers such as Mini-Circuits<sup>1</sup>.

When transmitting signals from port 1 to the output ports 2 and 3, an equal amount of current flows through each of the second transformer's windings. Since the currents flow in opposite directions through the two transformer windings, there is no magnetic flux produced. Without flux there is no voltage

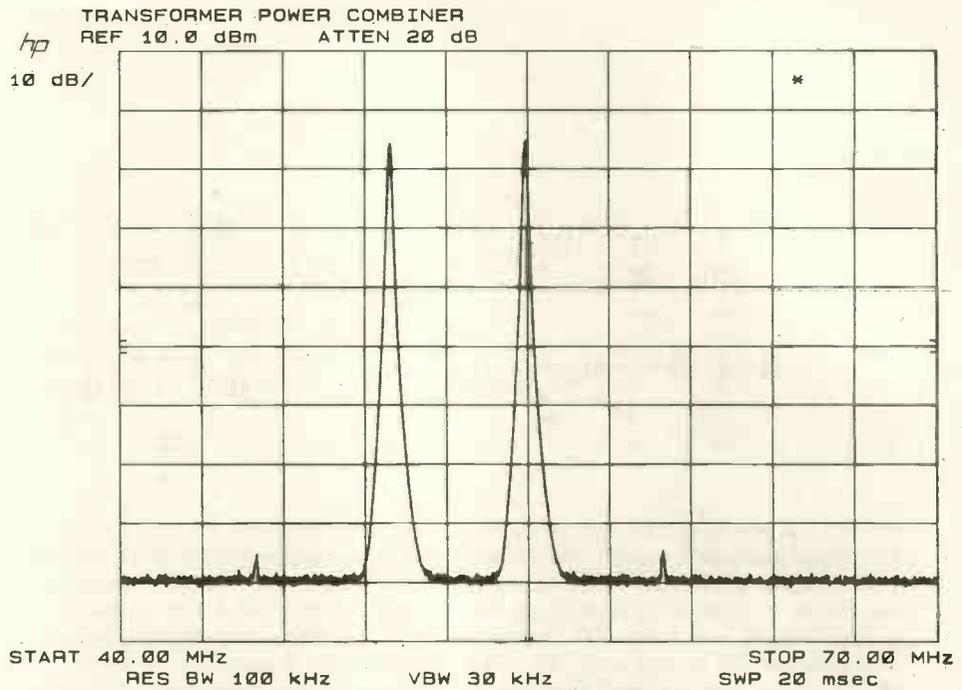


Fig. 3. The spectral plot delivered by the practical implementation of Fig. 2 demonstrates the splitting action, giving equal outputs at ports 2 and 3.

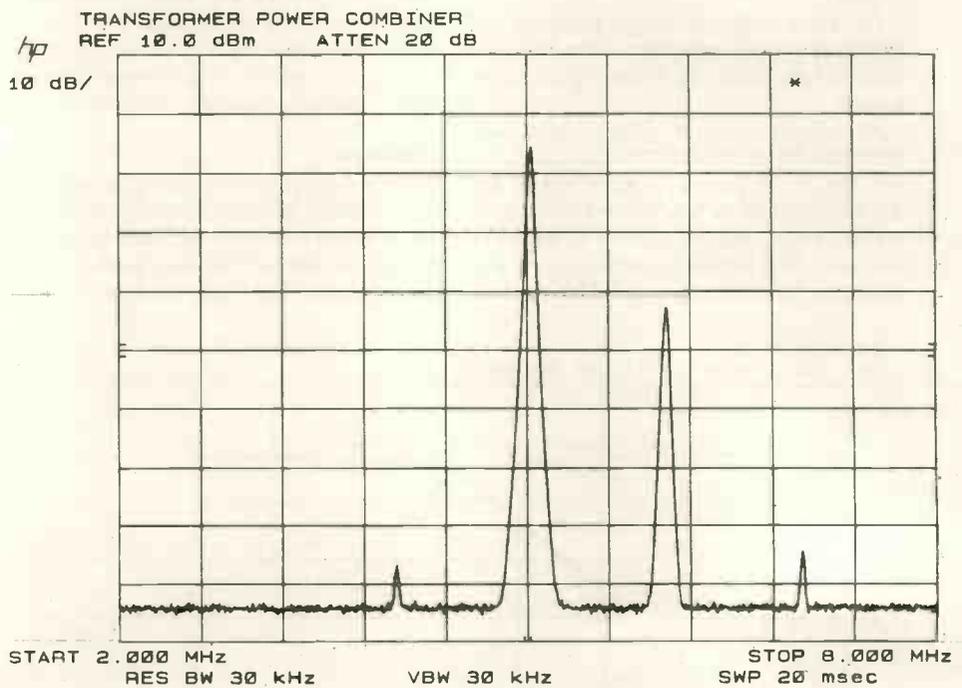


Fig. 4. Demonstrating the isolating action. Port 3 is connected to the spectrum analyser while port 1 has 5MHz signal applied (centre) and, at the same time, port 2 has a 6MHz tone applied. Both signal generator output levels are equal. The isolation achieved is about 30dB.

induced into the windings; the voltage at the auto-transformer tapping point appears equally at the two outputs. Imbalance in the output ports causes a current to flow in the 100Ω resistor connected between them: this limits the swr mismatch to 2:1 in worst case open circuit or short circuit on one port.

A model of the splitter omitting the auto-transformer – the source resistance given as 25Ω for clarity – is shown in Fig. 2b. This illustrates what happens when a signal is

applied to port 2. The ports all have 50Ω impedance, so the current flowing into port 2 with a one volt signal applied is 20mA. Since no signal should appear at port 3, it has a potential of zero volts, and it can be considered open circuit.

All the current must flow into the 25Ω resistor, producing 0.5V across it. Clearly the other 0.5V must be induced across one of the transformer windings (the upper winding in the diagram). By transformer action, the other

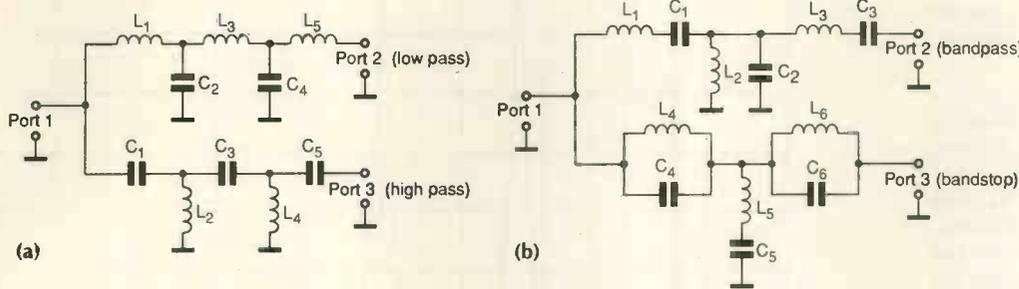


Fig. 5. Configuration (a) is a highpass/lowpass diplexer arrangement suitable for frequency splitting, for instance to a common aerial system. Diagram (b) demonstrates a bandpass/bandstop arrangement which could deliver a predictable termination to an impedance sensitive diode mixer.

winding will also have 0.5V induced into it. The voltage produced is negative with respect to the centre tap point. However, the centre tap potential is at 0.5V so, by subtracting the winding voltage, port 3 is at 0V. This also means that one volt appears across the 100Ω internal resistor, drawing 10mA from port 2. This accounts for the 3dB power loss when transmitting signals into ports 2 (or 3) and out of port 1.

The autotransformer simply translates the 50Ω system input impedance to the 25Ω presented at the centre tap of the second transformer.

An equivalent power splitter circuit was built from standard Mini-Circuits transformers. The circuit diagram is given in Fig. 2c. An autotransformer was not available, so a 1.414:1 turns ratio transformer (TMO2-1T) was used. The second transformer used just one centre tapped winding of a T1-1, the sec-

ondary winding being left open circuit.

With no adjustments, this simple equivalent circuit produced a useable performance, as shown in Figs 3 and 4. The spectral plot in Fig. 3 demonstrates the splitting action, giving equal outputs at ports 2 and 3. In Fig. 4 the isolating action is shown; port 3 is connected to the spectrum analyser while port 1 has 5MHz signal applied (centre) and, at the same time, port 2 a 6MHz tone applied. Both signal generator output levels are equal. The isolation achieved is about 30dB. Carefully designed splitters often achieve 40dB or more isolation. This is generally frequency dependent.

**Diplexers**

Diplexers are power combining and splitting circuits which have the advantage of producing no power loss. They use filters to separate or combine different frequency bands, and this limits their use. The filters may be either high-

pass and lowpass, or bandpass and bandstop. These two circuit configurations are illustrated in Fig. 5.

It is important that the -3dB frequency points are the same for both filters; a diplexer using a bandpass filter with a 1MHz to 1.5MHz passband must be combined with a bandstop filter that has a passband of 1MHz and below, and 1.5MHz and above. The case where lowpass and highpass filters are combined is illustrated by the frequency response given in Fig. 6. The two curves cross at the frequency where they are both at their cut off frequency (-3dB point).

Diplexers are valuable because the impedance at the common port remains constant across the band. This is true of a diplexer input, because if the signal frequency is outside the passband of one filter, it must be inside the passband of the other. The source sees a constant impedance, so its voltage does not change with frequency.

Applications for diplexers include mixer output impedance matching. Mixers need to see a constant 50Ω across a wide bandwidth at their output terminal. This constant bandwidth should extend beyond the input signal and local oscillator frequencies. If the mixer output terminal is connected to a low frequency amplifier stage, the impedance that it sees may be nothing like 50Ω at radio frequencies. A constant impedance can be realised by placing a diplexer with a lowpass filter between the mixer and amplifier, and a highpass filter to a 50Ω dummy load. Signals at frequencies outside the amplifier's range will be terminated, via the highpass filter, in the dummy load.

A diplexer can be designed using tables of normalised values, or by using filter synthesis computer programs such as *Filtech Professional*<sup>2</sup>.

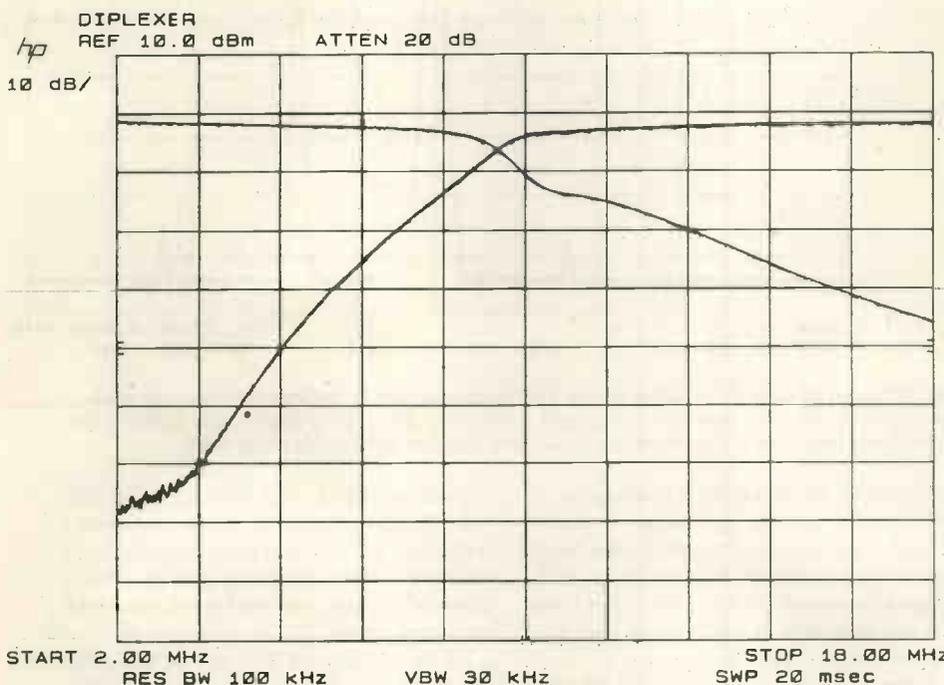
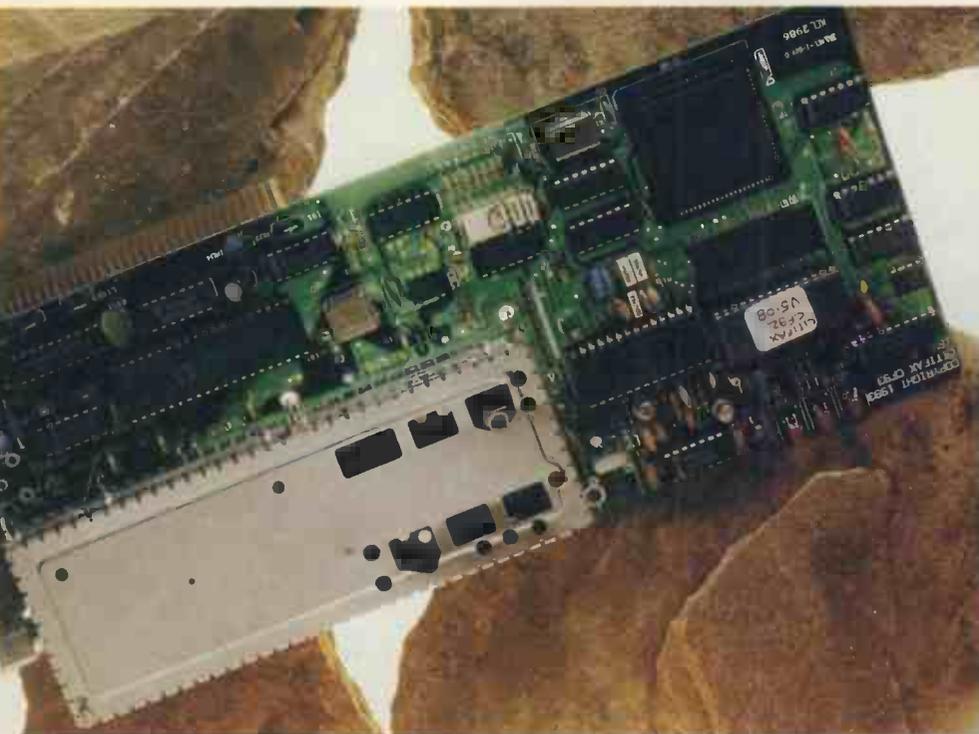


Fig. 6. Combined response of lowpass and highpass filters. The two curves cross at the frequency where they are both at their cut off frequency (-3dB point).

**References**

1. Mini-Circuits Ltd. Tel. 01252 835094
2. Number One Systems Ltd. Tel. 01480 461778





*Having described his tuner and decoder card for loading teletext data directly into a pc, Laurence Cook now discusses the software needed to access the teletext information via the pc's com port.*

## PCBs and software

Silk-screened teletext adaptor pcbs together with eeprom firmware and terminal software on disk are available. Details of the circuit for the adaptor, which features a video output for use with a monitor or multi-media, were presented in the August 1994 issue.

The board is double sided, with plated-through-holes, and has a gold-plated edge connector to fit an 8-bit pc isa bus slot). It has a silk screened component legend. Terminal software on disk allows teletext data to be displayed on the pc's monitor.

Originally, the pcb and software package was £65 inclusive, but the author has reduced this price to £45 inclusive due to demand. Send your cheque or postal order to L. Cook, 9 Goose Cote Hill, Bolton, Lancashire BL7 9UQ. Overseas readers please add £5 to cover extra postage. All enquiries relating to the adaptor should be accompanied by an sae and sent to Laurence at the above address – not to *EW+WW*. Laurence's telephone and fax number is +44-01204 307293.

# Programming for teletext on the pc

**T**he design published in August *EW&WW* for a pc card teletext adaptor has created a good deal of interest. This follow-up is intended to provide you with enough information to program the adaptor, by describing the command set and formats used to control the tuner and teletext processor.

Before describing how to communicate with the adaptor, it is worth mentioning the answer to the most frequent query I have had relating to the adaptor – “How do I set up the card for optimum reception?” The answer is simply “you don't”. It is supplied already set up for optimum results. There are no adjustments on the card, apart from the selection of com port and irq line. Even programmers can build the card, provided they can solder properly.

So how is teletext information made available to the pc? Simply via a standard serial com port using conventional asynchronous data on a row-by-row basis. It has a default bit rate of 9600baud (more on this later) just as if it had come from a modem or any other similar type of device.

The only quirk is the protocol used to specify and to encapsulate the data. Why does the data have to be supplied in this way, you may well ask? The answer is that the transmission system deals with ‘rows’ – which most people would call lines – of data. A ‘page’ of data, i.e. a screenful is 24 rows, cannot contain blank rows. This practice is sensible, since empty rows of forty spaces are a waste of valuable bandwidth.

So whereas, for example, a vdu might send two line-feed characters to include a couple of blank rows on a page, the broadcast teletext system simply omits to send those rows.

Normal pages, as seen on a domestic television receiver, comprise twenty four rows, numbered 0 to 23. The top row on the screen is special, and is called a header. It is row 0, and contains 32 characters for display (1 to 23 contain 40 characters each), the other 8 characters being used to identify the page to which the row belongs.

A ‘page’ of teletext is part of a ‘magazine’. What we see as page 601 for example is actu-

## Teletext adaptor command set

This is the command set and responses for the adaptor using the following convention:

- STX, EOT, etc, are ascii control codes, e.g. STX is 01<sub>16</sub>.
- Strings in quotes are literal, e.g. 'BC' is 42<sub>16</sub>, 43<sub>16</sub>.
- Commas are included for clarity only and are not used.

Function	Command	Response	Comment
RESET	SOH,'Z'	CF2000	Ver. 5.19
CHANNEL	SOH,'T',NN	SOH,ACK,'T',EOT	NN = 21 to 69
HEADERS ON	SOH,'H+'		
HEADERS OFF	SOH,'H-'		
SELECT PAGE	SOH,'P',pppssss, EOT	SOH,ACK,'P',EOT	s ls subpage

Function	Command	Response	Comment
SEND TESTPAGE	SOH,'P', FFFFFF',EOT	<Reset sequence>	page from ROM
SET BAUDRATE	SOH,'B',n		n=1 for 1200 n=2 for 2400 n=3 for 4800 n=4 for 9600 n=5 for 19200

Received data formats are as follows:

HEADERS	SOH,'H',N,STX,<40 characters>,ETX	n=magazine represented as an ASCII digit
PAGES	SOH,'P',pppssss,STX,' ',<40 chars.>for row zero r,<40 characters>r is (row+20) 7F HEX,EOT	page ends here

ally page 01 of magazine 6. This means that there are up to eight broadcast magazines. Zero to seven, you might think. No. They are referred to as magazines 1 to 8, but magazine 8 is number 0 while magazines 1 to 7 are numbered 1 to 7. This was chosen because the 3-bit binary number of a magazine, combined with the 5-bit binary number of a row, nicely fits an 8-bit byte which is contained at the start of each row transmitted. By choosing this scheme, the broadcasters can, and do, interleave rows from different magazines in a time division multiplex way, so improving page access time.

Luckily for the applications programmer, the adaptor sorts all this out and delivers a row prefixed with the appropriate number, as will be clear later. Content formats of rows, particularly those containing text and graphics, are best understood by referring to the Broadcast Teletext specification of September 1976.

Commands accepted by the adaptor are shown in the panel.

### Flow control

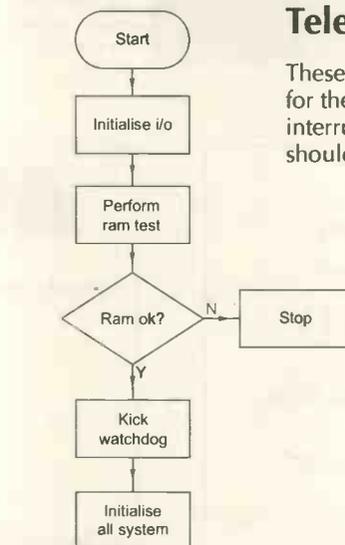
The adaptor uses out-of-band flow control. That simply means that instead of embedding XOFF (transmission off) and XON (transmission on) characters in the data stream, the modem control lines rts (request to send) and cts (clear to send) are used. These are hardware handshake lines that ensure data is not lost in either direction.

When programming the uart on the card, remember to have regard to the appropriate command and status bits in the relevant registers; unless of course you are using a comms library from C, in which case just call the pertinent functions.

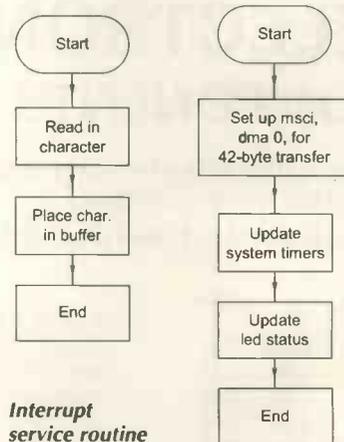
On the subject of uarts, note that the default rate is 9600baud and the data is asynchronous with 8 bits plus 1 stop bit and no parity. Also, you might have problems if you reprogram the

## Teletext adaptor software flow

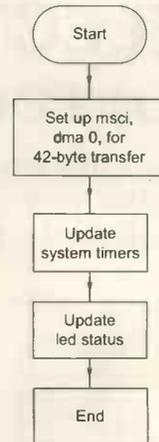
These simplified diagrams outline software flow for the teletext adaptor. Since the system is interrupt driven, the charts are discontinuous and should be treated as a guide only.



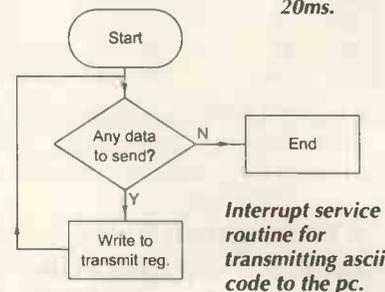
Main code flow for the teletext adaptor.



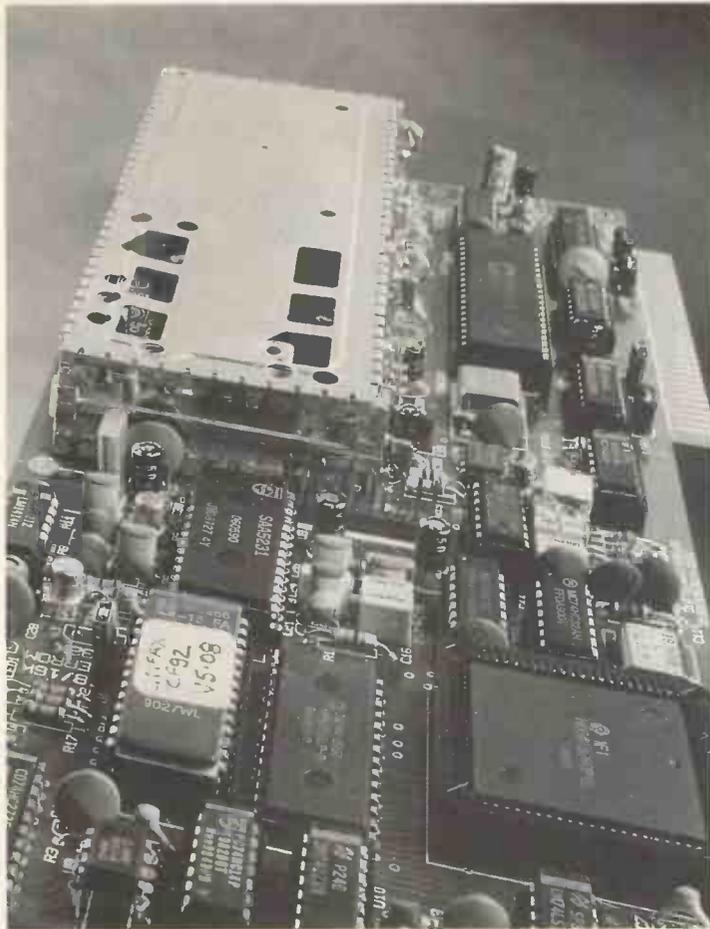
Interrupt service routine for receiving ascii from the host pc.



Frame synchronisation routine, used once every 20ms.



Interrupt service routine for transmitting ascii code to the pc.



Teletext adaptor for the pc incorporates a synthesized tv tuner and takes single 8-bit slot.

adaptor baud rate but forget to do the same for the uart. Of course, the two must always be in step for correct operation.

Throughout this short article, references to broadcasters can of course be widened to include cable operators. I have just subscribed to cable tv and I am amazed at the amount of worldwide data entering my home via teletext. I wonder how long it will be before the opportunities for distribution of information via tele-software are fully realised

**Hardware update**

Since I wrote the article describing my teletext adaptor for the pc in the August issue, Hitachi has introduced a new addition to the 64180 microprocessor family called the *HD64180S2CP10*. This is cheaper than the *HD64180SCP10* used in the original adaptor design. It should be a drop-in replacement for it, but I have not yet evaluated it.

The antenna socket on the *FQ844* tuner module needed a small flat filing on it in order for it to fit through one of the pc enclosure's expansion board access holes. There is now a new tuner called the *FQ916ME/LIF* whose antenna socket fits without modification. I am currently evaluating this part, and so far, all indications are that this is a drop-in replacement. As a bonus, it is also cheaper. ■

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# LETTERS

Letters to "Electronics World + Wireless World" Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

## Compulsive comment

As a compulsive audio-project constructor and potential user of ideas offered by Graham Nalty (*Letters*, December) I should like to say that a critical feature that distinguishes Douglas Self's amplifier from Graham's and other offerings is that of simplicity.

Exclusion of power supply regulation is a good start. Projects invariably provide the opportunity for the designer to offer components, multiple pcbs and often an upgrade service involving high-priced items and materials from what is becoming an array of audio boutiques. Moreover, these entrepreneurs tender a variety of 'improvements' to other people's designs with a similar range of expensive bits and pieces.

I can say that few projects have given as much satisfaction during construction and practical use as the D Self amp. The pcb alone is far superior to most encountered previously and the audible results are very acceptable.

I qualify that last point by stating that I do not enjoy the acute aural sensitivity of many of my contemporaries and judge by the effect of listening for substantial periods without fatigue.

Certainly, the generously rated components used at moderate domestic levels barely get warm and to that extent I am now installing four of the five boards built to date into one case. Current fashion dictates separate amps for each driver (plus the sub-woofer in the floor) and it is sensible to have them all at the same sensitivity. Moreover, the existence of logic decoders, and the novelty of effects available calls for even more amps and speakers, so there is no end in sight at present.

*EW + WW* has maintained its long-standing pre-eminence in sponsoring only projects that will stand the critical test of time. Long may it continue so to do.

**Hugh Haines**  
Sunderland

## Unacceptable terms

I note that Graham Nalty (*Letters*, December) sees fit to offer a set of suggestions for improving the class-

B amplifier design that I presented in my series *Distortion in Power Amplifiers, EW + WW passim*. All these modifications are supposed to give 'improved sound quality' a phrase which Mr Nalty chooses not to define, but which is left so vague as to be wholly unacceptable.

If Mr Nalty is unable to specify what effect his recommendations will have more closely than this, then it suggests he does not know himself.

The methodology used in the design in question, has, I think, been explained in greater detail than for any amplifier ever before. This has not dissuaded Mr Nalty from choosing to ignore the theoretical arguments, practical data and logical reasoning that I have tried to set out; he has simply sailed straight into a tinkering session based on reciting the usual old dogmas of subjectivism, which I suggest are by now becoming very tired indeed.

I do not think my amplifier is incapable of improvement. Indeed, having given the conventional architecture a fairly thorough workout, I have thought up some very interesting improvements in the last few months. But I am quite certain that stolidly replacing components with more expensive ones – apparently for no other reason than because they are expensive – is not the way to do it.

Firstly, all the exhortations about replacing perfectly good carbon-film resistors with expensive fetish items may be disregarded.

Improvement of the sound by this sort of thing is flatly impossible. I wonder if Mr Nalty just happens to sell Holco resistors, as he apparently sells expensive speaker cables? As for the rest of it, I have raked through the comments looking for any nuggets of truth, with the results I give below:

Nalty suggests: "Try replacing  $R_1$  with a polyester capacitor; or place a similar capacitor in parallel with  $R_2$ ."

Better not, as  $T_{r13}$  must have a source of base current to work, so replacing  $R_1$  with any sort of capacitor will stop the amplifier dead. A capacitor shunting  $R_2$  will almost certainly cause hf oscillation. On the other hand,  $R_1, R_2$  may be a typographical error for  $C_1, C_2$ , which has at least the advantage of making some sort of sense. However no

performance improvement can possibly result.

"I would prefer the feedback resistor to be high enough to use a polypropylene capacitor for  $C_2$ , if this can be done without creating a high dc offset at the output... remove  $C_4$  and  $C_5$ ..."

I would prefer to see Mr Nalty refrain from this sort of irresponsible tampering, which he cannot have thought through. I do not know what sort of polypropylene capacitors he has in mind, but the largest size stocked by RS Components is 0.47 $\mu$ F. This implies that  $R_8$  would have to be increased to about 2M $\Omega$  to maintain the same if response, and the resulting output dc offset would roam between  $\pm 10V$ , depending on transistor beta. Since a good deal of electronic design goes toward making sure that things do not depend on beta – which is often a wholly untrustworthy parameter – this hardly seems the way forward.

It is quite untrue that dc servos always sound awful, though it is certainly the case that they require some care in design. In the case of power amplifiers, however, I think I have demonstrated that if a  $\pm 50mV$  offset range is acceptable, neither servo nor dc trimmers are necessary, even if high-voltage low-beta devices are used for the input pair.

As for removing  $C_4$  and  $C_5$ , please don't do it.  $C_4$  provides signal feed-

forward around the  $V_{be}$ -multiplier, and without it there may be enough phase-shift in  $T_{r13}$  to cause instability.  $C_5$ , as I carefully explained, gives cleaner switch-off of the output devices and so reduces hf distortion. Am I to understand that Mr Nalty thinks that clean switch-off is a bad thing?

"There are audio transistors available which have smoother gain vs current graphs..."

Tell us what the part numbers are then. The basic transconductance linearity of a transistor (ie the  $I_c/V_{be}$  law) is fixed by physics and is about as negotiable as the law of gravity. If, however, this refers to beta variation with  $I_c$ , then the *MJ802/4502* have pretty much the same beta-variation as most large power devices; Normalised  $h_{FE}$  varies by a factor of about 4.5 over the  $I_c$  range 30mA-10A. As the published series of articles showed, this materially affects thd into 4 $\Omega$  loads or lower only. Mr Nalty's *MJ11016* shows a variation factor of 7.7 over the same  $I_c$  range, so his point seems somewhat obscure.

"The Darlington output arrangement is the best..."

In some respects it certainly is, but such a flat statement is misleading. *Distortion in Power Amplifiers* showed unequivocally that the CFP output stage has significantly better linearity and much more precise

## Sine post

Noting that Ian Hickman is interested in the study of waveforms ("Harmonising theory with practice", *EW + WW*, December, pp.98-1002), I wonder what he might have to say regarding the initiation of sinusoidal waveforms in simple series  $L, C$  and  $R$  circuit, by application of a step function of voltage to the combination.

Why are the voltage waveforms generated across the elements sinusoidal?

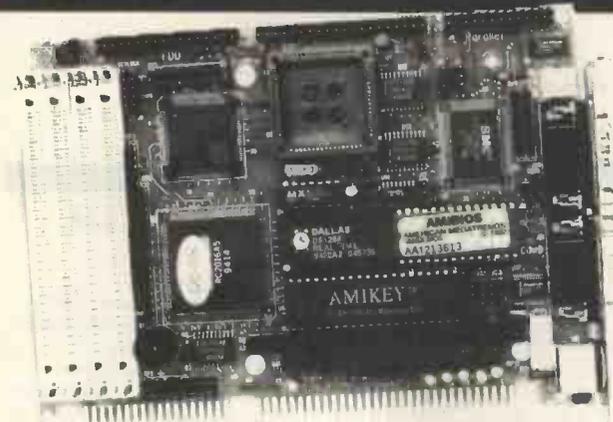
When a coil of wire is rotated between the poles of a magnet, it is obvious that the generated waveform should be sinusoidal, because the rate of change of magnetic flux linkage producing the voltage, is proportional to the sine of the angle of rotation of the coil. But there seems no easy physical property of the series  $L, C$  and  $R$  combination explaining why their voltage waveforms should be sinusoidal.

I know of course that we can prove, mathematically, via the solution of a differential equation for the circuit, that they should be so. But I am left wondering about the physical reasons.

**Peter Dawe**  
Oxford

*Ian is to reply to these comments in a future article.*

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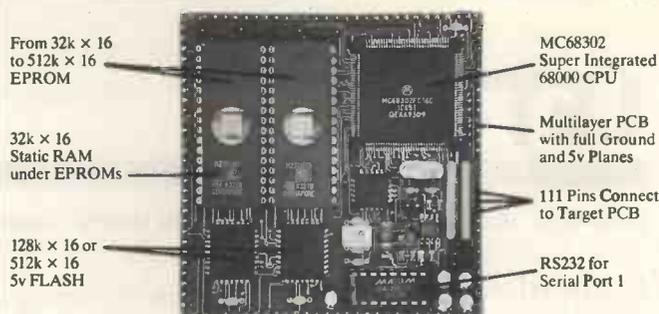


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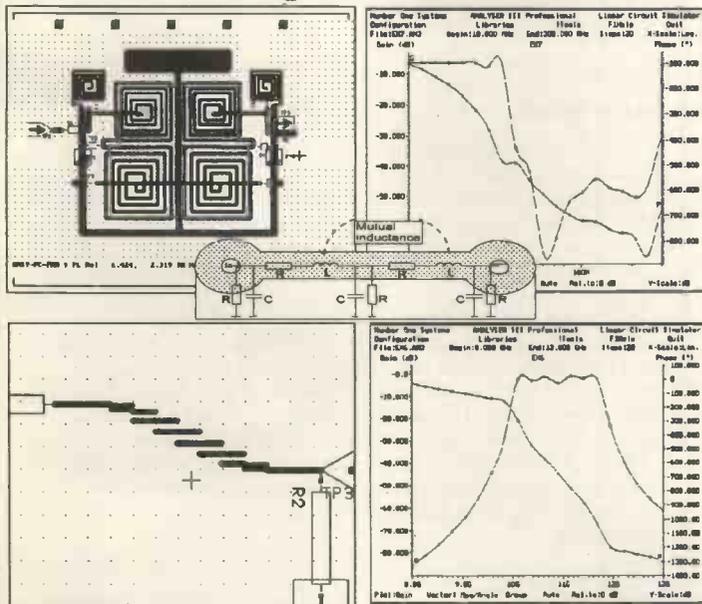
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quiescent control, and in many applications would probably be the preferred approach. Integrated Darlingtons give truly awful quiescent stability, and I have yet to hear of a good reason for using them in audio amplifiers.

"Constant current sources are good design." The simple RC decoupling applied to bias chain  $R_{21}$ ,  $R_{22}$  could indeed be replaced by a constant-current diode. One minor snag is that the cost is about thirty times that of doing it my way. In addition, an RC network can be relied upon to maintain effectiveness up to high frequencies, which is in general not true of 'electronic' decoupling methods.

"Cascode transistors above the collectors of  $Tr_2$ ,  $Tr_3$  and  $Tr_4$ ... offer further scope for improving sound quality"

Wholly untrue, I'm afraid. Cascoding the input pair  $Tr_{2,3}$  has the potential to improve negative-rail psrr, but the best rationale for doing it is to allow the input pair to be low- $V_{ce}$ , high-beta types, which will reduce the range of dc offset at the output. DC is not usually considered an audible phenomenon.

Cascoding the vas  $Tr_4$  is redundant as it is already linearised by  $Tr_{12}$ , and in any case cascoding alone is pointless because the vas collector impedance cannot be much raised when it is subject to the non-linear loading presented by the output stage. This was discussed at some length in the articles, and I rather wish Mr Nalty had read it.

I am unable to comment on the Apex and Virtuoso designs, as we are not told where or when they were published. Perhaps Mr Nalty might like to make the circuitry available for critical examination.

Finally, Mr Nalty remarks that "every single component in the audio chain degrades the signal" which is simply not true. From this point of view most of them are blessedly irrelevant. Even in the power amplifier there are very few components that have a major effect on the audio performance. This is, after all, the glory of negative feedback.

Mr Nalty's criticisms are couched entirely in the language of subjectivism. There is now a growing consensus that a school of thought which in 20 years has been quite unable to offer a single piece of solid evidence in support of its beliefs need not be taken too seriously.

Douglas Self  
London

## Slew review

Douglas Self seems to believe I was criticising his work (*Letters*, December) rather than simply

responding to anomalies raised in his articles.

I did not think it out of context to blame the input stage for the disparity in positive and negative slew since it is clear from the article that the primary effect of charge and discharge of  $C_{dom}$  is derived from the input stage.

It may be true that loading effects are influencing  $C_{dom}$  from the vas stage. But this was not really my point. I had prefixed my comment with the proviso "unless". Unless there were other mechanisms not alluded to, then the slow output devices would be responsible for the ultimate rate of slew.

In "Further complications" (High speed audio power, *EW + WW*, September, pp. 760-764) Self states that with certain values for  $R_4$  and  $R_{13}$ , the slew rate is +37 and  $-52V/\mu s$  for class B operation and an  $8\Omega$  load.

So I assume that the output devices are now connected in place of the test circuit and that slew rate figures refer to an actual completed amplifier. If such is the case then, as I quoted before, the condition for slew rate is: maximum frequency at which the amplifier can deliver a voltage corresponding to its rated full power.

Given the condition stated in my letter, Self's amplifier would need to maintain full power to a frequency exceeding 200kHz.

An alternative way of expressing the bandwidth of an amplifier, without using slew rate, is that if 20kHz is to be the accepted upper frequency response then the amplifier should be capable of delivering this *open loop* ie without feedback. With feedback applied, the upper frequency cut-off should be at least 20kHz multiplied by the feedback loop gain so that for 30dB of feedback, the cut-off should be 600kHz or more.

I have made as close an estimate as possible of slew rate and the alternative expressions based on the *The Audio Handbook* by Gordon J King, pp. 47 and 103.

On a personal note, I have auditioned bi-polar amplifiers over the years and where possible obtained the frequency/power curves. For those amplifiers with 100W and over, the rate of roll-off – with few exceptions – occurs around the 15kHz mark. It has in fact become endemic, and some amplifiers from reputable stables with multiple output devices start to roll off as low as 12kHz, while claiming slew rates of 30V/ $\mu s$  or more. Clearly this is an anomaly, and yes, they do sound dull compared to the mosfet amplifiers – which is hardly surprising.

Self's amplifier is being advertised as usable up to 100W

output. I had naturally assumed a high frequency penalty and had expected a qualification. As it stands the amplifier must remain the substance of things hoped for, and for me, the evidence of things not seen.

The mechanisms concerning  $C_{dom}$  can be found in J Linsley Hood's book *The Art of Linear Electronics*. Effects of the vas stage on  $C_{dom}$  – increasing its value by the stage gain to  $(A + 1)$  – and an alternative position for  $C_{dom}$  to overcome the problems experienced by Self's circuit topography can be found here: pp.119 and 163 refers.

With respect to the question of linearity and mosfets, Self will be able to find in both the above reference books categorical statements that mosfets are more linear than bi-polars – and perhaps more importantly why this is so.

I believe that the statistical evidence is in favour of the mosfet since even the worst examples of the Hitachi cloned circuits exhibit very low levels of distortion, with little effort. By contraindication, many of the bi-polar circuits, using more devices, are barely managing distortion levels 10 to 100 times worse. In fact many amplifiers could hardly be classified as *hifi* by the agreed standard, almost as if distortion is to be welcomed. Self's amplifier is very much the exception I'm afraid.

Dealing with the question of psychoacoustics, Self must be privy to information that most of us are not, since with all the work done over the years, we could have expected some of it to have trickled down into the audio world. I observe that no internationally agreed upper frequency hearing response has been established – or lower frequency response for that matter.

Sine wave expressions continue to be used for complex waves as does a criterion for testing amplifiers that has caused a division so great that there are now two distinct schools of thought (objectivity and subjectivity). That criterion is the double-blind listening test. Yet there seems to be no relevant technical paper to back it up, and main evidence appears to have been adapted from medical and scientific research methods using double blind tests. But the substance of those tests is based on an entirely different context. Exponents of this philosophy seem to believe that all properly-designed amplifiers sound the same, thereby at a stroke undermining all attempts by designers to improve circuit topographies, reduce distortion, and improve components.

I see no advance in the cause of psychoacoustics. But I do see the revival of the Zanzibar fallacy, the

story of which can be found in King's book.

Self's last comment is rather terse. I suspect we have all played with circuits. Surely that does not put expertise into question.

VJ Hawtin  
Middlesex

Douglas Self replies

I can reassure Mr Hawtin that I am in no way outraged by his comments, but I am saddened that he found my writings somewhat less than lucid.

I stand by my statement that while an amplifier input stage is the primary determinant of slew-rate capability, it is vital to remember that current sourced or sunk by this stage must also be sunk or sourced somewhere else if it is to complete its circuit. This typically adds limitations that render the maximal slew-rates asymmetrical. Similarly, it is easy to overlook how current siphoned off into other stages e.g. via the input impedance of the output stage, which is naturally load dependent can degrade slewing.

Output stages do not have an inherent slew-rate limit, although they naturally have bandwidth limitations. Mr Hawtin is unclear about the difference between these two parameters; they are in no way interchangeable, and can be altered independently of each other in most cases. An output capability that changes with frequency is no guarantee that there is a slew-limiting problem; there are other ways this can come about.

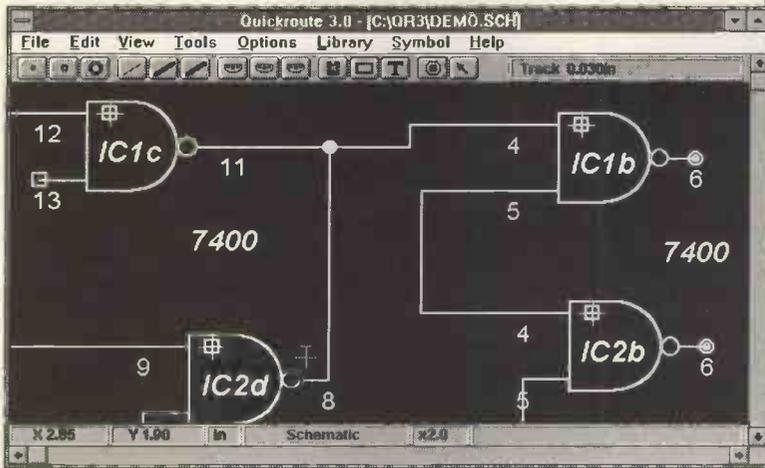
To move on to power fets, I am having some difficulty believing in the large body of "statistical evidence" that Mr Hawtin claims to have in support of his statement that mosfets are just as linear as bipolar, if not more so. Perhaps he will be good enough to let us see it, so we do not have to rely on his word? Or perhaps, like some other correspondents, he feels that to actually present evidence is in some way distastefully vulgar, and that it should merely be alluded to with a knowing smile.

'Hitachi-clone' mosfet amplifiers rarely seem to do better than 0.01% thd at 1kHz in practice, and I believe this to be determined very largely by the intrinsic Class B crossover non-linearity of these power devices. I think it would be very difficult to obtain less than 0.001% thd from mosfets in relatively straightforward circuitry, but this can be done routinely with bipolar if the correct design methodology is used.

Why Mr Hawtin should think that psychoacoustics is some sort of secret science I cannot imagine. Textbooks on the subject are available in any decent academic bookshop. A good introduction is

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the work, *An introduction to the psychology of hearing* by Brian Moore (Academic Press). It includes a section on psychoacoustics and *hifi* that illustrates why subjectivist notions are not necessary to explain the ear/equipment relationship.

Finally, I can make no sense of Mr Hawtin's repeated claim that all bipolar amplifiers have a bandwidth limited to 12 or 15kHz; since this is patently untrue, there is no point exploring the topic.

Doug Self

## Heartfelt criticism

Recently one of my customers wrote to thank me for the immense pleasure and satisfaction he was experiencing when listening to music reproduced through an amplifier manufactured by my company. Immediately, not to fall foul of the Trade Descriptions Act, I had to contact him to tell him that, though my test equipment had confirmed that his amplifier did indeed meet its published spec, what he had said was a) merely his opinion, b) not supported by any form of scientific evidence or research, and c) almost certainly invalid anyway because his hearing was far too unreliable, non-uniform and too easily fooled to be relied upon for a meaningful value judgement.

He faxed me back to say that if I was going to talk such a load of old twaddle then, thank you very much,

but he'd rather get on with listening to his own choice of music at his own choice of level. So would I.

Now a question for Phil Dennis and Douglas Self: what about the great symphonies, the playing of a lovingly-crafted violin, early morning bird song? You can't use a scope to explain why or how such things come in through our ears and touch our hearts... but it happens.

I tell you what I hear, you tell me I don't – although to 'prove' your prejudices you have to resort to derision, selective argument and syllogistic reasoning.

Presumably, as true objectivists, you must also deny the existence of beauty, or harmony, or perhaps even hunger as not being scientifically provable? Anyway, no more. I give in. Goodbye.

**Jerry Mead**  
Mead & Company  
Herts

## Hand-shakey?

In the scores of pc-related I<sup>2</sup>C implementations, nearly no one takes into account the fact that both the scl and the sda lines must be bidirectional, for handshake purposes.

The simple interface given in John Davies' article (*I<sup>2</sup>C via the PC*, December, pp.994-996) might well work in some simple cases, but you will hit trouble as soon as you try to use e<sup>2</sup>proms, or a-to-d/d-to-a converters; or the PCD8584 bus controller (even as a slave).

Hardware must provide an input to read the state of the scl line, and the software must provide a way of reading that input.

**Jean-Paul Brodier**  
Cuvry  
France

## Electrolytic detection

I noted the various electrolytic detectors described in George Pickworth's article (*Detection before the Diode*, pt 1, December, pp.1003-1006). The early pioneers seemed to use platinum or lead in various diluted acid solutions. But I wonder if other materials and solutions were tried as detectors.

Some electrolytic rectifiers which may be of relevance are:

Al Pb	10:1	forward:reverse current
Al Fe	10:1	
Cu Pb	12:1	
Zn Pb	4:1	

Also usable is zinc-aluminium alloy rod or cylinder with an iron tube.

In a saturated solution of phosphate of ammonium, current direction, alloy-to-iron results in film deposit on the surface of the alloy. Current direction, iron-to-alloy, results in reduction of film.

I think the claim was 85% efficiency, but would appreciate any comment related to this matter.

*Unsigned*  
Middlesex

## Coil question?

I am searching for several formulas involved in constructing inductors and coils using different configurations. I am familiar with formulas for calculating the number of turns, inductance, dimensions, etc, for single and multi-layer round inductors. However the project I am presently working on involves miniature coil construction using three-dimensional square and rectangle cores and two-dimensional pcb etched inductors.

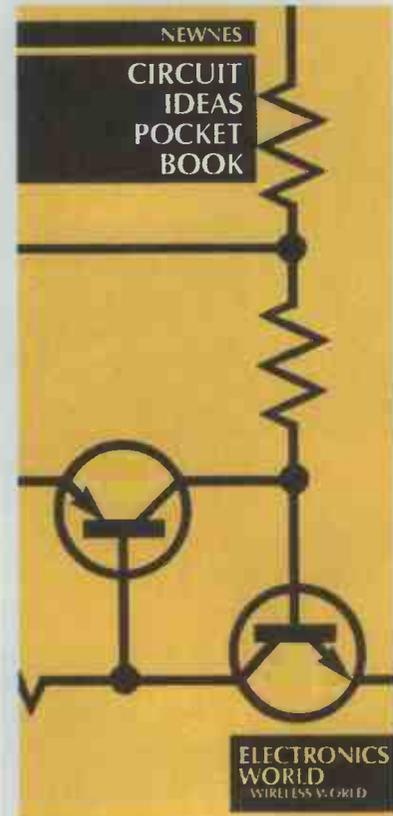
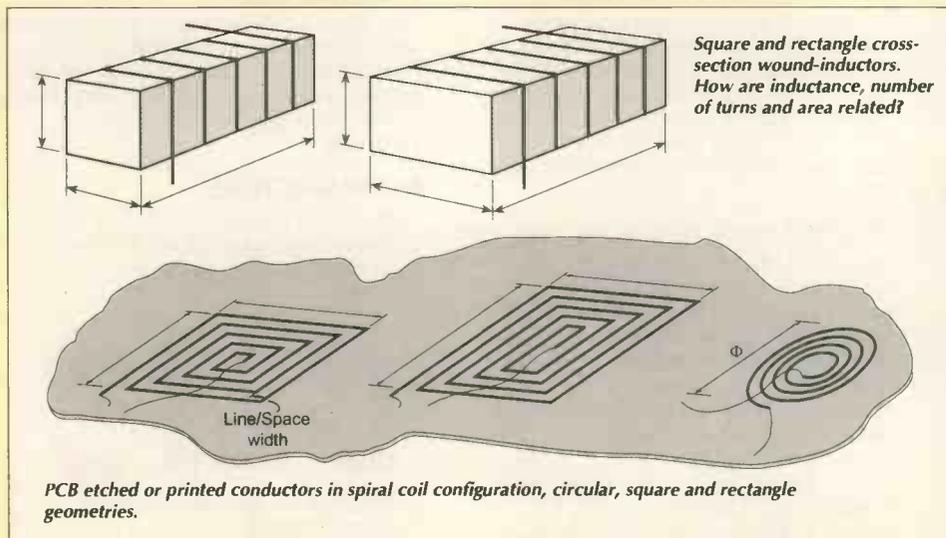
I have searched through numerous reference books

without success and would be most grateful for any help.

Formulas required are indicated in the accompanying figures. For example, I may wish to calculate the inductance from the number of turns etc or the coil design from a given inductance.

Any information will be greatly appreciated.

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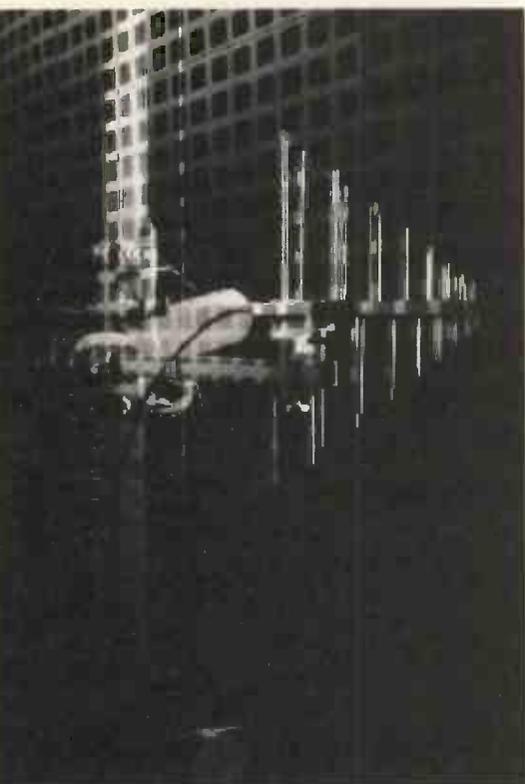
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*Norm Dye and Helge Granberg have shown how to develop the tools needed to design a small signal amplifier and how to determine source/load impedances. Here they bring all the tools together and suggest circuits that will realise those impedances. From the book RF Transistors: principles and practical applications.*

## Determining circuits for small signal amplifiers

**I**nput impedance for the narrow band example where the match is for optimum gain and  $S_{12} = 0$  (example 1) is plotted in Fig. 1. The Smith chart shows  $\Gamma_S = 0.6$  at an angle of  $-156^\circ$ . This is the value already determined from plots of optimum source and load reflection coefficients – see *EW + WW* January issue for more information – and the problem now is to take the  $50\Omega$  source and make it look like  $\Gamma_S$ .

The most common circuit is a low-pass filter configuration consisting of a shunt  $C$  and a series  $L$ . When using the Smith chart with shunt elements it can be used as an admittance chart, and when using it for series elements it can be used as an impedance chart. For this reason, Fig. 1 is plotted on a special Smith chart graph that simultaneously displays both normalised impedance and admittance circles.

The plot also demonstrates that a shunt capacitor having a susceptance value of  $+j1.75$ , which is the arc length from  $O$  to  $A$ , and a series inductance having a positive reactance value of  $+j0.225$ , arc length from  $A$  to  $B$ , will rotate the  $50\Omega$  source into the normalised value of  $0.6$  at  $-156^\circ$ .

Calculating the actual values that give these normalised results at a frequency of  $1\text{GHz}$  is a simple matter and the values turn out to be  $5.6\text{pF}$  and  $1.8\text{nH}$ .

Similarly, Fig. 2 shows the output load

impedance, also previously plotted. Again a network of shunt  $C$  and series  $L$  will be used to transform the  $50\Omega$  load to the desired load impedance represented by point  $D$  in Fig. 2. The arc from  $O$  to  $C$  calls for a shunt capacitor having a normalised susceptance value of  $+j0.5$ , and the arc from  $C$  to  $D$  needs a series inductance with a normalised value of  $+j0.45$ .

At a frequency of  $1\text{GHz}$ , these can be realised by a shunt capacitor of about  $1.6\text{pF}$  and a series inductance of  $3.6\text{nH}$  (Fig. 3).

### When $S_{12}$ is not zero

For the narrow band, match-for-optimum-gain instance, where  $S_{12} \neq 0$  (example 2) the output matching circuit is of course identical to example 1.

But slight changes in  $\Gamma_S$  and the shunt susceptance and series inductance are needed to transform the normalised  $50\Omega$  source impedance (centre of the chart) to the normalised value of  $\Gamma_S$  (Fig. 4).

Arc between  $O$  and  $A$  represents a shunt susceptance, i.e. parallel capacitor of  $+j1.6$  and arc  $AB$  represents a series reactance, in this case a series inductance of  $+j0.275$ . Component values at  $1\text{GHz}$  are  $5.1\text{pF}$  and  $2.2\text{nH}$ . The circuit is identical to that of example 1, shown in Fig. 3, except for the slight change in component values in the input matching network.



cuit is also identical to that shown in Fig. 2. Figure 6 is the final circuit for this example.

**Broadband design**

Gain circles for the broadband design, example 4, have been plotted previously but here they are re-plotted in Fig. 7 using the special

Smith chart with normalised impedance and admittance circles. The objective is to devise a matching network that sits on the  $-4.5\text{dB}$  gain circle at  $500\text{MHz}$  and at the same time on the  $+1.1\text{dB}$  gain circle at  $1\text{GHz}$ .

In this instance, the *Mmicad* optimiser program can determine an appropriate network,

consisting of a shunt  $C$  and a combination of series  $C$  and  $L$ . A similar solution could have been achieved by an iterative process of adjusting the circuit values until a combination is determined which meets both objectives.

At  $500\text{MHz}$  the shunt susceptance moves along arc  $OA$ . The same capacitor will move

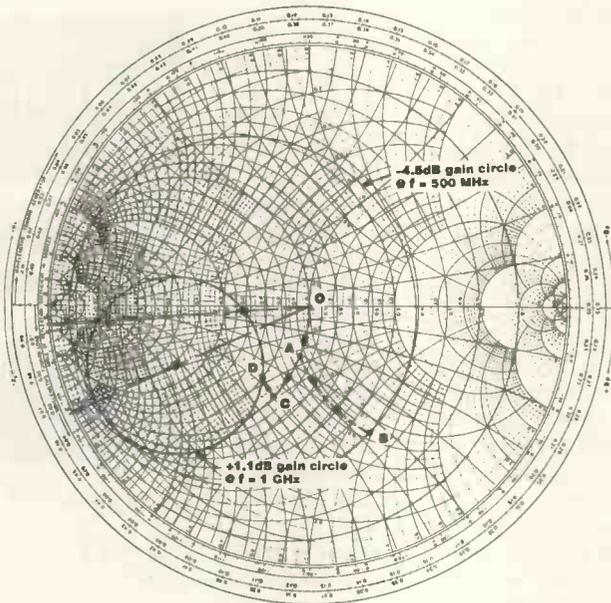


Fig. 7. Input matching for broadband design.

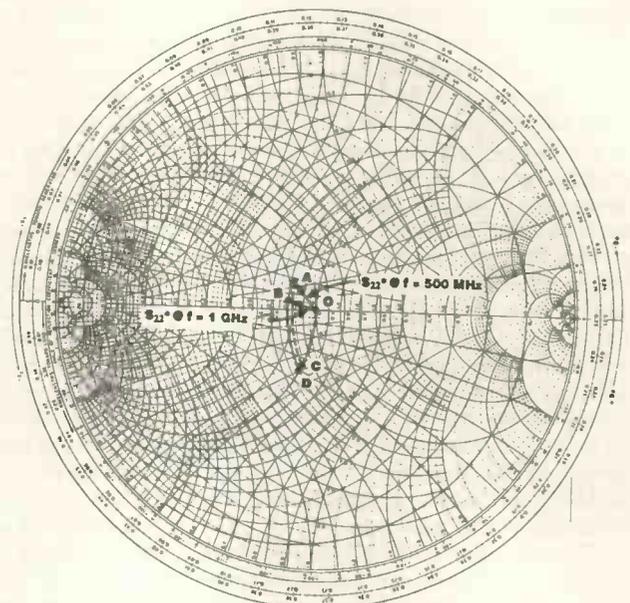


Fig. 8. Broadband design output matching.

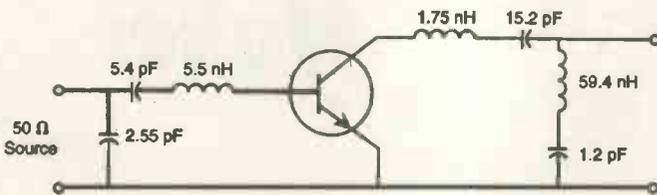


Fig. 9. Broadband design circuit.

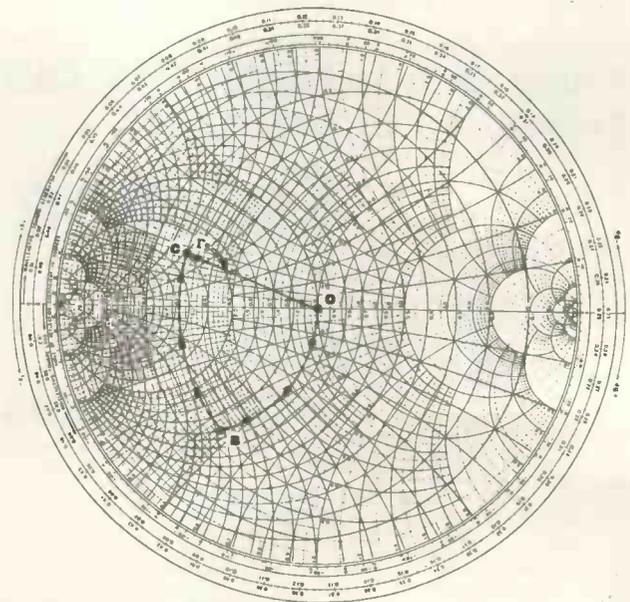


Fig. 10. Input matching for low noise.

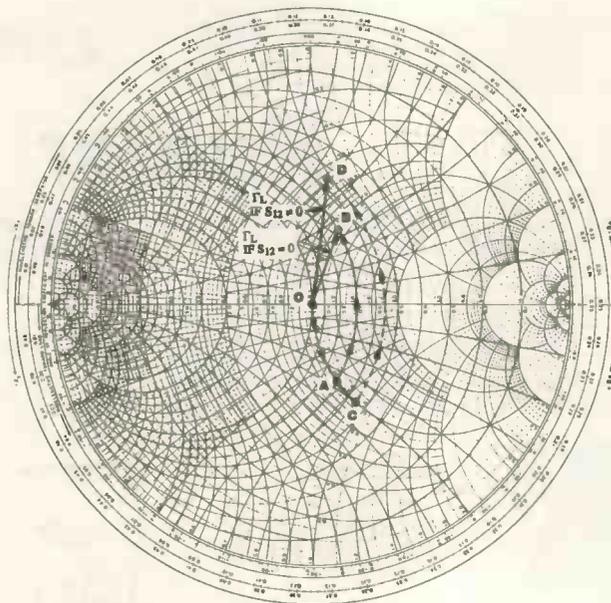


Fig. 11. Output matching for low noise.

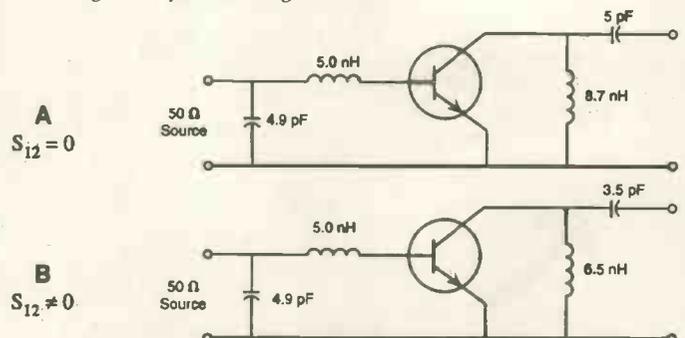


Fig. 12. Circuit configurations for low noise.

to *C* when the frequency increases to 1GHz. When the frequency is at 500MHz, the series network must move along arc *AB* while the same network must move from *C* to *D* at 1GHz.

Values that will achieve these objectives simultaneously are a shunt susceptance of  $+j0.4$  at 500MHz, increasing to  $+j0.8$  at 1GHz and an overall series reactance that is  $j0.84$  at 500MHz but  $+j0.09$  at 1GHz.

Component values that give the desired susceptance and reactances are a shunt capacitor of 2.55pF and a series capacitor of 5.4pF along with a series inductance of 5.5nH (Fig. 9). Because all the gain correction is placed in the input network, the output is the conjugate match, or  $S_{22}^*$  at each frequency, points plotted in Fig. 8. The problem now is to find a network that will simultaneously locate the matched load (*O* in the centre of the chart) at point *A* when  $f=500\text{MHz}$  and at *B* when  $f=1\text{GHz}$ . A shunt LC network with a series LC network is chosen such that the shunt network in Fig. 8 moves along the arc *OC* at 500MHz, but only moves from *O* to *D* at 1GHz.

Likewise, the series network moves the normalised line impedance from *C* to *A* at 500MHz, and moves the normalised line impedance from *D* to *B* at 1GHz.

Values of *L* and *C* that accomplish these objectives can be determined by solving relatively simple simultaneous equations relating

the susceptances and reactances at the two frequencies of interest.

Actual values, along with the input network, are shown in Fig. 9.

#### Designing for low noise

The value of  $\Gamma_S$  taken from a previous plot of the 2dB noise figure and 12dB gain has been re-plotted on the special impedance and admittance Smith charts used to determine graphically the impedance matching networks needed for a particular application (Fig. 10).

Matching is readily achieved by using a shunt capacitor to transform the 50Ω source from *O* to *B*. A series inductance finishes the match by transforming the value of *B* to the desired  $\Gamma_S$ , shown as *C*. Values of the shunt capacitor and series inductance that achieve the desired source reflection coefficient are 4.9pF and 5.1nH, Fig. 12, and output impedance matching is shown in Fig. 11.

The matching network, if  $S_{12} = 0$ , is a series capacitance and a shunt inductance that transforms the matched load first to *A* (series capacitor) and then to *B* (shunt inductance), or in the case of  $S_{12} \neq 0$ , to *C* (series capacitor) and then to *D* (shunt inductance).

So for  $S_{12} = 0$ , arc *OA* is a reactance of  $j0.64$  which translates into a series capacitor of 5pF. Likewise, arc *AB* is a susceptance of  $+j0.91$  which can be realised by a shunt inductance

having a value of 8.7nH as shown in Fig. 12a.

If the measured  $S_{12}$  value is used, matching requires a reactance represented by arc *OC* and a susceptance represented by arc *CD*. These values are realised by a series 3.5pF capacitor and a shunt 6.5nH inductance, Fig. 12b. ■

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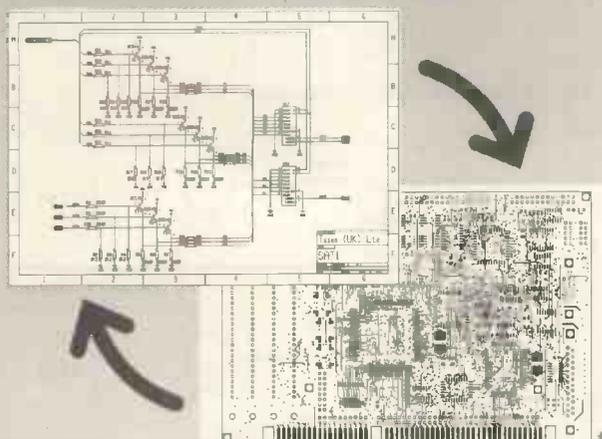
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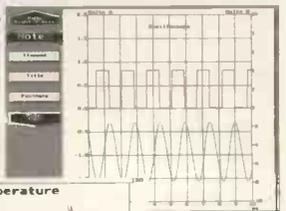
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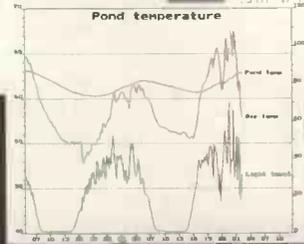
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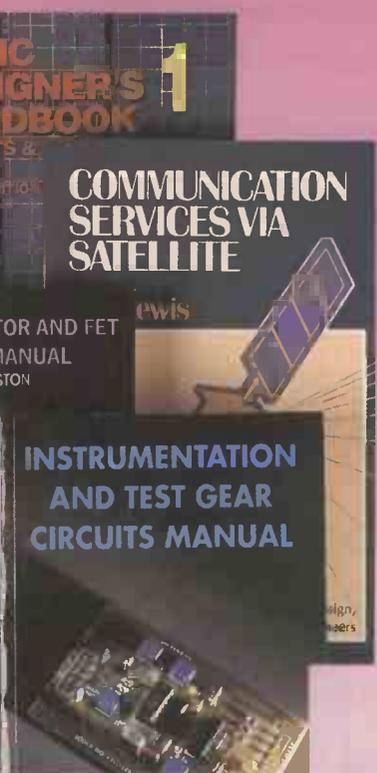
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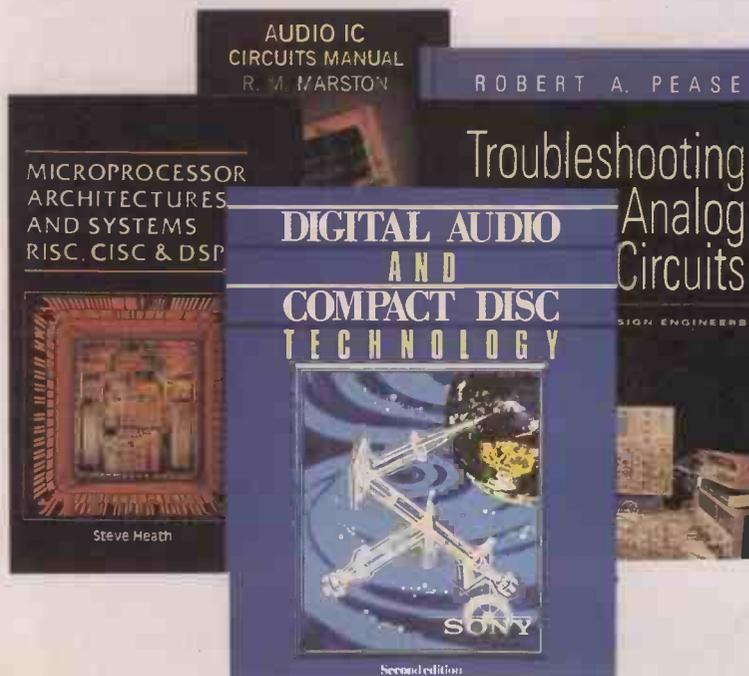
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# Analogue design with a 5V supply

*Op-amp selection can minimise the impact of single-supply design. Walt Jung and James Wong\* explain why careful choice is so important.*

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\*Walt Jung and James Wong are with Analog Devices Inc. This article first appeared in EDN.

System designers are turning to single-supply power not just in digital applications but, increasingly, in analogue circuits. Some loss in performance can result. But using amplifiers expressly designed for the job can help minimise this loss, and among the performance specs that will be enhanced are supply-power and operating ranges, dynamic range, and input and output ranges. Better linear operation overall should result too.

Op-amps operating from +15V supplies generally accept an ample input common-mode (cm) range and provide a wide output range – both typically +10V or more. But single-supply systems often handle much smaller voltage ranges: 12V down to 5, 3, or even 1.5V. At these levels, most dual-supply op-amps simply cease to function. Those that will operate suffer performance degradation, and so choice of single-supply amplifiers is much more limited than that for dual-supply.

General-purpose amplifiers such as 741s and 1558s – and most fet-input amplifiers – can operate from supplies as low as ±5V, or 10V total. Precision-amplifier families such as the OP-07 and OP-27 do not work below 5V, though a few selected family types may be useful for some regions of this voltage spectrum. For example the OP-97 proves useful down to ±2.25V (4.5V).

But, typically, only amplifiers specifically designed for single-supply or low-voltage operation can function without major performance compromises below 10V total.

Devices designed for ±15V operation typically require 2-5V of headroom, with respect to both supply rails and at both inputs and output. Even the more-flexible dual-supply amplifiers only come within 1V or so of the rails. Need for headroom in single-supply applications severely limits linear operation

over wide signal swings. The limitations are most acute at lower supply voltages and, for this reason, single-supply amplifier input stages usually remain linear even when the common-mode voltage applied is right at the negative rail (ground) or even slightly below ( $\approx -200\text{mV}$ ). Op-amp input stages that operate in this way include pnp bipolars, cmos or pmos types, and n-channel jfet types.

## Common-mode stresses

Obviously, devices with the above various input stages will have differing input-bias currents, noise voltage and currents, and offset voltage and drift. Spec sheets will delineate these characteristics, but a subtle (and potentially important) characteristic for a given device is how it reacts to common-mode stresses.

Over-driving amplifiers below the negative rail (pnp inputs in particular) can cause a non-destructive output-phase reversal. Over-driving may not actually damage the device if the input-fault current is limited, but the circuitry around the amplifier can react violently. If the circuitry is a polarity-sensitive servo, for example, it may lock up or oscillate.

Few single-supply amplifiers have internal protection against over-drive. Fortunately, preventing it is easy: simply clamp the input voltage to no more than 300mV below the negative rail with a low-threshold Schottky diode.

Protection of cmos- and pmos-input stage amplifiers is more difficult because exceeding their supply rails can trigger a parasitic scr within the device – with possible destruction. As with any cmos device, possible transients should be kept in mind and data-sheet recommendations followed.

At the positive end of the input-cm range, some devices can operate to within 1V of the rail while others may need 1.5V of headroom.

Generally, over-driving an op-amp's common-mode inputs above the positive rail causes non-linearity (as opposed to phase reversal) and, again, cmos- or pmos-input devices should not be driven beyond the positive rail.

With the exception of rail-to-rail output-stage devices, most op-amps can swing only to within 1-2V of the positive rail and when these devices operate from supplies of 5V or less, output swing is greatly reduced. Even if the input's noise floor remains constant - it seldom does - the s/n ratio and dynamic range suffer for devices operating from low voltages.

### Ultra-low quiescent current drain

A fundamental reason for using single-supply designs is to conserve power, so most single-supply op-amps draw low stand-by current. But very low current designs may involve fundamental trade-offs, and band width, slew rate, and input noise voltage are sometimes all sacrificed.

Stand-by current drain per amplifier channel can often make or break a device when a system's power consumption is critical. (There is no industry standard definition of low power. We suggest that devices drawing quiescent currents of  $\leq 1\text{mA/channel}$  are low power, and those drawing  $<100\mu\text{A/channel}$  are micro-power.)

Given a requirement for low-power designs, engineers face not just slower speeds but also band-width restrictions arising from higher circuit impedances.

Table 1. Noise vs supply current trade-offs.

	Noise (nV/√Hz)	supply current/channel
OP-295	51	160μA
AD820	12.5	750μA
AD822	12.5	750μA
OP-213	6	1.45mA

Table 2. Current vs band width trade-offs

	Current drain (μA/ch.)	Bandwidth	Slew rate (V/μs)
OP-295	160	75kHz	0.03
AD822	750	2MHz	3.5
324/358	500	1MHz	0.6

fully-buffered op-amp follower stages for the lowest wide-band dynamic impedance. When the dynamic current is low, a well by-passed noise-free divider is suitable for high-impedance loads. When high, a simple op-amp follower (Fig. 1) can do the job. In the circuit shown, a wide range of supplies can be used to produce a low-noise output of  $V_{\text{supply}}/2$ . Choice of  $IC_1$  determines the circuit's standby current.

If a precise pseudo-ground voltage is needed, a reference IC that can source and sink current (Fig. 2) should be used.

Operating from a 5V supply, the suggested

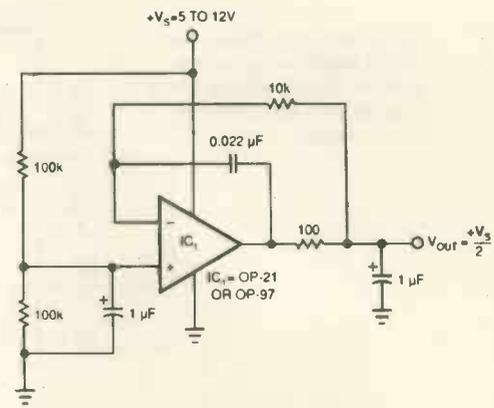


Fig. 1. False or "pseudo" grounds can be made any value desired. When dynamic currents are high, use this simple op-amp follower.

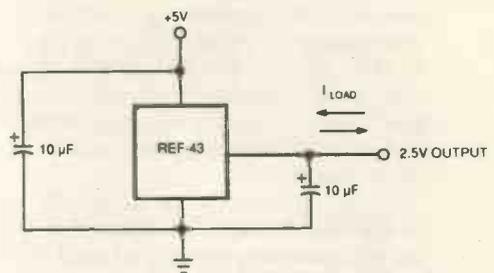


Fig. 2. Use a reference IC that can source and sink current if a precise pseudo-ground voltage is needed. This circuit can handle load currents both into and from the 2.5V source. Low-esr by-pass capacitors help keep the circuit's ac impedance low.

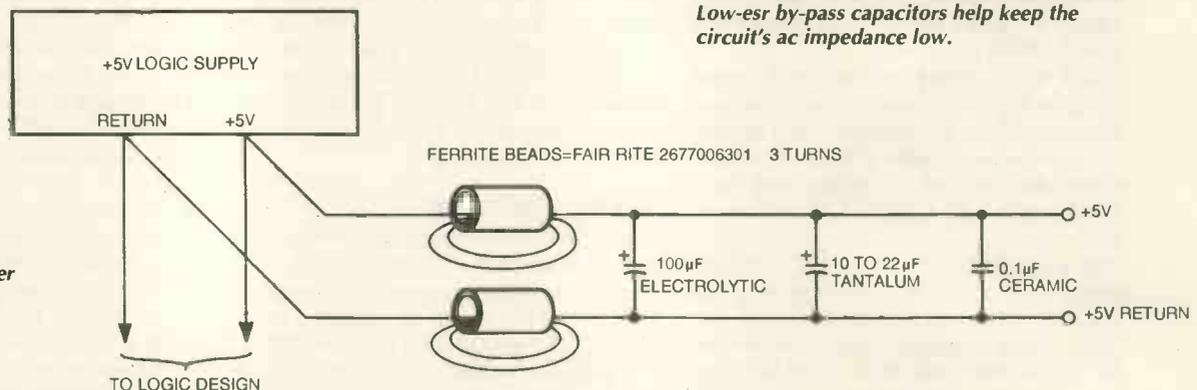


Fig. 3. Balanced IC filter suppresses glitch noise by 40dB or more.

Table 1 indicates trade-offs in noise versus supply current and Table 2, band width. The dual amplifiers in the Table 2 bracket the industry standard 324 and 358, with their  $500\mu\text{A/channel}$  current drain, 1MHz bandwidth, and  $0.6\text{V}/\mu\text{s}$  slew rate. Greater speed also comes at the price of increased current.

### Ground rules

Assigning ground is a task that can become important in single-supply, ac-coupled op-amp circuits. A false or pseudo ground can be anything required, the best choice depending on the individual application. But maximum amplifier flexibility helps make signal referencing less problematic, particularly if the design has substantial dynamic current into an elevated ground.

Possible choices for false ground range from simple ac-by-passed resistive dividers, to

circuit can handle load currents both into and from the 2.5V source. Low-esr (equivalent series resistance) by-pass capacitors help keep the circuit's ac impedance low.

### Power supply noise

Power-supply noise, if not dealt with correctly, can shred even the best paper design. Low-power circuits tend to have poorer supply rejection, and the noise from commonly used 5V digital supplies is nearly worst-case. Simple power-line filters may not be adequate for high-performance analogue stages. A common analogue and digital supply may be attractive for size and cost reasons, but a common supply should be avoided if at all possible. Power taken from the middle of a logic layout contains huge amounts of high-frequency noise - 100mV or more. Worse yet, logic supplies are typically switching types,

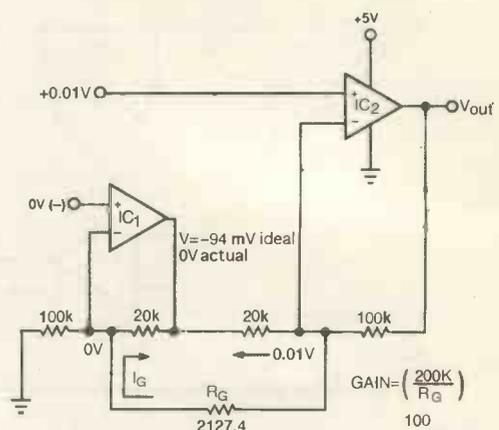
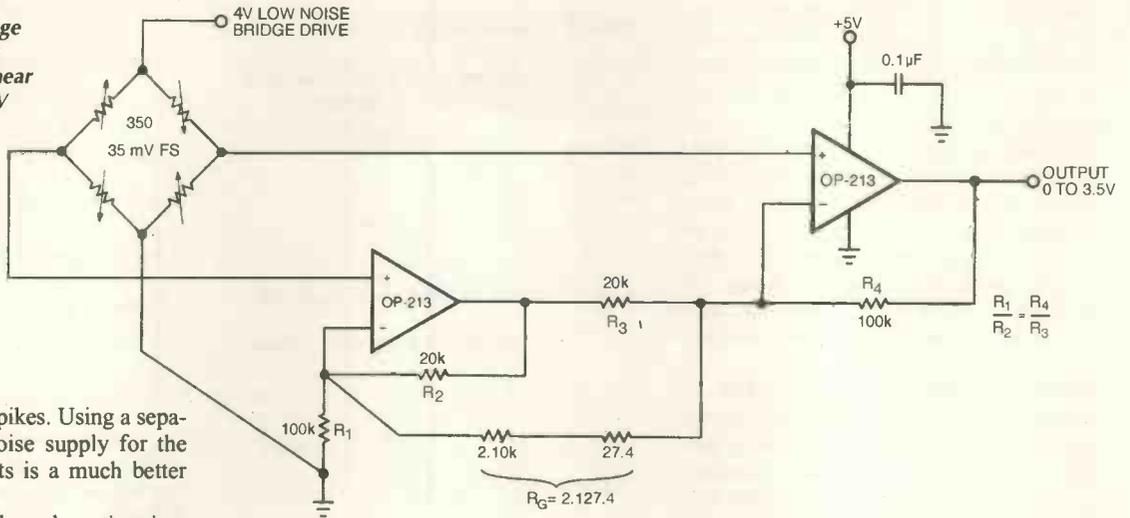


Fig. 4. Two op-amp instrumentation amplifier normally uses two supplies. Using a single supply constrains the design's common-mode range.

Fig. 5. Accurate strain gauge bridge amplifier operating from 5V and producing linear outputs to as low as 600µV from ground.



which have large output spikes. Using a separate, linear mode, low-noise supply for the sensitive analogue circuits is a much better idea whenever possible.

If a 5V logic supply is the only option, isolation and circuit partitioning, as well as optimised decoupling and filtering, can help greatly. Make sure the analogue power is tapped right at the supply terminals, not from the logic stages, so avoiding the voltage drops of the logic-supply runs and allowing maximum supply regulation.

By-pass capacitors alone are usually not adequate for filtering switch-mode glitches, so take additional steps such as a balanced IC filter (Fig. 3). The filter comprises two high-frequency inductors – each made from a three-turn, toroidal ferrite bead – and a large, composite output capacitor. Capacitors should be low-esr switching types for best performance. As shown, the filter suppresses glitch noise by 40dB or more.

Even with a device that has a wide output swing, an application's configuration can limit that swing. Take, for example, a two op-amp instrumentation amplifier (Fig. 4). It normally uses dual supplies, but in this case, has a single supply and a gain of approximately 100.

At first glance, the amp will faithfully amplify a 10mV differential input to 1V at  $V_{OUT}$  with a low common-mode voltage.

But a close examination proves otherwise. The output from amplifier  $IC_1$  must be a neg-

ative 94mV to satisfy the loop's requirements and produce 1V at the output. Obviously,  $IC_1$  cannot produce a negative output from a single supply. Indeed, if the circuit attempts such operation,  $IC_1$  saturates, and  $V_{OUT}$  becomes non-linear.

As it turns out, this input-cm limitation is a function of the instrumentation amp's gain as well as the output swing of the op-amps within it. So, low negative- $V_{sat}$  or rail-to-rail output-stage op-amps help improve common-mode handling. In Fig. 4 the worst-case minimum cm voltage is 0.4V (assuming a 5V output swing and each op-amp swinging to within 100µV of each rail). Op-amps that cannot swing to 0V will have a worse input-cm voltage minimum. The conclusion is that even though an op-amp might be designed for single-supply operation, a particular configuration could still constrain it.

**Sensible application**

But despite its basic limitations, the topology can still be useful if it is carefully applied. For example, consider an accurate strain-gauge bridge amplifier circuit in which the bridge and amplifier are powered from 5V (Fig. 5). The amplifier produces linear outputs to as

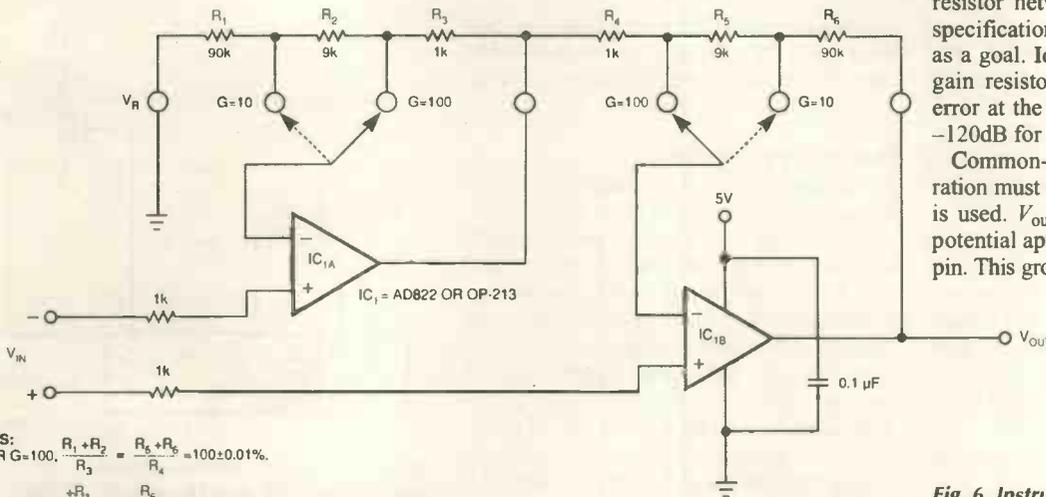
low as 600µV from ground. In this case, the bridge itself supplies the 2V common-mode bias, keeping the op amps in the middle of their common-mode range. Driving the bridge with a low-noise 4V supply further minimises noise.

Design of an instrumentation amplifier that shrugs off switch resistance is also possible (Fig. 6). Compared with the circuit in Fig. 4, this circuit's primary advantage is that it does not exhibit extreme common-mode limitations. Also, its topology switches divider taps for gain changes, so moderate switch resistances (10 to 100Ω) do not cause major errors. A circuit's gain can easily be set with jumpers or dpdt cmos switches. However, adding the same gain setting switches in series with  $R_G$  in Fig. 4 would produce serious errors.

In Fig. 6, amplifier  $IC_{1A}$  is a follower for signals at terminal  $-V_{in}$ ,  $IC_{1B}$  is a follower for signals at  $+V_{in}$ , and an inverter for signals at its negative input. Linear subtraction of cm signals and amplification of differential signals provides the circuit with a precise gain that can be varied between 10 and 100.

Performance keys to this circuit are the resistor network and the amplifier. For best performance with premium amplifiers, the resistor network should have a ratio-match specification of 0.1% minimum, with 0.01% as a goal. Ideal amplifiers and 0.01%-match gain resistors will have the common-mode error at the output on the order of -100 and -120dB for gains of 10 and 100, respectively.

Common-mode limitations of the configuration must be observed when a single supply is used.  $V_{out}$  in the circuit is referred to the potential applied to the resistor network's  $V_R$  pin. This ground assignment implies that if the



- NOTES:  
 1. FOR G=100,  $\frac{R_1 + R_2}{R_3} = \frac{R_5 + R_6}{R_4} = 100 \pm 0.01\%$ .  
 2. FOR G=10,  $\frac{R_1}{R_2 + R_3} = \frac{R_5}{R_4 + R_6} = 10 \pm 0.01\%$ .  
 3. R<sub>1</sub> TO R<sub>6</sub> ARE PART OF THE OHMTEK (NIAGARA FALLS, NY) 1043.

Fig. 6. Instrumentation amplifier that shrugs off switch resistance, enabling gains to be switched.

circuit has to exhibit high linearity for even small  $V_{in}$  difference voltages, the op-amps should be able to swing close to ground.

**Output-stage swing**

Not all single-supply amplifier outputs can swing to the negative rail (ground). Some may swing to within a diode drop (0.6V); others may swing within a few tens of millivolts. Only a very few swing to less than 1mV from the negative rail.

To understand which op-amp functions best for output-swing-critical applications, calls for detailed knowledge of output stages.

Figure 7 shows a sampling of output-stage topologies used in single-supply op-amps. At best, the bipolar complementary emitter-follower stage is only active to within a diode drop of each rail. It may or may not use a Darlington for transistor  $Tr_1$  – but usually has a single pnp for  $Tr_2$  – and cannot swing to ground without outside help, such as the added resistor  $R_{pull-down}$ . This resistor allows output linearity to ground for source-type loads only, while also raising overall dissipation for sustained high-level outputs.

Amplifiers with this type of output stage can be useful for single-supply applications, but must be carefully applied for all trade-offs to be optimised.

The nmos bipolar emitter-follower common-source output stage (Fig. 7b) has, by nature, an asymmetrical voltage swing. However, it exhibits low saturation voltages to ground because of the nmos pull-down transistor  $N_1$ . Saturation voltage to the positive rail is about 1V (or more), and the advantage of the stage is that the nmos pull-down transistor can drive current-sink loads to within a few millivolts of ground – enhancing the linearity in many single-supply applications.

The cmos stage (Fig. 7c) is, by definition, fully complementary and offers a resistive

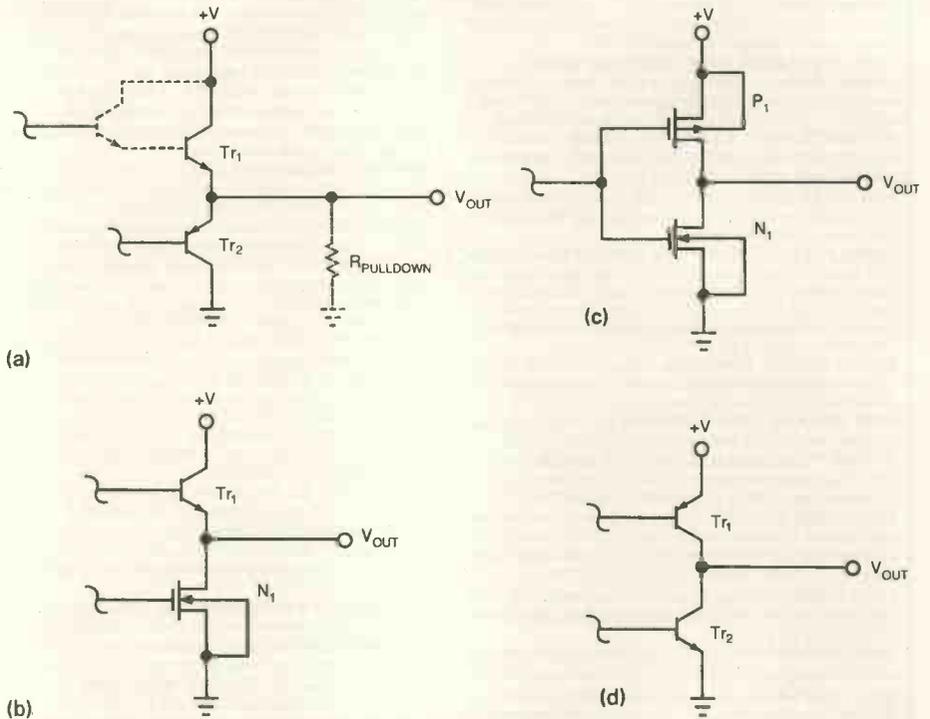
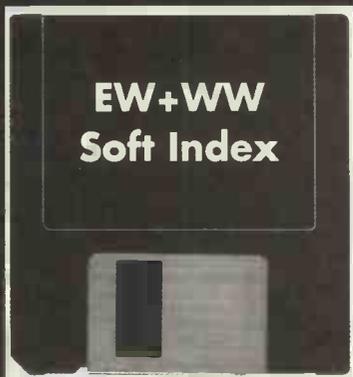


Fig. 7. Output-stage topologies used in single-supply op amps. The bipolar complementary emitter-follower stage (a) is active only to within a diode drop of each rail at best. The bipolar emitter-follower, nmos common-source output stage (b) has by nature, an asymmetrical voltage swing. The cmos stage (c) is, by definition, fully complementary. Complementary-bipolar common-emitter stage (d) is another rail-to-rail output stage.

connection to the supply rail for a high or low output, rail to rail. With appropriate low- $R_{on}$  transistors for  $P_1$  and  $N_1$ , saturation drops to either rail can be a millivolt or less at low currents. The stage is inherently class A so amplifier design must carefully control the static currents in  $P_1$  and  $N_1$  for low quiescent current. But for output currents of a few milliamps, this type of output stage is effective and quite versatile because of its rail-to-rail nature.

The complementary-bipolar common-emitter stage (Fig. 7d) is another rail-to-rail output stage. Saturation-drops to the rails range from a few millivolts to hundreds of millivolts over current ranges up to 20mA. Like the cmos rail-to-rail output stage, this bipolar counterpart is both effective and versatile, but avoidance of punitive trade-offs in power is critical to design of the bipolar stage.

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**A4 DTP MONITORS** Brand new, 300 DPI. Complete with diagram but no interface details. (so you will have to work it out!) Bargain at just £12.99 each!!!

**OPD MONITORS** 9" mono monitor, fully cased complete with raster board, switched mode PSU etc. CGA/ATL Invert (15watt D), IEC mains. £15.99 ref DEC23. Price including kit to convert to composite monitor for CCTV use etc is £21.99 ref DEC24

**LOW COST LASER NIGHT SIGHT JUST £79!!!**

Cheaper version of our £245 Cyclops model, works ok in low light levels or in complete darkness with the built in Laser. Ref 95/79 (please allow about 1 month delivery for these due to import difficulties)

**PC CONTROLLED 4 CHANNEL TIMER** Control (on/off times etc) up to 4 items (8A 240v each) with this kit. Complete with software, relays, PCB etc. £25.99 ref 95/26

**LOW COST RADAR DETECTOR** Built and tested pocket radar detector, ideal for picking up speed traps etc. Why pay £70 or more? ours is just £24.99 ref 95/25.

**COMPLETE PC 300 WATT UPS SYSTEM** Top of the range UPS system providing protection for your computer system and valuable software against mains power fluctuations and cuts. New and boxed, UK made Provides up to 5 mins running time in the event of complete power failure to allow you to run your system down correctly. SALE PRICE just £119.00.

**RACAL MODEM BONANZA!** 1 Racal MPS1223 1200/75 modem, telephone lead, mains lead, manual and comms software, the cheapest you onto the net! all this for just £13 ref DEC13.

**HOW LOW ARE YOUR FLOPPY'S?** 3.5" (1.44) unbranded. We have sold 100,000+ so old! Pack of 50 £2.99 ref DEC16

**BRITISH TELECOMM MULTIMETERS SA9083** These are 'return's so they may have faults but look ok. Complete with new leads and leather case. Price for two meters & 1 case is £10 ref DEC89.

**6mw LASER POINTER.** Supplied in kit form, complete with power adjuster, 1-5mw, and beam divergence adjuster. Runs on 2AAA batteries. Produces thin red beam ideal for levelling, gun sights, experiments etc. Cheapest in the UK! Just £39.95 ref DEC49

**SHOP WOBBLERS!** Small assemblies designed to take D size batteries and 'wobble' cardboard model signs about in shop windows! £3.99 ref SEP42.

**RADIO PAGER'S** Brand new, UK made pocket pagers clearance price is just £4.99 each 100x40x15mm packed with bits! Ref SEP5.

**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

**STEREO MICROSCOPE** 155x195MM, up to 600mm high, so items up to 10" will fit under lens. Rack and pinion focusing, 6 interchangeable rotating objective lenses, interchangeable eye pieces, +scaled eyepiece for accurate measuring etc. Powerful low voltage illumination system with green filter and variable intensity, 100mm black/white + ground glass stage plate, 70mm swivel mirror, adjustable eyepieces (both focus and width). Magnification range 4.6-100.8, field of view 39-2.4mm. Price is £299 for complete setup. Ref 95/300.

**3D 35MM CAMERA SYSTEM** Complete kit to convert a standard 35mm camera into a 3D version, enable you to take 3D colour slides with your own camera. Kit contains a prism assembly for the front of your existing lens, a sample 3D slide, a 3D slide viewer and 2 different lens mounts 49mm and 52mm, (other sizes available from photo shops at about £3 ea) all you need is standard slide film. Price for the complete kit is £29.99 ref 95/30.

**COMPUTER RS232 TERMINALS. (LIBERTY)** Excellent quality modem units (like wyse 50, s) 2xRS232, 20 function keys, 50 thro to 38,400 baud, menu driven port, screen, cursor, and keyboard setup menus (18 menus). £29 Ref NOV4.

**OMRON TEMPERATURE CONTROLLERS (E6C2).** Brand new controllers, adjustable from -50 deg C to +1,200 deg C using graduated dial, 2% accuracy, thermocouple input, long life relay output, 3A 240v o/p contacts. Perfect for exactly controlling a temperature. Normal trade £50+, ours £15. Ref E5C2.

**ELECTRIC MOTOR BONANZA! 110x60mm.** Brand new precision, cap start (or spin to start), virtually silent and features a moving outer case that acts as a fly wheel. Because of their unusual design we think that 2 of these in a tube with some homemade fan blades could form the basis for a wind tunnel etc. Clearance prices! Just £4.99 FOR A PAIR!! (note-these will have to be wired in series for 240v operation Ref NOV1.

**MOTOR NO 2 BARGAIN 110x90mm.** Similar to the above motor but more suitable for mounting vertically (ie turntable etc). Again you will have to wire 2 in series for 240v use. Bargain price is just £4.99 FOR A PAIR!! Ref NOV3.

**OMRON ELECTRONIC INTERVAL TIMERS.**

**Miniature adjustable timers, 4 pole o/p output 3A 240v,**  
HY1230S, 12VDC adjustable from 0-30 secs. £9.99  
HY1210M, 12VDC adjustable from 0-10 mins. £9.99  
HY1260M, 12VDC adjustable from 0-60 mins. £9.99  
HY2460M, 24VAC adjustable from 0-60 mins. £5.99  
HY241S, 24VAC adjustable from 0-1 secs. £5.99  
HY2460S, 24VAC adjustable from 0-60 secs, £5.99  
HY243H, 24VAC adjustable from 0-3 hours. £8.99  
HY2401S, 240v adjustable from 0-1 secs. £9.99  
HY2405S, 240v adjustable from 0-5 secs. £9.99  
HY24060M, 240v adjustable from 0-60 mins. £12.99

**PC PAL VGA TO TV CONVERTER** Converts a colour TV into a basic VGA screen. Complete with built in PSU, lead and sware. £49.95. Ideal for laptops or a cheap upgrade.

**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.

**GUIDED MISSILE WIRE.** 4,200 metre reel of ultra thin 4 core insulated cable, 28lbs breaking strain, less than 1mm thick! Ideal alarms, intercoms, fishing, dolls house's etc. £14.99 ref MAG15P5

**300v PANEL METER 70x60x50MM.** AC, 90 degree scale. Good quality meter. £5.99 ref MAG 6P14. Ideal for monitoring mains etc.

**ASTEC SWITCHED MODE PSU BM41012** Gives +5 @ 3.75A, +12 @ 1.5A, -12 @ .4A. 230/110, cased, BM41012 £5.99 ref AUG6P3.

**TORRODRIVE TX 30-0-30 480VA.** Perfect for Mosfet amplifiers etc. 120mm dia 55mm thick. £18.99 ref APR19.

**AUTO SUNCHARGER** 155x300mm solar panel with diode and 3 meter lead fitted with a digar plug. 12v 2watt. £9.99 ea ref AUG10P3.

**FLOPPY DISCS** DSDD Top quality 5.25" discs, these have been written to once and are unused. Pack of 20 is £4 ref AUG4P1.

**ECLATRON FLASH TUBE** As used in police car flashing lights etc, full spec supplied, 60-100 flashes a min. £9.99 ref APR10P5.

**24v AC 96WATT** Cased power supply. New. £13.99 ref APR14.

**MILITARY SPEC GEIGER COUNTERS** Unused an straight from Her Majesty's forces. £50 ref MAG 50P3.

**STETHOSCOPE** Fully functioning stethoscope, ideal for listening to hearts, pipes, motors etc. £6 ref MAR6P6.

**OUTDOOR SOLAR PATH LIGHT** Captures sunlight during the day and automatically switches on a built in lamp at dusk. Complete with sealed lead acid battery etc. £19.99 ref MAR20P1.

**ALARM VERSION** Of above unit comes with built in alarm and ptr to deter intruders. Good value at just £24.99 ref MAR25P4.

**CARETAKER VOLUMETRIC Alarm,** will cover the whole of the ground floor against forced entry. Includes mains power supply and integral battery backup. Powerful internal sounder, will take external bell if req'd. Retail £150+, ours? £49.99 ref MAR50P1.

**TELEPHONE CABLE** White 6 core 100m reel complete with a pack of 100 clips. Ideal 'phone extns etc. £7.99 ref MAR8P3.

**MICRODRIVE STRIPPER** Small cased tape drives ideal for stripping, lots of useful goodies including a smart case, and lots of components. £2 each ref JUN2P3.

**SOLAR POWER LAB SPECIAL** You get TWO 6"x6" 6v 130mA solar cells, 4 LEDs, wire, buzzer, switch plus 1 relay or motor. Superb value kit just £5.99 REF: MAG6P8

**SOLID STATE RELAYS** Will switch 25A mains. Input 3.5-26v DC 57x43x21mm with terminal screws £3.99 ref MAG4P10

**300DPI A4 DTP MONITOR** Brand new but shop soiled so hence bargain price! TTL/ECL. Inputs, 15" landscape, 1200x1664 pixel complete with circuit diag to help you interface with your projects. JUST £14.99. REF JUN15P2.

**BUGGING TAPE RECORDER** Small voice activated recorder, uses micro cassette complete with headphones. £28.99 ref MAR29P1.

**ULTRAMINI BUC MIC** 6mmx3.5mm made by AKG, 5-12v electret condenser. Cost £12 ea, Ours? just four for £9.99 REF MAG10P2.

**RGB/CGA/EGA/TTL COLOUR MONITORS** 12" In good condition. Back analysed metal case. £79 each REF JUN79

**ANSWER PHONES** Returns with 2 faults, we give you the bits for 1 fault, you have to find the other yourself. BT Response 200's £18 ea REF MAG18P1. PSU £5 ref MAG5P12.

**SWITCHED MODE PSU** ex equip, 60w +5v @ 5A, -5v @ 5A, +12v @ 2A, -12v @ 5A 120/220v cased 245x88x55mm IEC Input socket £6.99 REF MAG7P1

**PLUG IN PSU 9V** 200mA DC £2.99 each REF MAG3P9

**PLUG IN ACORN PSU** 19v AC 11w, £2.99 REF MAG3P10

**POWER SUPPLY** fully cased with mains and o/p leads 17v DC 900mA output. Bargain price £5.99 ref MAG6P9

**ACORN ARCHIMEDES PSU** +5v @ 4.4A, on/off sw uncased, selectable mains input, 145x100x45mm £7 REF MAG7P2

**GEIGER COUNTER KIT** Low cost professional twin tube, complete with PCB and components. Now only £19 REFAUG19.

**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.

**AA NICAD PACK** encapsulated pack of 8 AA nicad batteries (tagged) ex equip, 55x32x32mm. £3 a pack. REF MAG3P11

**13.8V 1.9A** psu cased with leads. Just £9.99 REF MAG10P3

**PPC MODEM CARDS.** These are high spec plug in cards made for the Amstrad laptop computers, 2400 baud dial up unit complete with leads. Clearance price is £5 REF: MAG5P1

**INFRA RED REMOTE CONTROLLERS** Originally made for hi spec satellite equipment but perfect for all sorts of remote control projects. Our clearance price is just £2 REF: MAG2

**200 WATT INVERTER** Converts 10-15v DC into either 110v or 240v AC. Fully cased 115x36x156mm, complete with heavy duty power lead, cigar plug, AC outlet socket. Auto overload shutdown, auto short circuit shut down, auto input over voltage shutdown, auto input under voltage shut down (with audible alarm), auto temp control, unit shuts down if overheated and sounds audible alarm. Fused reversed polarity protected, output frequency within 2%, v/dage within 10%. A extremely well built unit at an excellent price. Just £64.99 ref AUG6S.

**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

**MAINS CABLE** Precut black 2 core 2 metre lengths ideal for repairs, projects etc. 50 metres for £1.99 ref AUG2P7.

**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 PCs over a long distance. Complete kit £8.99.

**MINICYCLOPS PIR** 52x62x40mm runs on PP3 battery complete with shrill sounder. Cheap protection at only £5.99 ref MAR6P4.

**ELECTRIC MOTOR KIT** Comprehensive educational kit includes all you need to build an electric motor. £9.99 ref MAR10P4.

**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

**SOME OF OUR PRODUCTS MAY BE UNLICENSEABLE IN THE UK**

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**\*FM CORDLESS MICROPHONE** Small hand held unit with a 500' range! 2 transmit power levels. Reqs PP3 9v battery. Tuneable to any FM receiver. Price is £15 REF: MAG15P1

**LOW COST WALKIE TALKIES** Pair of battery operated units with a range of about 200'. Ideal for garden use or as an educational toy. Price is £8 a pair REF: MAG 8P1 2 x PP3 req'd.

**\*MINATURE RADIO TRANSCEIVERS** A pair of walkie talkies with a range of up to 2km in open country. Units measure 22x52x155mm. Including cases and earpieces. 2xPP3 req'd. £30.00 pr. REF: MAG30

**COMPOSITE VIDEO KIT.** Converts composite video into separate H sync, V sync, and video. 12v DC. £8.00 REF: MAG8P2

**LQ3600 PRINTER ASSEMBLIES** Made by Amstrad they are entire mechanical printer assemblies including printhead, stepper motors etc etc In fact everything bar the case and electronics, a good stripped etc REF: MAG5P3 or 2 for £8 REF: MAG8P3

**LED PACK** of 100 standard red 5m leds £5 REF MAG5P4

**UNIVERSAL PC POWER SUPPLY** Complete with flyleads, switch, fan etc. Two types available 150w at £15 REF: MAG15P2 (23x23x23mm) and 200w at £20 REF: MAG20P3 (23x23x23mm)

**\*FM TRANSMITTER** housed in a standard working 13A adapter! the bug runs directly off the mains so lasts forever why pay £700? or price is £26 REF: MAG26 Transmits to any FM radio.

**\*FM BUG KIT** New design with PCB embedded coil for extra stability. Works in any FM radio. 9v battery req'd. £5 REF: MAG5P5

**\*FM BUG BUILT AND TESTED** superior design to kit. Supplied to detective agencies. 9v battery req'd. £14 REF: MAG14

**TALKING COINBOX STRIPPER** 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 original use or used for something else?? Price is just £3 REF: MAG3P1

**TOP QUALITY SPEAKERS** Made for HiFi televisions these are 10 watt 4R Jap made 4" round with large shielded magnets. Good quality. £2 each REF: MAG2P4 or 4 for £6 REF: MAG8P2

**TWEETERS** 2" diameter good quality tweeter 140R (ok with the above speaker) 2 for £2 REF: MAG2P5 or 4 for £3 REF: MAG3P4

**AT KEYBOARDS** Made by Apricot these quality keyboards need just a small mod to run on any AT, they work perfectly but you will have to put up with 1 or 2 foreign keycaps! Price £6 REF: MAG6P3

**HEADPHONES** Ex Virgin Atlantic. 8 pairs for £2 REF: MAG2P8

**DOS PACKS** Microsoft version 3.3 or higher complete with all manuals or price just £5 REF: MAG5P8 Worth it just for the very comprehensive manual! 5.25" only.

**DOS PACK** Microsoft version 5 Original software but no manuals hence only £5.99. 3.5" only.

**PIR DETECTOR** Made by famous UK alarm manufacturer these are hi spec, long range internal units. 12v operation. Slight marks on case and unboxed (although brand new) £8 REF: MAG8P5

**MOBILE CARPHONE** £5.99 Well almost! complete in carphone excluding the box of electronics normally hidden under seat. Can be made to illuminate with 12v also has built in light sensor so display only illuminates when dark. Totally convincing! REF: MAG6P6

**ALARM BEACONS** Zenon strobe made to mount on an external bell box but could be used for caravans etc. 12v operation. Just connect up and it flashes regularly! £5 REF: MAG5P11

**6"X12" AMORPHOUS SOLAR PANEL** 12v 155x310mm 130mA. Bargain price just £5.99 ea REF MAG8P12.

**FIBRE OPTIC CABLE BUMPER PACK** 10 metres for £4.99 ref MAG5P13 ideal for experimenters! 30m for £12.99 ref MAG13P1

**HEATSINKS** (finned) TO220, designed to mount vertically on a pcb 50x40x25mm you can have a pack of 4 for £1 ref JUN1P11.

**INFRARED LASER NIGHT SCOPES**

Image intensifier complete with hand grip attachment with built in adjustable laser lamp for zero light conditions. Supplied with Pentax 42mm camera mount and normal eye piece. 1.6kg, uses 1xPP3, 3xAA's Please allow us about 1 month for delivery on this product due to import problems

**NEW HIGH POWER LASERS**

15mW, Helium neon, 3 switchable wavelengths. 63um, 1.15um, 3.39um (2 of them are Infrared) 500:1 polarizer built in so good for holography. Supplied complete with mains power supply, 790x65mm. Use with EXTREME CAUTION AND QUALIFIED GUIDANCE. £349+Vat

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**3FT X 1FT 10WATT SOLAR PANELS**

**14.5v/700mA**

**£44.95**

(PLUS £2.00 SPECIAL PACKAGING CHARGE)

**TOP QUALITY AMORPHOUS SILICON CELLS HAVE ALMOST A TIMELESS LIFESPAN WITH AN INFINITE NUMBER OF POSSIBLE APPLICATIONS, SOME OF WHICH MAY BE CAR BATTERY CHARGING, FOR USE ON BOATS OR CARAVANS, OR ANYWHERE A PORTABLE 12V SUPPLY IS REQUIRED.**

**PORTABLE RADIATION DETECTOR**

**WITH NEW COMPUTER INTERFACE.**

**£59.00**

A Hand held personal Gamma and X Ray detector. This unit contains two Geiger Tubes, has a 4 digit LCD display with a Piezo speaker, giving an audio visual indication. The unit detects high energy electromagnetic quanta with an energy from 30KeV to over 1.2MeV and a measuring range of 5-9999 UR/h or 10-99990

Nr/h. Supplied complete with handbook. Ref. NOV18.

# NEW PRODUCTS CLASSIFIED

Please quote "Electronics World + Wireless World" when seeking further information

## ACTIVE

### Asics

**Mixed-signal, 50V asics.** AMS can now supply mixed-signal asics that take a 50V supply voltage and offer the option of a second polysilicon layer. Integrating bipolar transistors in a cmos device makes for lower power consumption and fast switching. Since high voltage is needed to drive fast power fets, this process allows a low-cost cmos driver to control fets handling up to several hundred amps. It also enables the construction of floating polysilicon-polysilicon capacitors, removing voltage-dependence problems and parasitics. Austria Mikro Systeme International. Tel., 01276 23399; fax, 01276 29353.

### A-to-D and D-to-A converters

**Modulator.** Crystal Semiconductor has introduced the CS5321, a delta-sigma modulator meant for use with the CS5322 filter to form a 24-bit analogue-to-digital converter. CS5321 implements a fourth-order, 256 times oversampled modulator to give 123dB dynamic range at a thd of 118dB.

**Measuring amplifier.** To overcome the limitations of purely analogue measurements from mechanical transducers, such as inferior accuracy, lack of set-up storage and size, HBM introduces the *MGC Measuring Amplifier System*, which uses gate-array techniques to produce a new analogue-to-digital converter and flexible dsp to do the signal conditioning. Analogue signal is digitised at high speed and at high resolution to allow low-error processing over the whole frequency range providing a three-times improvement in accuracy over purely analogue methods. Measurement set-ups are stored on amplifier cards or in external memory modules and the *MGC* gives either analogue or digital output. Control is by front panel or by computer via RS-232C, RS-485/422 or IEEE488 interfaces. The compact version is a two-channel instrument, but models are available with six, twelve or sixteen channels. HBM United Kingdom Ltd. Tel., 0181-420 7170; fax, 0181-420 7336.

while consuming only 70mW for both chips. Use of the switched-capacitor technique reduces the effects of clock jitter without the use of plls and vcocs. Sequoia Technology Ltd. Tel., 01734 258000; fax, 01734 258020.

### Discrete active devices

**High-voltage mosfet.** High-voltage p-channel mosfets, the *Supertex TP2635* and *TP2640* are rated at maximum drain/source voltage of 350V and 400V, have low gate/source thresholds of 2V maximum and a guaranteed  $R_{ds(on)}$  of 15 $\Omega$ . 2001 Electronic Components Ltd. Tel., 01438 742001; fax, 01438 742002.

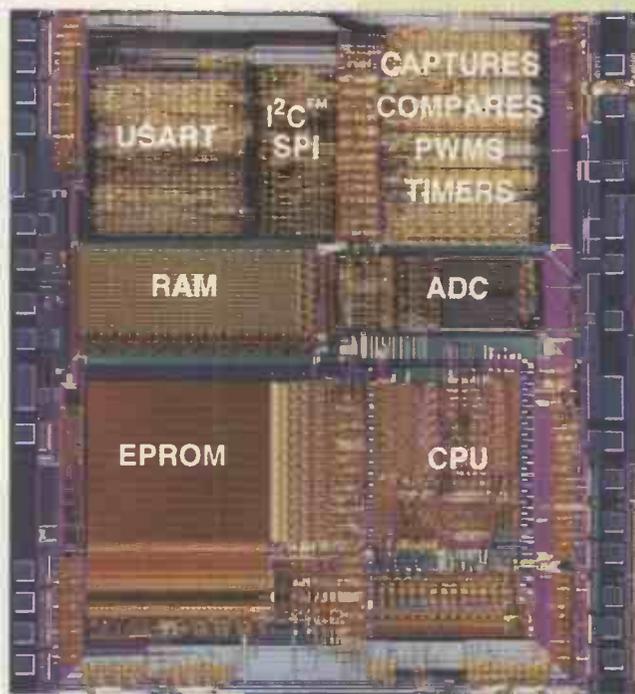
### Digital signal processors

**Serial i/o filter.** Harris's *HSP43124* is a digital filter engine designed to take the burden of filter processing from a digital signal processor. Signal is input digitally and a second input takes the mix factors which are multiplied by the input samples. A multiply/accumulator performs the sum-of-products needed for the filter format selected by further inputs. Outputs go to an output formatter which rounds or truncates them to a selected width and generates timing and sync. signals for serial output transmission. The device will also serve in digital tuning. Dynamic range is over 100dB, operating on 24-bit words with 32-bit coefficients. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.

### Linear integrated circuits

**Quad power drivers.** In one package, each of four power drivers in the Elantec cmos *EL7412CM* sources or sinks a continuous 100mA into 4 $\Omega$ . Response time is 20ns, with transient times of 10ns, matched to reduce timing errors. The chip takes 2mA from a single 5-15V supply. METL. Tel., 01844 278781; fax, 01844 278746.

**Video buffer.** *MAX4005* is a high-speed buffer with an internal, thin-film 75 $\Omega$  resistor to prevent reflections when driving 75 $\Omega$  cables. The input stage is a jfet, which takes only 10pA input current. Features include gain and phase errors of 0.11% and 0.03°, 60MHz, 0.1dB gain flatness, better than -60dB third-harmonic distortion, 1000V/ $\mu$ s slewing and 2ns settling to 0.1%. Rise and fall times are 350ps. Maxim Integrated Products UK Ltd. Tel., 01734 845255; fax, 01734 843863.



### Memory chips

**64Mb drams.** Engineering samples of Micron Technology's 64Mb drams, organised as 16Mb by 4, are available. They are made in a four-poly, double-metal cmos process with channel lengths of 0.35 $\mu$ m. The *MTL4LC16M2A7* is available with access speed of 60ns, 8K refresh and fast-page mode in a 500mil 34-lead SOJ package. Micron Europe. Tel., 01344-360055; fax, 01344-869504.

**Eeproms for SPI.** Xicor's 8-64K serial eeproms in the *X25000* series support the Serial Peripheral Interface bus protocol, interfacing directly to the serial port of microcontrollers when in synchronous mode or to a separate serial interface for inter-chip transfers on some controllers. Selected areas of the memory can be protected. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261678.

### Microprocessors and controllers

**Twin-Pentium processors.** *TwinStar* multi-chip modules based on two 100MHz *Pentium* processors are being sampled by MicroModule. Each module is in a 347-pin package with a pinout that is a superset of the Intel Overdrive socket, needing only a 3.3V supply. *TwinStar* is validated with both Intel *8243nx* and

**OTP microcontroller.** *PIC16C74* by Microchip is a 40-pin, field-programmable 8-bit microcontroller that has an a-to-d converter and extensive comms facilities; it also has a 4K one-time-programmable program memory. Its risc core uses Harvard architecture with a 200ns cycle time and 35 single-cycle, 14-bit instructions. The a-to-d converter acquires in 16 $\mu$ s. Thirty-four pins are concerned with i/o functions, including an 8-bit asynchronous slave port for other controllers, a synchronous serial port for SPI or I<sup>2</sup>C/Access buses, a 5Mb/s usart, two pwm outputs at 80kHz and a 16-bit capture/compare facility with 200ns resolution. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Intel *82497/82492* chipsets and is in the process with another set. It complies with the MPS 1.1 standard. MicroModule Systems Inc. Tel., +1 408 864 7437; fax, +1 408 864 5950.

**Frugal 16-bit microcontroller.** Affording a saving of 50% in battery power over standard 5V microcontrollers, Toshiba's *TMP93CM40F* 16-bit chip has

## NEW PRODUCTS CLASSIFIED

Please quote "Electronics World + Wireless World" when seeking further information



**Laser diode.** Made by EG&G in Canada, the C86137E InGaAs pulsed laser diode uses a multiple quantum well design to give a peak output of 8W at a peak forward current of 25A and 10V; wavelength is 905nm. Price is under £10.00. Pacer Components Ltd. Tel., 01734 845820; fax, 01734 845425.

optimised CPU, 16Mbyte of internal memory and i/o cells that operate on 2.7-5.5V supplies at currents as low as 5mA at 3V and 12.5MHz. It has four-channel direct memory access. Toshiba Electronics (UK) Ltd. Tel., 01276 694600; fax, 01276 691583.

**Cooler Pentia.** Cirrus Logic offers the two-chip set *Golden Gate*, which reduces power consumption and heat generation of a *Pentium* by up to 80%, without impairing the microprocessor's performance. The company's subsidiary, PicoPower, uses the Power-on-Demand technique to reduce system power when the *Pentium* is idle, cycling power to the processor on and off in increments of 1ns. Cirrus Logic Inc. Tel., 01727 872 424; fax, 01727 875919.

### Mixed-signal lcs

**Motor controllers.** IR's new intelligent power modules lower the cost of motor control below that of discrete components or existing modules. Opto-isolators are eliminated, as are three of the four isolated supplies normally needed. First in the family of devices, which contain an IR2130 power IC with *Hexfred* diodes and IGBTs, are the IPM1560 and IPM2060 15A and 20A devices rated at 600V, to drive 1hp and 2hp motors. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

**Video encoder.** Philips's SAA7187 digital video encoder takes MPEG compressed data or digital YUV video and encodes it into ntsc or pal composite video or S-video signal. It generates subcarrier and colour modulation and inserts sync. signals; luminance and chrominance are filtered to RS-170-A and CCIR-624 requirements. Pixel data in various formats goes to one of three 8-bit wide ports, one of them also taking a microprocessor input when free from video. Philips Semiconductors (Eindhoven). Tel., +31 40 722091; fax, +31 40 724825.

### Optical devices

**Blue leds.** Nichia gallium nitride leds provide a bright blue light at a luminous intensity of 1000mcd, output being 1200mW at 20mA. Peak wavelength is 450nm with a viewing angle of 15-45°, depending on type. Packaging is in standard 3mm and 5mm types. Hero Electronics Ltd. Tel., 01525 405015; fax, 01525 402383.

### Oscillators

**Crystal oscillator.** IQD's 60 series of surface-mounted oscillators measure 7.5 by 5mm by 1.6mm high and the thermal coefficient of the ceramic package matches that of the substrate in multi-chip modules, for which the oscillator is designed. Frequency coverage is 1.5-60MHz at stabilities of either ±50ppm or ±100ppm. Compatibility is with ttl or CMOS/LS-ttl, with or without tri-state operation. IQD Ltd. Tel., 01460 77155; fax, 01460 72578.

### Power semiconductors

**600V mosfets.** New power mosfets from Toshiba are rated at 400-600V and have power loss reduction of 30% over earlier techniques. They are meant for use in lighting and switched-mode power supplies and offer an  $R_{ds(on)}$  15% lower and switching speed 25% faster than before. The 500V, 15A 2SK2150 has an on resistance of 0.29Ω, the 20A 2SK2057 showing only 0.24Ω. Both are in TO3P(N); there is also a 600V, 6A type in TO220. Toshiba Electronics (UK) Ltd. Tel., 01276 694600; fax, 01276 691583.

**Power fets.** New *Hexfet* power transistors by IR exhibit an  $R_{ds(on)}$  of less than 50% that of existing devices of the same size; the 55V n-channel IRF3205, for example, has an  $R_{ds(on)}$  of 8mΩ, the IRL3803 in the 30V series showing only 6mΩ in both 4.5V and 10V,  $V_{GS}$  states, all in TO-220. P-channel devices show an  $R_{ds(on)}$  only twice that of the n-channel types. International Rectifier. Tel., 01883 713215; fax, 01883 714234.



## PASSIVE

### Passive components

**Metal-film resistors.** Two ranges of Philips metal-film resistors cover resistance ranges of 24Ω-100kΩ (MPR24) and 4.99Ω-1MΩ (MPR34) to tolerances of ±0.05, 0.02 and 0.01% for the former and ±0.5, 0.25 and 0.1% for the latter. Those in the 24-200Ω range to ±0.1% tolerance are of low inductance. Dissipations of 125, 250 and 400mW are available. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

**Tantalum capacitors.** Low equivalent series resistance is the feature of AVX's new series of tantalum chip capacitors. With esr in the range 0.15-1Ω, the number of capacitors used to filter the output of switched-mode power supplies can be reduced. Ripple current rating is higher, which makes for lower temperatures and greater reliability. In 3V logic, any reduction in ripple and noise on the supply is helpful; the capacitors help to ensure that. AVX Ltd. Tel., 01252 770000; fax, 01252 770001.

**Inclinometers.** Two kinds of inclinometer by Control Transducers use optical and force-balance techniques. Optical types are either of the incremental variety, giving 7200 pulses/rev with ttl output and, optionally, an index pulse, or the absolute type providing 4096 pulses/rev, which have either 0-4096mV output or RS-232 serial interface port to drive up to 15 sensors on one SEI bus. Force-balance models are in five ranges from ±5.75° to ±90°, with an output of 0 to ±5V full scale. Non-linearity is ±0.005% and non-repeatability ±0.005%. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

**Resistor network.** Using the Vishay-Sfernice CNS range of sip resistor networks, LSW offers a custom network production service in non-standard values from 50Ω to 1MΩ, the resistors being single or multiple with a common lead or in chains. Resistors exhibit a temperature coefficient of 10ppm/°C and 0.1% tolerance. LS Wiggins Ltd. Tel., 01243 826824; fax, 01243 860493.

### Displays

**Thin, mono vga.** A new black and white vga display from Hitachi, the LMG5378XUFC offers the full vga 640 by 480 dot resolution and measures only 5.5mm thick, weighing 330g. That, together with a single 5V supply at 20mA, makes the device very suitable for portable equipment. Its features include a fluorescent backlight, an anti-glare screen and a wide viewing angle. Overall size is 257.7mm by 174mm, with a display area of 196mm by 150mm. The recommended controller is the HD66841. Elger Technologies Ltd. Tel., 01928 579000; fax, 01928 579123.

**Bright lcds.** Three graphic lcds from Citizen use the super-twisted nematic (stn) passive-matrix technique, Citizen's 'chip-on-glass' technology and dual-scan addressing for faster refresh. K6481L-FF is a 4.7in unit for palm-top equipment; K6483LFF an upgrade of the 7.9in K6480L-FF with a thicker metal casing for better rf performance and a brighter backlight; and G6489L-FF, a 9.4in monochrome type measuring 5mm thick, with twin retardation control films for improved contrast. All provide 640 by 480 resolution and use a fluorescent backlight. Citizen Europe Ltd. Tel., 01753 584 111; fax, 01753 582 442.

### Filters

**Motor-drive filter.** Schaffner has introduced a new type of filter, the FN500, to follow the output of 4-16A, 6-20kHz motor-drive inverters which, in the raw state, can cause a shortening of motor life by up to 50% due to high peak currents; leakage losses due to fast edges and stray parasitic capacitance in shielded cables; interference and therefore compatibility precautions; and acoustic noise. Schaffner says the filters provide a smooth output waveform to reduce or eliminate all these effects, for example reducing output pulse slewing rate from 8kV/μs to 400V/μs. Schaffner EMC Ltd. Tel., 01734 770070; fax, 01734 792969.

### Hardware

**Enclosures.** Equipment enclosures from Briticent come in ranges of polycarbonate, ABS, polyester, steel, stainless steel for the food industry and aluminium and are of standard sizes or custom-made, complete with accessories. Sizes range from small boxes for push-button switches to floor-standing types. Options include inspection windows, vents, cover plates and specified machining and

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drilling. A catalogue is offered. Briticent International. Tel., 01425 474617; fax, 01425 471595.

### Instrumentation

**Graphical multimeter.** Fluke's 860 Series Graphical Multimeters combine a dual analogue/digital readout with waveform display up to 1MHz, a 30h by 1s plot of readings with a graph of the stored result, in-circuit test of components and logic test to show transitions to 10MHz, frequency and dc average. Parameters determined by the meter readout are ac and dc voltage, current, resistance, conductance, capacitance, frequency, duty cycle, pulse width, period and decibels. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225067.

### Radio communications products

**Low-power RF links.** LPRS has a matched pair of low-power radio telemetry transmitter and receiver modules for the transmission of analogue or digital data up to 200 metres. Both the TXM-433-A transmitter and SILRX-433-A receiver are pcb-mounted and operate at the standard European ETS 300-220 frequency of 433.92MHz (418MHz in the UK). Transmitter modulation bandwidth is 0-10kHz with an internal first-order low-pass filter, while the Rx needs only an antenna. Power to the transmitter is 10mA from 9V, although it accepts 6-12V, and that to the receiver is 15mA from 5V. Carrier detection signal appears in less than 3ms to allow power-saving circuitry to operate. Low Power Radio Solutions Ltd. Tel., 01993 709418; fax 01993 708575.

**Mains dropout testing.** Schaffner EMC offers the NSG642 variable ac source, intended for use with, and control by, the NSG1003 mains dropout generator, running the Windows-based WIN1003 software. The 642 enables under and over voltage testing between 1V and 280V ac, changing output voltage at up to 120V/s to produce timed voltage dips as required by EN61000-4-11. Its two motor-driven transformers are rated at 16A continuous or 80A short-term. The combination also functions as a stable ac source for sensitivity and reliability testing. Schaffner EMC Ltd. Tel., 01734 770070; fax, 01734 792969.

**Photometer.** Tek's LumaColor II handheld photometer makes colour light measurement for colour matching and balance in television studios, crt manufacture and services such as street lighting, using the existing range of measuring heads. With available single-sensor heads, it measures luminance, illuminance, radiance or irradiance. It has a 10-channel memory for reference colours, rgb bargraphs helping to match colours with the reference. There is RS-232 control, output being either analogue or by RS-232. Tektronix UK Ltd. Tel., 01628 486000; fax, 01628 474799.

**GSM test sets.** Racal has two portable test sets, the 6103 for digital radio telephones and the 6113 for base transceiver stations. They retain the features of earlier instruments, but offer new features. Two PCMCIA cards in the 6113 allow test sequences to be stored and recalled and enable operating code to be loaded into the instrument to lengthen its operating life. 6103 has manual and automatic operation and will check a GSM telephone in under 30s. Both can be remotely controlled and



**Multimeter/recorder.** Hioki offers the 8220 Multi HiPlay combined digital multimeter and recording instrument, which presents changes in dual-channel voltage and current readings on an lcd graph. The instrument simultaneously displays on the 5in lcd and stores readings for up to six days, giving a graph of variations over time. A cursor gives a numerical reading of values from the graph, including minimum, maximum and average values, date and time. 01734 786911; fax, 01734

contain an average of around 320 pages and describe how a particular market is composed and who the market leaders are, growth or otherwise of a market, company profiles, competitive influences and technology trends. Prices vary from around £650 to £3000. Frost & Sullivan. Tel., 0171 730 3438; fax, 0171 730 3343.

### Materials

**Green degreaser.** Vaposol by Croftshaw is a replacement for chlorinated hydrocarbon vapour degreaser, such as 1.1.1 trichlorethane, which is not only more friendly to the environment but also saves time and money. It cleans by condensation in one tank, so that parts are clean and dry in five minutes at low temperature. Vaposol can be used with existing equipment. Croftshaw Solvents Ltd. Tel., 0181 6985556; fax, 0181 6974614.

the 6113 runs a Windows control package. Racal Instruments Ltd. Tel., 01734 669969; fax, 01734 262121.

### Literature

**Crystal products.** IQD's 1995 Crystal Products Data Book is now available, having 330 pages. Additions since the last one include a digital temperature-compensated oscillator, a range of surface-mounted filters and ceramic-packaged SM crystals. IQD Ltd. Tel., 01460 77155; fax, 01460 72578.

**PCB standard.** British Standards has published BS EN 60249, which is concerned with base materials for pcb manufacture. Spec. 6 and 7 are on phenolic cellulose paper copper-clad sheet burning tests, horizontal and vertical. The standard supersedes BS 4584 and costs £44 or £22 to subscribing members. BSI Customer Services, Publications, 389 Chiswick High Road, London W4 4AL.

**Consumer ICs.** In consumer ICs, GEC Plessey concentrates on rf front ends for broadcast, cable and satellite television, vcrs and teletext. Its Consumer IC Handbook describes the company's products in these areas in detail and contains application notes with practical examples. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

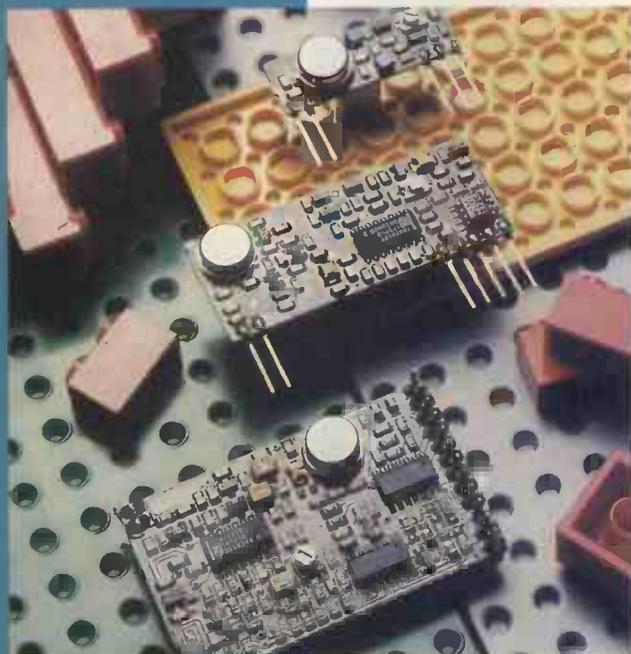
**Market intelligence.** Frost and Sullivan has new catalogues of market research reports on Sensors, instrumentation and control and Electronics and components. Reports

### Navigation systems

**GPS from GPS.** GEC Plessey Semiconductors has made available a complete Global Positioning System development environment that

### Interfaces

**PCI-PCMCIA interface.** From Cirrus Logic, the CL-PD6729 is claimed to be the first mixed-voltage PCI-to-PCMCIA host adaptor to control two independent PCMCIA sockets, allowing the use of 5V and 3.3V peripherals simultaneously. The dual-slot host adaptor gives a direct connection to the PCI system bus, supporting both PCI interface for x86 systems and PowerPC systems. Low power consumption is achieved by a suspend mode and an automatic low-power dynamic mode which stops transactions on the PCMCIA bus, stops clock distribution and turns most of the internal circuitry off. There is full PCMCIA compatibility and bios-level software to interface host adaptor to host. Cirrus Logic Inc. Tel., 01727 872424; fax, 01727 875919.



## NEW PRODUCTS CLASSIFIED

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converts a 486 pc into a 12-channel GPS receiver for learning the system and product development. It includes a chipset and a set of GPS tracking and acquisition software for 12-channel parallel processing in C, available as source and executable code. The chipset includes the GP1010 rf downconverter, DW9230 35.4MHz IF saw filter and two six-channel GP1020 correlators, all on a daughter card mounting on the main card with an ISA interface to the host. The system also includes a patch antenna, cables, software and a manual. GEC Plessey Semiconductors Ltd., Tel., 01793 518510; fax, 01793 518582.

### Production equipment

**Marking machine.** Rejafix has the *Hotmarker Model H.402*, a manually operated bench machine for the permanent identification marking of cable sleeving and tube to the BS.3858 standard. The required mark is dialled into the machine, which gives hot-foil, instant-drying print and avoids the need to hold stocks of specific markings. A number of available print wheels are either flat or curved for tubing and some models are able to cut the tube or cable to length after printing. Rejafix Ltd., Tel., 0181 5602224; fax, 0181 5608774

### Power supplies

**Pentium power.** First in a range of dc-to-dc converters designed to supply power to Intel's 3.3V P54C-VR Pentium processor, the Semtech *MP54C-1B* is a 68% efficient 15W device operating over the 4.75-5.5V input range and needs no heat sink. It provides linear regulation, overvoltage, short-circuit and thermal protection and a signal to indicate when the output is within limits. Semtech Ltd., Tel., 01592 773520; fax, 01592 774781.

**Step-down controllers.** *MAX796-799* high-power step-down controllers drive external n-channel mosfets to power cpus at 95% efficiency, logic-controlled shut-down current being 1.2µA. *MAX797/8* have a logic-controlled, synchronisable, fixed-frequency pwm mode to reduce noise in communications equipment, while *MAX796/9* use a secondary feedback input for transformer-based application, to improve cross-regulation of auxiliary outputs. Input range is 4.5-30V and outputs are 2.9V, 3.3V and 5V at up to 10A. Maxim Integrated Products UK Ltd., Tel., 01734 845255; fax, 01734 843863.

**Intelligent SMPS.** Vero's *ISI-POWER* switched-mode power supply is claimed to be the first 19in rack-mount psu with an intelligent controller, which allows programming, control and interrogation from a pc via an RS-232 serial link. It is a 180W unit in a 6U by 6HP module, modules available providing 3.3V, 5V, 12V, 15V and 24V from an autoranging universal input. Monitoring by pc includes input, output and sense voltages and output current; a temperature probe can be fitted, with temperature limits set remotely to

activate fans. Any problem causes an alarm box to open on the computer screen. Vero Electronics Ltd., Tel., 01489 780078; fax, 01489 780978.

**DC-DC converters for lcds.** Surface-mount dc-to-dc converters from *Murata* are designed to power lcds, voltage from the external power source being converted within the display unit. They are designed for use with simple monochromatic lcds and are in a metallic case to reduce radiation. Output on/off switching is provided. *Murata Electronics (UK) Ltd.*, Tel., 01252 811666; fax, 01252 811777.

**Power and protection.** Fiskars' *PowerRite Plus* uninterruptible power supplies provide protection for small systems, incorporating a communications interface that allows the use of Lansafe III power-management software to guard pcs and workstations, point-of-sale equipment, remote lan nodes and others against power problems. The units also have the company's Advanced Battery Management functions to extend battery life by up to 30%, battery test and warning of impending battery trouble. Three models provide 250VA, 400VA and 600VA. *Fiskars Electronics Ltd.*, Tel., 01734 306600; fax, 01734 305868.

### Television components

**Image enhancement.** *TDA9170* by Philips is a single-chip video processor that improves the quality of the picture on television receivers using the standard luminance and colour-difference interface (yuv). It uses adaptive, non-linear amplification of the y signal, based on real-time analysis of the picture content to enhance contrast to give an apparently sharper picture and better detail in dark areas. There is also adjustable gamma correction, automatic u and v correction and control over the picture area in which black-level detection and the picture analysis are carried out. Philips Semiconductors (Eindhoven), Tel., +31 40 722091; fax, +31 40 724825.

### Transducers and sensors

**Monostable solenoid.** BLP's *PE*

*Series 68* direct acting solenoids are now available in a monostable version having an international standard frame size. Power rating is 3.5W at 240V dc maximum, the coil pulling a load of 0.4kgf at a 10mm stroke. Custom-designed types are offered. *BLP Components Ltd.*, Tel., 01638 665161; fax, 01638 660718.

**Vislon systems frame grabber.** *DataCell* has upgraded its *J-Tool* colour Sbus frame grabber to provide 4416 by 3456 pixel resolution at 24 bits/pixel. When used with a *Sun SPARCstation* and a *JVC TK-F7300* camera, the *SS2200 J-Tool* card allows small areas to be analysed at this very high resolution. *J-Tool* supports TIFF image storage and capture, providing live image display in an X-window for manual focus and set-up. *DataCell.*, Tel., 01628 415415; fax, 01628 415400.

## COMPUTER

### Computer board-level products

**Graphics for VME.** Syntel's *SYN-B301* is a graphics controller with a *GDS5422* core, capable of 24-bit true-colour imaging. Its video interface is programmable so that video and sync. signals can be arranged to cope with a range of requirements under

**Servo set up.** *Servo Tuner* is an add-on software module to work with *Compumotor's* Windows-based *Motion Architect* software, together providing graphical feedback of real-time motion information and an easy method of setting tuning gain and other characteristics in the 6000 Series of servo controllers. The package allows the drive system to be tuned without the position loop and the position loop tuning with graphical feedback to show exact system motion, automatically generating set-up code, editing and executing motion-control programs and creating a test panel. *Parker Hannifin plc*, *Digiplan Division.*, Tel., 01202 699000; fax, 01202 600820.

**Small pc board.** DSP Design offers the *GCAT 6000*, an update on earlier *GCAT7* boards with more functions and still the size of a credit card. Functions include PCMCIA Type II interfacing, an a-to-d converter, 1Mb of rom and 8Mb of dram. It forms a reasonably priced alternative to custom boards in embedded microcontroller applications. Engineers can configure i/o, peripheral and graphic processor modules with the help of a development system. Two PCMCIA devices, for example an Ethernet/modem and hard disk, may be simultaneously connected to the double-stack connector. Ports include keyboard, two serial ports and lcd and crt interface, higher-resolution and electroluminescent displays also being supported. The board draws 75mA from 5V on fast run and 50µA suspended. *DSP Design.*, Tel., 0171 482 1779; fax, 0171 482 1773.

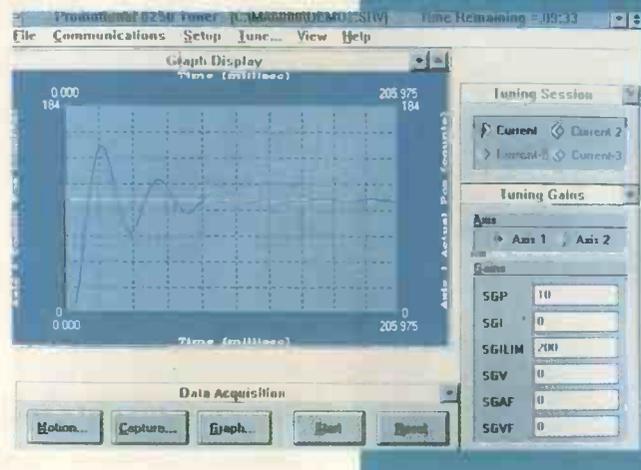
software control. Resolution up to 1280 by 1024 is offered and video frequencies are generated by phase-locked loop, programmable between 7MHz and 80MHz. Video memory is 1Mbyte, expandable to 2Mbyte. There is audio output and the serial mouse is Logitech-compatible. *Syntel Microsystems.*, Tel., 01484 535101/2/3; fax, 01484 519363.

### Data communications

**200Mbyte/s for VMEbus.** Using PEP's *AutoBahn Starter Kit*, designers of VMEbus systems can obtain a data transfer rate of up to 200Mbyte/s. The kit includes *AutoBahn Spanceivers*, 15-slot backplane, host cpu, drivers, notes and test data, and may also be used for non-VMEbus hardware. Processing power is provided by PEP's *VM30 68EC030/68302* single-board computer giving 16Mips. *PEP Modular Computers.*, Tel., 01273 441188; fax, 01273 441199.

**Plug-and-play modem chips.** Users of *Sierra's SC11140* fax/modem chipset will have no need to spend time on modem and computer adjustments, since the chips automatically configure an ISA modem. The controller chip queries system resources and determines which com ports and interrupts are available, also incorporating the logic to add voicemail and Windows MCI-compatible business audio to a V.32bis modem. It is suited for Windows 95, when that is available. *Sierra Semiconductor bv.*, Tel., +31 73 408 888; fax, +31 73 423 155.

**Dual-port serial boards.** *Amplicon Liveline* has six *200 Series* serial communications boards for PC XT or AT(ISA) bus computers, each having two serial ports conforming to RS-232, RS-422 or RS-485 and the



choice of opto-isolation. All have flexible base address and interrupt to allow each port to be independently addressed at locations from 000 to 7F8<sub>16</sub> at any 8-bit boundary. Each serial port has a 16-byte input fifo and a 16-byte output fifo in the uart, both being software-selectable. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

**Wireless modem.** Newest member of Rockwell's family of wireless communications products is the RC32ACC, a cellular digital packet data (cdpd) V.32bis data modem two-chip set, allowing landline, cellular or digital packet radio communications. It uses the same cellular network for voice and data and delivers services including e-mail and fax in one device and is pin-compatible with the earlier RC144ACL wireline data/fax modem. Rockwell International Corp. Tel., 0181 577 2800; fax, 0181 577 2257.

**Crash protection for networks.** GEC Plessey offers the PCA873, an IC for the protection of networks against low-frequency disturbances on network signals caused by losses in transformers. This is a particularly severe effect on fast signals such as those in Ethernet at 100Mb/s or ATM at 155Mb/s. The IC also protects against the kind of data frame that accentuates the 'base-line wander' and allows high-speed networks to

use unshielded and shielded twisted-pair copper cables. GEC Plessey Semiconductors Ltd. Tel., 01793 518510; fax, 01793 518582.

**Mass storage systems**

**Sram PCMCIA card.** Centennial has the first sram PCMCIA cards with a rechargeable lithium battery, which is charged each time the card is inserted. A range of 3.3V and 5V flash, mask-rom and otp cards is available. The company also offers a GPS receiver housed in a Type 2 PCMCIA enclosure. Trident Microsystems Ltd. Tel., 01737 765900; fax, 01737 771908.

**Software**

**Speech compression.** GSM speech-compression software is obtainable as source code for the AT&T DSP32C signal processor working in VME or Sbus systems from Bores. The software costs £4000, without further royalty payments. It will run on a multi-processor VME system with multi-channel speech i/o. Bores Signal Processing. Tel., 01483 740138; fax, 01483 740136.

**Operating system simulator.** A Windows version of Enea Data's OSE/SIM is now available, which allows developers of embedded applications using the OSE real-time

operating system to debug and test software independently of hardware and even before the hardware exists. The package does the same as the existing version, but has Hypertext help and simplified signal and process entry. It supports 8-bit, 16-bit and 32-bit microprocessors. When combined with a native C compiler and source-level debugger, the package forms a complete development environment and is also available for pc-dos, Unix and VAX/VMS hosts. Reflex Technology Ltd. Tel., 01494 465907; fax, 01494 465418.

**Network power management.**

Fiskars' LanSafe III and FailSafe III are uninterruptible power supply software packages that monitor network power and control shutdown and rebooting of networks. It shuts down workstations, servers, bridges and the other devices to a defined schedule, saving data and flushing system caches to disk. LanSafe III provides network functions on all major operating systems and FailSafe III on single-user environments such as Windows. Fiskars Electronics Ltd. Tel., 01734 306600; fax, 01734 305868.

**6805/HC05 C cross-compiler.** Pentica claims its true C compiler for the Motorola 6805 processor to be the first available. Hi-Tech C produces compact and efficient code, is Ansi-compliant and creates complex interrupt-driven applications in C. It supports prototypes and variable argument lists, function pointers, long arithmetic, floating points, bit fields, structures, unions and a full C code printf(). There is full control over variable placement including i/o ports at fixed addresses, and internal or extended ram. Interrupts may be declared in C and called from hardware interrupts; generated code saves and restores the processor state, handles the interrupt and returns. Macros initialise, enable and disable interrupts from C and interrupt vectors can be changed 'on the fly'. A compiled stack preassigns static memory addresses to local variables and function arguments, overcoming the 6805's limitation of having no addressable stack, which makes it normally impossible to run a C run-time model with stack-based functions. Pentica Systems Ltd. Tel., 01734 792101; fax, 01734 774081.

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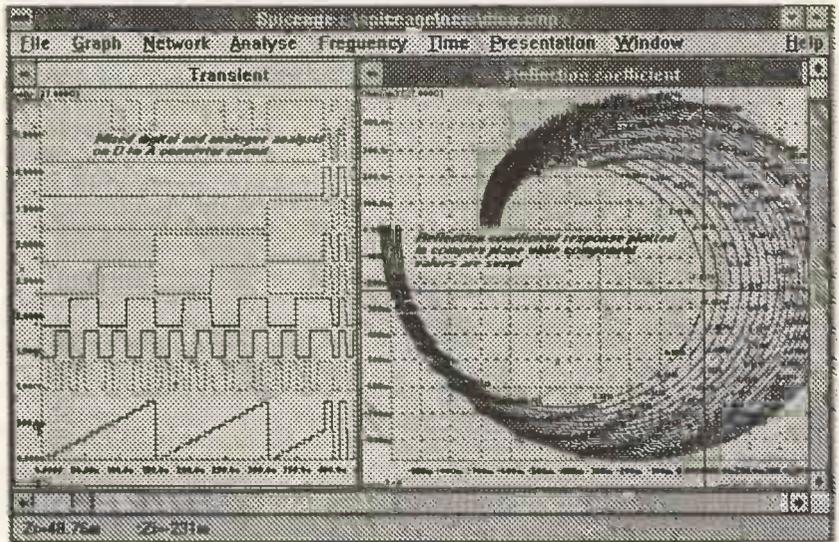
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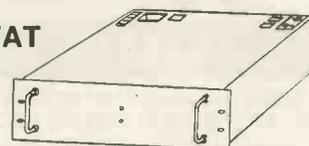
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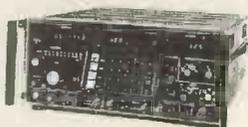
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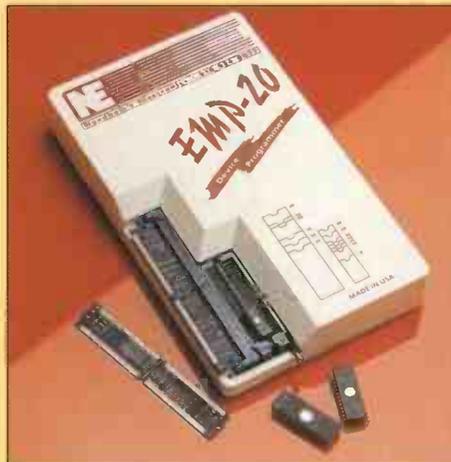
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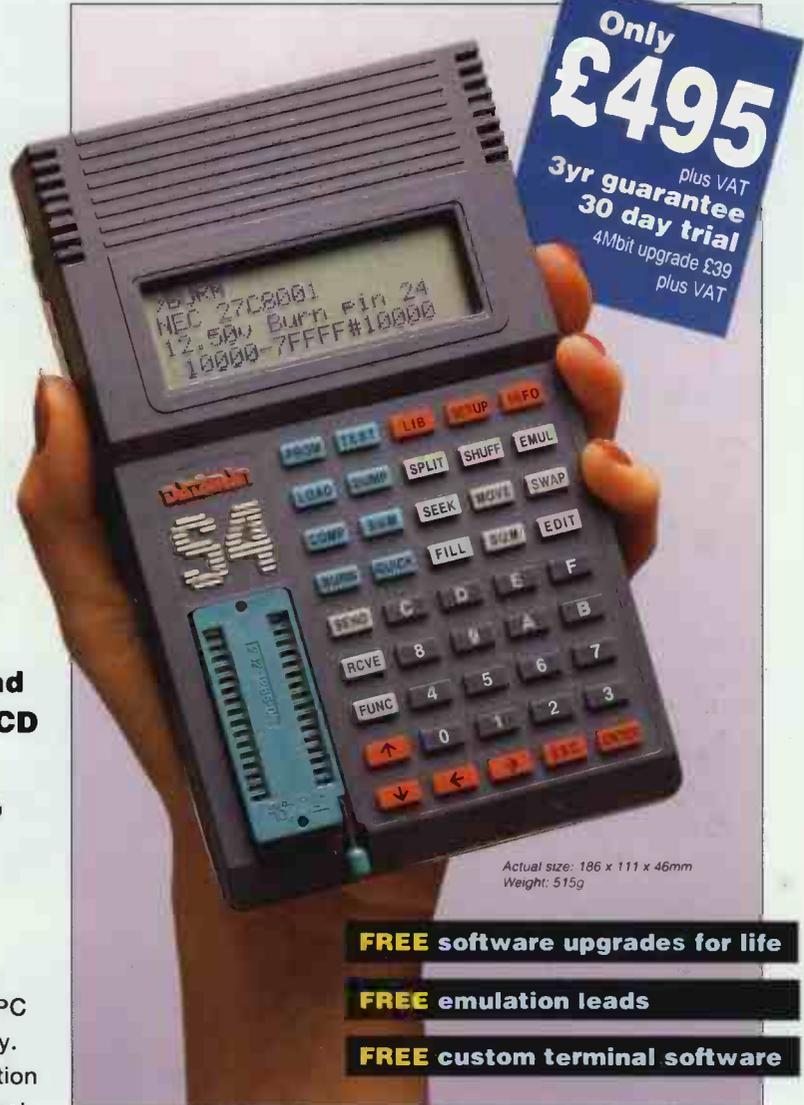
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